Dated: Nov. 1986

TECHNICAL NOTES FOR EME COMMUNICATION

These notes were written as a technical aid to radio amateurs involved in EME communication at the UHF and SHF amateur bands.
The purpose is to provide information for the understanding, measurement, and construction techniques of EME radio systems in that portion of the spectrum. These notes are not intended to be a primer or text book of EME communication but rather a set of stand-alone advanced reports on specific topics.
The ultimate objective is to provide information leading to a most efficient EME radio system for reliable communication without the use of expensive test equipment. We are fortunate in these times that solid state technology has developed very low noise devices required for reception, and that the space communications industry has honed antenna technology and related techniques which are applicable to EME communication.

Although many of the reports were written in the mid-1960s, much of the information is still applicable today.
The original set of technical notes was prepared while the author was a member of the Crawford Hill VHF Club, W2NFA, associated with the Bell Telephone Laboratories at Holmdel, New Jersey, U. S. A.

This particular set of dated notes has been partially revised, updated and supplemented from the original set, which received wide distribution over the years.
The ordering of the subjects remains as in the original notes and represents only the chronological order in which they were written. The table of contents should be consulted to determine both the subject material and an indication of its applicability at this time.

I reserve the right to make corrections, revisions and additions to these notes in future issues. Each set of notes will be dated as of the last revision issue.

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Technical Report # 1
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From: The Crawford Hill VHF Club
Date: (1968) rewritten Sept 1986

Subject: <u>Circular Polarization for EME Communication</u>

Faraday rotation of linearly polarized EME (Earth-Moon-Earth) signals can produce deep fades in the receiver-detector output, which vary over considerable range in periodicity. (See Technical Report # 20) The periodicity of polarization rotation under enhanced conditions can, at low VHF (144 mc/s), be as short as a second or less and can cause additional difficulties in reception of EME CW or modulated signals. The effect diminishes gradually with increasing frequency until about 3000 to 4000 mc/s where it is virtually non-existant.
The slowing down of the fading rate with increasing frequency can be particularly annoying at, for example, 1296 mc/s where a period can be measured in tens of minutes. Under these conditions an EME signal may remain essentially cross polarized, and consequently undetectable or very weak, for some time in a linearly polarized system.
Linear polarization control has been used (currently used on 432 mc/s) by physically (or electrically) rotating the antenna polarization orientation to off-set the Faraday rotation. Such a scheme introduces an additional variable for the operator to adjust and manage because Faraday rotation is a non-reciprocal process in nature. This means that the correct alinement of antenna polarization for one direction of the same EME path is not necessarily correct for the reverse path. For rotation periodicities of around a few seconds, this scheme is unmanageable without sophisticated control.

Fortunately the fading caused by Faraday rotation of linearly polarized transmission can be virtually eliminated by simply employing circular polarization exclusively for EME communication. The purpose of this report is then to urge and recommend that ALL EME communication be carried out with circular polarization. Even at microwave frequencies where Faraday rotation is not a problem, there is still a linear polarization time dependent tracking problem due simply to the geometry of the EME path. **

In addition to the above benefits, circular polarization offers the unique property that if a convention of ALWAYS using RIGHT circular polarization on transmit and LEFT circular polarization on receive is adopted, then all EME stations will be compatible with each other for two-way communication, AND for echo testing. This is so because reflection of a circularly polarized wave also reverses the sense of polarization, whether it be off the Moon or any reflecting process. In reflector antennas the feed antenna sense must therefore be opposite to the ultimate radiated sense required.
Another benefit which is very valuable to an EME system is that

the two senses, right and left circular, are in nature uncoupled modes of transmission. This means that a radiating system, antenna, may radiate each sense independently. Theoretically then, it is possible to construct a single antenna system with two ports (connections), one for left and one for right circular polarization, which are physically and electrically independent. In practice, electrical isolation between the ports of - 20 db can be achieved readily and higher isolation can be achieved by careful electrical balancing (nulling out) at one frequency. The implication of this feature is that the transmitter may be permanently connected to one port with no need for switching at high r-f levels, and with reduced losses; while the receiver preamp may be connected to the opposite sense port with a overload protection (crowbar) of modest requirements, again minimizing losses at the antenna-preamp interface where reducing losses is most beneficial to the overall operating system noise temperature. (See Tech. Report # 3 and # 11)
See also Tech. Rept. # 5 on limitations of port isolation for a front fed parabolic reflector antenna. See Tech. Rept. # 18 for an improved off-set fed reflector antenna system which has no limitation on port isolation.

Elimination of Faraday fading and of the need for a high-power T-R switch, as well as eliminating the need for linear polarization tracking and adjustment, surely justifies the use of circular polarization.

Reports # 2 and # 9 describe hardware methods of implementing circular polarization for an EME system.

** "Polarization of an Az-El Mounted Antenna Viewing Celestial Objects", IEEE Transactions on Antennas and Propagation, September, 1966, p. 650.

TECHNICAL REPORT # 2
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From: The Crawford Hill VHF Club, W2NFA
Date: 1963 (revised January 1987)

Subject: <u>Quadrature Hybrid</u>

In Technical Report # 1 on the need for circular polarization for <u>all</u> EME communication to avoid Faraday rotation fading and linear polarization rotation, this report describes a quadrature hybrid device which may be used to implement one method of achieving circular polarization. There are other methods which may be implemented (one such method is described in Technical Report # 9) which will not be described here.

<u>Circular Polarization</u>
A circularly polarized propagating radio wave is one in which the electric field, which defines the polarization is rotating (twisting) as it propagates forward. The angular (rotational) rate of rotation is simply 2 f. Where f is the frequency of the radio wave. This means that the field has rotated one complete revolution at the r.f. rate in one wavelength of propagating distance. Although the wave polarization is linear at any instant in time, viewed over any reasonable period of time such as one or more r.f. periods, the wave appears to have any polarization orientation. For communication purposes then, the radio wave is twisting so fast that it may be considered having all polarizations, i.e., circular polarization.
In actuality, if the r.f. wave is improperly launched or suffers distortion in propagation, the circularity may become modified into an elliptically polarized wave. Radio frequency waves may therefore propagate as pure linear polarized, pure circular polarized, or any form of elliptic polarized wave.

The method suggested in this report to obtain circular polarization employs two linearly polarized antennas arranged with their polarizations at right angles to each other (orthogonal), and their phase centers coincident. That is, the radiating elements lie in the same physical plane. Examples are crossed dipoles or nested crossed Yagis. It is also a fact of nature that linearly polarized antennas or fields that are at right angles to each other, are independent and do not couple to each other. A fortunate situation, for it means that each linearly polarized antenna may be adjusted, tuned and impedance matched independently of the other.
If then, each linearly polarized radiator is fed equal amplitude (power) but with a 90 degree difference in phase, the vector addition of the two radiated wave fields will always be equal in total amplitude but the rotational orientation will change at a rate comparable with the angular frequency = 2 f. This then is the need for a <u>quadrature hybrid</u> to feed the two orthogonal linearly polarized radiating antennas to produce a circularly polarized radaited wave. Quadrature implies a 90

degree differential between output ports, and hybrid means an equal power split between output ports. In this report only a 4-port quadrature hybrid will be described.*
To repeat, there are other methods of launching a circularly polarized radio wave. The methods suggested here has the feature of permitting both Left and Right circular polarization rotation to be implemented without physical changes or switching in the system.
By IEEE definition, a wave receding away from the antenna (transmit mode) which is rotating clockwise, is RIGHT CIRCULARLY polarized. While a wave receding away from the antenna which is rotating counterclockwise, is LEFT CIRCULARLY polarized.
It is also a fact of nature that a circularly polarized wave which is reflected by a conductive sheet will reverse its sense of rotation. This fact is of particular interest in parabolic (or any single reflecting) antenna system where the feed antenna must have the opposite rotational sense from the final radiated wave sense. More simply, if the feed antenna is arranged for RIGHT circular polarization, then the reflector radiated wave will be LEFT circularly polarized !

The Quadrature Hybrid

The device described here is a four port (port in this particular case meaning coaxial connections) matched network which performs the function of power splitting (3 db coupler, or hybrid) with a phase differential between output ports of exactly 90 degrees. The form of this hybrid uses closely coupled transmission lines. The operation of a matched (impedance matched) 4-port quadrature hybrid may be explained with the diagram below.

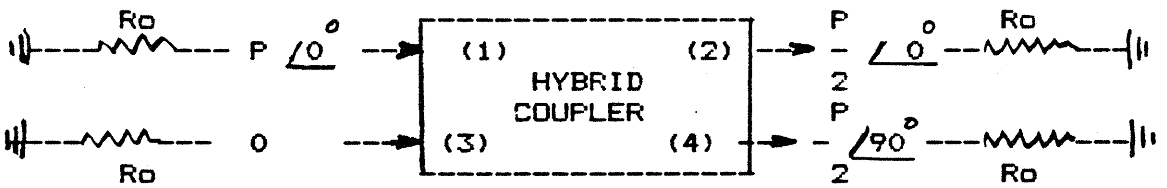
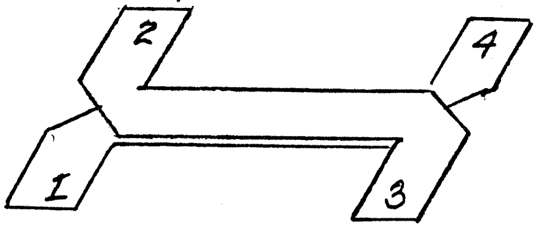


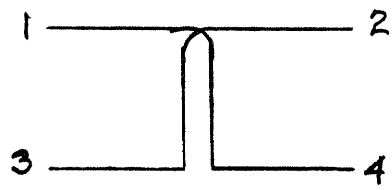
Figure 1

Ideally, if power enters port (1) as shown above, half the will appear at each of the two ports (2) and (4), and port (3) will receive no power (a null port). Note however, that the phase at ports (2) and (4) will have a 90 degree difference. In addition, because this network is a lossless, reciprocal, matched 4-port, it is also symmetric. That is to say, if power enters any one port there will always be a null port and the other two remaining ports will receive half the input power and with a 90 degree phase differential.
In particular we are interested in what happens when power enters port (3). In this case port (1) becomes the null port and ports (2) and (4) receive equal half power levels, but one with one very significance difference, the output port phases are reversed. Port (3) will now lag port (4) by 90 degrees.
This property of output phase reversal, dependent on which input

port is driven, is what makes the quadrature hybrid useful to implement independent LEFT and RIGHT circular polarization. The fact that the input ports are essentially isolated under impedance matched conditions performs the function of a T/R switch automatically. In practice, the isolation is difficult to obtain comparable with a good T/R relay, however, sufficient isolation < -20 dB is easily achieved permitting the transmitter to be permanently connected to one port, while the receiver preamp (connected to the other input port) will require a protective relay or 'crowbar' to prevent burn-out of the preamplifier.

The physical ordering of the port numbers is determined by its construction and cannot be assumed from external appearance. The ordering used and shown for this hybrid is the usual arrangement encountered and is also the most desirable since it permits the ports to be paired on opposite sides of the coupler.
To obtain this ordering of the ports, the coupled line structure is as shown below together with the symbolic designation of a quadrature hybrid.





Another feature of the coupled line hybrid is that the phase differential between output ports is by definition 90 degrees, but the power output balance must be trimmed (adjusted) to obtain a good - 3 dB split. In addition, the isolation between input ports is also dependent, not only on the impedance match at the output ports but also the electrical balance of the coupled lines. Adjustment procedures are described in the section on construction notes.

Hybrid construction

The drawings at the end of this report contain sufficient information for the construction of three different coupled line quadrature hybrids. The choice of which one to build is left to the reader and materials available. The first two designs, I and II are symmetrical in physical construction and complete dimensions are not always shown. The striplines are always centered between the ground planes. No critical machine work is required and the dimensions which are important are the spacing between the coupled lines and the line section lengths themselves. N-type panel receptacles are used for all ports.

Design No. I

This design is very rugged mechanically and provides edge mounting of the N-type receptacles, which is desirable for good impedance transition. When completed, the unit does not include sidepanel shields which may be added. Nylon screws are used to hold the strips centered. A good substitute can be made from strips of low density expanded foam commonly found in commercial packaging. The small capacitor tabs on this unit are soldered to the strip lines and can be adjusted for best directivity (null between input ports) and best balance between output ports. In general directivity of less than - 20 dB is desirable, while balancing of the output ports is of secondary portance.

Design No. 1

This unit is built around commonly available copper clad printed circuit board. It is available from many sources and is not necessary to get high quality, low loss material, since the functional part is only the copper inner surface. Many sources of 'mud' PC boards are available. A thicker board can provide better mechanical stability to the unit.

The boards are conveniently held apart by 20 - 1/4 inch diameter by 1/2 inch long commonly available spacers. They may be mounted with screws extending through the center hole or tapped at both ends for short screws.

The stripline sections are best cut as a pair with a nibbling tool or other cutting device which does not warp or distort the edges. Hand shears tesd to warp the sheet as it is cut, requiring extensive work to flatten the material.

Drill the copper clad PC boards as a pair for good registration. The N-type receptacle is soldered directly to the copper cladding around its inside shoulder, as indicated, to provide food electrical continuity with the ground plane (inside plane). The small copper tabs soldered to the ends of the strip lines for adjustment may be adjusted empirically with a non-metalic tool, reaching through the spacers. The adjustment procedure is the same as with the first design No. I. This unit provides all the ports on one side of the unit, although they may be arranged in other fashion on either side of the unit for convenience in application, to shorten inter-connect leads. The unit may also be enclosed on the sides for environmental protection.

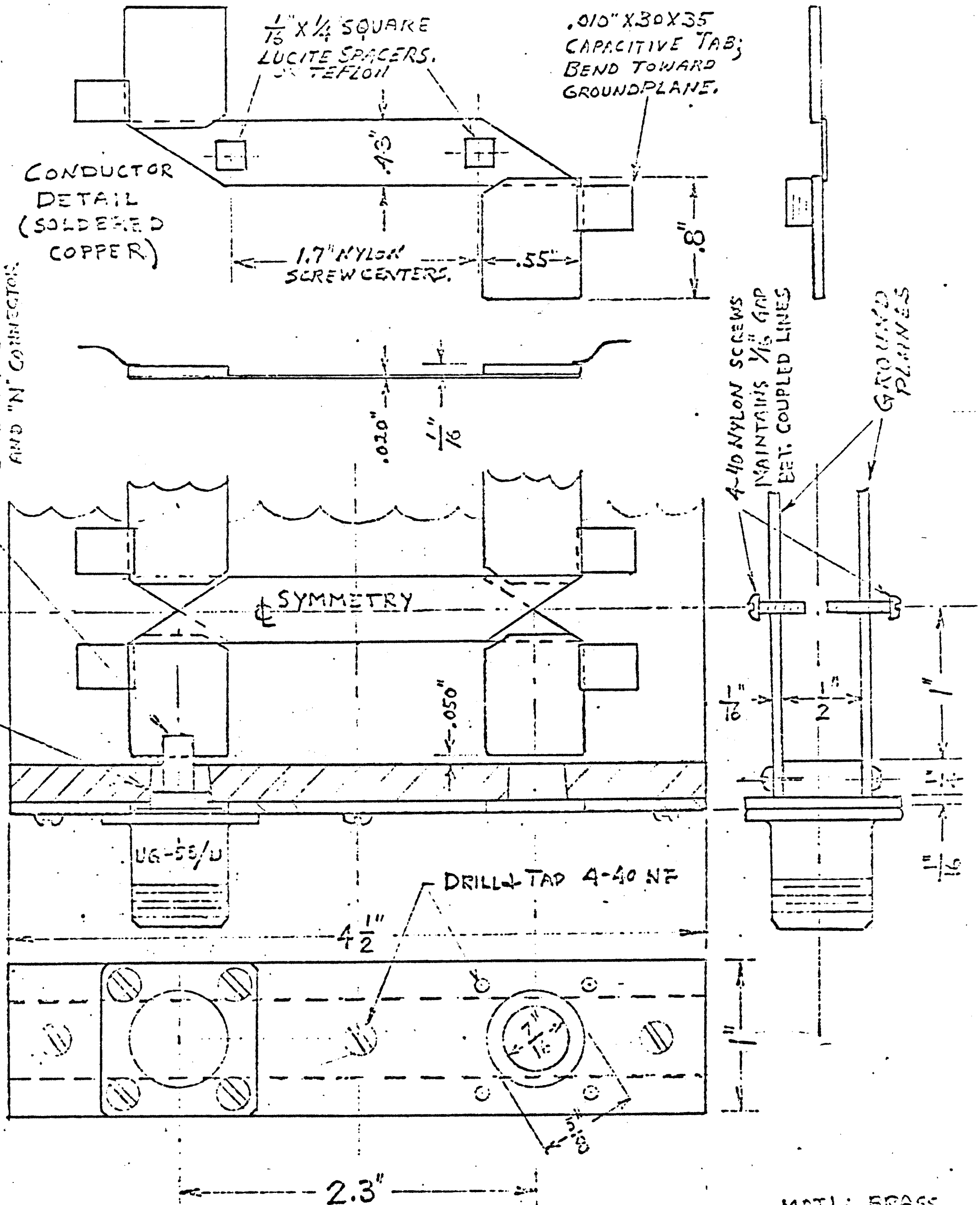
Measuring and Tuning

Upon completion of the coupler hybrid, it is highly desirable to make sure that the device is working properly by making some measurements and adjustments. The by-word or credo of all home made hardware (and even bought or borrowed hardware) is, DO NOT trust anyone elses word, measure it to your own satisfaction. Many well designed electronic systems have been reduced to utter disaster by interfacing hardware which was erroneously specified. In order to make measurements on your hybrid, it will be necessary to have some equipment. A low r-f power source (a watt or less) at 1296 mc/s, a pair of matched terminations, see Report # 14, and a suitable r-f detector which can, in conjunction with your r-f source, measure at least a 20 dB range. A set of calibrated attenuators would be nice but are not necessary. Any 50 ohm attenuator with at least 10 dB loss is also required.

Such an attenuator may be made of a long section of RG 8 coaxial cable, but is expensive.
The measurement procedure is to connect the source to port (1), a 50 termination to the null port (3), and a matched 50 ohm termination to one output port, say port (2). The 10 dB attenuator and detector are then connected to the other output port (4). Making sure that the measuring system is not saturated, read the output r-f level at port (4).
Next switch the output ports, and note the r-f reading again. If these readings are withing less than 0.5 db, the output balance is for most purposes satisfactory.
If you have no calibrated attenuator or method to measure differences less than 1 dB, sections of RG 8 or better still RG 214 coaxial cable may be used as fixed attenuators. Consult handbook tables of cable loss per hundred feed as a guide.
The output port balance may be adjusted by changing the spacing between the coupled line sections. Make every attempt to keep the spacing uniform along the coupled section.
When the output balance is within a few tenths of a decibel, the idrectivity or null port may be checked by first placing both matched terminations on the two output ports and the r-f detector (with of without the attenuator) on the null port (3).
It should now be possible to adjust (bend) the capacitive tabs at the ends of the coupled line sections until a good null is obtained on the r-f detector at port (3).
A measure of the actual coupler directivity may be made if r-f calibrated attenuators are available. The procedure is to first obtain a measure of the r-f source power output with at least a 30 dB attenuator (pad) between source and detector. Then connect the source directly to port (1). Now connect the detector with appropriate attenuators to measure the ratio between the total input power to the actual power at the null port (3).
If the ratio of these two levels is greater than 20 dB, the directivity can be considered satisfactory for most purposes.
Careful adjustments can result is much better directivity.
If all active parts are carefully smoothed and free of sharp burrs and corners, this coupler should perform satisfactorily at power levels at least through 500 watts.
These couplers tend to be correct at one frequency but are not particularly critical or vry narrow band.

Reed Fisher, W2CQH has developed similar coupled line transformers for lower power using tightly twisted insulated wire augmented with capacitors, which give excellent results up to the UHF range.
Similar quadrature hybrids can also be built using similar techniques wound on Ferrite toroidal cores which also give excellent results through any region of the radio frequency spectrum.

QUADRATURE COUPLER 1296 mc
3db - 90° HYBRID - 50 OHM

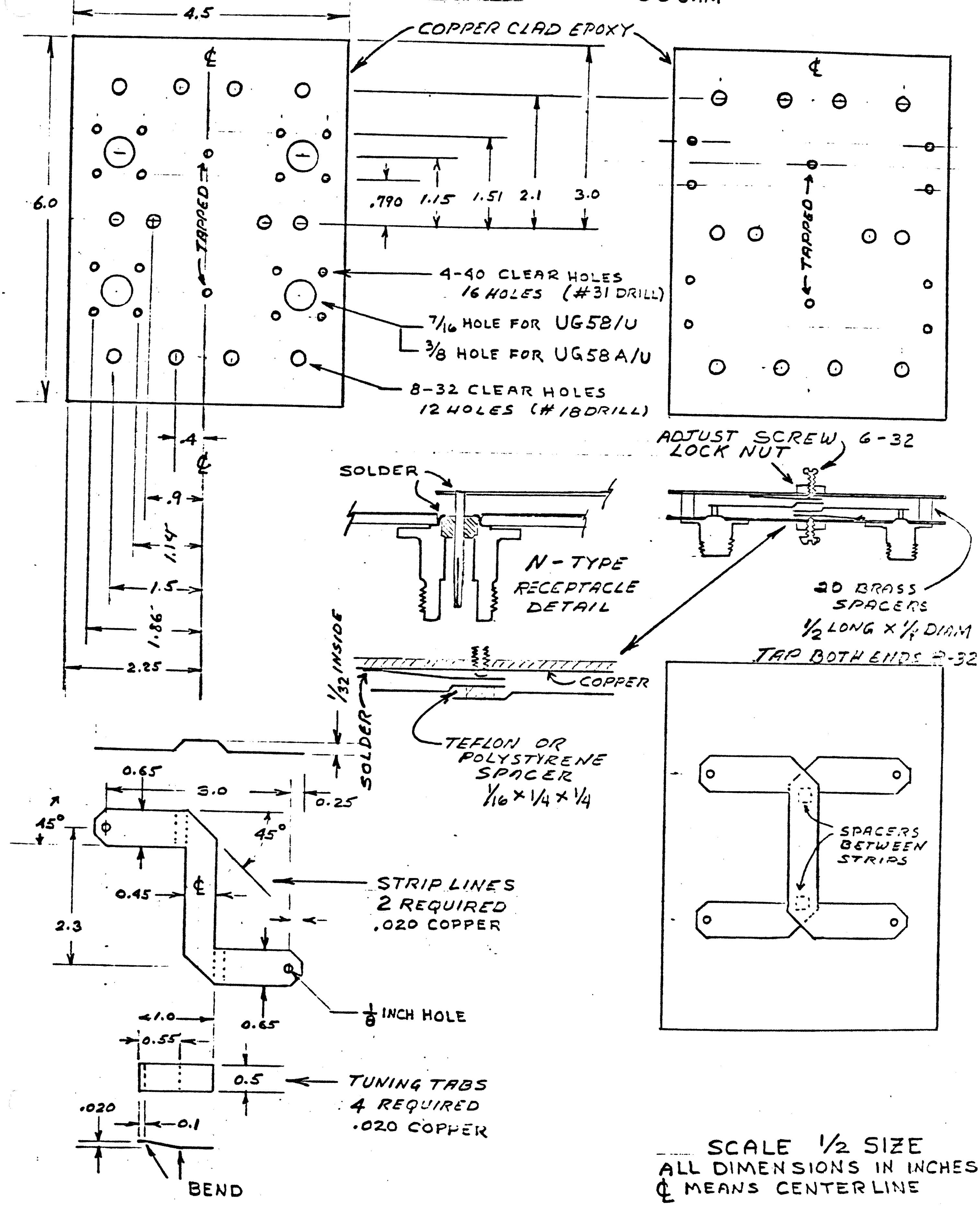


λ/4 = 2.3"

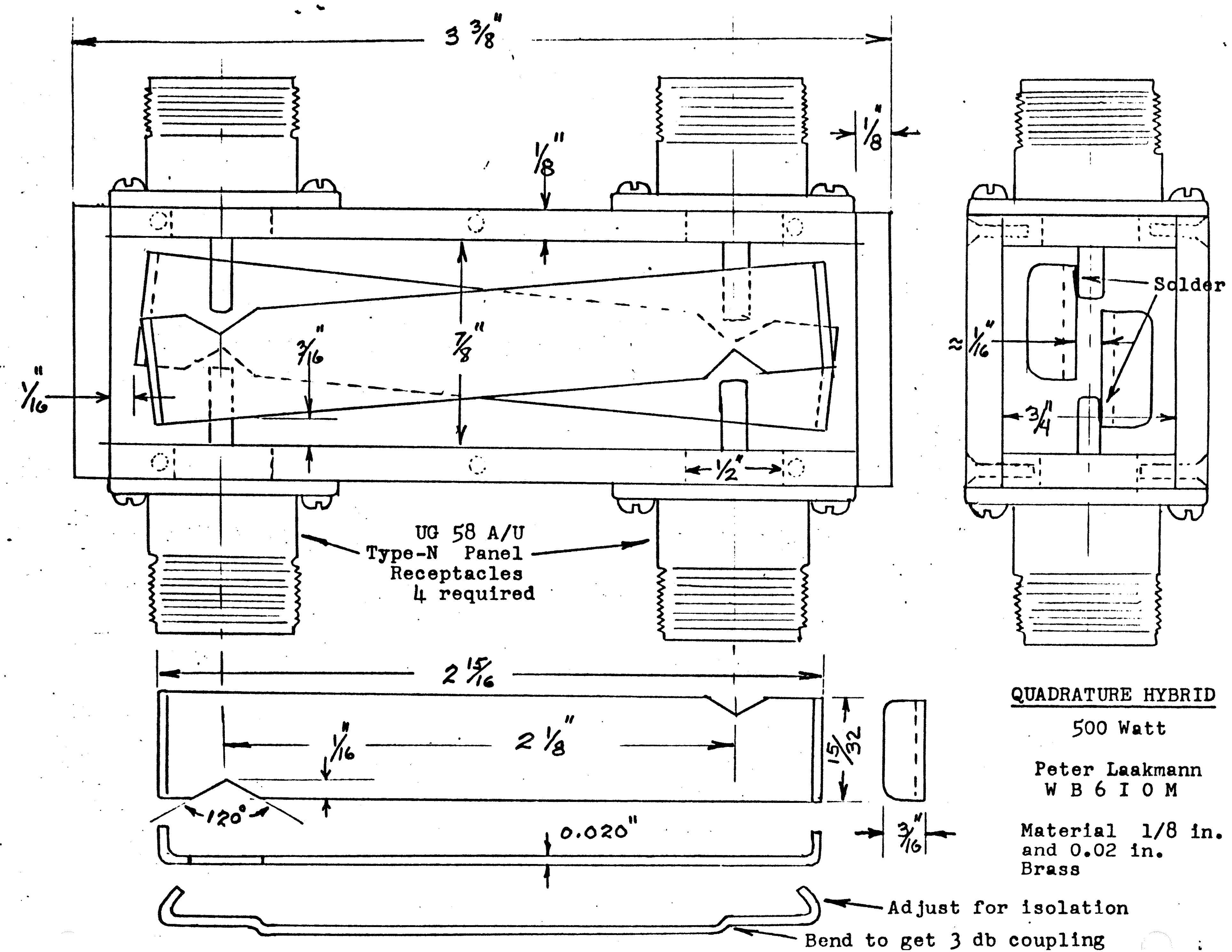
MATL: BRASS
FULL SCALE

QUADRATURE COUPLER 1296mc

3db - 90° - HYBRID - 50 OHM



SCALE 1/2 SIZE
ALL DIMENSIONS IN INCHES
℄ MEANS CENTERLINE



QUADRATURE HYBRID

500 Watt

Peter Laakmann
W B 6 I O M

Material 1/8 in.
and 0.02 in.
Brass

Technical Report # 3 #3

From: The Crawford Hill VHF Club
Date: (1970), slightly modified Oct. 1986

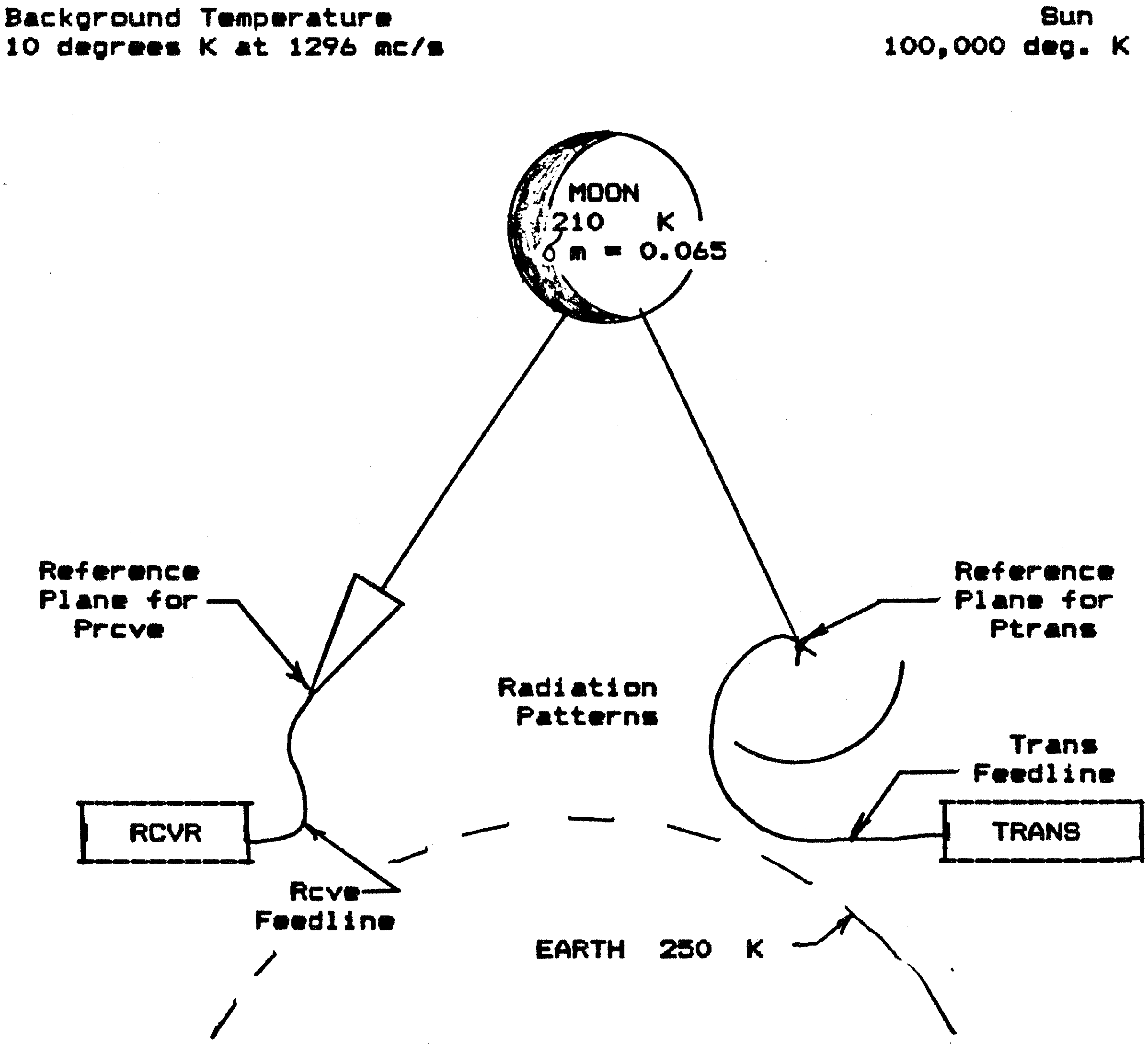
Subject: System Consideration for the E M E Path

Of all the radio propagation paths used by radio amateurs, few can be accurately predicted in terms of path loss and dependability. In the VHF-UHF spectrum, only the line-of-sight path which includes the active and passive (EME) satellite paths, can be reliably predicted since most of the path is in free-space. For the Moon as a passive satellite, the amount of reflection from the Moon has been adequately documented, and the use of high-gain antennas virtually eliminates ground reflection problems on the EME path.
By contrast, all point-to-point VHF-UHF Earth surface paths commonly used by amateurs make use of some freak of nature such as sporadic-E, ducting, temperature inversions and fronts, Aurora, meteor trail scattering, etc., which are unpredictable, sporadic and not at all useful for 'real time' amateur communication. Even the so called line-of-sight paths generally involve ground reflections which though constant for a given situation, are difficult to predict.
The EME path is particularly noteworthy for UHF because of the accuracy with which path loss calculations can be made and for its dependability. These factors permit the EME path to be used for reliable DX communication and as a means of antenna and system evaluation.
The great distance to the Moon (high path loss) and the Moon's poor reflective properties make this propagation path appeal to those interested in advanced work on receivers, transmiters and antennas for the UHF spectrum. Indeed, because of the great path loss involved and the license power limitations, every part of the station facilities must be optimized to achieve success. The first major step in individual station success occurs when reception of echoes is achieved. The echo property of the EME path is unlike any other propagation path, except cross-band active satellite paths, since achievement of echoes means that your station is now capable of communicating with any other station that can receive echoes.
The EME path therefore provides a challenge for the amateur who wishes to upgrade his technical ability and explore new horizons in DX communication.

In this report, the total system aspect of EME transmission will be considered briefly. The system includes both the transmitting and receiving equipment, as well as the EME (Earth-Moon-Earth) path. Little will be indicated about the components of the system except their characteristics as related to the system. For instance, the single important system characteristic of the transmitter is its power output; it is assumed the frequency stability, and r-f purity, d-c input limitations, etc., are all

satisfactory and meet license requirements.
No modulation is considered, and the system operates CW with the final detector being the human ear.
In presenting this material, a number of objectives are indicated; a better understanding of the total system, to bring into focus those characteristics of a system which are most important, to indicate where trade-offs can be made in an individual system, and to provide <u>all</u> the relations necessary for complete system evaluation on a mathematical basis.

THE E M E SYSTEM



To illustrate the EME system, a one-way path is chosen and shown by the above drawing in which a paraboloidal reflector type antenna is shown for transmitting and a horn type antenna is shown for receiving. The components of the system are characterized as follows:

Po is the transmitter power output,
Lt is the transmitting feedline loss,
Gt is the effective gain of the transmitting antenna,
Gr is the effective gain of the receiving antenna,
Ta is the receiving antenna temperature in degrees Kelvin,
Lr is the receiving feedline loss,
Tl is the temperature of the receiving feedline in $^\circ$ K,
and Tr is the receiver input temperature in $^\circ$ K.

In addition, the EME path may be characterized by:

R is the distance between the center of the Moon and the center of the Earth (238,636 s. miles +/- 1 db),

dm is the diameter of the Moon (scattering object) 2160 s. miles,

λ is the free space wavelength for the desired operating frequency, in statute miles, to be consistent with R and dm units, and

δ_M is the reflection coefficient of the Moon (0.065 in the VHF-UHF portion of the spectrum).

Using the antenna terminals as convenient reference points for calculating power levels, the total path loss between <u>antenna terminals</u> is

$$\text{Path Loss} = \frac{\text{Prcve}}{\text{Ptrans}} = \text{Gt Gr} \left| \frac{dm^2 \lambda^2 \delta m}{(16\pi)^2 R^4} \right| . \qquad (1)$$

The bracketed term is the 'free space' path loss between antenna terminals with Gt = Gr = 1 (isotropic antennas), and is tabulated below for convenience,

Frequency in mc/s(mHz)	Free Space Path Loss +/- 1dB between isotropic antennas
50	- 242.9 dB
144	- 252.1 dB
432	- 261.6 dB
1296	- 271.1 dB
2300	- 275.5 dB

The +/- 1 dB variation is to account for variations in the range R, since the Moon orbit is not exactly circular. At 1296 mc/s where atmospheric and ionospheric attenuation is almost always negligible, the free space path loss of - 271.1 dB is quite accurate for all cases where the antenna is uneffected by the presence of the Earth (high elevation angle). The purpose of these calculations is to show that if all components of the system can be accurately characterized, then an accurate determination of signal-to-noise ratio at the receiving

#3

detector can be made. And this in turn determines whether of not the system is adequate for EME communication.
The procedure employed here to determine S/N is: (a) compute the received power level, Ps, at the receiving antenna terminals given the antenna gains, transmitting power and path loss, (b) next a system noise power, Pn, referred to the receiving antenna terminals is computed based on the overall system operating temperature, Tsys, and the receiver effective noise bandwidth, and finally (c) form the ratio of Ps-to-Pn, which is the signal-to--noise ratio. It should be mentioned here that when measuring the output of a receiver the ratio of (S+N)/N is actually measured. (A convenient curve relating S/N to (S+N)/N can be found in Tech. Report # 11.)

In the calculations to follow, the receiving system effective noise bandwidth employed will be 50 cycles per second (Hz). This is based on human 'ear' detection and does <u>not</u> imply that the receiver have a 50 cps filter included. It has been adquately demonstrated that the receiver bandwidth can be as great as 3 kc/s without impairing the ability of the ear-brain filter-detector from detecting a single CW signal in noise. It has also been adequately demonstrated that a signal-to-noise ratio of unity is extremely marginal for the ear-brain system to detect. For reliable CW communication, a minimum S/N of + 6 dB is desirable (ear-brain 50 cps in a 3 kc/s band of noise). The reason for selecting the ear-brain system as the final detector is that the ultimate objective of the EME program is 'real-time' communications. While more sophisticated detection processes can be employed to dig the signal out of the noise, the information rate will suffer (slower keying speed).

<u>System Noise</u>

The system noise power <u>referred to the receiving antenna terminals</u> is

$$Pn = k \; Tsys \; B \qquad (2)$$

where k = 1.38 x 10E-23 watts/cycle-°K, Boltzmann's constant,
B = 50 cycles per second (Hz), the receiving system effective noise bandwidth,
and, Tsys is the system operating temperature in degrees Kelvin.

Tsys includes the antenna noise temperature as well as the overall receiver noise and any interconnecting line losses between antenna and receiver.

$$T_{sys} = T_a + (L_r - 1)\,T_l + L_r\,T_r \qquad (3)$$

where the various terms have been defined on page 2. Lr, the receiving feedline loss is a pure number greater than 1 in this context. For a loss of 1 dB, Lr = 1.26. Loss in decibels (dB) = 10 log Lr.

Tsys therefore acounts for all the noise in the system referred to the receiving antenna terminals rather than the receiver front end, arbitrarily and for convenience.

Tr, the receiver overall noise temperature, preamp plus receiver system, is related to noise factor, NF, by

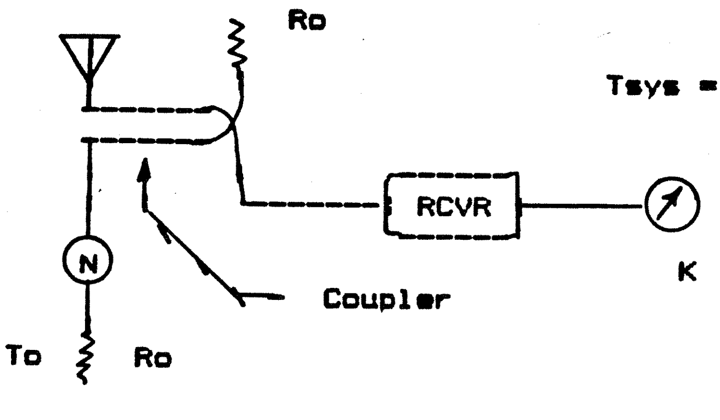
$$T_r = (NF - 1)\,290^{\circ}K \quad \text{(IEEE definition)}$$

A good estimate of overall receiver noise factor can be made from just knowledge of preamplifier NF and gain, and post amplifier NF. This two stage amplifier has a combined NF of

$$NF_{sys} = NF_{preamp} + (NF_{post\ amp} - 1)/G_{preamp}.$$

Ta, the antenna noise temperature is a measure of the noise power which the antenna picks up from sources such as the warm Earth or noisy regions of the Galaxy and should not include man-made noises. Ta varies with the physical positioning of the antenna and is usually a difficult quantity to determine (see Tech. Reports # 5 and # 11).

Fortunately, Tsys may be measured directly by introducing a calibrated noise source into the feedline at the antenna terminals by means of a weakly coupled directional coupler, with coupling factor less than − 20 dB.



$$T_{sys} = \frac{C}{K - 1}\,(T_n - K\,T_o)$$

Where C is the coupling coefficient of the directional coupler,
TN is the "ON" temperature of the calibrated noise source,
To is the temperature of the noise source termination Ro,
and K is the ratio of the receiver output noise power for conditions of noise source "ON" to "OFF". This scheme is very

Sample of System Calculations for 1296 mc/s

It is convenient and customary to compute power levels in milliwatts (mw.), and also relative to one milliwatt when using decibels. By definition then, a power level of one milliwatt is referred to as 0 dBm.

To illustrate a system calculation, assume that all the necessary numerical characteristics of the system are known with some degree of accuracy. The ultimate question is, can a signal be detected on a one-way transmission basis using audible detection methods (your ears). Let:

P_o = 250 watts or + 54 dBm (transmitter output power)

L_t = 1.26 or 1 dB of feedline loss for transmitting

G_t = + 32 dB (transmitting antenna gain - this is equivalent to a fully illuminated 10 foot circular aperture)

G_r = + 32 dB (receiving antenna gain)

T_a = 20° K (receiving antenna temperature - typical of a good horn antenna aimed near the zenith at 1296 mc/s)

L_r = 1.26 or 1 dB receiving feeline loss at a temperature of 290 K (room temperature)

T_r = 75.4 K, equivalent to a 1 dB preamp-post amp temperature. $T_r = (1.26 - 1)\ 290^\circ$ K

Calculate the system operating temperature using equation (3):

$$T_{sys} = (20 + (1.26 - 1)\ 290 + 1.26\ (75.4))\ \text{degrees Kelvin}$$
$$= 20 + 75.4 + 95$$
$$= 190.4^\circ \text{ K}$$

Calculate the receiving system noise power, P_n, referred to the receiving antenna terminals using equation (2). A more convenient form of this equation is

$$P_n = k\ T_{sys}\ B = (k\ T_o)\ \frac{T_{sys}}{T_o}\ B, \quad \text{where } T_o = 290^\circ \text{ K.}$$

In decibel form (k To) computes to be − 174 dBm/cycle, and

$$Pn \text{ (in dBm)} = -174 \text{ dBn/cycle} + 10 \log_{10} \left(\frac{Tsys}{To}\right) + 10 \log_{10} (B)$$

For B = 50 cycles (Hz) and Tsys = 190.4 K,

$$Pn \text{ (in dBm)} = -174 + 10 \log (0.656) + 10 \log (50)$$
$$= -174 - 1.82 + 17$$
$$= -158.8 \text{ dBm}$$

Now consider the received signal power level in dBm

Ps (in dBm) = + 54 dbm − 1 db (feedline loss) + 64 dB (antenna gain) − 271 db (path loss)

Ps (in dBm) = − 154

Finally, the S/N ratio

$$\frac{Ps}{Pn} \text{ (in decibels)} = Ps \text{ (dBm)} - Pn \text{ (dBm)}$$
$$= -154 \text{ dBm} - (-158.8 \text{ dBm})$$
$$= +4.8 \text{ dB !}$$

Since the S/N is only 4.8 dB, this system must be considered to be marginal for communication, but detectable signals should be obtained.

Let us now remove the feedline losses for both receiving and transmitting, and recompute the S/N. This situation would occur perhaps in a station where a common feedline would be used between feed antenna and the equipment, and the signal were of echoes. In this case

$$Tsys = (20 + 75.4) \quad K$$
$$= 95.4 \quad K$$

$$Pn(dBm) = (-174 - 4.8 + 17) \quad dBm$$
$$= -161.8 \text{ dBm}$$

$$Ps(dBm) = (+54 + 64 - 271) \text{ dBm}$$
$$= -153 \text{ dBm}$$

and,

$$S/N \text{ (dB)} = -153 - (-161.8) \text{ dB}$$
$$= +8.8 \text{ dB !!}$$

An increase of 4.0 dB in S/N by removing a 1 dB (common to both Tx and Rx) lossy feedline. This surprising improvement was only possible because both the receiver and antenna temperature were low. This also illustrates the sensitivity of losses in the system, especially between the antenna and preamplifier.

It is interesting and obvious to note that nowhere in the system S/N calculations does the transmitting antenna temperature appear. This leads to the obvious conclusion that for transmitting, the antenna aught to have maximum gain, regardless of other factors. However, for receiving, the antenna temperature is an important consideration as well as its gain. In general, if the receiver temperature is known and receiver feedline losses are very small, then an antenna temperature comparable or slightly smaller than the receiver temperature is desirable.

An antenna quality factor for maximizing signal sensitivity is the ratio Geffective/Tantenna. The higher G/T, the better the antenna will perform in receiving signals, especially weak EME signals. This paradox in antenna requirements can be accomodated in the case of the parabolic reflector antenna by providing a feed which has adjustable illumination properties.

<u>Summary</u>

Since the maximum transmitter power is fixed by license regulations, it is highly desirable to take advantage of the low background temperature of the Universe at 1296 mc/s and higher, and employ receiving antennas with low temperature and the best state-of-the-art preamplifiers to improve S/N.

TECHNICAL REPORT #4

FROM: The Crawford Hill VHF Club - W2NFA June 1969

SUBJECT: A Horn Antenna for EME Communication

This report will consider a horn antenna for 1296 mc EME work. Comparisons will be made with the more popular paraboloidal reflector antenna. The purpose of this report is to present all the characteristics of the most simple horn antenna in order that the reader may be better able to decide which antenna is more suitable for his requirements.

The particular horn antenna presented here is a square aperture pyramidal horn with a square waveguide feed. The square aperture and feed are chosen so that circular polarization of the radiated energy may be achieved at 1296 mc by the method of crossed linearly polarized feeds which are energized 90 degrees out of phase electrically by means of a quadrature hybrid.

The following information is for an optimum horn design. An optimum horn is by definition one in which the length is fixed and the horn is permitted to change in flare angle (aperture size) until the gain is maximum. The optimum condition is therefore a practical compromise which results in the minimum over-all length for a desired gain. This horn is therefore the most simple and practical case. Variations using focusing reflectors or aperture lenses are not simple and will not be considered here. Horns of different geometry such as conical horns or horns flared with curved walls cannot improve the gain over the optimum design for the same length horn and so they are also omitted. Figure 1 shows the square aperture horn design considered here.

Any antenna for 1296 mc EME work may be characterized by the following factors, gain, radiation pattern, polarization, bandwidth, impedance match at the feed, physical size, construction tolerances, noise, temperature, and cost. These factors will now be considered.

<u>Bandwidth</u>: Horn antennas are basically high pass filters by virtue of the waveguide like construction and have a lower cut-off frequency determined by the smallest inside dimension of the waveguide feed. A horn is therefore a very broadband device and susceptable to out-of-band interference. With the 6 inch square waveguide feed shown in Figure 1, the cut-off frequency is 985 mc. A further modification of the bandwidth characteristics is obtained from the method of exciting the waveguide feed.

<u>Impedance match</u>: This is a circuit problem and does not belong to the antenna proper but since the waveguide must be excited by a device or other type of feedline with different characteristic impedance, a suitable transition device or coupler must be employed. Our requirements at 1296 mc are very narrow band and so there are many ways to effect a matched coupler from a feedling to the dominant mode impedance of the waveguide. It should be done however in as low loss a manner as possible in order not to degrade the gain and noise temperature. A suitable coupler to go from coax line to waveguide will be described later.

<u>Gain and size</u>: These important factors are related by the design curves shown on Figure 1. For a desired gain, the horn length and aperture size are read directly from the curves. For example, a 30 db gain horn is 50 feet long and has an aperture size of 10 by 10 feet. These dimensions are strictly for 1296 mc. The gain values are referenced to an isotropic or point source radiator. Gain is simply defined as the ratio of radiation intensity (power if you like) in a particular direction usually the main beam of a directive antenna, to the radiation

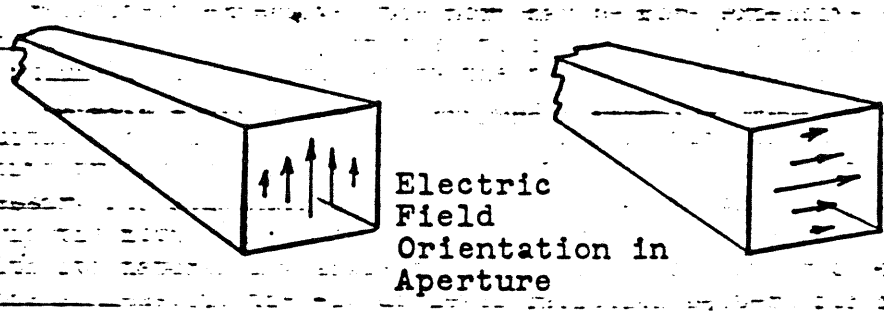
intensity from an isotropic antenna with the same input power. An isotropic antenna is a ficticious antenna which radiates equally in all directions of space.

Now a slight digression will be made into aperture theory for those who desire more insight. For a particular aperture size (area), the maximum possible realizable gain which can be achieved is numerically equal to $4\pi A/\lambda^2$, where A is the aperture area in the same length units as (λ) Lambda the free space wavelength, 9.13 inches or about 3/4 foot at 1296 mc. Furthermore, a term called aperture efficiency is used to indicate the ratio between the actual realized gain of an aperture type antenna and the gain value computed from the above formula. An optimum horn achieves just about 50 percent aperture efficiency for a wide range of gain values. This value of aperture efficiency has already been included in the design curves of Figure 1. The optimum horn 50 percent aperture effeceincy is due to both the distribution of energy over the aperture area resulting from the constraintof the dominant mode waveguide feed, and to the phase error over the aperture resulting from the horn flare angle. If the phase error over the aperture were corrected by means of a loss less lense or focusing-reflector, or the horn length increased many many times longer than optimum, the aperture efficiency would increase to a maximum of 80 percent, the remaining 20 percent loss in aperture efficiency is due simply to the amplitude distribution of the dominant mode field over the aperture.

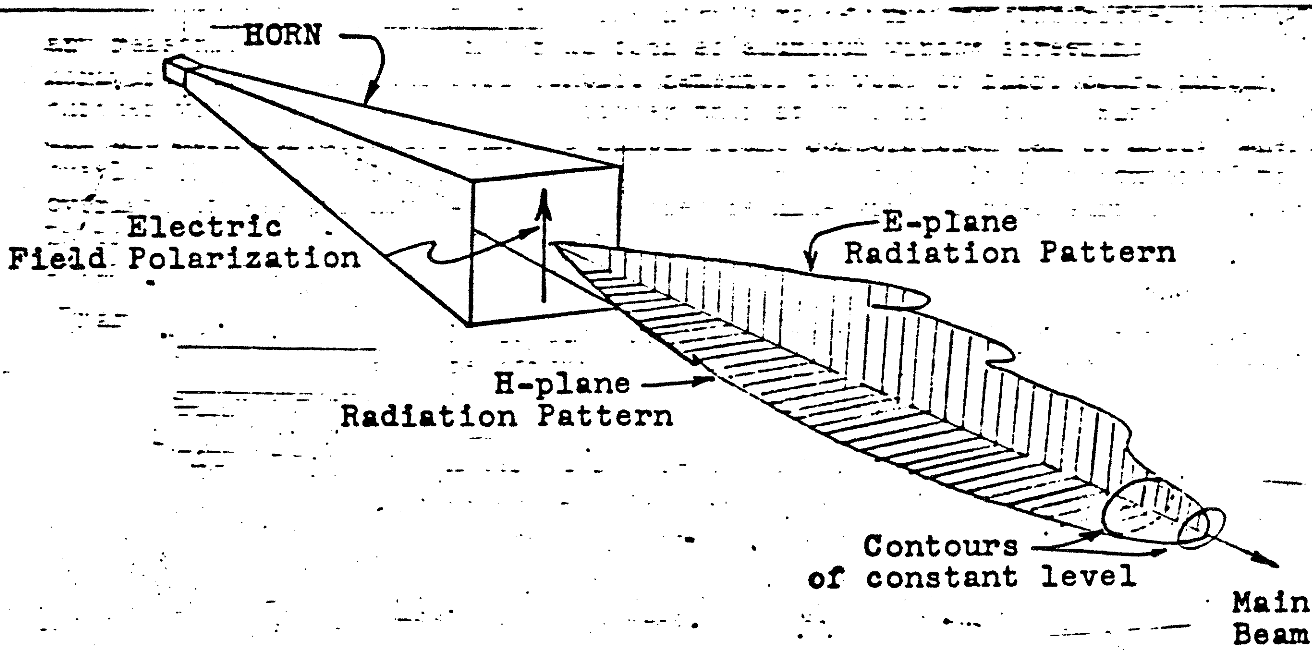
This report is too brief to consider other more complicated forms of higher order mode excitation of the horn aperture. Therefore, it is stressed here that this entire analysis is based on constricting the smaller end of the horn so that only the fundamental or lowest order mode, TE_{10}, is excited. Because of the complexity of exciting controlled higher order modes, virtually all practical application handbooks treat this fundamental case only.

Polarization: Let it be understood that the gain values read from the curves are the same for linear or circular polarization provided that the antennas used for comparison are all consistent in polarization. It is desirable to use circular polarization for all EME work, and this will be assumed for all practical cases. However, it is sufficient to consider the linear polarization characteristics of an antenna and then by geometric construction arrange the antenna so that it will have the same characteristics when excited by a linearly polarized wave which is oriented 90 degrees from the first in space. This is why the square aperture was chosen. It is geometrically the same when rotated 90 degrees about its axial center. The polarization scheme suggested in Technical Report #1 is implicit.

Radiation Pattern: Figure 2 shows universal radiation patterns normalized with respect to the aperture size. The patterns are for the two principal planes and for linear polarization. It is customary to refer to the principal planes of radiation as those planes which are parallel with the electric and magnetic fields, E-plane and H-plane respectively. These planes are always at right angles to each other with their intersection line usually along the direction of the main radiation beam. The E-plane is also the reference plane of linear polarization and so the radiation planes are intimately related to the aperture field orientation. For the square aperture horn considered here, the electric field orientation in the aperture can be either of the two cases shown as follows:



The principal planes E-plane and H-plane for which detailed radiation patterns are given by Figure 2 for linear polarization, are illustrated below. The actual radiation beam is sort of cigar shaped in volume while the drawing below only shows an axial slice of the pattern in the principal planes. Only half of the patterns are shown below and by Figure 2 since the full pattern is symmetric on either side of the center of the main beam.



One of the most important characteristics of an antenna radiation is its half power (-3 db) beamwidth. This is the angle within which half of the total radiated power is contained. This angle is important because it is a measure of the antenna gain for high gain antennas especially.

For circular polarization an approximate half power beamwidth angle related to aperture size is given under Figure 2.

A logical question which may arise is of what use are the linearly polarized radiation patterns when the antenna will in all probability be used only in the circularly polarized mode. The reason for not including circularly polarized radiation patterns is that they are not available for this particular horn geometry and are difficult to compute exactly. However, for practical purposes the approximate circularly polarized mode radiation patterns can be assumed to be similar to the E-plane case for linear polarization. The radiation can only be circularly polarized over a limited region of space when the proper feed conditions are met to obtain circular polarization at the center of the main beam. For the optimum horn as with most other beam antennas, circular polarization with tolerable ellipticity can be obtained from the center of the main beam to where the E and H plane linear polarized radiation patterns differ in amplitude. This occurs at about -5 db level for the case considered here. At other directions in the space pattern all degrees of ellipticity are possible. This bit of information is of little practical significance for EME since the target (moon) will be illuminated by only a small portion of the main beam and so will be totally illuminated by circularly polarized radiation.

Antenna Noise Temperature: Since a large horn has very low side and rear lobe radiation, the noise temperature of the antenna will be essentially the temperature of the region into which the main beam is pointed. Therefore, at 1296 mc, a large aperture optimum horn will have an effective temperature of no greater than about 20 degrees Kelvin when pointed away from the warm earth. This is very small when compared with a typical paraboloidal antenna which may have a noise temperature of 75 to 100 degrees Kelvin when aimed in the same direction. The reason for the higher noise temperature of a paraboloidal antenna is that the feed for the paraboloid is looking in the opposite direction which is towards the warm earth and spill-over or radiation beyond the outer periphery of the paraboloid will be directed towards the warm earth and so contribute noise. In the horn the feed is essentially imbedded in the horn and can contribute nothing extra to the antenna noise.

It should be kept in mind, however, that unless a very low noise receiver is employed with the horn, the system operating temperature may not be significantly improved. See Technical Report #43. For instance, a typically well constructed parametric amplifier will have a noise temperature of about 100 degrees Kelvin. The improvement in sensitivity using a horn with 20 degrees as compared with a paraboloid with 100 degrees will be the ratio of 100° + 100°/100° + 20° or 2 decibels.

When the horn antenna is directed parallel with the surface of the earth, half of the radiation beam will be directed toward the warm, 300 degrees Kelvin, earth and the other half into a relatively cold background temperature at 1296 mc. This means that the effective antenna temperature for this particular pointing direction can be no less than 150 degrees.

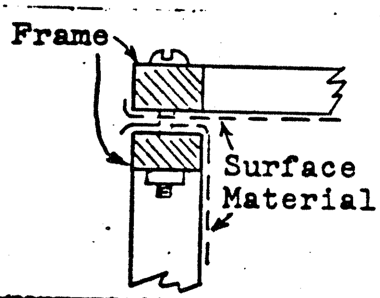
Comparison Between Optimum Horn and Paraboloidal Reflector Antenna: It is interesting and coincidental that the aperture efficiencies of both the optimum horn and paraboloidal antennas are about the same, 50 percent. Because of this, some useful comparisons can be drawn since both antennas will have the same physical aperture size for a given value of gain. Although one is circular in aperture and the other square, each will have the same number of square feet in area.

The paraboloidal surface must be constructed with great accuracy and maintained accurate whereas the walls of the horn may be constructed an order of magnitude (10 times) less accurate without impairing the horn efficiency. However, for the same gain, about 8 times more surface material is required with the horn. Since conductive material (screening, etc) is relatively expensive, the horn may be more expensive but permits more simple construction techniques to be employed. Another consideration is the volume or space occupied by the horn as compared with a paraboloid.

A horn lends itself to low angle radiation while a paraboloid is most suitable for high angle pointing. This latter consideration presents a mounting problem but has interesting implications. Low angle radiation is necessary for long haul DX work but does not restrict short haul or local paths via the moon. High angle radiation restricts the long haul path, the horn with its natural low angle mounting must also have good foreground clearance at least to the -10 db level radiation pattern for reasonably low noise performance.

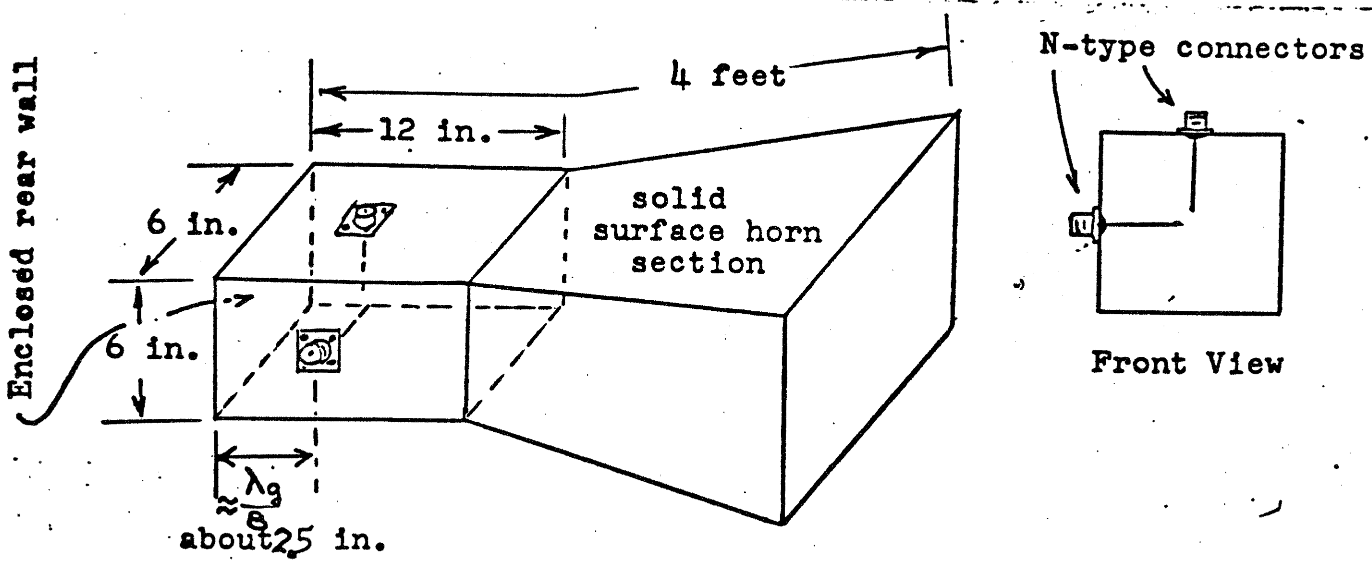
One of the greatest advantages of the horn is that the feed point is readily accessible near the ground so that equipment may be mounted directly at the feed with little or no feedline losses. Although feed mounted equipment can be physically realized with a front fed paraboloid, the equipment is directly in the aperture area causing blockage and scattering in addition to being difficult to reach for adjustment.

Construction Comments: The horn surfaces may be constructed of any reasonably conductive material such as aluminum window screening, hardware cloth, aluminum foil contact cemented to wood or fibre board panels, chicken wire with hole sizes no loarger than about 1 inch, etc. Since all the horn surfaces are flat, a simple frame construction can be used. Where overlapping of material is necessary and at frame joints it is recommended that overlaps of essentially $\lambda/4$ or 2.25 inches be made with clamped surfaces. In the horn corners the surfacing material may be brought over the framing around the frame so that the overlap is compressed between frames.



It is also suggested that the small end of the horn be constructed entirely of sheet material well bonded at all seams and joints with similar overlapped and clamped areas where joints between sheet and other material are required. The length of this first solid sheet section should be at least several wave-lengths or perhaps a minimum of about 4 feet long.

The square waveguide or feed section should be made of heavier material since in all probability a transition from waveguide to coax-line will be required. The heavier wall permits direct mounting of a type-N receptacle for probe excitation or the guide. This is the most simple form of transition and is illustrated below:



The probes may be number 10 wire or larger diameter rod soldered directly to the connector. Type-N panel receptacles have captive pins so that no further support will normally be required to maintain the probe mechanically fixed in position. The length of the probe and the position with respect to the short circuit or rear wall should be adjusted for minimum S W R on the coax line. Although for best matching results the complete horn should be available, only the first solid section will be adequate for initial adjustments. The rear or short circuit wall should be well bonded electrically to the waveguide. In making impedance matching adjustments the horn must be aimed into a region free of obstacles or straight into the sky.

The two probes are in the same transverse plane in the guide and their tips may become close together. Nevertheless, the coupling between the two probes should be -20 db or less. Should this not be the case by direct measurement, the probes may be carefully bent or skewed slightly until minimum coupling exists. Under these conditions the two probes will be mutually independent and permit circular polarization to be implemented as indicated earlier.

Concluding Remarks: Although the optimum horn has some appealing and useful characteristics, the primary ones being construction intolerance and low noise, the required gain of a single horn for EME work results in a very large space consuming structure which will be difficult to steer. A more practical solution to the use of the horn design is in an array of perhaps four 24 db gain horns all fed in parallel through a suitable low loss interconnecting feedline harness.

Careful consideration should be given to all the advantages and disadvantages of the horn antenna with respect to your individual situation before construction. It is this author's opinion that while the horn has great merit for EME work, the paraboloidal reflector type antenna is more practical.

This report has been presented in response to requests for horn design information.

As an addendum to this report, a one page design sheet for optimum horns of 15 and 20 db gain at both 1296 and 2390 mc has been included. The dimensions are accurately computed in order that these horns when accurately constructed and impedance matched will serve as standard gain horns with gain tolerance of ± 0.2 db for comparison measurements with higher gain antennas. For best gain accuracy, the horn should be made of conductive sheets material with all joints well bonded. These designs are for linear polarization only and the feed sections are standard size rectangular waveguides.

These horns also serve very well as relatively easy to construct low gain antennas for terrestrial communications. A thin sheet of polyethelyene, vinyl, or mylar sheet covering the aperture can be used for weatherproofing the horn.

SQUARE APERTURE OPTIMUM HORN DESIGN FOR 1296 MC

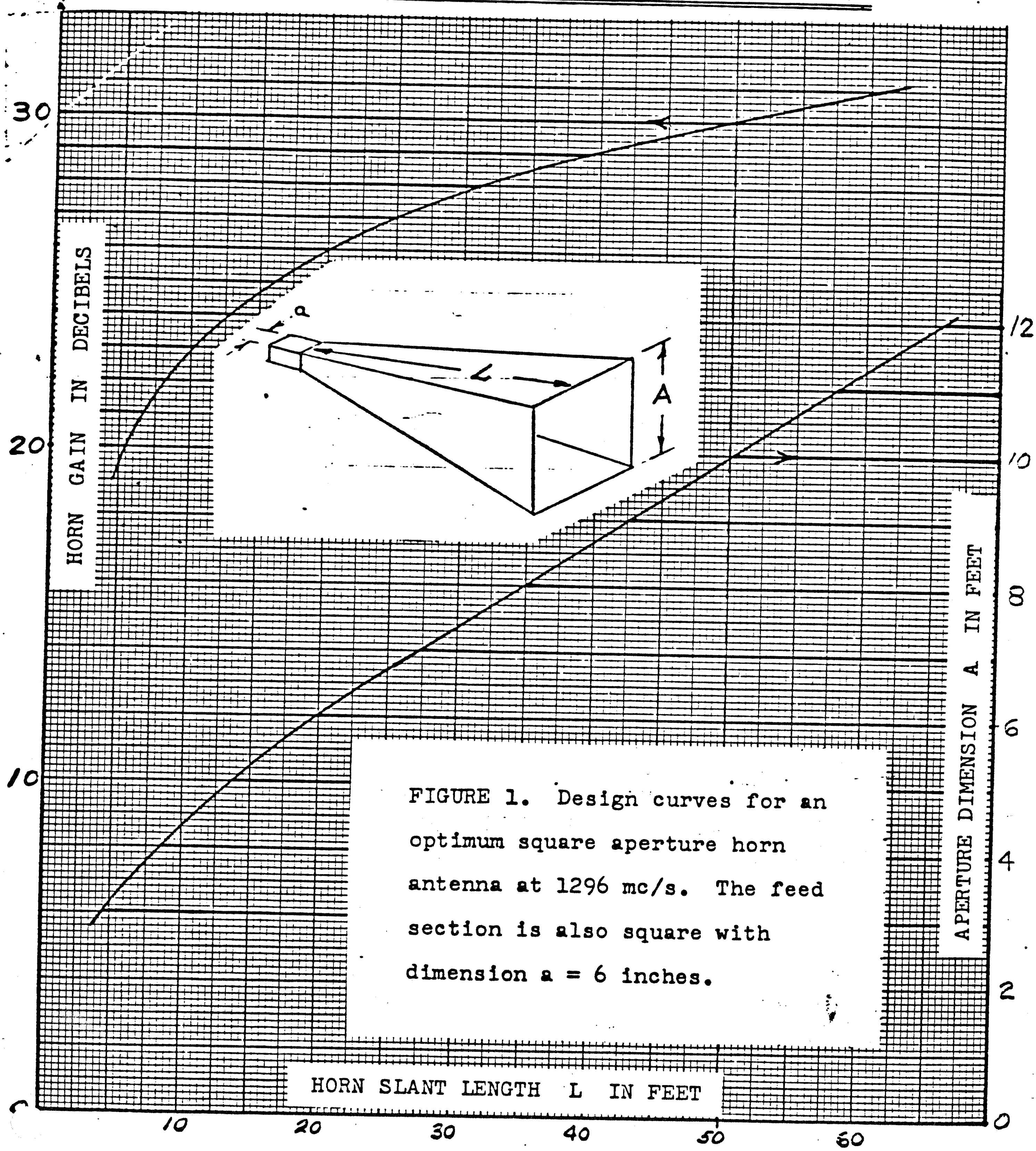
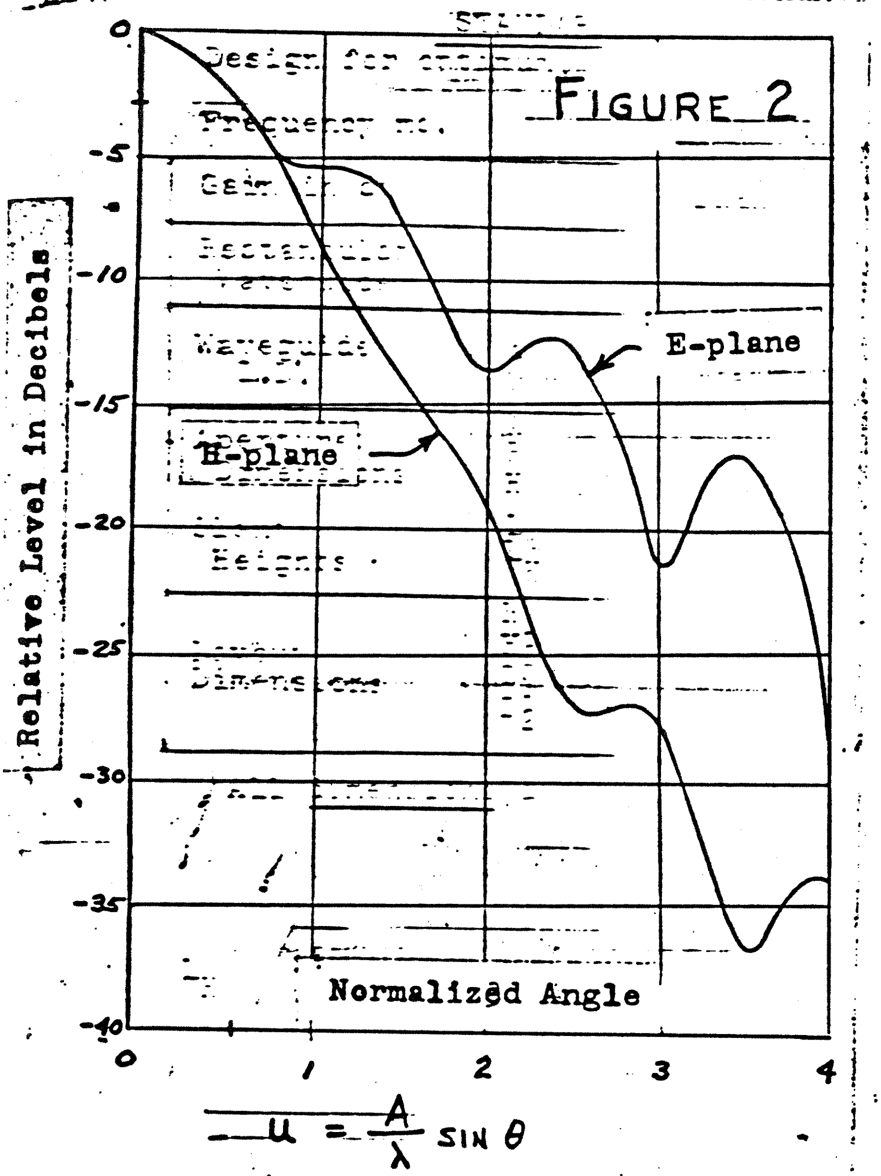
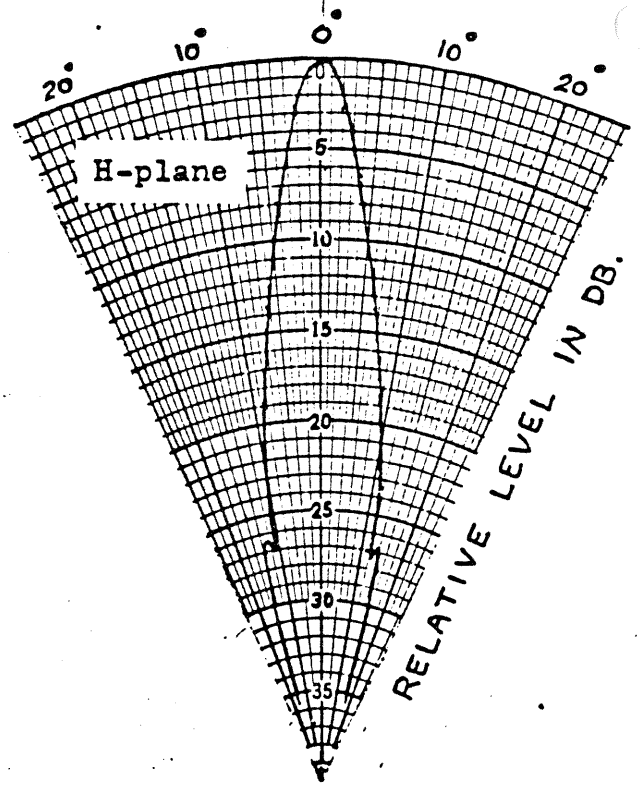
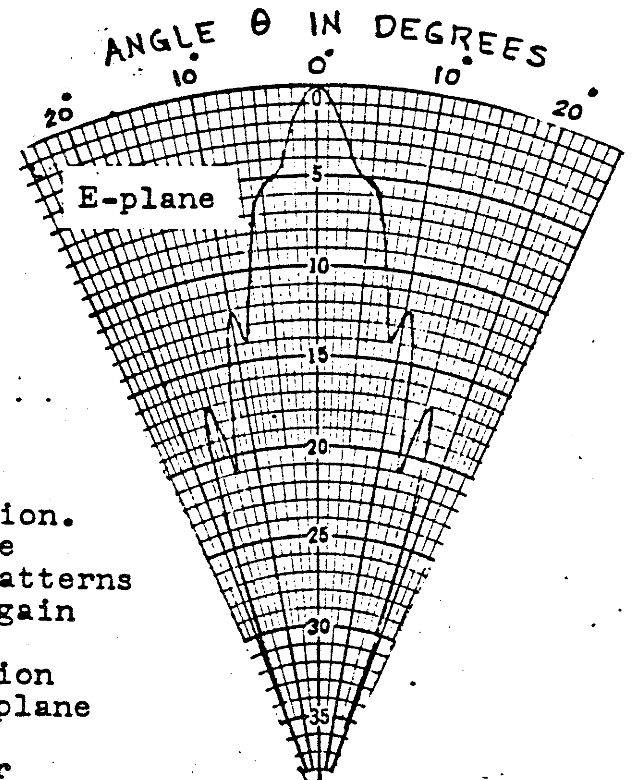


FIGURE 1. Design curves for an optimum square aperture horn antenna at 1296 mc/s. The feed section is also square with dimension a = 6 inches.







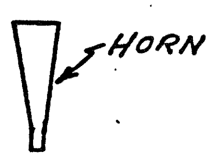


Figure 2. Shown above are the principal plane, E and H plane, radiation patterns for an optimum horn with linear polarization. The angle scale has been normalized to the aperture dimension A, which makes these patterns universal for any size optimum horn with gain greater than 10 db.

For circular polarization the radiation pattern will more closely resemble the E-plane linearly polarized pattern shown above.

The half-power beamwidth for circular polarization is in degrees approximately

$$\theta^{\circ} = 2 \arcsin\left(\frac{0.44}{A}\right) \quad \text{for 1296 mc.}$$

where A is the square aperture size in feet.

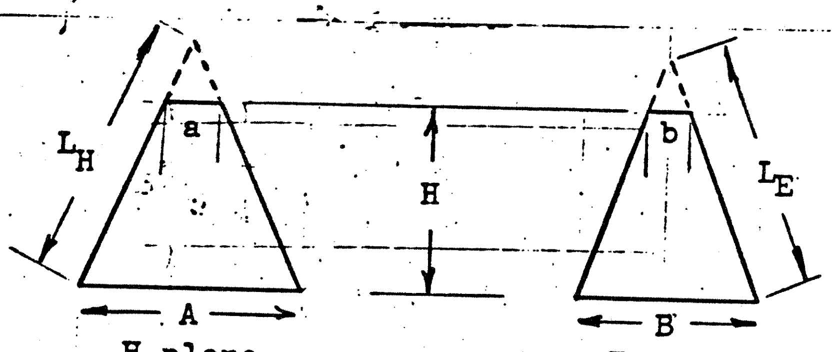
The linearly polarized patterns to the right are shown in polar form and are for a specific horn having 30 db gain and a 10 X 10 foot aperture.

STANDARD GAIN HORN DIMENSIONS

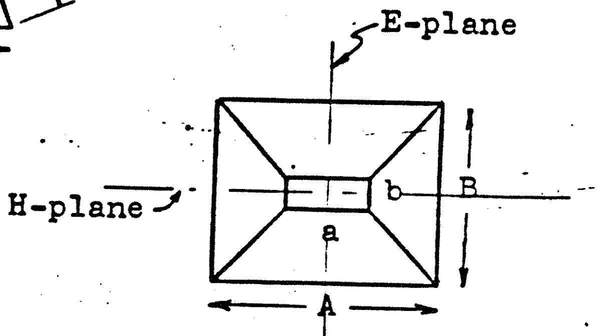
Design for optimum gain pyramidal horn is 3.1 db below area gain.

Frequency mc.		1296	1296	2390	2390
Gain in db		15	20	15	20
Rectangular Waveguide		WR650	WR650	WR340	WR340
Waveguide I.D.	a	6.50	6.50	3.40	3.40
	b	3.25	3.25	1.70	1.70
Aperture Dimensions	A	21.47	41.67	11.62	22.56
	B	15.78	32.52	8.55	17.62
	H	8.19	47.45	4.49	25.83
Slant Heights	L_E	12.99	55.17	7.05	29.91
	L_H	15.92	59.96	8.60	32.44
Layout Dimensions	H_1	10.31	49.66	5.65	27.03
	H_2	11.10	50.60	6.09	27.55
	D_1	14.79	58.83	7.99	31.83
	D_2	13.98	56.22	7.59	30.49

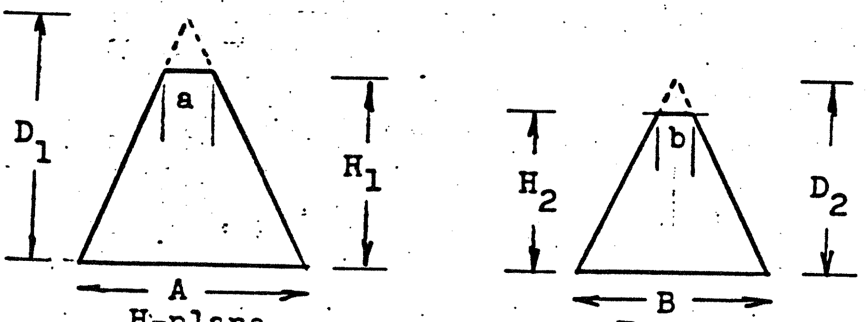
All dimensions are in inches to inside surfaces



Cross-section Dimensions



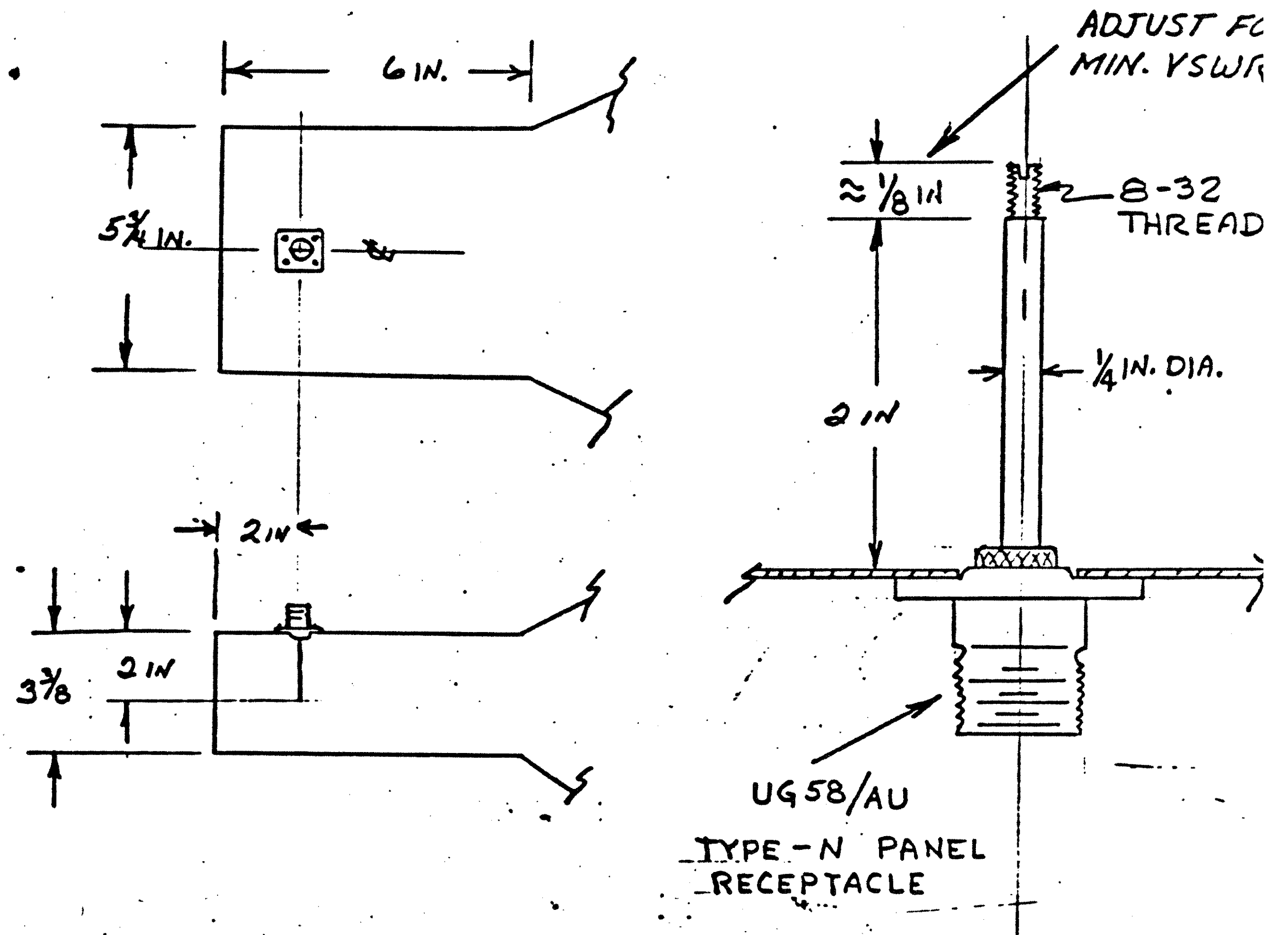
Front View



Layout Dimensions of actual sides of horn

Coaxial Line (50 ohm) to Rectangular Waveguide Transition

W2CCY



This coax to waveguide transition is suitable for the standard gain horn design at 1296 mc only. The short section of rectangular guide inwhich the coax probe is located may be fabricated of sheet material or from a standard minibox with one end removed and all joints thoroughly bonded with screws every inch or two.

These dimensions do not correspond exactly with the standard rectangular waveguide size and allowances must be made at the interface between horn and transition. The dimensions shown above are inside dimensions in inches.

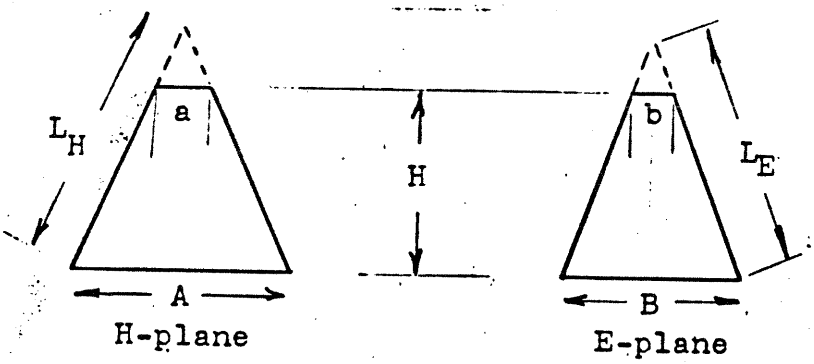
March 28, 196
W2IMU

STANDARD GAIN HORN DIMENSIONS

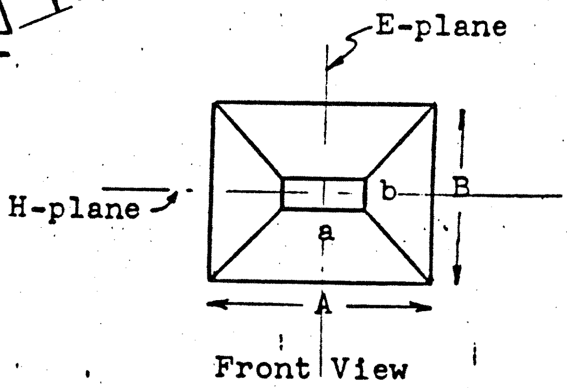
Design for optimum gain pyramidal horn is 3.1 db below area gain.

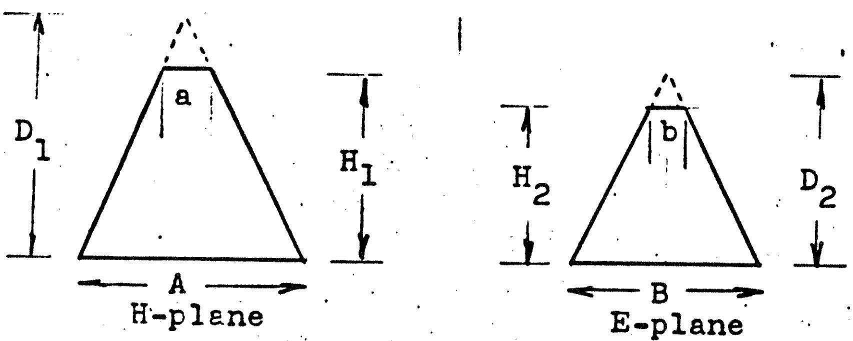
Frequency mc.		1296	1296	2390	2390
Gain in db		15	20	15	20
Rectangular Waveguide		WR650	WR650	WR340	WR340
Waveguide I.D.	a b	λ=12.8 6.50 3.25	6.50 3.25	3.40 1.70	3.40 1.70
Aperture Dimensions	A B H	21.47 15.78 8.19	41.67 32.52 47.45	11.62 8.55 4.49	22.56 17.62 25.83
Slant Heights	L_E L_H	12.99 15.92	55.17 59.96	7.05 8.60	29.91 32.44
Layout Dimensions	H_1 H_2 D_1 D_2	10.31 11.10 14.79 13.98	49.66 50.60 58.83 56.22	5.65 6.09 7.99 7.59	27.03 27.55 31.83 30.49

All dimensions are in inches to inside surfaces



Cross-section Dimensions





Layout Dimensions of actual sides of horn

TECHNICAL REPORT # 5

From: The Crawford Hill VHF Club
Date: (1970) revised November 1986

Subject: A Paraboloidal Reflector Antenna for 1296 mc/s

This report will consider a paraboloidal reflector type antenna suitable for EME communications at 1296 mc/s. Basic design considerations and construction techniques will be presented.
A 20 foot (6.1 meters) diameter, f/D = 0.5, design will be detailed. This size reflector antenna should provide a minimum of 35.0 dB effective gain at 1296 mc/s and make possible reliable echo reception with 500 watts of available r-f transmitting power and a receiving system operating temperature under 300 degrees Kelvin.
The design presented here is slightly different from the original Notes. A more detailed analysis and fine-tuning of the design is included here.

Basic Design Considerations

A paraboloid, parabola of revolution or "dish" can be used as a focussing reflector when suitably illuminated by a feed antenna which has a well defined phase center of radiation, and radiates principally over the area of the reflector surface.
Figure 1a depicts the basic components of a parabolic reflector antenna, the feed and the reflector, along with support struts.
Figure 1b is a cross-sectional view through the center of the reflector showing the various factors of interest in the design of any paraboloidal reflector antenna.
The most important consideration of this type antenna for EME purposes is that it provides maximum effective gain for transmitting and maximum gain/antenna temperature ratio for receiving. As will be seen, these requirements are somewhat paradoxical.

Effective Gain

Maximum effective gain of an aperture antenna, of which the parabolic reflector antennas is a member, can be expressed as the product of an efficiency factor times the maximum possible aperture area gain.

$$\text{Effective Gain} = \eta \left(\frac{4\pi A_o}{\lambda^2}\right)$$

Where the same length units (feet, inches, meters, etc) are used to compute the physical aperture area, and for the free-space wavelength λ. Gain in these Reports will always be referenced to an isotropic radiator whose gain is defined as equal to 1.
For a circular aperture, the same gain expression may be written as

$$\text{Effective Gain} = \eta \left(\frac{\pi D}{\lambda}\right)^2$$

where D is the reflector diameter as shown by Figure 1b.
Gain in decibels above isotropic is $10 \times \log_{10} G$ (above).

For a paraboloidal reflector antenna with front, prime focus, feed, the total efficiency, , is essentially a product of two factors: (1), the fraction of the total power radiated by the feed which is intercepted by the paraboloid and is thus made available to the aperture to generate gain; and (2), the efficiency with which the paraboloidal reflector concentrates the available power in the desired forward direction into a pencil beam of radiation.
The significance of the first factor will be clearly understood from the cross-sectional drawing of Figure 1b. Here most of the radiated energy from the feed impinges upon the reflector surface and is focussed by reflection into parallel rays of energy in the forward direction (to the right in the Figure). A small portion of the radiated energy from the feed extends beyond the edge of the reflector as shown by the shaded area in the Figure. This energy is lost and cannot contribute to the gain of the antenna in the forward direction. This lost energy or power is called spillover. It will however contribute to unwanted rear radiation in the transmit mode and to the antenna temperature in the receive mode.
The first factor then is simply the ratio of energy intercepted by the reflector to the total energy radiated by the feed.

The second factor, (2) is termed aperture illumination efficiency and is a term commonly used with high-gain large-aperture antennas to denote how effective the aperture area is in producing gain. The actual aperture area of a paraboloidal reflector antenna is the plane area enclosed by the rim of the reflector. Aperture theory tells us that the effectiveness of this area in producing gain is related only to the way in which the available energy is distributed over that aperture, in both phase and amplitude.
Ideally, for maximum possible gain, the aperture distribution of energy should be uniform in both amplitude and phase. The geometrical properties of a parabolic reflector fed by a source which itself has a well defined phase center located at the focal point of the reflector, results in a constant phase of all energy from the feed over the aperture area, i.e., the distance from the focal point to anywhere on the aperture plane is constant. And the amplitude distribution of energy over the aperture plane is largely determined by the radiation characteristics of the feed antenna. For best efficiency, the feed antenna is always aligned with its direction of maximum radiation into the center (vertex) of the paraboloid. The resultant energy distribution will therefore always be maximum at the center of the aperture area, decreasing in some fashion towards the edge (rim) of the reflector. This unavoidable tapering of energy distribution results in an aperture illumination efficiency factor of slightly less than unity. There is an additional tapering of the feed energy which is due simply to the geometry of the paraboloid.

Numerically, this additional tapering is the ratio of the distance, R, from the focal point to the center (vertex) of the reflector (i.e., the focal length), to the distance to any other point on the reflector surface, or (1 + cos(O))/2. Figure 2 shows a plot of the additional space tapering v.s. f/D, the focal length to diameter ratio of the reflector. The total energy tapering is the combined feed radiation and space tapering.

Efficiency Computations

To bring into focus these efficiency factors relating to the gain of a paraboloidal reflector antenna, a specific feed must be selected. This report will use only the dual-mode (IMU) type feed which is the highest efficiency small aperture feed antenna known to date. All results obtained will be for this particular feed, whose measured radiation characteristics are shown by Figure 3. For computational purposes, an approximate mathematical model is also shown. Note that the model feed radiation pattern is taken to be circularly symmetric which is correct at least over the reflector intercept region and only very slightly in error in side and back radiation.

All mathematical relations used to determine efficiency factors and spillover temperature are in Appendix A.

Spillover and illumination distribution efficiency have been computed and plotted separately on Figure 4, together with the combined total efficiency. Note the broad maximum in efficiency around an f/D ratio of 0.56. These calculations predict that a maximum possible effective antenna gain will be the aperture area gain reduced by the efficiency factor, 0.785 (- 1.05 dB), in this case. Such high efficiency is never realized in practice owing to at least the following detrimental effects: inaccurate reflector surface, poor reflecting surface material (transmission leakage), blockage and scattering from the feed and feed support struts, cross polarized radiation of the feed and heat losses mainly in the feed itself.

In practice a total efficiency factor of -1.8 dB or 66% can be achieved with care and good materials. The discrepancy of 0.75 dB can be accounted for among the above mentioned detrimental effects.

Figure 4 may also be used to determine the penalty to be paid when using the dual-mode feed with a reflector having an f/D other than the optimum 0.56. For example, if you have a reflector with an f/D of 0.4, the expected loss of antenna gain will be very nearly 0.5 dB. Because of antenna noise considerations, it is always best to err on the side of lower than optimum f/D with virtually any feed arrangement. Antenna Gain/Temp ratio will be considered in the section on antenna temperature. Note that Figure 4 also includes an antenna temperature curve for spillover to the warm Earth only.

Antenna Temperature and Optimization of Gain/Temp.

The effective antenna temperature presented at the feed connection (port) is the aggregate of several sources. The most significant are spillover from the feed, scattering from feed support struts, or any other radiation blockage, diffraction

around the rim edge of the reflector, and heat losses mainly in the feed antenna itself. In addition there can be noise leakage through imperfect surfacing material.
All noise sources except heat losses are in some way related to the position (pointing) of the antenna. This is especially true of discrete celestial sources such as Galactic noise, Radio Stars, the Sun and the Moon. Of these the Sun is by far predominant, having an effective temperature of about 100,000 degrees Kelvin at 1296 mc/s compared with 210 deg K. for the Moon. Man made noise sources, which can be troublesome at times, are not considered in this report. In general most man made noise sources such as automobile ignition and motor commutator sparking have a very small noise component at 1296 mc/s.
At low elevation pointing angles there is also a significant increase of antenna noise because the main beam of the antenna is partially in view of the Warm Earth. Also at low elevation angles more Earth atmosphere is intercepted by the main beam which causes and additional small increase in antenna noise. For these reasons EME communications with the Moon at rise or set times can be compromised by this additional noise.
Spillover noise can be easily computed and may be controlled by proper selection of the reflector f/D ratio. For a given feed radiation pattern, decreasing f/D will also decrease spillover noise, see Figure 4. In view of the low noise receiver devices available at present times (1980s) it is important to minimize all antenna noise without greatly compromising antenna gain.
Feed antenna heat losses can be decreased by the use of highly conductive material in construction, copper instead of brass, care in construction to minimize solder on all active antenna areas along with good designs where r-f current paths are not interrupted by soldered seams where possible. Broad band feed designs (low-Q) are also potentially less lossy.
Strut scattering can be minimized by the use of dielectric struts instead of metallic ones. Reflector surface leakage can be virtually eliminated by using material with transmission loss greater than 30 dB, and with minimum seams and joints.
A realistic approach to optimizing the paraboloidal reflector f/D ratio is to assign some reasonable estimate to all other antenna noise except spillover, and then by computation maximize the antenna gain squared divided by Tsys, the system operating temperature. For the EME echo path the same antenna is usually used for both transmitting and receiving, therefore the square of the antenna gain (or just efficiency squared) should be used. The system operating temperature consists of the sum of the total antenna noise plus the receiving system total noise. The factor gain squared divided by Tsys appears in the EME system receiver output S/N ratio (see Report #3). Maximizing this factor will therefore optimize the S/N for best reception of echoes.
If the system operating temperature, Tsys is written as the sum of Tsp (spillover temperature) + all other sources (call this Tc), then direct evaluation of the optimum f/D v.s. Tc can be made, and is graphed by Figure 5.
A very useful example is for Tc = 110 deg. K. where an optimum f/D =0.5 is indicated. Tc includes a receiving system with about

1 dB NF (75 deg K) and the remaining 35 deg K mainly due to heat losses in the feed (35 deg. K corresponding to 0.5 dB loss, a very reasonable estimate considering interconnect between feed and receiver as part of the system). The loss in antenna efficiency (one way) at f/D = 0.5 compared with the maximum possible efficiency at f/D = 0.56 is only 0.1 dB. The total improvement in S/N is only of the order 0.1 dB, but represents a positive improvement.
This f/D optimizing process clearly indicates that the choice of f/D for the paraboloid is not critical but should be made towards a lower value to obtain the meager benefits. It does point out however that decreasing other sources of antenna noise as well as receiver noise effects the choice of an optimum f/D ratio.

The Dual-Mode Feed Antenna

The dual-mode (W2IMU) feed design embodies all the best attributes of a high efficiency feed antenna; single circularly symmetric beam radiation, virtually no side or back radiation, a well defined phase center and a small aperture to minimize blockage. This design in circular waveguide permits linear or circular polarization to be readily implemented.
Figure 6 contains complete information for constructing this feed for circular polarization at 1296 mc/s. It may be readily scaled to other bands except for the 30 degree conical flare angle which remains fixed. The term dual-mode refers to the excitation of two circular waveguide modes in the structure, the dominant TE11 mode and the TM11 mode. These modes when combined in proper phase and amplitude at the small feed antenna aperture result in the desirable radiation characteristics obtained. Other type feeds can give similar results but may be more difficult to adjust and construct.

Reflector Surface Error Effects

An exact paraboloidal surface can be described mathematically as shown by Figure 1a, with dimensional coordinate system centered at the vertex of the reflector surface. By setting either X or Y equal to zero, the more familiar plane or two-dimensional parabolic arc is obtained which can be used to layout the required curve for constructional purposes, see Figure 7.
A large paraboloid is very difficult to construct accurately and so some determination of the effects of inaccuracies must be included in estimating the effective antenna gain. Any departure from an exact paraboloid will always result in loss of gain and increased sidelobes in the radiation characteristics of the antenna system. Figure 8 shows a set of curves which relate the loss of gain to r.m.s. surface deviation. This implies a surface, which on the average is an exact paraboloid but has bumps and depressions above and below the average surface. These bumps and depressions cause changes in phase length from the focal point to the aperture plane and thus cause a departure of phase over the aperture area from the desired uniform distribution. Aperture theory indicates that radiation from a large aperture is more sensitive to phase distribution than amplitude distribution which reinforces the need for accuracy of

the reflector surface.
R.m.s. means root-mean-square and the r.m.s. surface deviation may be obtained by the following procedure, though very difficult to perform in a practical situation.
Divide the reflector area into a large number of equal area sections. Then measure the deviation in inches at the center of each subsection from the exact surface. Take each deviation measurement, square it, add all the squares together, take the square root of the sum and divide by the total number of measurements. The result is the r.m.s. surface deviation in inches which may be used with Figure 8 to estimate the loss in gain at 1296 mc/s. Additionally, a measure of the correlation factor C should be made which is the extent or size of the bumps. This is an equally difficult measurement to perform but some idea of C can be found by careful scrutinization of the deviation data, provided that a large enough number of measurements were performed.
Figure 8 clearly reveals, however, that even severe bumps which are not large in area cause little effect, but as the bump size approaches a wavelength, the loss in gain becomes appreciable.

Another form of surface distortion which can be even more devastating is warpage, and deviation from an exact paraboloid curve in the radial direction. Both of these effect can result in severe loss in gain since large area phase errors are involved.
For example, a smooth departure from center to edge of the reflector by an eighth wavelength (1.13inches at 1296mc/s) will degrade the antenna gain by about 1 dB. If the surface error at the rim is a quarter wavelength, the gain will degrade by 4 dB !

It cannot be overstressed, that the performance of a reflector antenna is greatly dependent on the care and accuracy of construction of the reflector surface and material used, as well as the choice of feed antenna.

A very bad surface can be detected by poor gain performance as well as assymetric beam shape, even beam splitting and skewing off axis. Radiation pattern measurements of the antenna system, though difficult to perform with large EME antennas, can be very helpful for diagnostic purposes if antenna performance is below par.
A common condition which exists for home made reflectors is that the average surface may actually have a slightly different focal point than the design value, due to inaccuracies in construction. The result is an out-of-focus antenna system with poor gain. It is always advisable when testing a new reflector antenna to probe the focal point with the feed by moving it toward the reflector and away from the reflector by at least a wavelength, while monitoring the antenna gain. A single peak in gain should occur. If two peaks appear, or the peak in gain is smeared (no distinct peak observed), then the reflector surface most assuredly has inaccuracies and should be fixed or redone. A movement of the feed +/- one half wavelength should produce a noticeable (1 dB at

least) change in gain with a good reflector surface.
Misalignment of the radio beam with the expected centerline of the paraboloid (optical boresight) indicates a distorted reflector surface, assuming that the feed is centered on the reflector axis (centerline).

From the above considerations, it should be clear that construction of a large reflector should emphasize accuracy of the paraboloidal shape with surface roughness a secondary consideration, provided the roughness is fine grained (small area bumps).

Surfacing Material

A very important consideration in the construction of any reflector antenna is the actual reflector surface material and how it is applied. Ideally, the surface should be conductive and continuous over the entire reflector surface. It should have a thickness of several 'skin depths' at the operating frequency. One 'skin depth' of aluminum at 1296 mc/s is about 0.0001 inch. Any of the readily available aluminum foils is adequate.
The material need not have a high electrical conductivity because a reflector is a low-Q device with surface currents well distributed. Iron, copper or aluminum are acceptable for surfacing and readily available in foils and mesh form.
These include; window screening, hardware cloth, expanded metals sheet, perforated sheet, fencing, etc. Materials such as screening and hardware cloth which are constructed of wires forming a square mesh may be evaluated for transmission loss with the aid of Figure 9. This nomograph relates radio transmission loss through the material as a function of the physical dimensions of the mesh and wires. Since power transmission through the reflector is lost, like spillover, the degradation in antenna gain is readily evaluated. Figure 9 includes computed loss of gain, along with transmission loss.
In recent times, expanded aluminum sheet has become readily available and is used extensively in home TVRO systems. This material should not be overlooked as it is perhaps the only material which will deform in two dimensions to permit very accurate surface contours to be formed. Another material which offers possibilities for light-weight solid-surfacing is a building trade insulation siding material consisting of a sandwich of two layers of aluminum foil with a urethane foam core. It is available in 4 by 8 foot sheets either 1/2 or 3/4 inch in thickness, is easily cut with a knife and may be bent moderately in one plane only.
In addition to loss of gain from leakage through the reflector surface, there is also a penalty in antenna temperature because leakage also takes place in the reverse direction allowing the feed to 'see' the temperature behind the reflector attenuated by the transmission loss. The worse case would be for high elevation pointing angles where the warm Earth is behind the reflector. In this case for a transmission loss of 10 db, the antenna temperature will be raised by approximately 250/10 deg. K, or 25 degrees K. In view of modern low noise device

technology, the receiving degradation due to this additional noise temperature must be considered significant.

Materials whose mesh shape is not square, such as hexagonal fencing (chicken wire), may be evaluated by considering an equivalent geometric square, or by direct measurement of transmission loss through a large sample area. In general, if the largest dimension of the mesh hole exceeds one inch, the material may be marginal or unacceptable for a reflector at 1296 mc/s. Three quarter inch hexagonal chicken wire fencing, for example, has a measured - 10 dB transmission loss at 1296 mc/s and should be considered marginal for a high efficiency system.
It should also be pointed out that although cross over points in most mesh material are electrically bonded, it is not necessary.
It should also be pointed out that heavily corroded materials are good reflectors until the actual metal has been completely eaten away. Coatings such as paint, sealers or even corrosion cause no degradation because their thickness is a very small part of a wavelength, and even though the coating may be a lossy material as a dielectric, there is virtually no loss because the electric field at the surface of a reflector is virtually zero.

When applying surface material to the reflector support frame, additional precautions should be observed. In general, the smallest dimension of a section or panel should be <u>larger than a wavelength</u>, or at least 12 inches at 1296 mc/s. Also, because the feed illumination is strongest at the center of the reflector, this area should be given special attention as to accuracy and with the largest available pieces of material. In the sectioned (petal shaped) design described in this report, the central area should not be covered by extending the wedge shaped pieces to the center point. Rather, the wedges should be truncated where the width is no less than about 12 inches (for 1296 mc/s).
Inevitably there will be seams or joints where panels meet. In general an electrically bonded seam is desirable. However, since the free-space impedance is high (377 ohms) the actual seam need not be bonded provided that an overlap of one quarter wavelength (2.25 inches at 1296 mc/s) is formed. This overlap, if very close spaced, is essentially a low impedance parallel plate transmission line transformer which reflects a short circuit at the surface gap.
If the sections of surfacing material are very large then butt seams which are neither bonded nor overlapped will provide satisfactory performance. The butt joints should be reasonably close spaced to prevent excessive leakage through the slot.

<u>Feed Support Struts</u>
A necessary part of a front fed parabolic antenna design are the feed support struts. These struts must be located directly in the active radiating region of the antenna and thus must be given due consideration because of the potential aperture area blockage which reduces the effective gain, and scattering which increases antenna noise. The effect can be even more severe if the support

struts or any other members, such as feed line or brackets are permitted to block and scatter feed radiation.
For these reasons it is highly recommended that support struts be mounted well out of the region around the feed aperture, at least in the frontal hemisphere around the feed. The struts should also extend out to the rim of the reflector to effect as little blockage to the feed illumination as possible.
No more than three struts equally spaced are required. With an az-el antenna mount, the struts should be placed so that only one strut is in tension when the antenna is at low elevation angles.
Struts may be made from thin wall aluminum (hard grade) tubing just large enough in diameter to support the feed assembly. Alternatively, the struts can be made from rigid PVC water drain pipe material. The use of dielectric material for the struts is highly recommended to greatly reduce blockage and scattering from the struts. A larger diameter PVC tube can be used for strength and tapered at the ends to minimize wind loading.

Reflector Construction

The following information on reflector construction may be used as a guide. This particular method of surface design uses the gored pattern or petal design in which cylindrical parabolic sections (the petals) provide an easy way to used surface material which can only be curved in one dimension, such as sheet or mesh material, see Figure 7. This construction method leads to regular phase errors over the surface (mainly towards the rim) which degrades the antenna gain. Figure 10 can be used to determine the degradation from physical parameters. The design presented here limits the maximum degradation to 0.25 dB.
In the 20 foot diameter reflector design presented here, a total of 16 trussed ribs ($f/D = 0.5$) is suggested in order to accomodate available material size in the U.S.A. The individual petal layouts are limited to 48 inches in width by 96 inches in length (see Figure 7). The actual focal length of the reflector will be slightly less than the expected 120 inches due to the modified surface design. The expected nominal focal length will be approximately 117 inches for this design. As suggested earlier in this report, the focus of any new reflector should always be probed with the feed to find the optimum location of the feed antenna.
The backup support structure used here is the trussed rib and central hub method where the hub diameter is constrained to be about 48 inches in diameter. Figure 7 suggests the form of construction where aluminum tubing and gusset plates are used extensively, held together with pop-rivets. The hub may be formed into a 16 sided polygon from tubing bent with the aid of a tool called a "hickey", used by electricians to bend electrical conduit and which may be rented from most electrical supply houses. The rim of the reflector is not circular but made from straight 48 inch lengths of tubing to form a 16 sided polygon Additional straight spreaders may be used between the ribs and towards the rim to support the surface material from sagging.
The parabolically curved member of the rib should be made from a softer grade aluminum tube and carefully bent into the

approximate parabolic arc to reduce stress spring-back in the trussed rib assembly. A wooden jig should be made and used to assemble the trussed ribs to ensure that all are identical and of accurate shape. Those ribs assigned to support the feed struts may be strengthened with additional trussing and perhaps a double back member.

When petal sections are used with no overlaps at the rib seams, the exact shape of the section is not exactly triangular but has slightly curved edges along the long sides. The exact layout shape is shown by Figure 11, with necessary equations to compute dimensions.

Antenna Mounting and Drive Considerations

Large antennas primarily intended for EME communications should be mounted no higher above the Earth than necessary. Although elevation-over-azimuth mounts are popular, the polar mount is highly recommended for simplicity of the Moon orbital tracking drive mechanism, and accuracy of tracking.

For any given installation, the height of the antenna will be determined by foreground clearance in the Moon orbital direction. Foreground clearance usually considers buildings, trees or other obstacles which can block and scatter radiation from the antenna.

The radiation characteristics of a large aperture antenna can be described as a circular tube having a diameter equal to the largest extent of the aperture, and extending directly out infront of the aperture by at least 10 aperture diameters or more. This cylindrical volumetric space can be used to determine if an obstacle is in the radio beam.

The minimum height of the antenna should be approximately half the aperture diameter to permit easy access to the feed and also near horizon pointing at Moon rise and set, for maximum DX. In general, hill top locations are desirable for maximum DX conditions at Moon rise and set, but are also more vulnerable to radio receiver interference from nearby sources, TV or FM broadcast harmonics, RADARs, commercial mobile radio service, industrial r-f heating devices, etc.

A careful survey of available space should be made before installation of an EME mount and antenna to allow for maximum Moon orbital access with minimum blockage, and also minimum friction with family and neighbors. Even though r-f fields in the beam of a large aperture antenna are diluted, pointing such antennas at occupied buildings can be a serious matter with which to contend, and should be avoided.

Summary

Design information and construction suggestions have been presented for a 20 foot parabolic reflector antenna employing the dual-mode (IMU) feed. The reflector design presented is suggested to make best use of surfacing materials such as screen mesh and sheet which cannot be contoured in two dimensions but are very well suited to the cylindrical petal surface design.

The basic purpose of this material is for use in realizing a

high-gain low-noise antenna suitable for EME communication. As in all UHF design and construction, careful attention to detail and accuracy are the key to a successful hardware project.
For construction of a large antenna of the type described in this report, the major factors in the order of their importance are: physical rigidity and strength to withstand environmental effects and maintain reflector shape, accuracy of parabolic surface, low leakage surface material, minimum heat losses (especially in feed and interconnect cables), high efficiency feed (IMU type recommended), and minimimum strut blockage.

APPENDIX - A

For a paraboloidal reflector antenna geometry and feed which has a circularly symmetric radiation characteristic, the spillover and aperture illumination efficiencies may be computed thus:

$$\eta_{sp} = \frac{\text{Power from feed intercepted by reflector}}{\text{Total power radiated by feed}} = \frac{\int_0^{\theta_0} |E(\theta)|^2 \sin(\theta)\, d\theta}{\int_0^{\pi} |E(\theta)|^2 \sin(\theta)\, d\theta}$$

and,

$$\eta_{illum} = \frac{\text{Power radiated in direction of the main beam}}{\text{Total power radiated by aperture as a point source}} = 2\cot\left(\frac{\theta}{2}\right) \frac{\left| \int_0^{\theta_0} E(\theta) \tan\frac{\theta}{2}\, d\theta \right|^2}{\int_0^{\theta_0} |E(\theta)|^2 \sin(\theta)\, d\theta}$$

where θ_0 is the angle to the edge of the reflector (see Figure 1b). For any given f/D ratio;

$$\theta_0 = 2 \arctan\left(\frac{1}{4\, f/D}\right)$$

and, the normalized radiated electric field of the (IMU) feed can be represented with good accuracy by the expression ;

$$E(\theta) = 10^{-1.325\left(1 - \frac{\cos(\theta)}{|\cos(\theta)|}\cos(\theta)^{1.15}\right)}$$

Also the component of antenna temperature which is due to spillover alone when the background temperature is the constant Earth temperature is:

$$T_{a-sp} = T_{Earth} \frac{\int_{\theta_0}^{\pi/2} |E(\theta)|^2 \sin(\theta)\, d\theta}{\int_0^{\pi} |E(\theta)|^2 \sin(\theta)\, d\theta}$$

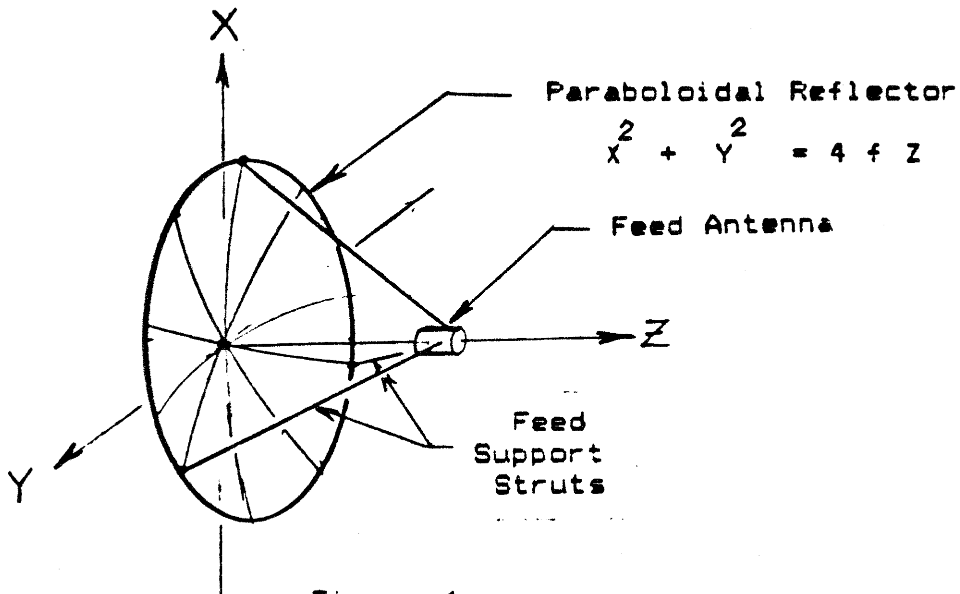


Figure 1a.

Parabola Geometry in Polar Coordinates

$$R = \frac{2f}{1 + \cos\theta}$$

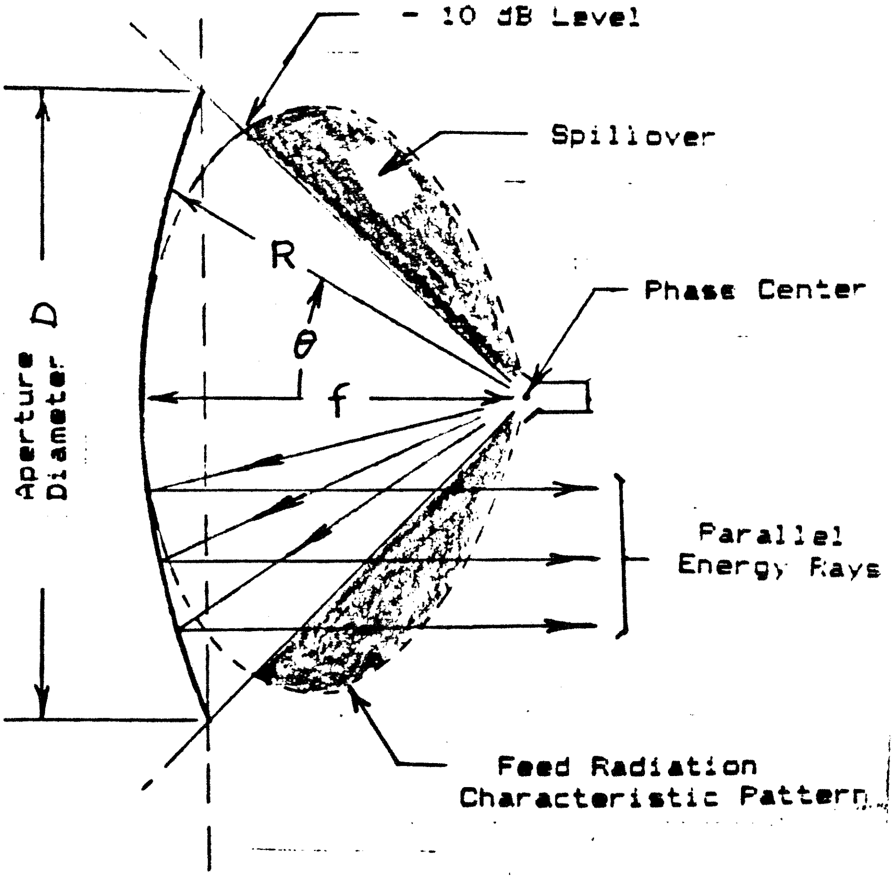
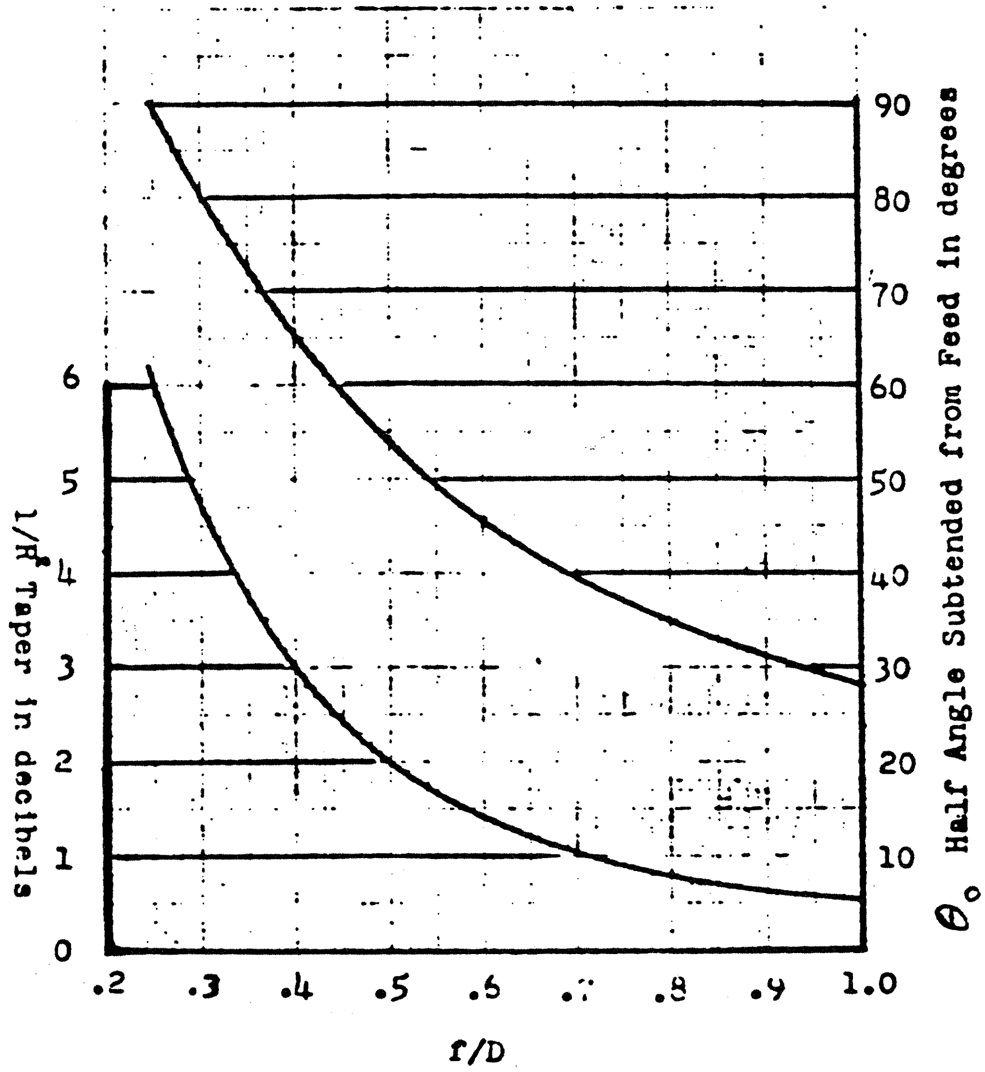


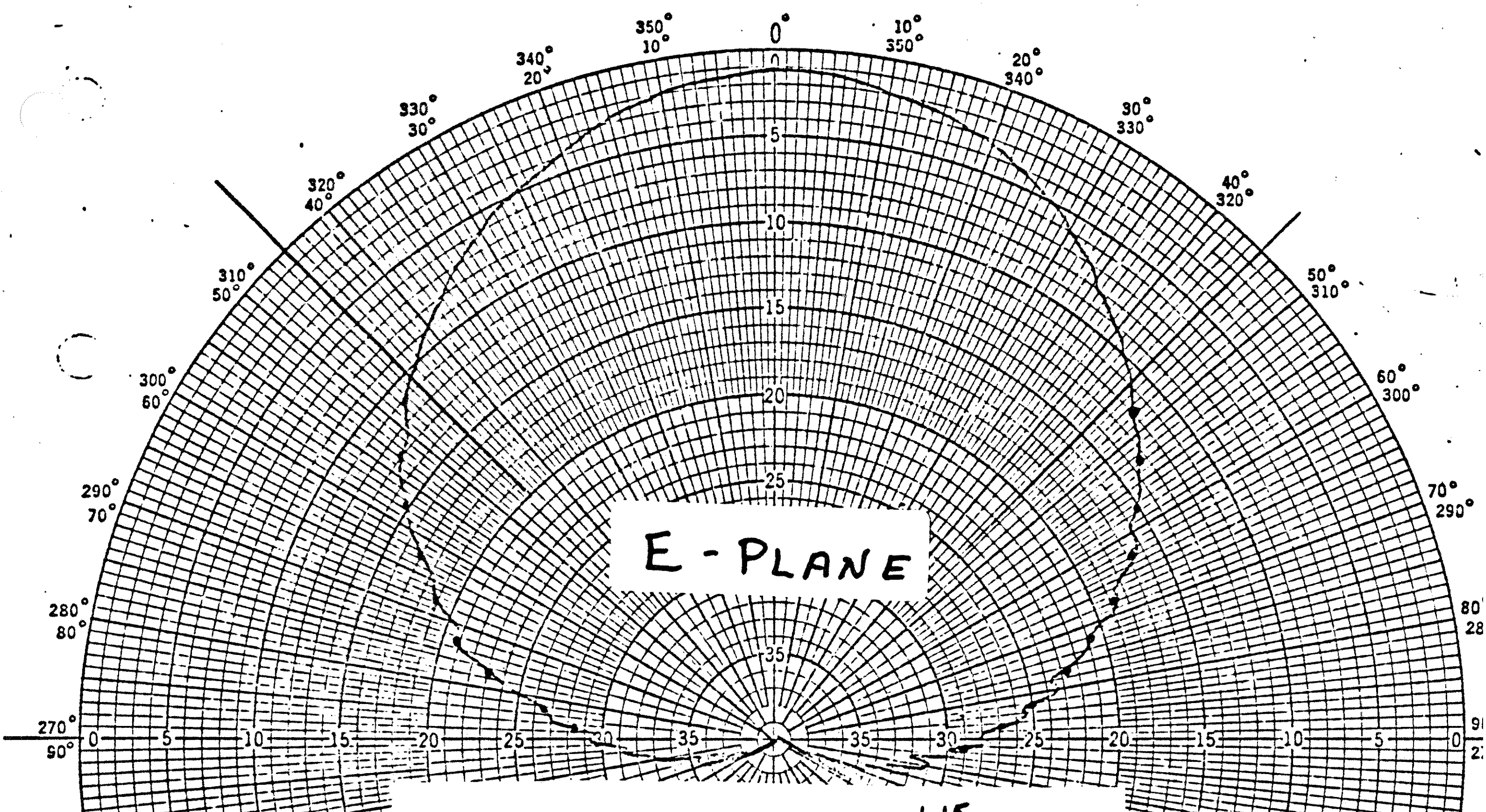
Figure 1b.

Figure 2 Subtended half-angle from feed to reflector as a function of f/D; and, $1/R^2$ illumination taper as a function of f/D.



$$\theta_o = 2 \arctan\left(\frac{1}{4\ f/D}\right)$$

$$\text{Space taper (dB)} = 20 \log_{10}\left(\frac{1 + \cos(\theta_o)}{2}\right)$$



$$E(\theta)\ \mathrm{dB} = -26.5\,(1 - \cos^{1.15}(\theta))$$

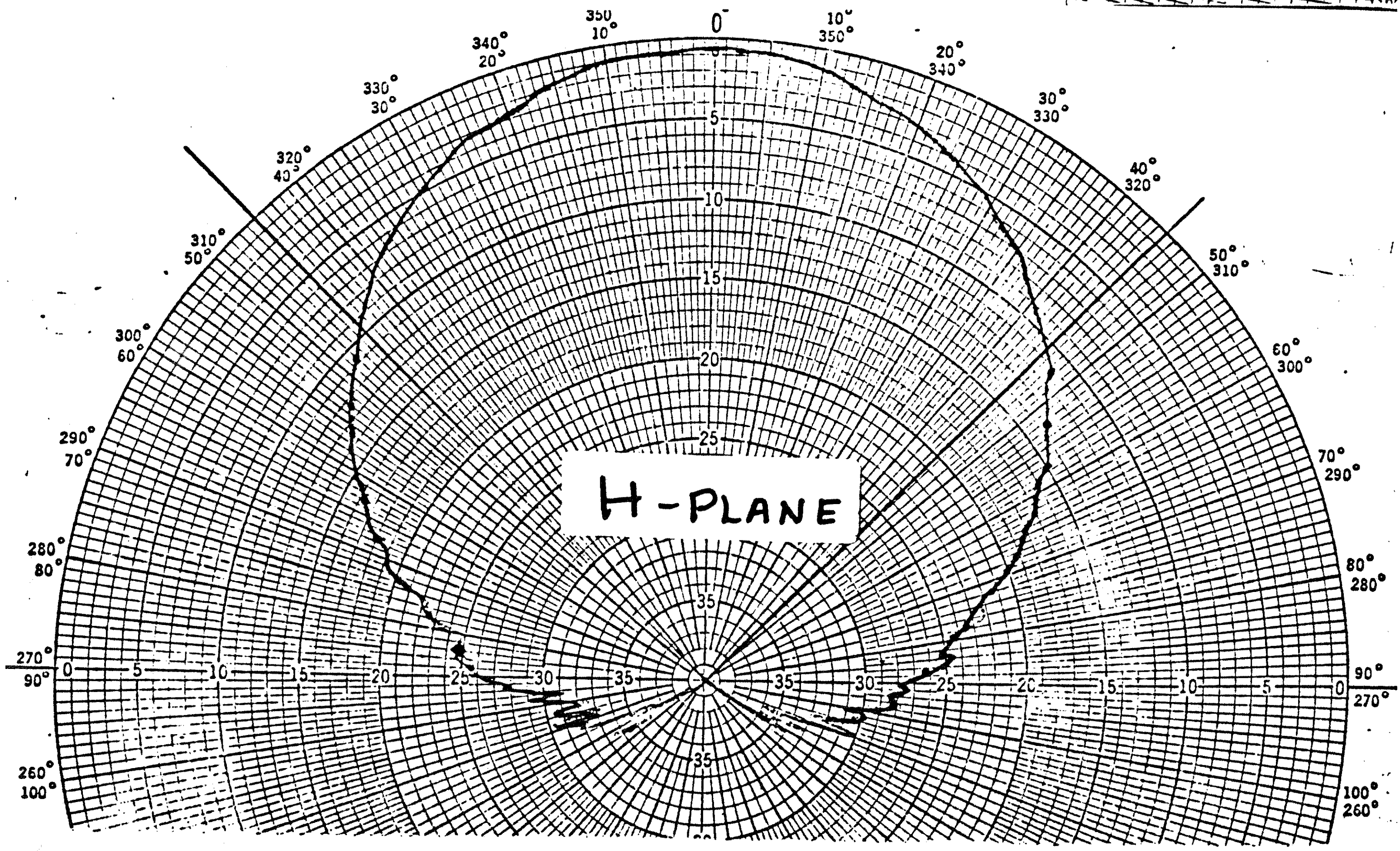


Figure 3. The measured radiation patterns, in the two principal planes, of a properly adjusted dual-mode (IMU) feed antenna, 1296 mc/s.

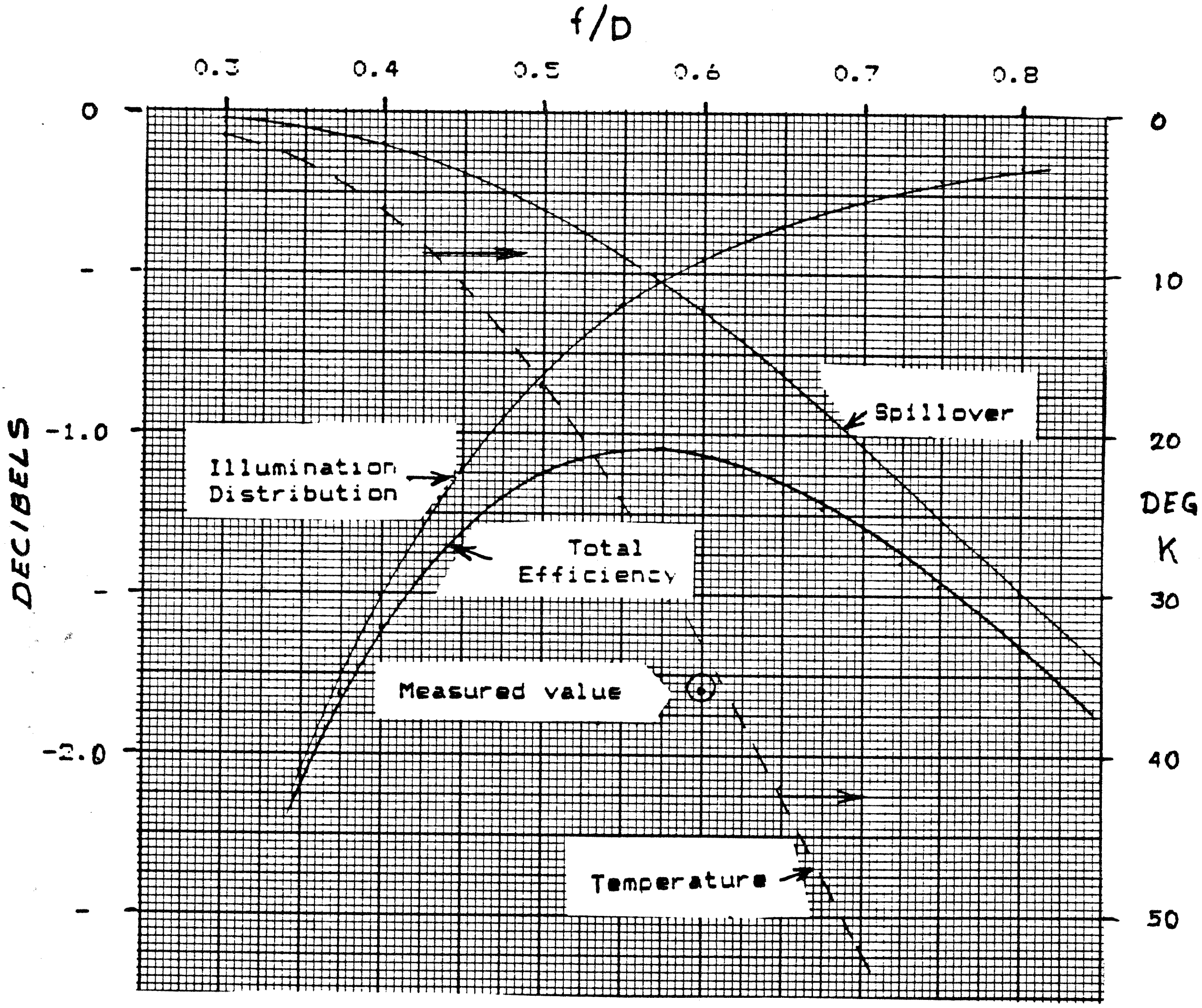


Figure 4. Computed efficiency and spillover temperature for a paraboloidal reflector with a dual mode (IMU) feed.
The total efficiency curve can be used to determine loss in antenna gain as a result of using a paraboloid with a different f/D ratio than the optimum 0.56.

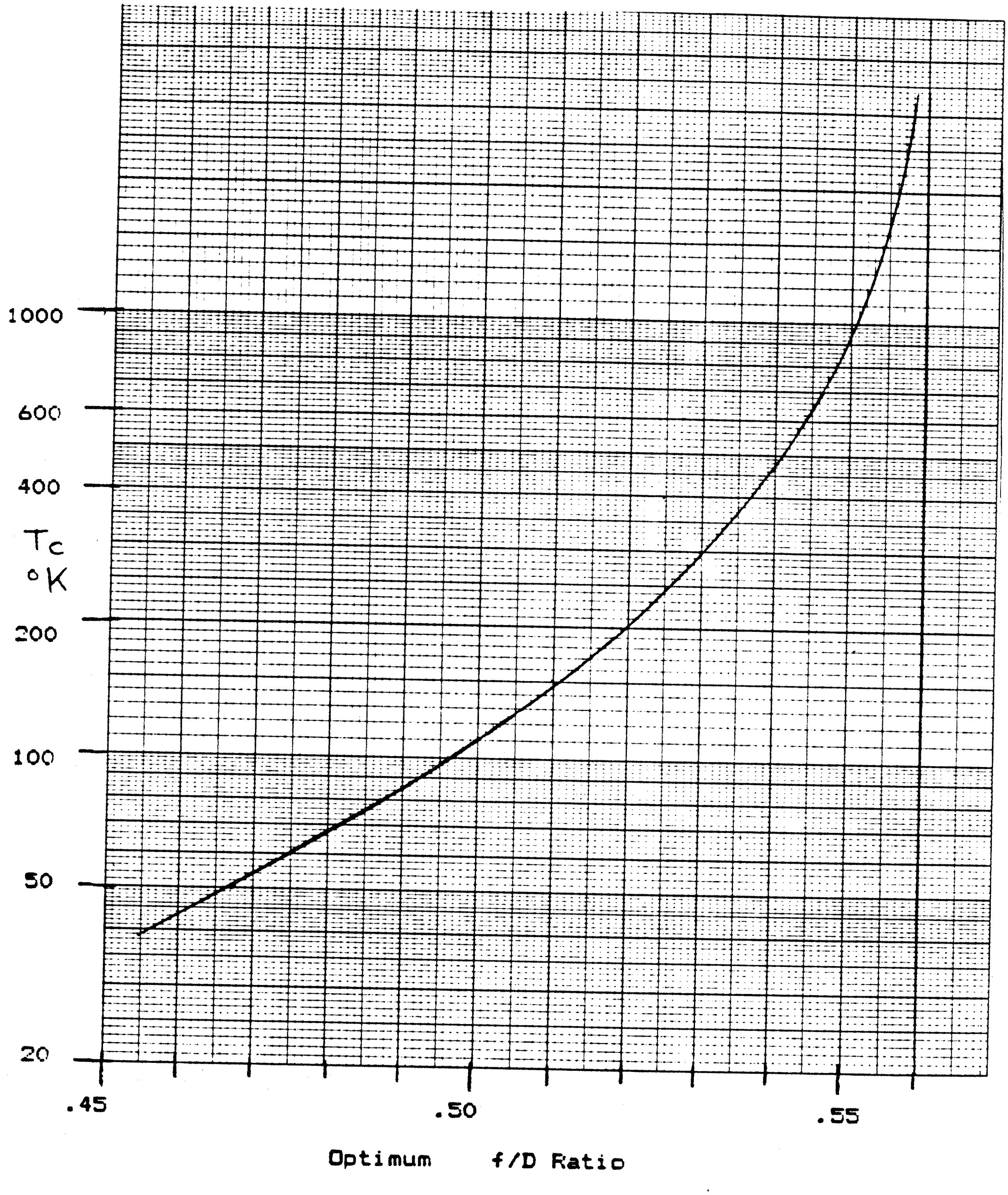
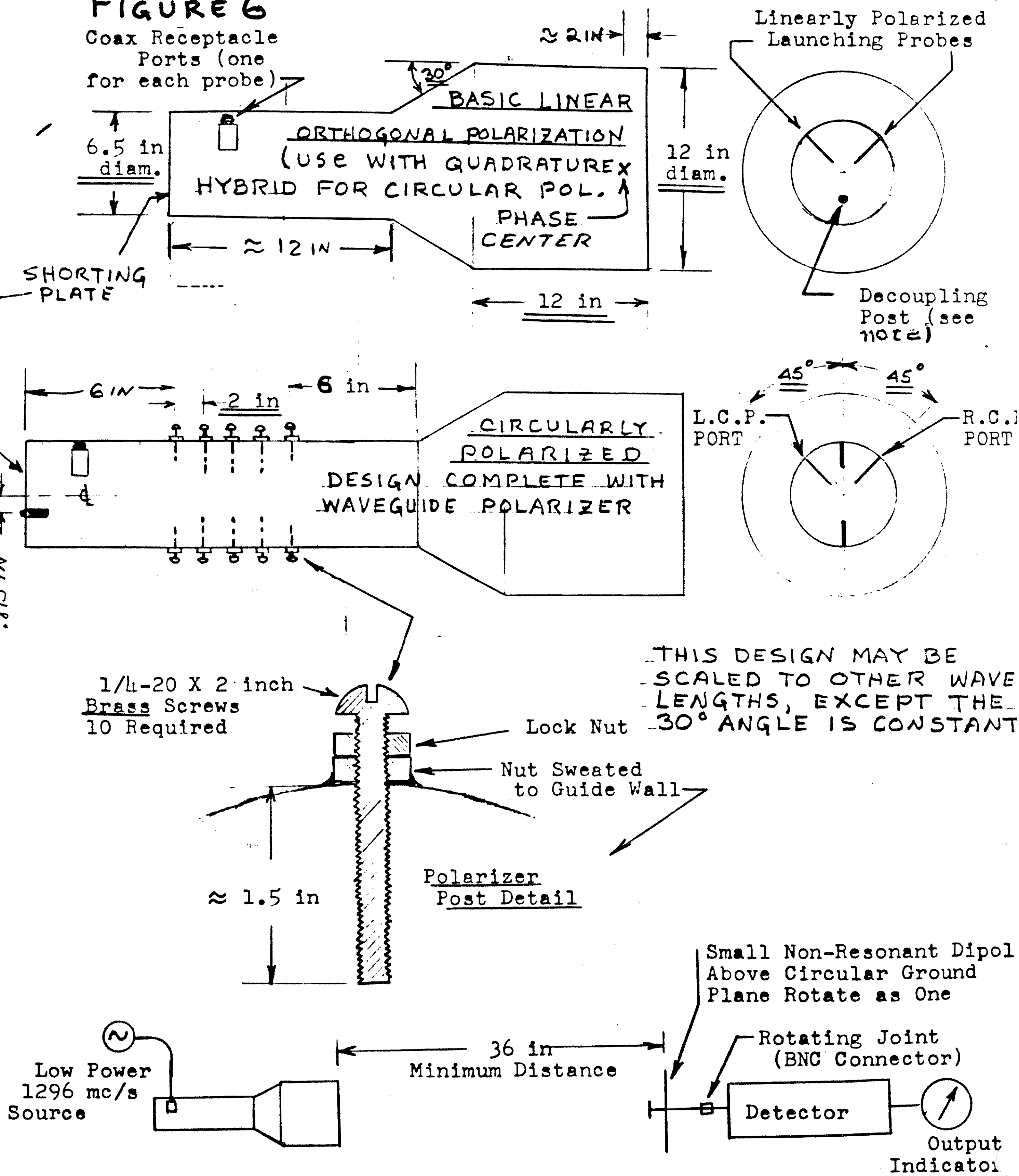


Figure 5. Computed optimum f/D ratio v.s. Tc, the component of Tsys not including apillover antenna temperature, Tsp. Tsys = Tsp + Tc . Optimum f/D computed for the maximum value of $(\eta_T)^2$ / Tsys.

DUAL-MODE FEED DESIGN FOR 1296mc/s

W2IMU

FIGURE 6



Rotate sampling dipole slowly and determine ratio of maximum to minimum signal detected level. Be sure that the output indicator can display a 1 db change in level.

DUAL-MODE SMALL APERTURE FEED NOTES

The dual-mode small-aperture feed antenna makes use of the TE11 (dominant mode) and the TM11 circular waveguide modes which when properly phased and amplitude controlled at the radiating aperture, minimizes edge currents around the aperture rim. The result is a very circularly symmetric single lobe radiation characteristic pattern with side and rear radiation levels below - 35 dB. This feed when used with a paraboloidal reflector having an f/D = 0.56, will result in a reflector antenna system with overall measured aperture efficiency of 66 %.

Construction of this feed from copper sheet is recommended for best performance, although brass sheet will give good results. Suitable copper sheet is available from building supply houses as copper roof flashing. This is a soft copper sheet and should be worked carefully to maintain the accuracy of the circular guide sections. One or more stiffening rings may be cemented or soldered to the outside of the circular guide sections for mechanical rigidity, especially near the aperture.
Underlined dimensions on the drawing should be adhered to accurately. Axial alinement of the circular sections should be done carefully. A wooden frame jig is recommended.
All joints should be tight fitted butt seams with overlay tabs or strips soldered in-place on the outside for physical strength and electrical integrity. Clean excess solder from the inside of the feed. Solder is a poor and lossy conductor compared with copper. The butt seams between conical and straight guide sections should be fitted carefully and secured with a multitude of bent tabs soldered on from the outside. Small brass screws may be used to initially hold a few tabs inplace, provided that the screws are cut off and filed flush on the inside of the structure.
Impedance matching at the coaxial drive ports is accomplished by adjusting the length of the probes incrementally and also bending the probes back or forth slightly, in an axial direction. The nulling post is required to minimize cross coupling between the orthogonal probes due the higher order assymetric modes stirred up by the unbalanced probes. Adjustment of the position of the post must be done carefully to obtain cross coupling less than - 20 dB.
The nulling post is a 7/16 inch diameter brass or copper rod, 1.31 inches long and is located 0.875 inches off-center on the inside of the shorting end plate, as shown. It is secured with a machine scres screwed into the post.
The probes are made of # 14 copper wire.
Adjustment of the multiple screw phase shift polarizer may be done using the set-up shown at the bottom of the drawing. The end pairs of screws should penetrate less than the others to effect a smoother impedance transition. There will be some interaction between all impedance and circularity adjustments.

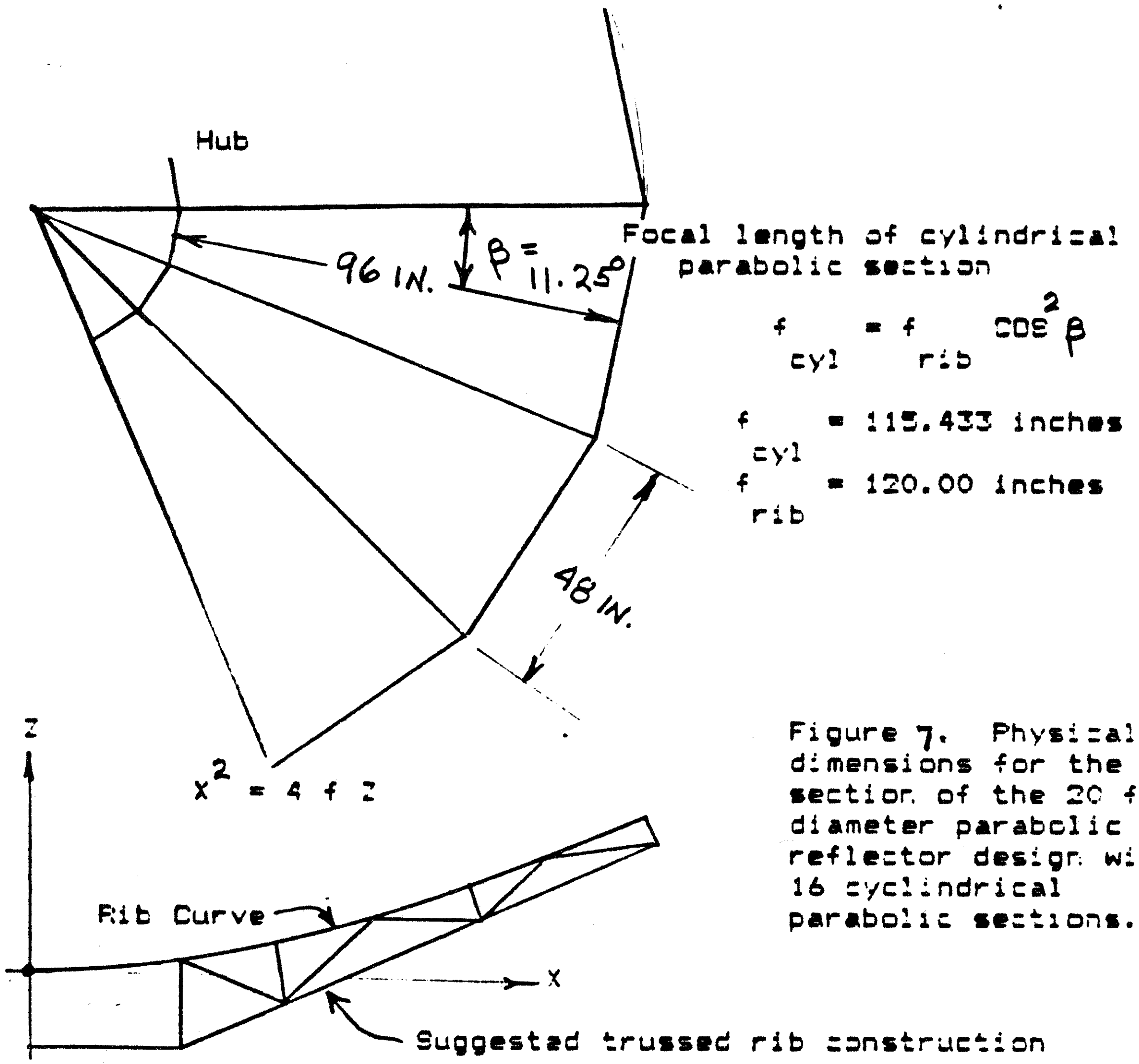
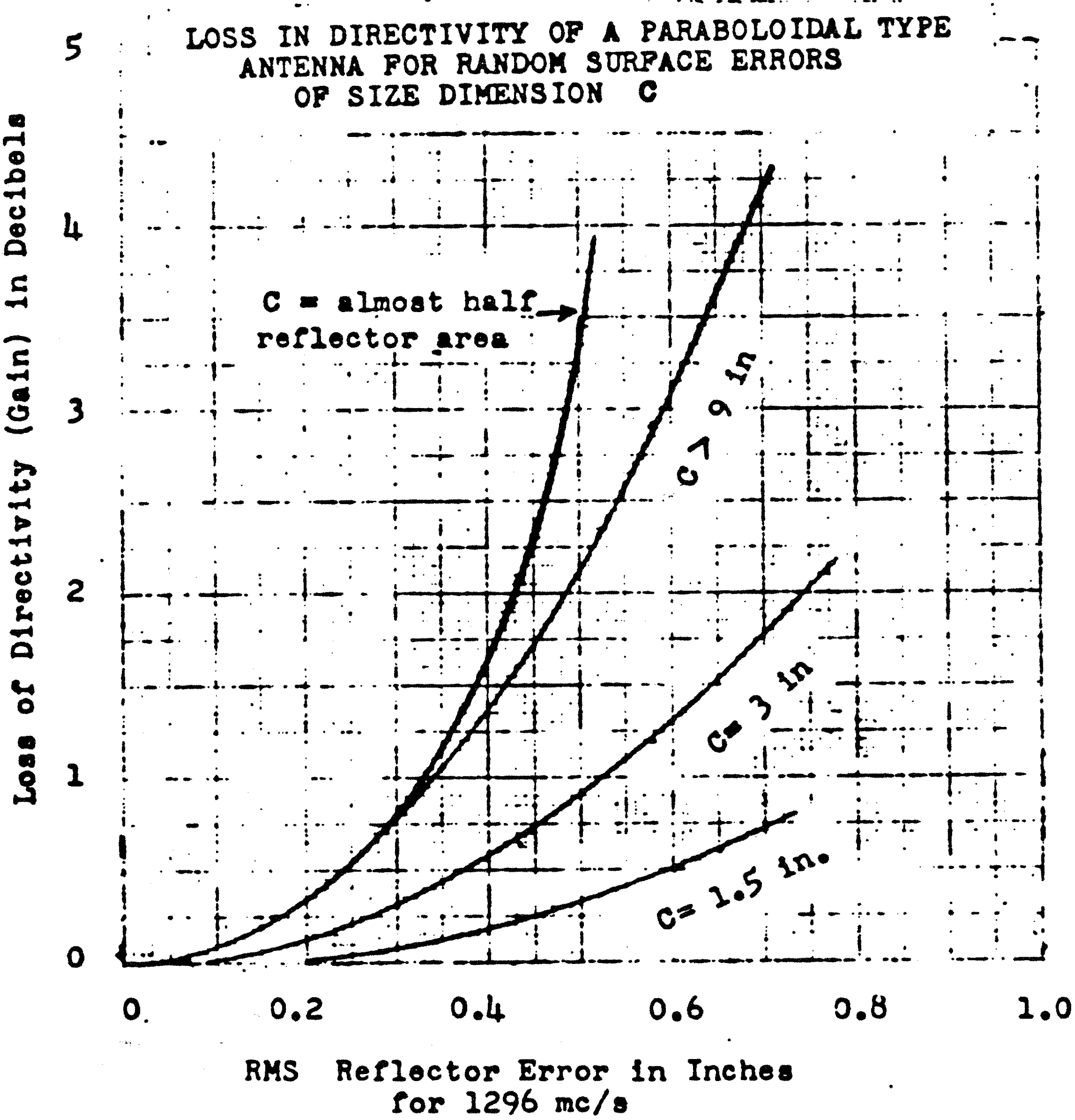


Figure 7. Physical dimensions for the rib section of the 20 foot diameter parabolic reflector design with 16 cyclindrical parabolic sections.

PARABOLIC RIB CURVE COORDINATES

inches		centimeters	
X	Z	X	Z
0	0	0	0
6	0.075	15.24	0.191
12	0.300	30.48	0.762
18	0.675	45.72	1.715
24	1.200	60.96	3.048
30	1.875	76.20	4.763
36	2.700	91.44	6.858
42	3.675	105.68	9.335
48	4.800	121.92	12.192
54	6.075	137.16	15.431
60	7.500	152.40	19.050
66	9.075	167.54	23.051
72	10.800	182.88	27.432
78	12.675	198.12	32.194
84	14.700	213.36	37.338
90	16.875	228.60	42.863
96	19.200	243.84	48.768
102	21.675	259.08	55.055
108	24.300	274.32	61.722

FIGURE 8


LOSS IN DIRECTIVITY OF A PARABOLOIDAL TYPE
ANTENNA FOR RANDOM SURFACE ERRORS
OF SIZE DIMENSION C
Loss of Directivity (Gain) in Decibels
5
4
3
2
1
0
C = almost half reflector area
C > 9 in
C= 3 in
C= 1.5 in.
0.
0.2
0.4
0.6
0.8
1.0
RMS Reflector Error in Inches
for 1296 mc/s

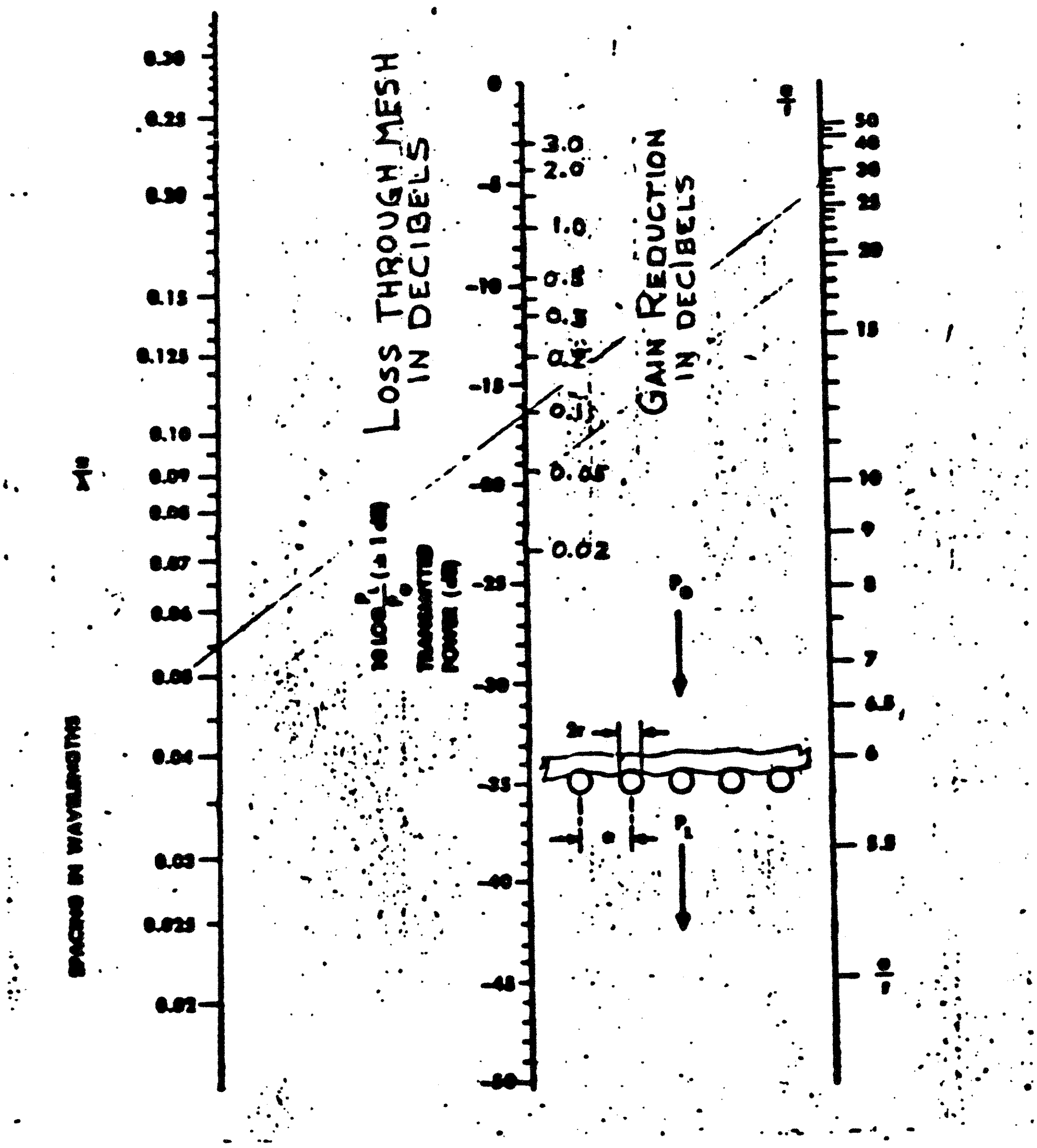


Figure 9. This nomograph relates transmission loss through a square mesh screen of conductive material, to the screen mesh and wire size. The resulting loss in antenna gain is also included for convenience.

As an example, suppose galvanized iron hardware cloth is selected for the surface material. The square mesh size is 1/2 x 1/2 inch and the wire diameter is 0.0340 inch. Therefore, a = 0.5 in., r = 0.0175 in. and λ = 9.13 in at 1296mc/s Compute a/λ = .055 and a/r = 28.5. Connect these values on the appropriate scales with a straight line as shown above and read the transmission loss as -15.3 db. The corresponding loss in antenna gain for this material will be 0.1 db.

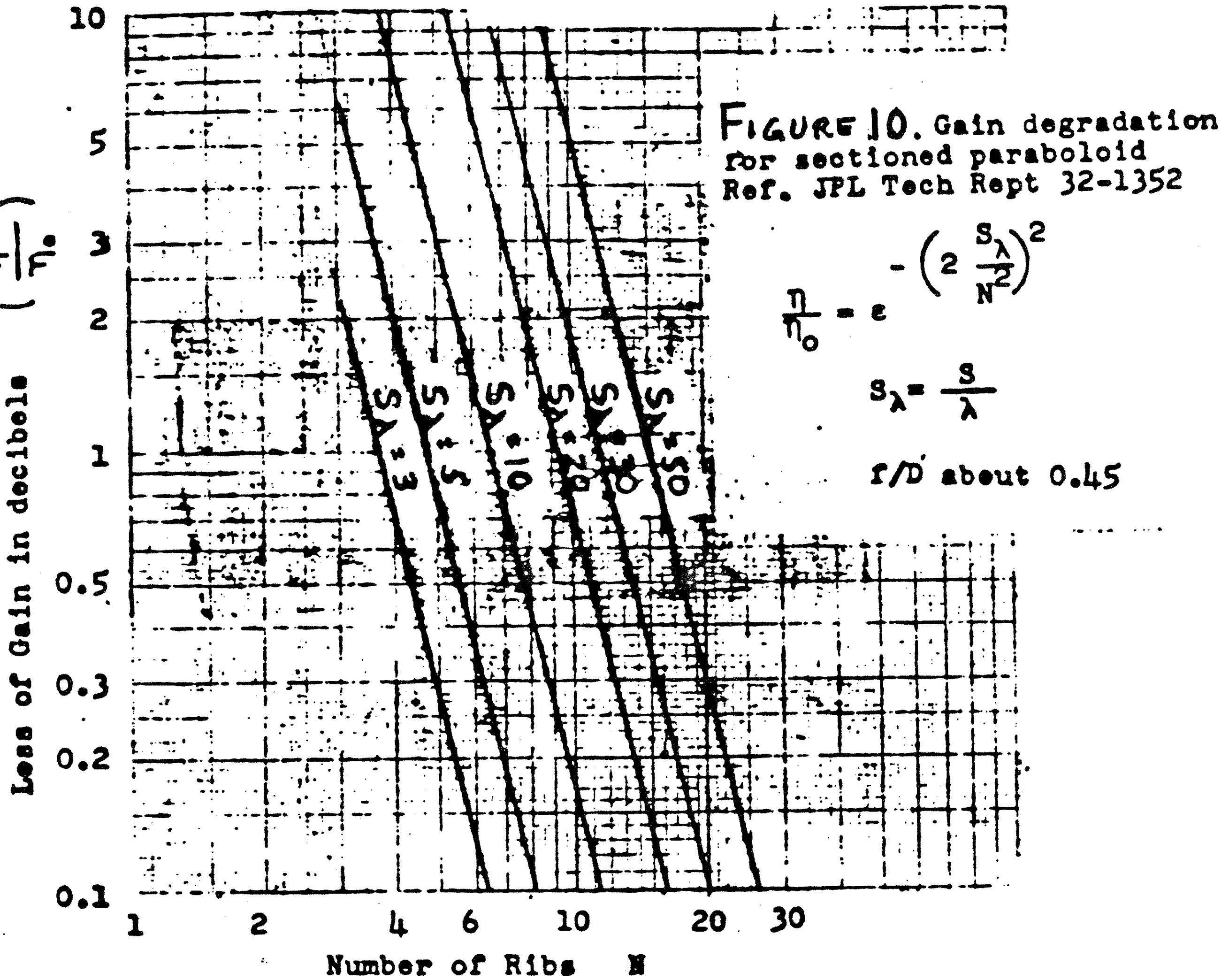
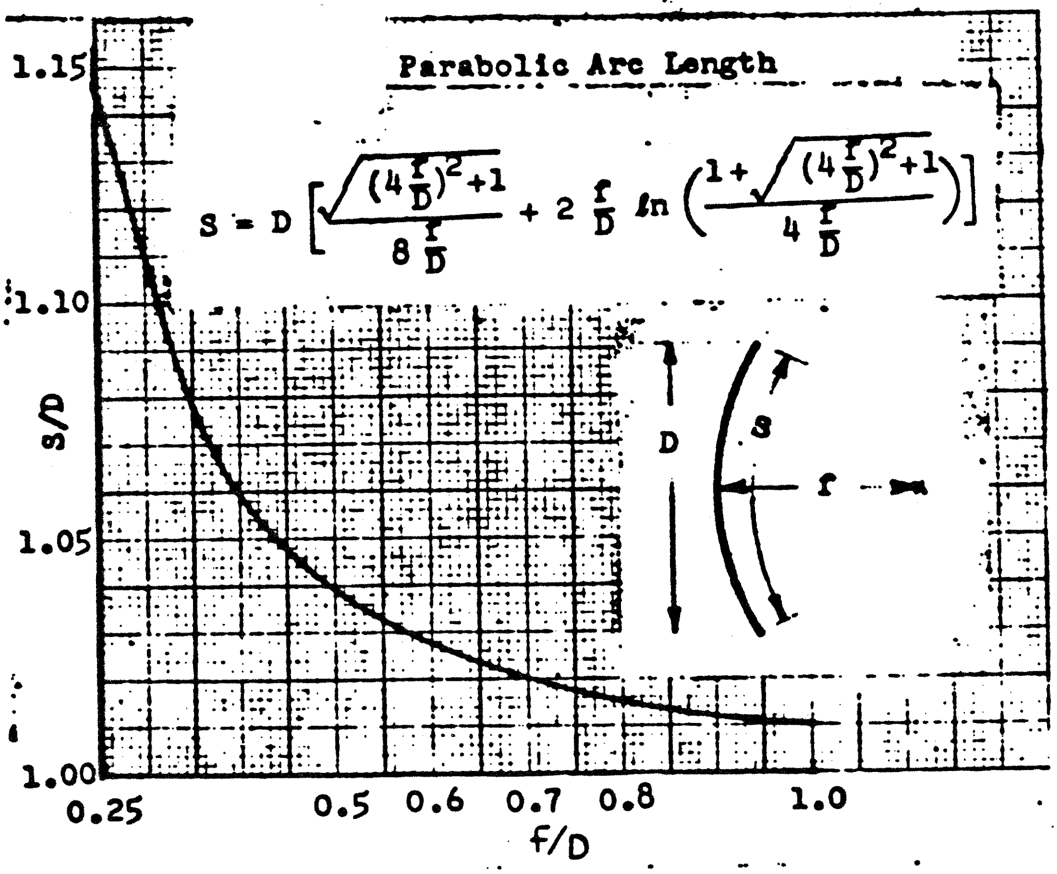
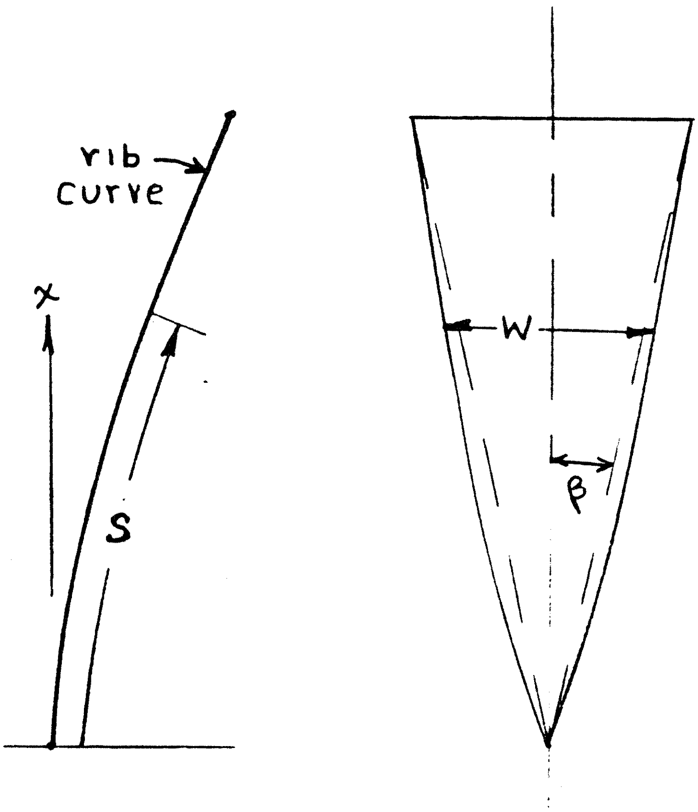


FIGURE 10. Gain degradation for sectioned paraboloid
Ref. JPL Tech Rept 32-1352





$$S = \frac{1}{4f}\left[X\sqrt{4f^2 + X^2} + 2f^2 \ln\left(\frac{X + \sqrt{4f^2 + X^2}}{2f}\right)\right]$$

$$W = 2\,X \tan(\beta)$$

Where β is half the angle between ribs, (for 16 ribs β = 11.25)

$\ln$ is the natural log (log to the base e)

Figure 11. Physical layout for petal surface section where sheet material with NO overlap is employed.

TECHNICAL REPORT # 6

From: The Crawford Hill VHF Club
Date: July 1971 (revised Oct 1986, see Additional Addendum, Oct 1986 (at end of this report).

Subject: A Water Cooled Power Amplifier for 1296 mc/s

General description

In this report a power amplifier is described which will deliver a CW output of at least 200 watts with a total plate circuit efficiency of better than 50%, and a power gain of 10 db at 1296 mc/s. This amplifier is water cooled and employs a pair of 7289/3CX100A5 planar triodes in parallel in a grounded grid circuit with provisions for individual metering and biasing of each tube. (See addendum notes on improved biasing method)

The input (cathode) circuit is an electrical halfwave resonant line section. Approximately half of this line is in stripline with adjustable capacitive end loading for tuning and adjustable input capacitive coupling (impedance matching). The remaining portion of the halfwave resonant line is the internal structure of the 7289s. The stripline is split in such a way as to provide tight r-f coupling between tubes, but d-c isolation for individual metering and biasing.

The anode (output) circuit is a halfwave cavity with the tubes plate-to-grid capacitance providing the reactive loading which foreshortens the electrical halfwave cavity to a physical length of 7.62 cm. (3 inches). The cavity can be considered as a section of reduced height, beyond cut-off rectangular waveguide with the tubes in the transverse plane. Tuning is accomplished by means of a sliding plunger on one side of the tubes, and output coupling is provided by an inductive coupling loop and coaxial line which protrudes through a fixed wall on the opposite side from the tuning plunger. Output loading (impedance matching) is obtained by rotating the coupling loop and moving the anode plunger.

The amplifier is constructed of standard brass bar, plate and sheet stock, and may be fabricated with a minimum of machine work. Since the resonant circuits are relatively low-Q, little improvement in performance can be obtained by silver plating. However, silver plating will enhance the appearance and greatly extend the life of the amplifier.

Water Cooling System

The water cooling system employed with this amplifier is a closed circulating system which NEVER permits the water to boil in the anode jackets. While this system is less efficient in extracting heat, it DOES provide the lowest operating temperature which is considered more important in this case. In addition, this closed system permits direct evaluation of power output by simple calorimetric means, as described in Technical Report # 7.

Each tube is fitted with a semi-permanent water jacket as shown by Figure 4. The jacket has two water connections which permit the jackets to be attached to the water reservoir and pump by means of 3/8 inch I.D. (inside diameter) plastic tubing as indicated in Figure 1.

The series connected water jackets maintain equal flow in each jacket and with reasonable flow (1/2 gallon per minute minimum) the differential temperature between the tubes will be negligible.

Distilled water should be used and a reservoir of 1 to 2 gallons should be adequate for intermittent amateur service. Sufficient flow rate can be easily obtained from a small fish tank or fountain pump.* Since the anode water jackets are at high voltage potential, it IS NECESSARY to provide a protective ground for the water supply and pump. This is provded for by two short metal nipples inserted in the water lines and grounded to the amplifier ground system (see Figure 1.). A minimum of one foot (30 cm.) of plastic tube should be allowed between the tube water jacket and nipple in each line. The d-c resistance of distilled water in a 3/8 inch I.D. clear vinyl tubing is about one megohm per foot. The nipples also provide a convenient disconnect point for the water supply when maintenance is necessary and should be located physically with the amplifier. Should water leakage current exceed 5 ma., as indicated by the anode current meters with H.V. applied but no heater current to the tubes, this indicates that the water has become contaminated. The system should be flushed and a new supply of distilled water installed.

The amplifier should NEVER be operated without water flow. Although no flow interlock device is shown, some means of indicating flow should be used. A simple visual device may be constructed as shown in Figure 1, and consists of a brightly colored thread or ribbon secured to a pin which is inserted diametrically through a rigid plastic nipple. If the device is placed horizontally, flow will be indicated by movement of the ribbon while lack of flow is indicated by the ribbon settling to the bottom of the nipple and remaining stationary.

A suitable reservoir may be made from a variety of readily available plastic containers. Choose one with flat sides so that tubing connectors may be installed with minimum trouble. Also choose one which can be closed or has a cover to prevent external contamination. Do NOT use metal or some plastics, such as Plexiglas which is hygroscopic and will eventually warp when in contact with water. With a metal reservoir, electrolysis and corrosion will contaminate the water.

It will be necessary to purge the system initially to clear all air from the lines and jackets, and it may be necessary to repeat the purge operation after long idle periods. Pinching the flexible tubing where air pockets occur, with the pump running, will remove most of the air.

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* Consult supply catalogues, Sears Roebuck and Company for one.

Construction Details

All important dimensions and most construction details should be self-evident from the drawings. While substitution of materials may be made, the important inside dimensions shown should be adhered to carefully.

When soldering in the finger stock it is advisable to first form the stock into rings to fit the hole size and then use a 'dummy' tube or an equivalent size aluminum rod to hold the finger stock in place while soldering. Install finger stock last and use a minimum of heat as the stock material may be annealed with too much heat, resulting in poor contact with the tube rings. If a substitute grid finger stock is used, no guarantee can be made that the anode circuit will resonate at 1296 mc/s since the physical shape immediately adjacent to the tube can change the reactive loading. Note that NO finger stock is required at the anodes. The resonant quarter wavelength anode chokes MUST use Teflon for a dielectric in order to maintain the correct electrical length. About 1/16 of an inch of the dielectric should be allowed to protrude into the anode cavity, and about 1/4 inch above the top of the outer cylinder. The Teflon is wrapped tightly around the water jacket as shown in Figure 5, and then the assembly is carefully slipped into the outer cylinder.

Special care should be observed to see that ALL tube socketing details are in good coaxial alinement so that no undue transverse forces are applied to the tubes when they are in place. A locking arrangement at the top of the water jackets forces the tubes down into the amplifier securely.

It is recommended that all parts which will be initially secured together, be sweated together permanently. This includes the basic parts shown in Figure 3, and also the complete anode cover plate. In this way the critical areas of high r-f current will be assured of good electrical conductivity, especially in the anode cavity immediately adjacent to the tubes. The various pieces should of course be fitted together before sweating to check and correct alinement and surface warpage. The grid ring finger stock should be installed last as indicated previously and with minimum heating. It has been found that in some amplifiers, poor anode circuit efficiency has been traced to loss of tension in the finger stock at the grids and thus poor electrical conductivity. This can be caused by over heating during installation, or from insufficient cooling during long periods of operation.

After sweating together all the basic parts and fabricating the remaining parts, the entire amplifier should be completely disassembled and thoroughly scrubbed, including the tubes, with a stiff brush, detergent soap and water. Each part should be carefully clean dried with absorbent towel and air blast where necessary. Be sure that all traces of solder flux and rosin are removed especially from inside the cavities. Round off all corners and sharp edges in the r-f resonant sections especially before the cleaning operation for electrical and personal safety. The heater chokes which are quarter wavelength resonant stubs are simply a means of physically supporting the heater chuck with negligible r-f distrubance to the cathode tank circuit. Other means may be used to support this chuck physically, and a simple wire wound RFC can be used for r-f decoupling.

Although adequate cooling of the anodes is provided by the water circulating system, the grid and cathode tube seals must be cooled. This is provided for by a small blower which forces air through the 3/8 inch holes in both the cathode and anode side walls on one side of the amplifier and exits the air through similar holes on the opposite side. A plenum between blower and air holes would be highly desirable for efficient use of available air flow.

In the water tubing system it may be desirable to have sharp bends in the flexible tubing. Most non-rigid tubing may be permanently bent into sharp radius curves without sacrifice of cross sectional area by the following method. First pack the section of tube to be bent with fine sand. Next, gently and uniformly heat the section to be bent on a heater or open flame. Do not overheat and check a small sample of the tubing for combustion hazard. When the material becomes noticeably softened while rotating and moving the section over the heat source, remove from the heat and quickly make the desired bend slightly less in radius than desired. Finally cool the section under running water while holding the shape. Remove and flush out the sand.

No mounting details are given here for this amplifier as individual circumstances and needs will suggest these details. There are no restriction in mounting position for this amplifier, as long as air pockets are removed from each water jacket.

Tuning and Operation

Initial adjustments of this amplifier are similar to any new amplifier, that is they should be done with caution and perception at reduced voltages and power level. Apply about 500 volts to the anode, circulating water and sufficient bias to permit 10 to 30 ma. of idling anode current on each tube. Have some termination (50 ohm) fo the amplifier to dissipate r-f power into, such as a dummy load or antenna fairly well matched and aimed away from all living things. Next apply sufficient drive to raise the anode current to about 50 ma. per tube and attempt to tune the anode cavity to resonance. The output link coupling should be in a position of nearly minimum coupling (plane of the loop parallel with top cavity wall is minimum coupling). The usual sharp dip in anode current should be observed as an indication of cavity resonance at about midrange of the tuning plunger. If the dip is not pronounced it may indicate faulty construction of the anode cavity which will greatly reduce the Q of this tank. Resonance at the extreme end of the tuning plunger range usually indicates that the cavity dimensions, tube seating or finger stock placement is not correct. A moderate correction in anode resonance can be achieved by moving the cavity wall which is opposite the plunger. It is advicable to initially secure this wall by means of two 'C' clamps and adjust its position so that the tuning plunger will resonate the cavity at midrange of the plunger extremes.

Next check the cathode line resonance to see that it can be obtained at a reasonable point in the tuning range of the disc capacitor. This resonance is quite broad and can be observed by peaking of the cathode currents at resonance. When resonances have been established, anode voltage may be increased and input and output couplings adjusted for maximum power output. Be sure to check both input and output resonance tuning with each change in coupling.

At full anode voltage and bias, no anode current should flow when drive is removed. When full operating conditions are reached with r-f drive, the cathode bias on each tube may be trimmed so that each tube draws about the same anode current. Depending on the condition of <u>your</u> tubes, the final operating conditions will be approximately:

Eb = 2000 volts (d-c)

Ib = 125 ma. (per tube)

Ik = 150 ma. (per tube)

Ekg = 45 volts (zener + resistor bias) per tube

Po = 250 watts (r-f)

Pin anode d-c = 500 watts

Pdrive r-f = 25 watts

Under these conditions and properly adjusted, the water in the tube jackets should not boil. Onset of boiling is indicated by a 'popping' sound in the jackets. Off resonance anode tuning or over coupling at the output can cause sufficient inefficiency which may cause boiling.

Although this amplifier has been operated with cathode current in excess of manufacturers maximum ratings (125 ma. per tube), which will probably shorten tube life, it is felt that for intermittent amateur service especially in the keyed telegraph mode, the strain on the tube may be justifiable.

No attempt has been made to modulate this amplifier or check its linearity at full power. At reduced levels the amplifier will deliver 100 watts output with 1000 volts on the anode and about 20 volts bias. Under these conditions tube life should be according to manufacturers specifications and the amplifier may be high level modulated for AM service.

Precautionary measures should be taken with tubes of the 7289 type to limit peak anode current in the event of internal flashover. A 300 ohm 25 watt resistor between the amplifier and 2000 volts supply is adequate to limit the peak current to about 6 amps with a loss in anode voltage under operating conditions of about 100 volts. In addition, a high voltage type fuse is recommended to prevent burn out of the current limiting resistor or the primary of the anode supply transformer may be fused to just above operating conditions.

This amplifier, as with any grounded grid amplifier, can become unstable without output loading. It is highly recommended that in station operation the antenna switching relay and high voltage control relay be sequenced or interlocked in such a way as to prevent the amplifier from having high voltage without r-f load at any time and especially during switching.
It is also possible for this amplifier to become unstable (oscillate) at lower r-f due to external leads in the cathode and anode returns. This latter type instability may be suppressed by suitable resistors and by pass capacitors expernal to the 1296 mc/s circuitry. A suitable suppressor is shown by Figure 1. and is highly recommended. The by-pass capacitors are high voltage mica (Sangamo). In addition, VHF parasitics may appear as a result of cathode RFC resonance. Should this occur, the cathode RFC may be altered to more or less turns. Little effect should be noted at 1296 mc/s for large variations in the size of this choke.

Remarks
The original design of this amplifier was by W2CCY in a convection water cooled form and later modified by W2CQH to an air cooled design which was published in March 1970 HAM RADIO magazine, page 43. The circulating water cooled system was suggested by K6MYC aqnd the amplifier described here was built and measured by W2IMU and the Crawford Hill VHF Club.
Some versions of the 7289 tube have screw-on air cooling fin assemblies rather than the split collar and set screw arrangement. The former may be separated from the tube by first filing or machining away the bottom fin which is adjacent to the top of the anode cup. This will release the strain on the threading and the fin assembly may be unscrewed easily (right hand thread).
Two amplifiers may be combined in parallel by means of quadrature hybrids (Tech. Report # 2) to obtain 500 watts output for EME application. See also Technical Report # 19.

Addendum (August 1973, revised Oct 1986)

It has been several years since this water cooled amplifier was built and some additional experiences have since been gained.

1. Although water cooling does permit higher anode dissipation to be achieved, the high power limit of the 7289 type tube may be fixed more by grid dissipation. It is therefore stressed that extra air flow at the grid finger stock be employed. One way to achieve this is by using the anode cavity as a pressurized plenum forcing air to pass through the grid finger stock and exit into the cathode enclosure. This method requires a better blower than indicated previously.

Excess grid dissipation is caused by over driving as well as a lossy anode cavity, incorrect tuning and loading of the anode cavity. An indication of a faulty anode circuit is when anode current 'dip' and maximum power output are not coincident with anode tuning.

Tube failures will occur when excessive grid dissipation is present by physical warpage of the grid screen mesh, eventually causing a grid-to-anode or grid-to-cathode short circuit.

Grid warpage due to excessive dissipation and or poor cooling of the grid ring is also evident by drift in anode circuit tuning as the amplifier runs.

Should any of the above effects be noted, the amplifier should be dismantled and examined carefully for faulty construction or materials.

2. The water system will contaminate over a period of a year or so. Contamination accelerates electrolysis and damages the anode cup and pinch-off tube in the cup. If allowed to persist, the metal wall may actually puncture with catastrophic results. Electrolysis may be slowed down by changing the distilled water frequently and increasing the hose length between amplifier and grounded water reservoir.

3. The cathode stripline resonator which includes a d-c blocking capacitance may be fabricated from 1/16 inch thick double sided copper clad Fibreglas printed circuit board. Using the Fibreglas as a dielectric, the part is made essentially with the same dimensions as the original shown by Figure 5. The copper may be either etched away at appropriate places with suitable masking, or may be removed mechanically by scraping or peeling. Since the resonator is low-Q, the dielectric losses should be negligible here.

Additional Notes (November 1986)

Additional experience and information now avilable indicates that water cooling has gained favor amongst EME enthusiasts. Numerous designs have emerged to modify the 7289/3CX100A5 for water cooling which are easier to fabricate that the type shown in Report # 6. See the K2UYH "432 mc/s and Above" newsletter technical section (1985-6) for some designs.

A great improvement in the power output can be obtained from recent developmental work by N6CA*, K6UQH and members of the Eimac Company.
The modification is to replace the zener/resistor cathode bias arrangement with an electronically regulated d-c cathode bias supply. This very "stiff" bias voltage supply permits much higher peak anode current in the tubes which results in a power output increase of at least 2.
The attached Figure shows one version of this cathode bias regulator for use with the two tube amplifier.

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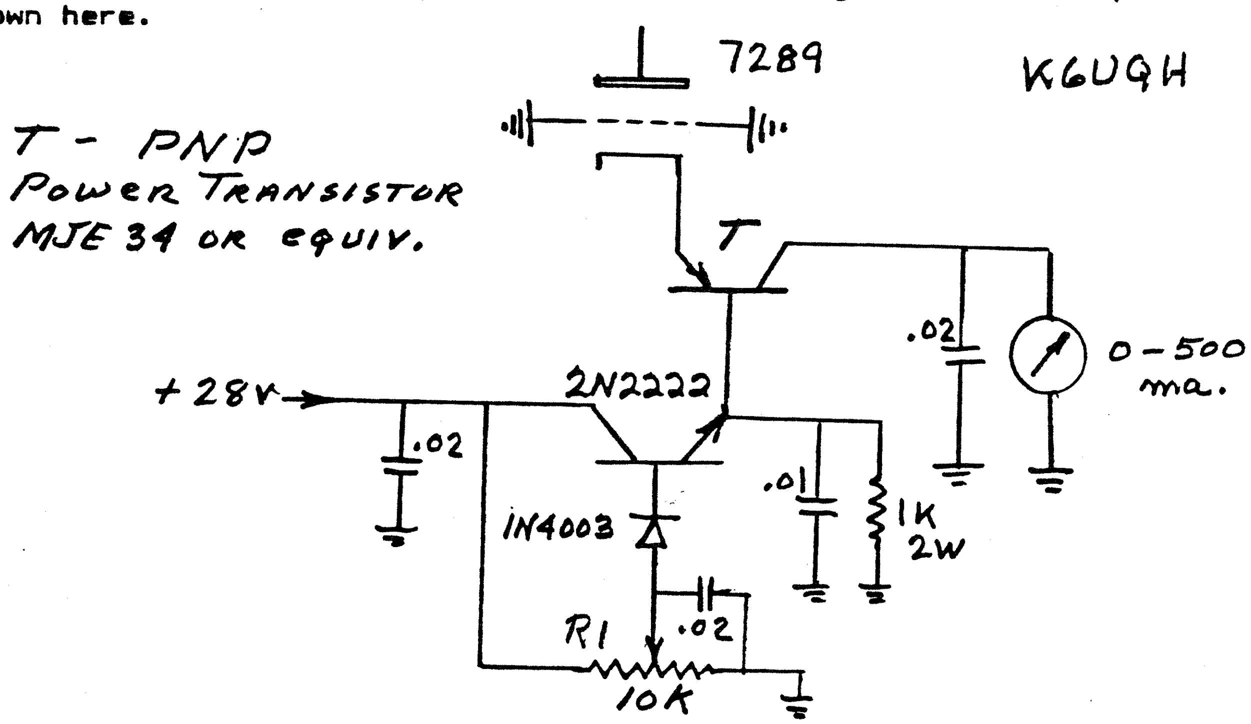
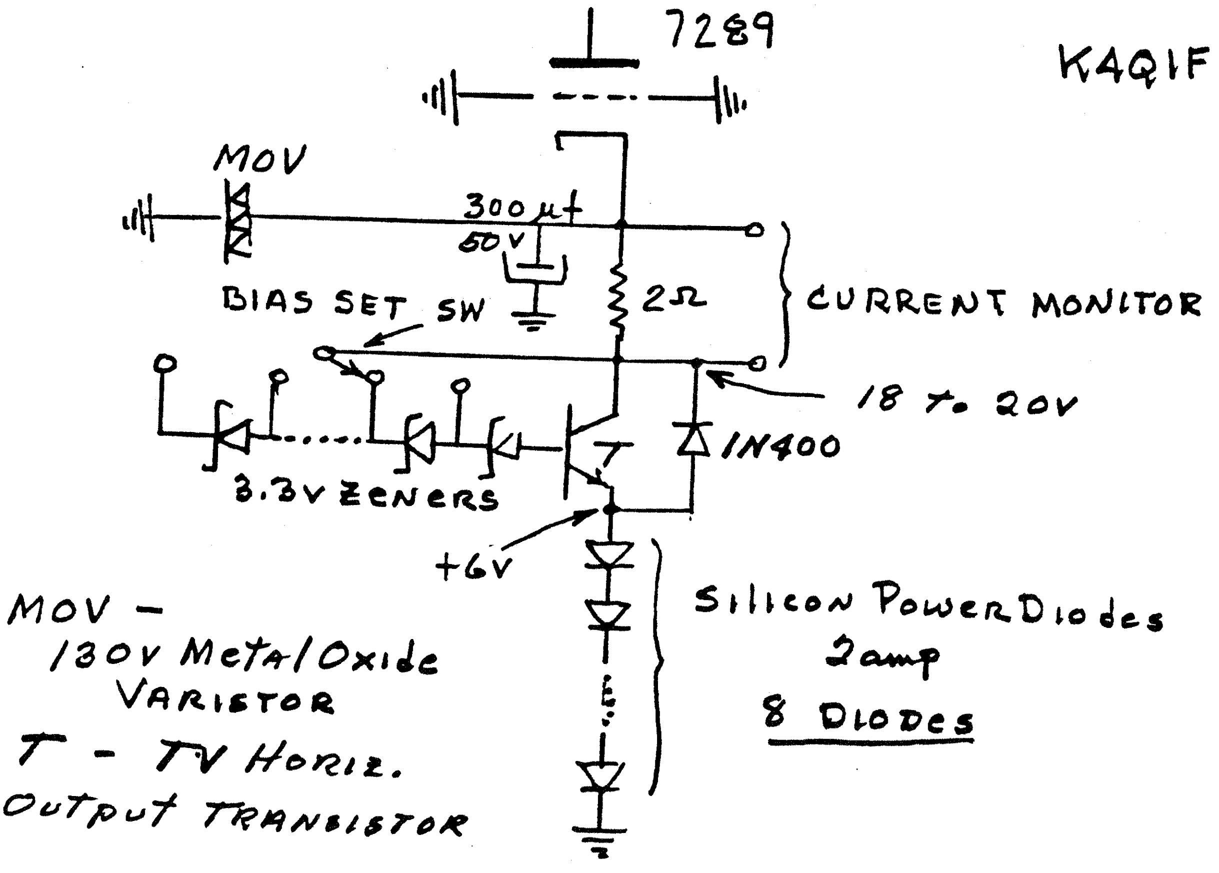
* See QST March-April 1985, " A Quarter-Kilowatt 23-CM Amplifier"

Improved cathode bias for grounded grid power amplifiers using planar triodes of the 7289/3CX100A5 type.
The upper circuit suggested by K6UQH requires a +28 volts low current supply to bias the regulator circuit. The potentiometer, R1, adjusts the cathode current level.
The lower circuit suggested by K4QIF requires no additional supply and is adjustable in steps of 3.3 volts.
Each 7289 tube requires a separate cathode bias regulator circuit, as shown here.





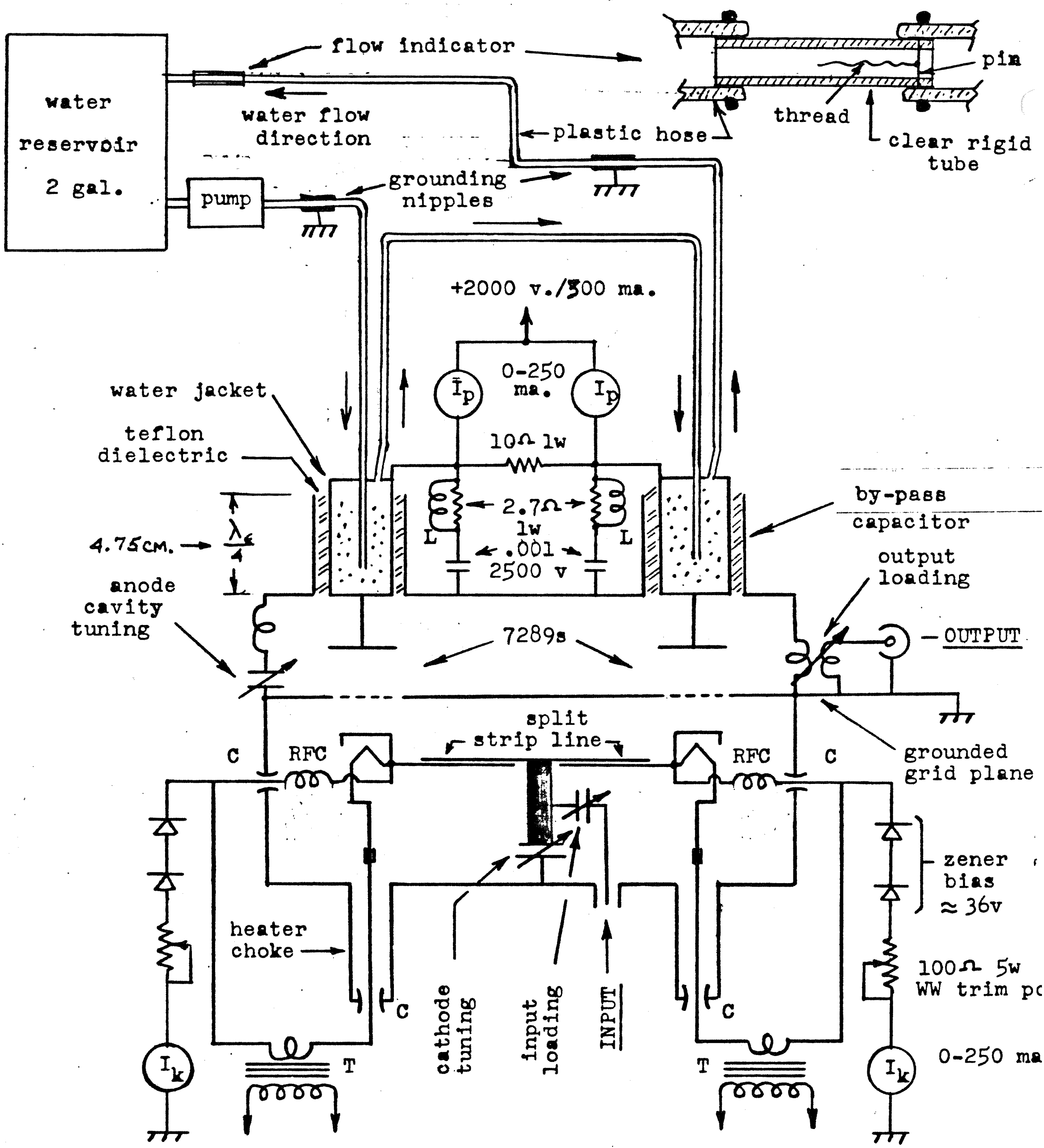


Figure 1. Electrical schematic and water system diagram of the dual 7289/3CX100A5 grounded grid power amplifier for 1296 mc/s. RFC is 6 t #18 Formex air wound on 1/8 inch diam. L is 3 t #22 wire wound directly on the 2.7 ohm 1 watt composition resistor. C is 500 pf feedthru capacitor. T is heater transformer 6.3 v at 1 amp.

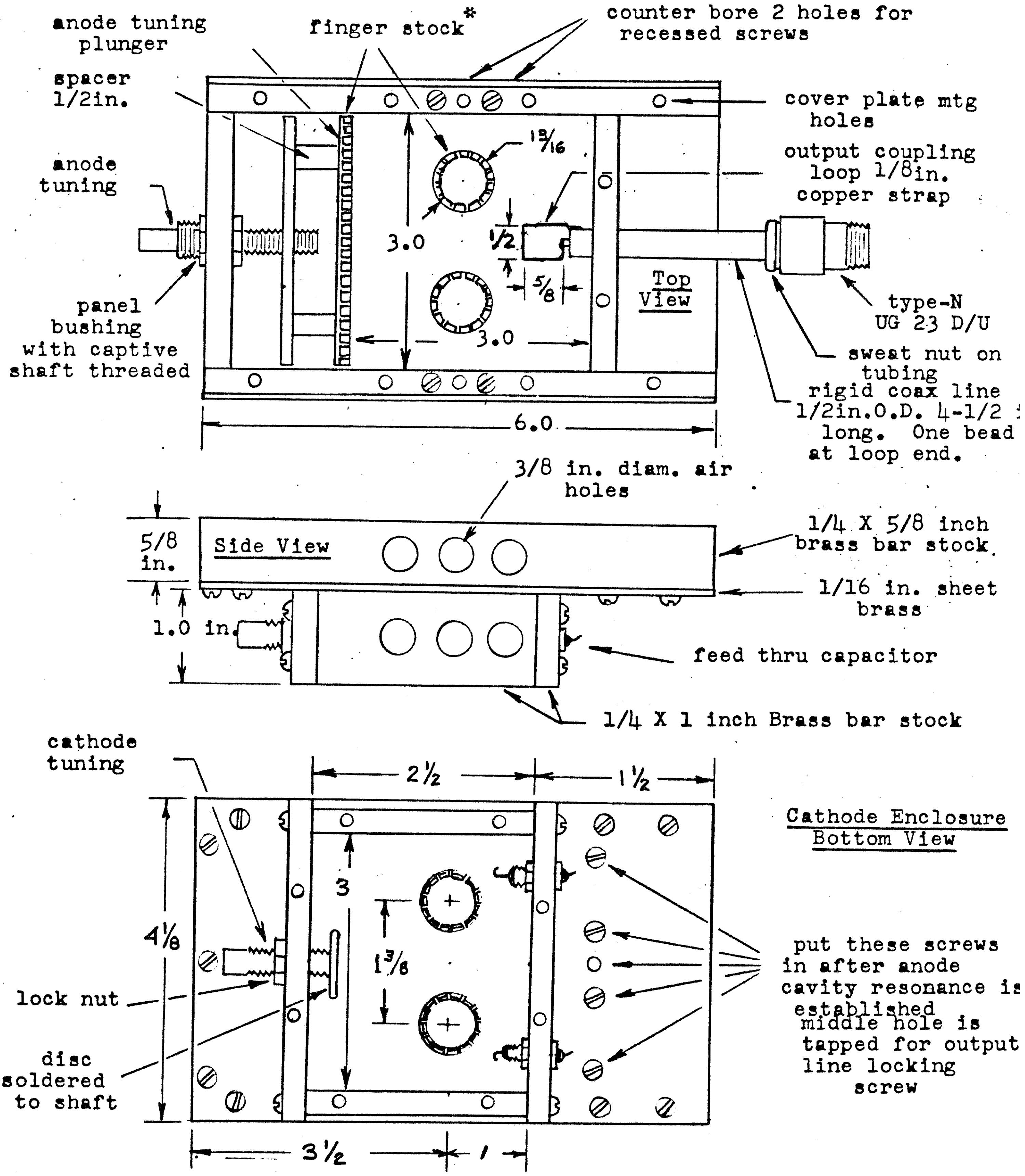


Figure 2. Detailed views of amplifier without top and bottom cover plates. All dimensions are in inches.

* All finger stock is Instrument Specialties Co. part # 97-251 Little Falls, New Jersey or equivalent.

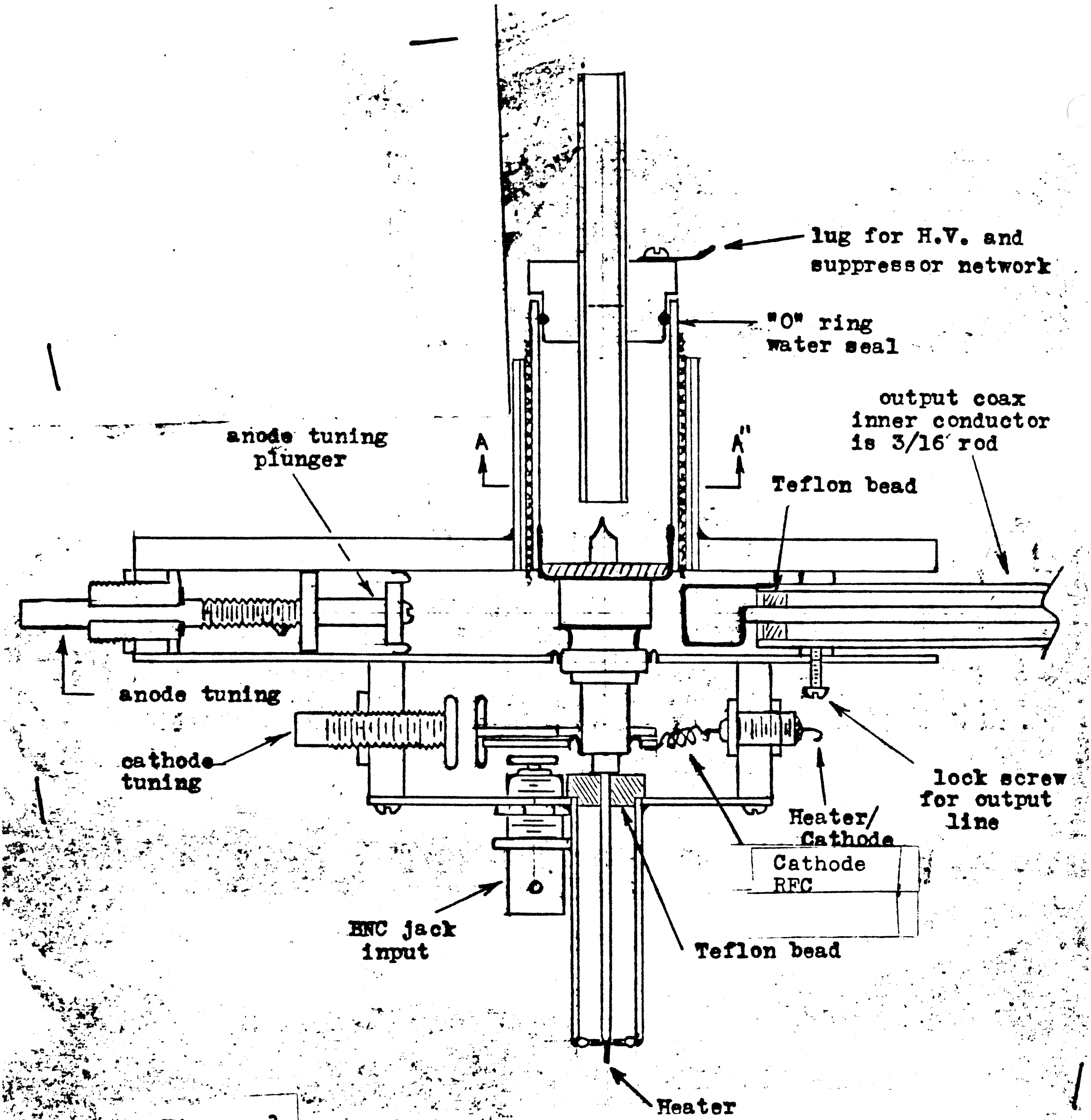


Figure 3. Exposed side view showing assembly details. See other drawings for dimensions. This drawing is approximately full scale.

The dielectric in the coaxial by-pass capacitor consists of sheet teflon, 0.001 to 0.005 inch thick, tightly wound around the water jacket to fill the space between the cylinders. There is about 0.030 inch of space to be filled as completely as possible. A small tab of clear Scotch tape may be used to secure the inner end of the teflon tape to the water jacket for winding convenience.

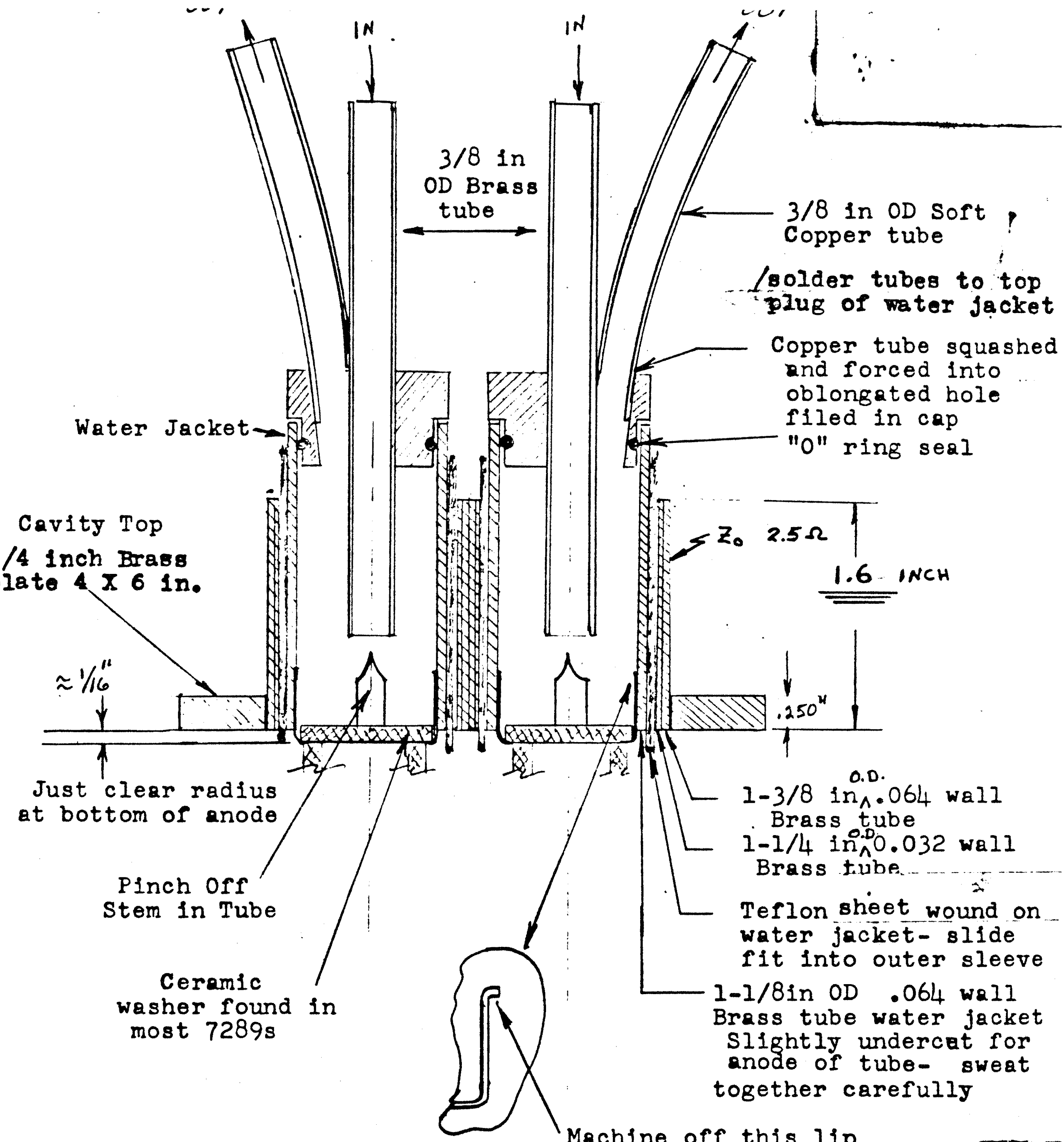


Figure 4. Cross sectional view of water jacket assembly and resonant coaxial by-pass capacitor. Standard brass tube is indicated for convenience. The 1-3/8 and 1-1/4 inch O.D. tubes are forced together and soldered.

The water jackets are soldered directly to the anode cups of the 7289s. The lip at the top of the anode cup on most tubes should be carefully machined or filed off. Bring the jacket with tube inserted slowly up to temperature on a hot-plate, solder, then allow to cool very slowly to avoid straining the ceramic to metal brazed seals of the tubes. A piece of aluminum foil wrapped around the ceramic part of the tube will prevent contamination of this area by flux splatter during soldering.

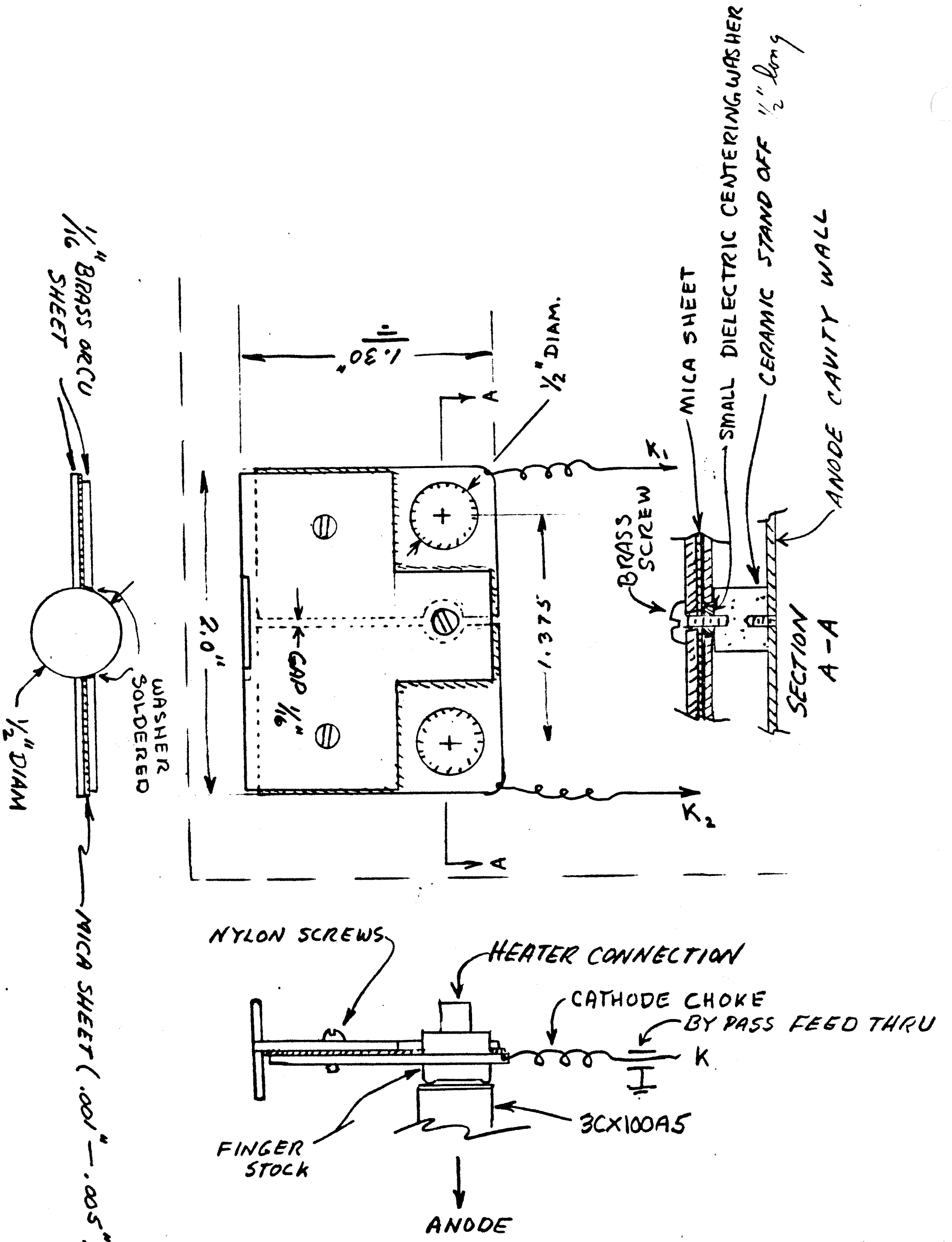
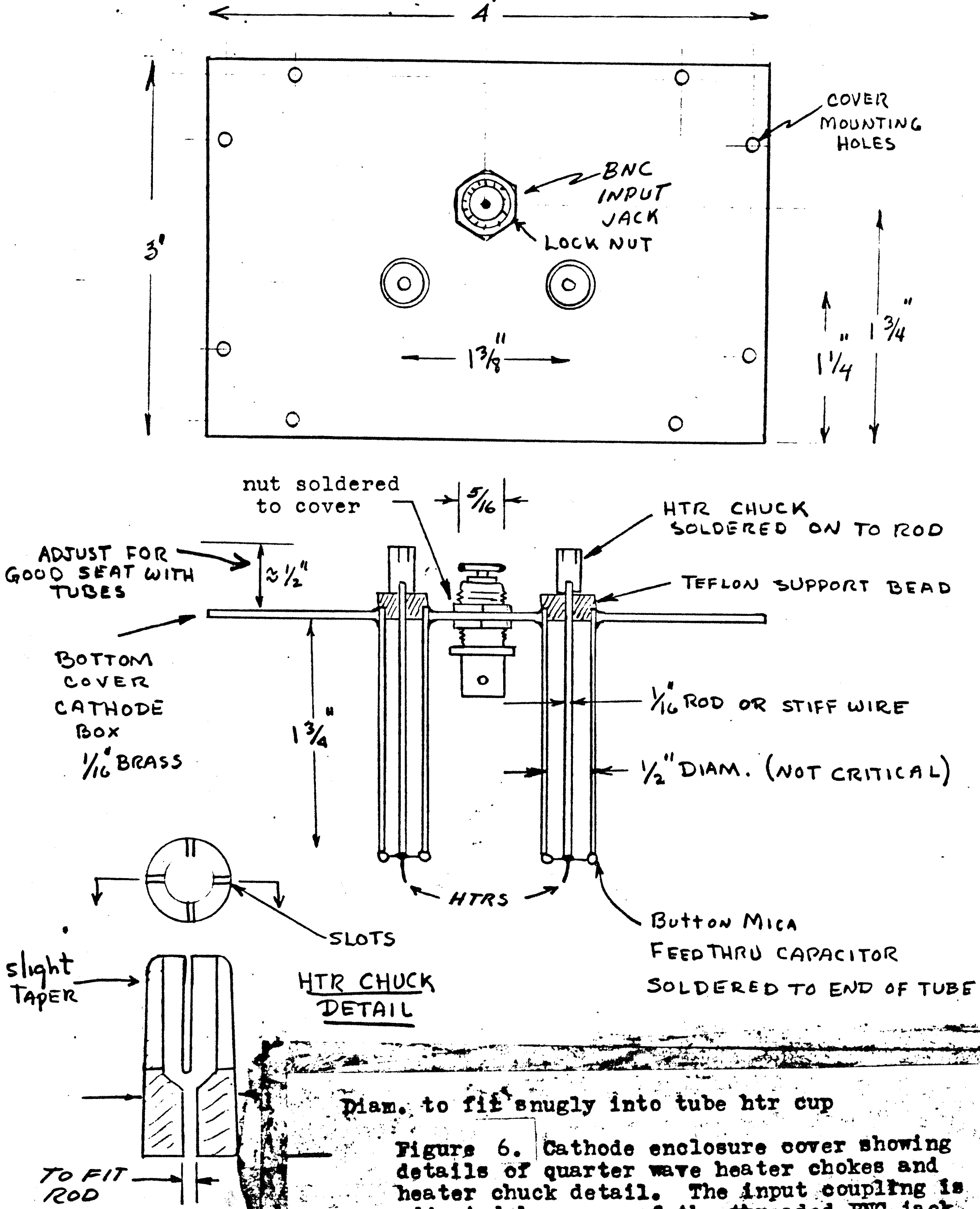


Figure 5. Cathode strip line details. The line is split down the center to permit separate biasing and metering of the two tubes. All dimensions are in inches. Mylar may be substituted for the mica dielectric. The line sections should be very flat so that maximum capacitance occurs in the dielectric sandwich.



Diam. to fit snugly into tube htr cup

Figure 6. Cathode enclosure cover showing details of quarter wave heater chokes and heater chuck detail. The input coupling is adjusted by means of the threaded BNC jack. A lock nut is used on the outside and a simila[r] nut may be soldered to the inside of the cove[r] plate, or the plate may be tapped 3/8-32.

Technical Report # 7

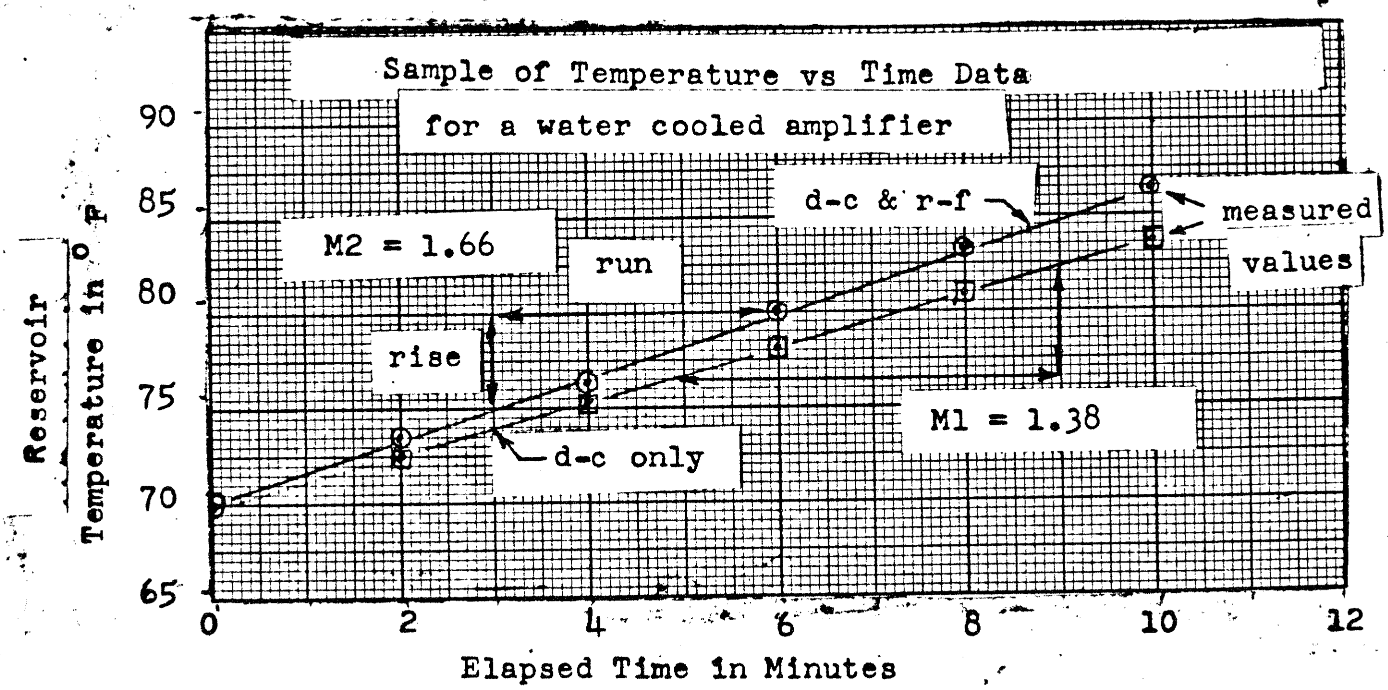
From: The Crawford Hill VHF Club, W2NFA Date: December, 1971

Subject: POWER MEASUREMENT at 1296 mc/s

Direct power measurement at 1296 mc/s in the range above a few watts has always presented a problem for the experimenter unless he is fortunate enough to have one of the few direct reading calibrated high-power watt meters for the L-band frequencies. This report presents a very inexpensive but accurate method of determining output power of an amplifier at 1296 mc/s by a heat equivalence method. The accuracy is largely determined by the accuracy of the d-c meters used to measure the d-c input power to the amplifier. The disadvantage of this method is that it is not instantaneous but is an integration process requiring a finite amount of time during which the amplifier must be maintained at constant input and output levels. The method can be applied to both air and water cooled amplifiers with suitable procedures for each. This method is expecially useful to those experimenters who must rely totally on their own resources for measurement of r-f power.

Water Cooled Amplifiers

In a water cooled amplifier where the flow rate is sufficient so that the water temperature in the anode jackets is essentially the reservoir temperature, the anode dissipation appears as a rise in reservoir temperature. If the cooling system is at room temperature to begin with and the anode dissipation is held at some constant value, for instance 500 watts d-c with no r-f drive, the temperature of the reservoir will rise in a very linear manner over a period determined mainly by the quantity of water in the reservoir. A 2 gallon reservoir will rise in temperature from 70°F (21°C) to 80°F (26.5°C) in about 6 minutes with about 500 watts of power dissipated into the cooling system. The slope of the line graph plotted from measured data, temp. v.s. time, is therefore a measure of the power going into the cooling system. Samples of this data are shown plotted below.



If the power dissipated into the reservoir is continued for a longer period, eventually the rise in temperature will no longer be linear but will tend to saturate mainly due to radiation and conduction heat losses.

Thus far d-c calibration for a constant known dissipation was measured. It is a simple extension of this notion to realize that with an amplifier operated AØ on a single frequency, the difference between the d-c anode input power and the power dissipated into the water cooling system must be the r-f power output delivered by the amplifier.

The only additional piece of equipment required is an inexpensive thermometer in the reservoir. The thermometer need not be accurately calibrated as long as it can be read in the temperature range of interest with about 1 or 2 degree steps. The thermometer is used to establish the slope of the linear temperature vs time graph and it is immaterial to the method what temperature scale is used.

The precautions which need be followed in this method are:

(a) Maintain the system unchanged during a series of measurements,

(b) Maintain the volume of water constant in the system, and

(c) Always start a measurement at or near room temperature. This latter precaution may be very time consuming and it may be expedient to add a sufficient quantity of ice to the reservoir between measurements to force the reservoir temperature down to room temperature correcting the water volume as required.

In order to estimate how your particular system water temperature will rise it takes 146.8 watts of power to raise one gallon of water $1^{o}F$ in one minute at standard Earth pressure, 14.7 pounds per square inch.

$$1 \text{ gallon} = 3.785 \text{ liters}$$

$$\text{Temp.}^{o}\text{ Centigrade} = \frac{5}{9}\left(\text{Temp. }^{o}F - 32^{o}F\right)$$

Procedure

Suppose a water cooled power amplifier for 1296 mc/s has the capability of the maximum U.S.A. legal d-c plate input power of 1000 watts, and its efficiency is expected to be approximately 50%. The expected r-f output power should be 500 watts. To actually determine the output power, start with the system water temperature at approximately $70^{o}F$ and operate the amplifier in a normal manner, tuned and loaded for maximum r-f output power with the d-c plate input power set and maintained at a constant 1000 watts.

Take a reservoir temperature reading at regular intervals say 1 minute apart and plot this data on rectangular graph paper as shown on page 1. Recording many points will allow a more accurate determination of the slope and also a check on the saturation temperature. The plot should be quite linear over at least the first 10^{o} rise in temperature. The slope of this linear plot of measured data is now determined in ^{o}F/minute (rise divided by run) as shown by the graph on page 1. The slope in this case is determined to be 1.66 ^{o}F.min.

Next, the reservoir temperature is returned to $70^{o}F$ and the amplifier bias is altered so that without r-f drive, the amplifier will draw sufficient total anode current so that the d-c plate input power will be 500 watts. Repeat the temperature recordings at 1 minute intervals and plot this data on the same graph paper. A typical set of measured values are also shown by the graph on page 1. Again determine the slope of this second linear plot to be $1.38^{o}F$/minute in this case.

The r-f output power is then

$$P_{r\text{-}f\ out} = 1000\ w. - \frac{1.66}{1.38} \times 500\ w. = 400\ w.$$

As a further check on the measurement, another d-c only run can be made with the d-c input power set for 600 watts, for this example. The resulting plot should have exactly the same slope as in the case when the amplifier was delivering r-f output power.

Air Cooled Amplifiers

A similar output power determining technique may be used with air cooled amplifiers. In this case the thermometer is placed in the exhaust air stream and the exhaust tempertaure is allowed to stabilize (saturate) for each measurement. Precautions similar to the water cooled case are:

(a) Sufficient air flow be maintained so that the tubes do no overheat in normal operation,

(b) Arrange the plenum so that all tubes (if multi tube amplifier) receive approximately equal air flow. And more important that all exhaust air streams are collected into one exhaust port where the thermometer is located.

(c) Locate the exhaust thermometer about six inches or so away from the amplifier proper so that direct radiation from hot metal parts does not effect the exhaust temperature readings.

Procedure

The procedure in the air cooled amplifier case is to first operate the amplifier in its normal mode delivering r-f power to the antenna or dummy load for a sufficient time to permit the exhaust temperature to stabilize. Typically, a few minutes depending on the air flow rate will be long enough. Record this temperature carefully and record the total d-c plate input power, $I_{dc} X\ E_{dc}$ total measured directly in the anode circuit. The absolute temperature is unimportant.

Next with NO r-f drive to the amplifier, adjust the bias so that the amplifier draws anode current and is stable (no oscillations). Keep the load on the anode output circuit to prevent possible oscillations. The amount of current drawn by the amplifier anodes should at first be set according to the expected approximate efficiency. That is, if the expected efficiency is 50% then set the bias so that the anodes receive about half the d-c current that was measured with d-c plus r-f operating conditions. Now note the exhaust temperature when stabilized. If it is higher than the recorded temperture measured under d-c plus r-f conditions, lower the d-c anode current (input power d-c) by adjusting the bias and vice versa. The object is to match the temperature reading obtained under the d-c plus r-f conditions with d-c alone (heat equivalence). When this has been done, record the total d-c input power.

The r-f power output of the amplifier is then the difference between the anode d-c power under normal operating conditions and the d-c anode power under no r-f drive conditions. The simplicity and directness of power determination by this method should now be evident and also that the thermometer calibration plays no part in the accuracy of the results only acting as a reference of exhaust heat equivalence.

Remarks

A special comment should be made regarding determination of anode current and voltage for computing total d-c input power. In a grounded grid amplifier with cathode bias, the anode voltage is NOT the supply voltage nor is the cathode current the anode current. The anode voltage is the supply voltage minus the measured cathode voltage at a given cathode current. Likewise the anode current is the cathode current minus the grid current in the r-f drive case. It is advisable therefore to measure anode current and voltage directly at the tube terminals with separate meters and with proper precautions for safety. The anode voltage is measured with a voltmeter connected directly from anode-to-cathode and the anode current should be measured with the meter directly in the anode supply line. Only from these direct readings should total anode input d-c power be computed.

One further comment should be made regarding absolute accuracy. Some of the r-f power generated in the anode circuit of the amplifier may be dissipated in the anode circuit, cavity walls, joints, etc., and will not appear as heat at the anodes of the tubes. This will result in slightly optimistic determination of useful output power. Note that the true output power is determined by this method but it can not all be recovered at the output port. In general; the amplifier with poor efficiency will produce the greater error of this type.

As an example of the magnitude of this type error, data taken on an amplifier at 1296 mc/s with poor total efficiency indicated an output power of 29 watts for 100 watts d-c input power while the true output power measured directly at the output port was 25 watts. The 14% (0.7 db) error can in part be accounted for by the d-c meter inaccuracies but is believed to be largely due to circuit losses in the anode cavity. The saturated exhaust temperature for this particular case was 124°F.

If other devices are used to measure temperature, such as a thermistor, it is advisable to determine the effects of stray r-f fields on the operation of such devices. Indeed, a mercury column thermometer may be effected if its column length is near half wave resonance at 1296 mc/s. It may be advisable to use alcohol thermometers in cases where stray r-f leakage is occurring.

The accuracy of your d-c meters to measure both anode current and voltage should be ascertained if absolute output power determination is desired. It is interesting to note however, that even with inaccurate meters, the plate efficiency may be determined with high accuracy.

While power determination at 1296 mc/s has been stressed in this report, it should be obvious that the heat equivalence method can be applied at any frequency.

Contributions by W2CQH and WA2VTR are acknowledged.

TECHNICAL REPORT # 8

Date: October 1971

From: Crawford Hill VHF Club, W2NFA

Subject: A Low Noise Preamplifier for 1296 MHz

A low noise transistor preamplifier for 1296 MHz is described in this report employing the Nippon Electric Company microwave transistor, V766A (NPN silicon bipolar transistor), in a grounded emitter circuit. A measured noise figure of 2.7 db (250°K) with 10 db gain has been achieved at 1296 MHz.

Fabrication is on glass-epoxy printed circuit board, $\epsilon = 5$, of the inexpensive 1/16 inch thick double clad type. Since the circuit impedances do not exceed several hundred ohms, and the electrical circuitry is short in terms of a wavelength, the dielectric quality of the board is of little concern. The basic structure is a 50 ohm strip line with the transistor inserted near the input connector and d-c blocking capacitors placed at the junction between coaxial connector and strip line.

No input tuning is provided since best noise performance is obtained with a 50 ohm source impedance. The actual input impedance of this transistor is about 14 + j3 ohms (series equivalent). This means that the input is mismatched for best noise performance, and that the antenna and feedline must themselves be reasonably well impedance matched around the operating frequency for optimum noise performance. The inclusion of a filter in the input line may not realize the expected filter response because the transistor input is not impedance matched.

The output impedance of this transistor is about 150 - j100 ohms (series equivalent) and is tuned and matched to 50 ohms by means of a sliding capacitor tuner. In the finished design, the capacitor is a tab of copper foil whose area (capacitance) and position along the output line are experimentally adjusted for maximum gain at 1296 MHz. No other tuning is required. Under conditions of 50 ohm input and output, the amplifier gain will be about 10 to 11 db at midband and exhibit a -3db bandwidth of several hundred MHz.

Biasing

An important feature of this circuit is its stability (freedom from spurious oscillations) which is the function of the elaborate bias network shown on the schematic. This network consists of a 56 ohm 1/4 watt resistor in series with a resonant choke for midband. The choke, L, consists of 6 turns of #30 wire air wound on a 1/16 inch diameter form with turn spacing of about one wire diameter. The overall length of the winding is about 1/8 inch.

The small resistor is first soldered to the strip line (straight upwards away from the strip) with as short leads as possible. The top of the resistor is now a terminal point for one end of L, the other end of L is connected to a good UHF 1000 pf bypass capacitor (low inductance). These three components provide a good high impedance chode at midband, 1296 MHz, and a relatively lossy impedance out of band.

At lower frequencies in the VHF region, the inductance is small and the 56 ohm resistor heavily loads the circuit preventing resonances of large enough Q to start oscillations. At still lower frequencies an additional RC network is provided which essentially loads the transistor with a total of 112 ohms well into the HF region. Below the HF region the 100 pf blocking capacitors roll off the external circuit response and the 5 uf capacitor prevents internal feedback. The combined action of these components in the bias network discourage external and internal circuit resonances which may cause self oscillations.

In addition, the bias network provides collector current stabilization by means of negative feedback through the 560 ohm collector dropping resistor. A 15 volt zener regulator must be used to back up the whole network since the V766A transistor has a V_{ce} maximum rating of 15 volts. The above described network is highly recommended for all UHF transistor amplifiers.

The operating conditions for minimum NF in this circuit were 5 to 6 ma. of collector current at about 12 volts. The 25 K ohm bias adjust resistor may be experimentally set for best NF.

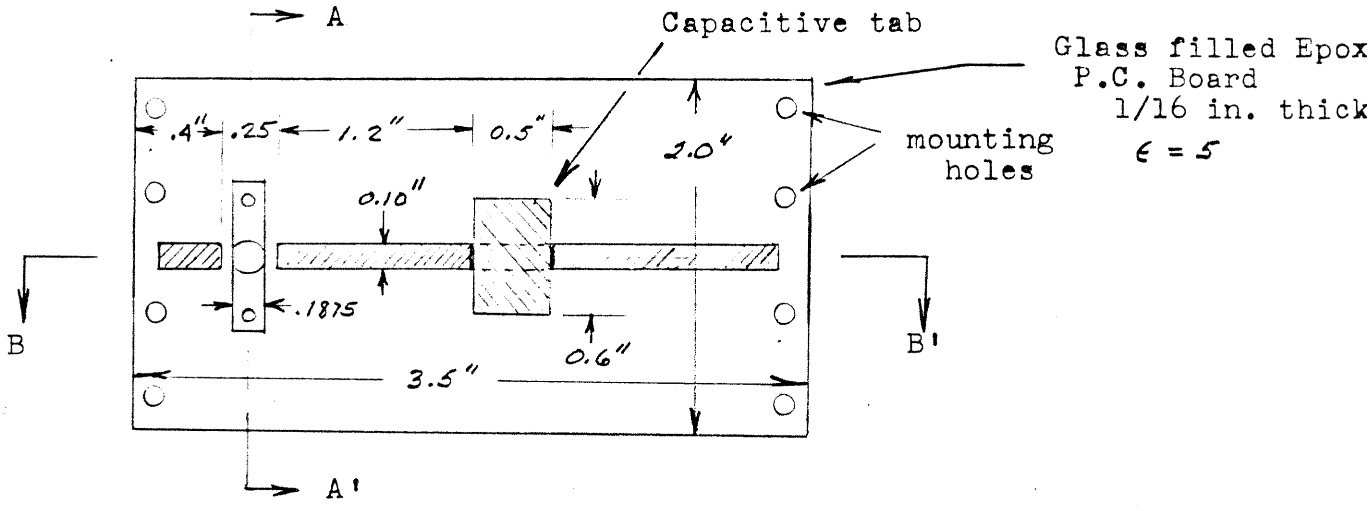
The transistor comes in a radial flat lead package with two emitter leads. The base lead is slant cut at its end for easy identification. The base and collector leads may be soldered to the strip line with a minimum of heat, low temperature iron. All leads may be shortened for convenience in mounting. The emitter clamp block should be carefully cleaned and free from filings so that the emitter can be grounded electrically secure with minimum inductance.

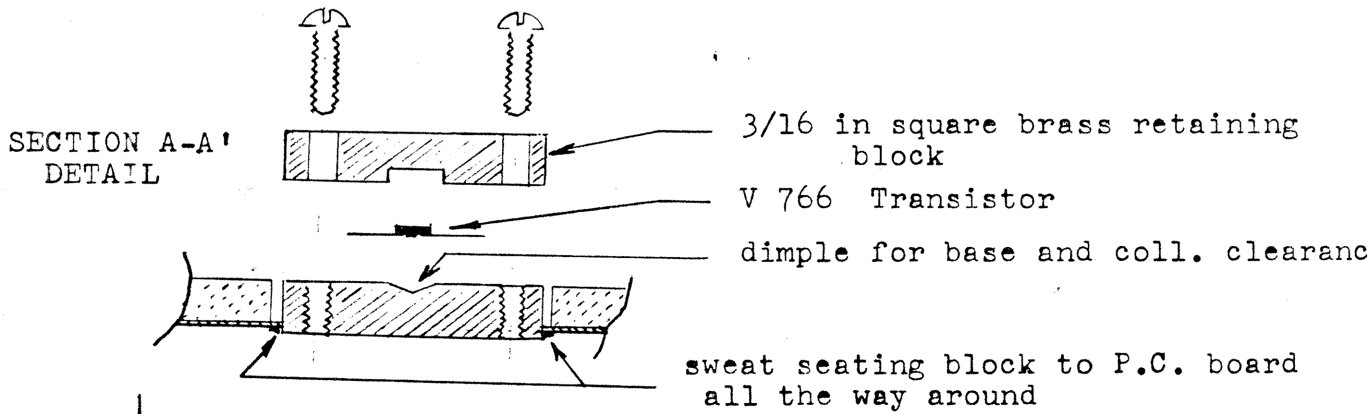
Because of the small physical size of this preamplifier, it may be conveniently weatherproofed and mounted at the feed of an antenna in order to minimize transmission line losses on receiving.

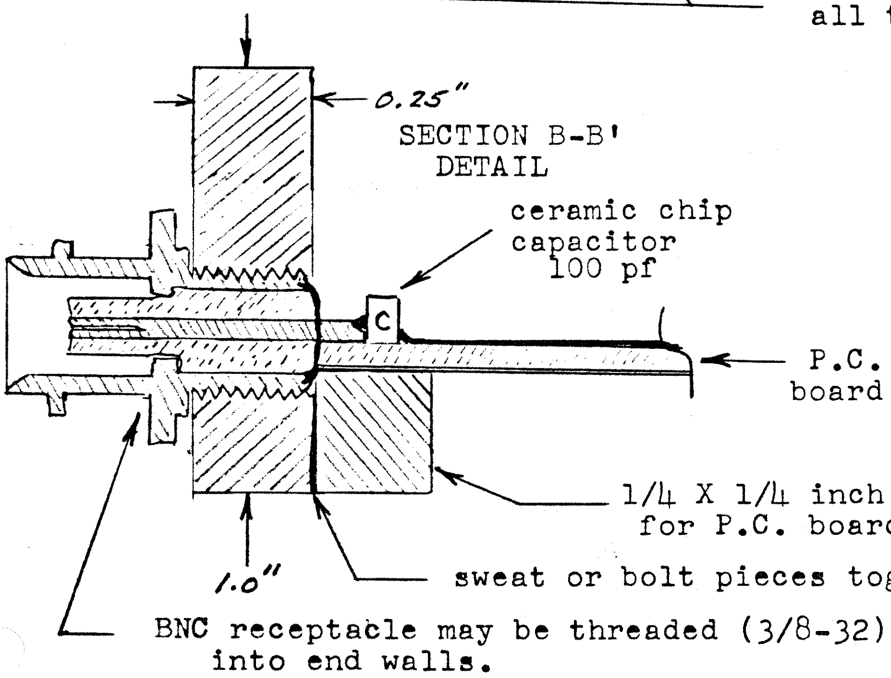
At this time, 1971, the V766A transistor may be obtained in small quantity at $26.80 each from CAlifornia Eastern Laboratories Inc., 1540 Gilbreth Road, Burlingame, California, 94010, or on the East Coast of the U.S.A. from Marv Groll, 87 Terrace Hall Ave., Burlington, Mass. 01803.

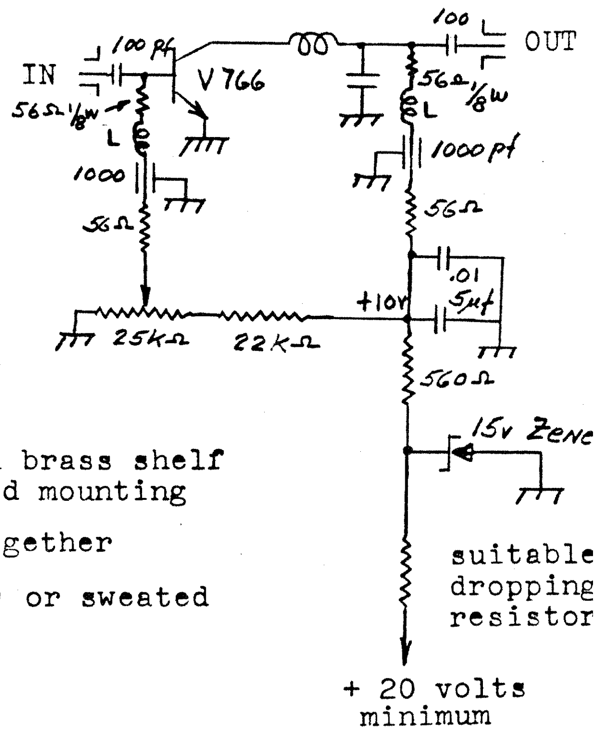
While the price of this transistor may seem severe, it must be pointed out that at this time, 1971, there are few if any other transistors capable of this measured NF at 1296 MHz. The only competitive preamplifier would be a mediocre parametric amplifier, which at best is far more difficult to construct and tune not to mention the need for a pump source and circulator.

A LOW NOISE PRE AMPLIFIER FOR 1296 MC/S USING THE NEC V766 TRANSISTOR





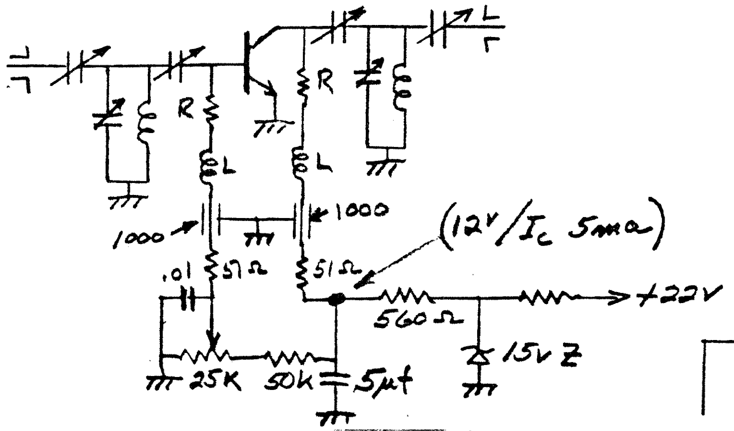


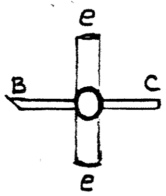


Dec, 1971
K H Turrin

1296 mc/s Narrow Band Pre Amp

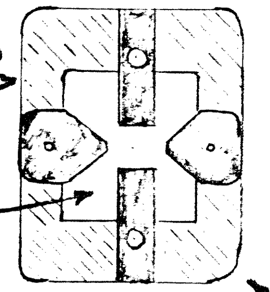
NF 3db GAIN 10 ↔ 12 DB B.W. 30 mc/s





V766A

P.C. BOARD ETCHED



CUT-OUT

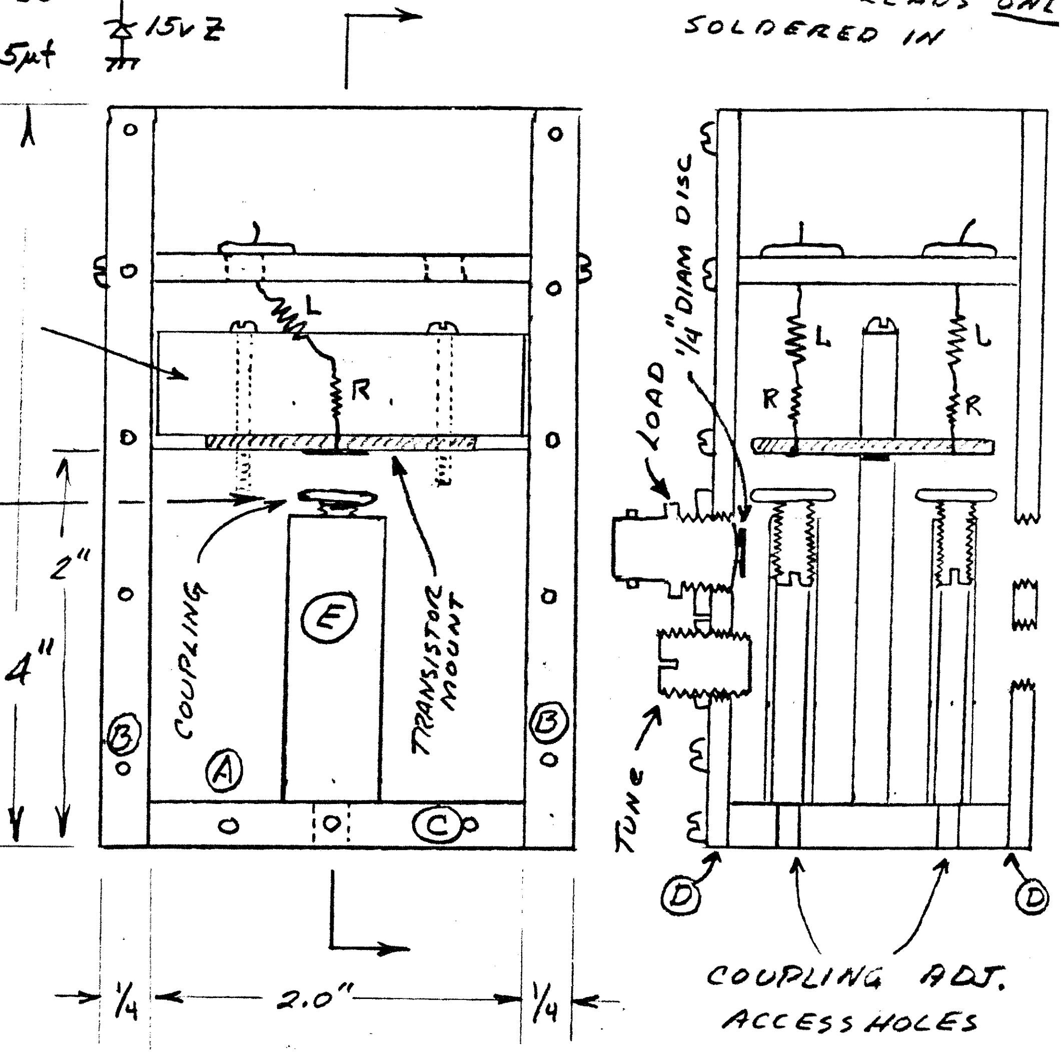
TRANSISTOR MOUNT

COLL. & BASE LEADS ONLY SOLDERED IN

FULL SCALE DRAWING

MOUNT CLAMP BLOCK

12-24 × 1/2 BRASS SCREW WITH HEAD FILED FLAT



R - 56Ω 1/8 w

L - 6t #30 tinned wire 1/16" I.D. 1/8" Long

(RESONANT CHOKE)

COUPLING ADJ. ACCESS HOLES

PARTS (A), (B), (C) & (E) ALL SWEATED TOGETHER

COVER PLATES (D) BOLTED ON

STOCK BRASS BARS

(A) 3/16 × 2

(B) 1/4 × 1 1/2

(C) 1/4 × 1 1/2

(D) 1/8 × 2 1/2

(E) 1/2 × 1/4 O.D. SILVER K-BAND WAVEGUIDE 1 1/2" LONG TAPPED 12-24 ON INSIDE of one end for coupling screw

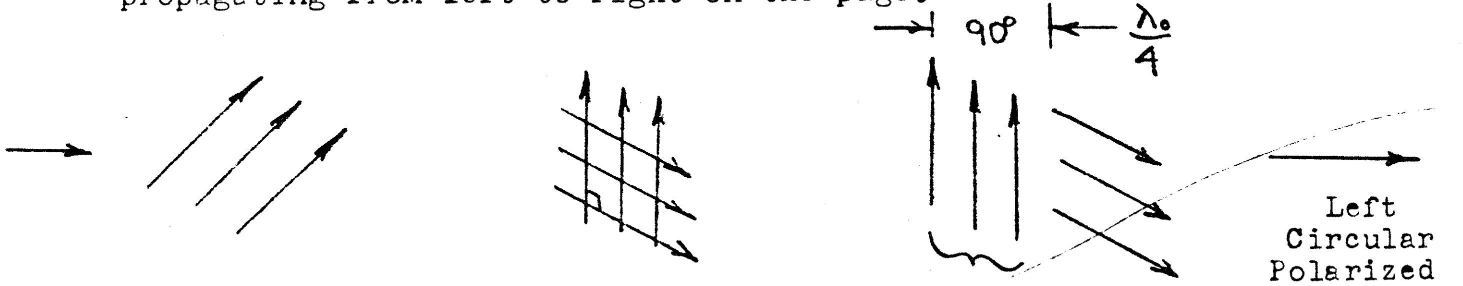
Technical Report #9

From: The Crawford Hill VHF Club, W2NFA

Date: December 1971

Subject: <u>A CIRCULARLY POLARIZED FEED ANTENNA FOR 1296 m c/s.</u>

To convert a linearly polarized radio wave into a circularly polarized wave the process is to decompose the linear wave into two components which are equal in magnitude and phase and are at right angles to each other. Then cause one of the components to be advanced or delayed in phase by 90 degrees with respect to the other. This process is illustrated vectorially step-by-step below. The wave is propagating from left to right on the page.

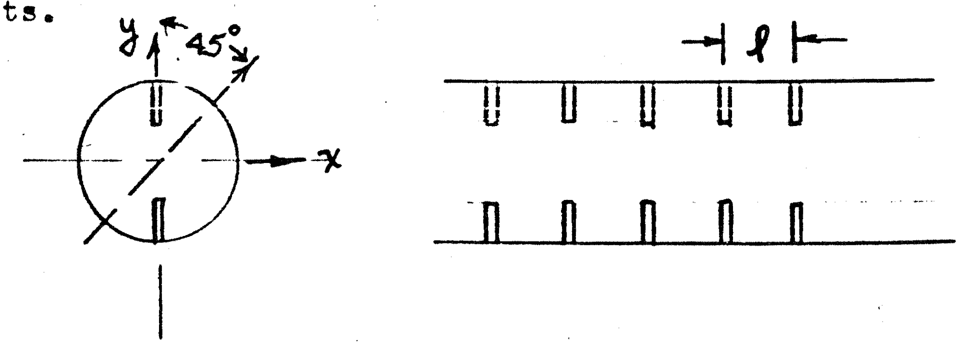


In Technical Report #2, it was suggested that a quadrature hybrid be employed to process the radio signal into the proper phase and amplitude components and then radiate the two components through linearly polarized orthogonal (at right angles to each other) antennas. In this report another method or processing the wave is described which can be applied to the high efficiency feed antenna described in Technical Report #5. The processing is done directly in the circular waveguide and no external hybrid is required. Both left and right circular polarization are available through the original orthogonal coaxial ports on the feed.

A reminder that the sense of circularly polarized wave is reversed upon reflection. Therefore, with a reflecting type antenna the sense of the feed polarization must be reversed from the desired radiated polarization. Using the convention suggested in Technical Report #1, transmit right circular and receive left circular, the feed antenna ports must be interchanged.

The Method

The method chosen here to obtain the phase difference between linear components is by means of a periodically loaded circular waveguide, loaded in one plane by means of an array of diametrically symmetric posts.



By introducing a linearly polarized propagating wave in the guide at an angle of 45 degrees with respect to the plane of the posts, this wave IS decomposed into two equal amplitude in phase components linearly polarized in the x and y direction. This constitutes the basic power splitting action.

Now the two waves must be caused to differ in electrical phase by 90 degrees in order for the original linear wave to be transformed into a circularly polarized wave. The phase shift is brought about by the post loaded guide which slows the y polarized component down while the x polarized component is uneffected by the post loading. The reason that the x polarized component does not interact with the posts is that the electric field of the dominant waveguide mode is everywhere at right angles to the posts and so cannot couple. For the y polarized component the posts essentially represent small periodic capacitive loading. (Delay line)

By suitably positioning a number of posts along the guide and adjusting their penetration into the guide, the required 90 degrees of difference phase between the x and y components can be achieved as the wave proceeds down the loaded section of guide. No further phase difference occurs after the wave leaves the loaded section, and since the guide is circularly symmetric the now circularly polarized wave will continue undisturbed.

Polarizer Design

The design of this type of polarizer is uncritical and easy to adjust when certain guidelines are followed. The number of pairs of diametrically opposite posts should be an odd number, 3 pairs being the minimum number. A larger number of posts will result in a physically longer polarizer but less critical and wider in bandwidth. The choice of an odd number of post pairs results in an even number of sections in the polarizer which gives the minimum VSWR.

The design procedure will be illustrated here for the case shown by Figure 1. Five pairs of posts were selected resulting in a four section polarizer. The phase shift per section in this case is therefore $90^{o}/4 = 22.5$ degrees.

The differential phase shift between x and y polarized waves in the waveguide for a single section of length ℓ is given by:

$$(1) \qquad \Delta\theta = \beta'\ell - \beta\ell$$

where β' is the y polarized wave propagation factor in the guide and β is the x polarized wave propagation factor in the guide.

And the relationship between β' and β in the loaded guide is given in terms of the capacitive susceptance loading (normalized to the guide admittance Y_o) by:

$$(2) \qquad \beta'\ell = \cos^{-1}\left(\cos\beta\ell - \frac{B}{Y_o}\sin\beta\ell\right)$$

$$\beta = \frac{2\pi}{\lambda_g}$$

In general values of B/Y_o, the normalized capacitive susceptance, below 1.0 are desirable with values below 0.5 preferred.

Equations (1) and (2) have been reduced to a family of curves in Figure 2 from which the phase distance, βl, between pairs of posts may be obtained for a practical range of $\Delta\theta$ and B/Y_o. Having determined βl from the equations or Figure 2 for a selected $\Delta\theta$ phase shift per section, the value of l , the physical distance between pairs of posts may be evaluated by means of the equations at the bottom of Figure 2.

For the design shown by Figure 1, with $\Delta\theta$ =22.5° per section, from Figure 2, βl = 45°, for B/Y_o = 0.45. With the 6.5 inch diameter circular waveguide the guide wavelength, λg at 1296 mc/s is 16 inches and the separation between pairs of posts is 2.0 inches.

Although the value of B/Y_o is now determined its actual value is of academic concern since it is quite satisfactory to simply adjust the penetration gradually inward (all posts the same penetration) and sample the emerging wave for circularity. (A method of doing this is suggested at the bottom of Figure 1.) The penetration depth will quickly converge to the proper value as the emerging wave goes from linear polarization through various degrees of elliptic polarization to near circular polarization. It is sufficient to stop adjusting when the max. to min. ratio of the elliptic polarized wave is less than 1 db, as this already means that the cross polarized wave (opposite sense polarized wave) is some 25 db below the level of the desired sense polarization.

At this point in the tuning procedure it is best to check the impedance match of the original linearly polarized launchers. In general this should be found to be satisfactory since the conditions of the design equations have been met empirically. If the match is not good enough (20 db return loss or 1.2 VSWR) it may be improved by trimming the end screws of the polarizer, symmetrically.

A further extension of the design equations permits an exact impedance matched polarizer to be designed. The polarizer loaded waveguide sections now become resonant and exhibit a fixed phase shift per section at one frequency of $\beta' l$ = 90°. The equations which specify this design are

$$\Delta\theta_{\text{per section}} = 180° - 2\beta l$$

and $$\cot \beta l = \frac{1}{2}\frac{B}{Y_o}$$

These design equations are from an article by A.J.Simmons, in the IRE Transactions on Microwave Theory and Techniques, December 1955.

Construction Notes

This polarizer has been applied to the dual mode feed horn described in Technical Report #5 but may be applied to any feed in circular waveguide geometry. With suitable modifications in the design equations for the guide cut-off wavelength, this method may also be applied to square waveguide feeds or antennas. The only precaution is to allow about a quarter of a guide wavelength or more space between the polarizer and any other obstacles in the guide.

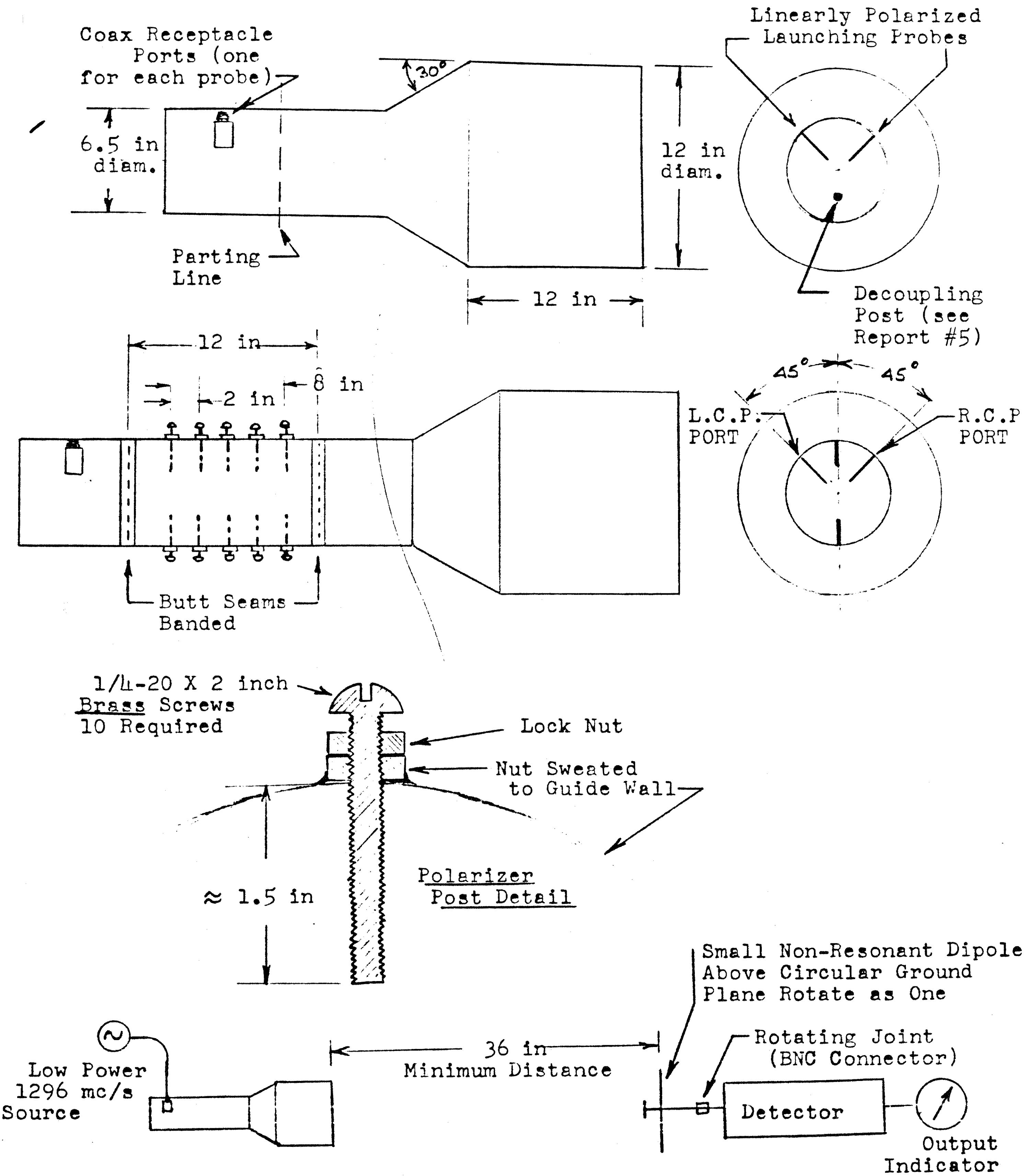
The modifications to the 1296 mc/s dual mode feed are shown by Figure 1 and consist of parting the small diameter section of circular waveguide about midway and splicing in the polarizer section. The choice of a 12 inch long polarizer section is purely from convenience of available sheet brass size. The splices are butt joined with external bands and completely soldered around the joint.

The posts in this design consist of 2 inch long 1/4 - 20 brass screws. A brass nut is first sweated to the sheet brass guide section at the appropriate locations determined from the design. The brass nut serves as a guide for drilling and tapping the sheet brass wall after the nut is soldered in place. A second brass nut is used as a lock to secure the screw adjustment electrically and mechanically.

Measurements of the completed feed as per Figure 1 showed the return loss at each port to be better than 25 db and the ellipticity of the circularly polarized wave for either sense to be about 1 db, with the post penetration as shown. Both the return loss and polarization ellipticity are preserved below 1296 mc/s but depart rapidly above. No attempt was made to optimize the bandwidth as operation over a narrow band ± 100 kc around 1296 mc/s is all that is contemplated.

The cross talk between coaxial ports on the feed was measured at -25 db with the "nulling" post unchanged from the original design.

Figure 1. Modified 1296 mc/s Dual Mode Feed for Circular Polarization.



Rotate sampling dipole slowly and determine ratio of maximum to minimum signal detected level. Be sure that the output indicator can display a 1 db change in level.

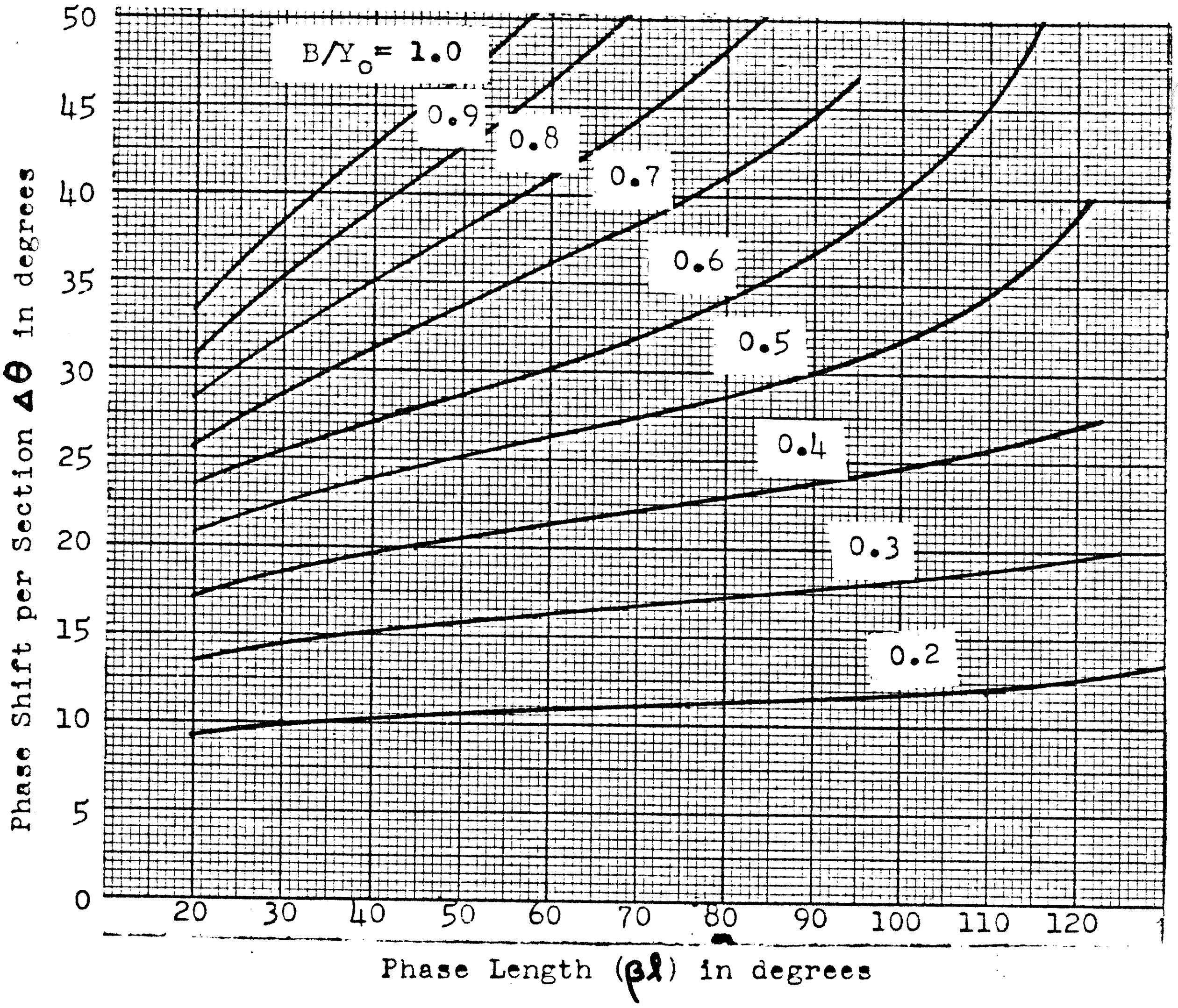


Figure 2. Phase shift per section, $\Delta\theta$ v.s. post separation, βl in degrees for specific values of B/Y_o.

The length, l can be determined from:

$$l = \frac{(\beta l)^\circ}{360^\circ}\lambda_g$$ where the guide wavelength

$$\lambda_g = \frac{\lambda_o}{\sqrt{1-\left(\frac{\lambda_o}{\lambda_{co}}\right)^2}}$$

and the cut-off wavelength

$$\lambda_{co} = \frac{\pi d}{1.841}$$ for the TE_{11} mode

d = diameter of the circular waveguide
λ_o = 23.148cm or 9.113 inches at 1296 mc/s.
π = 3.1416

Technical Report # 10

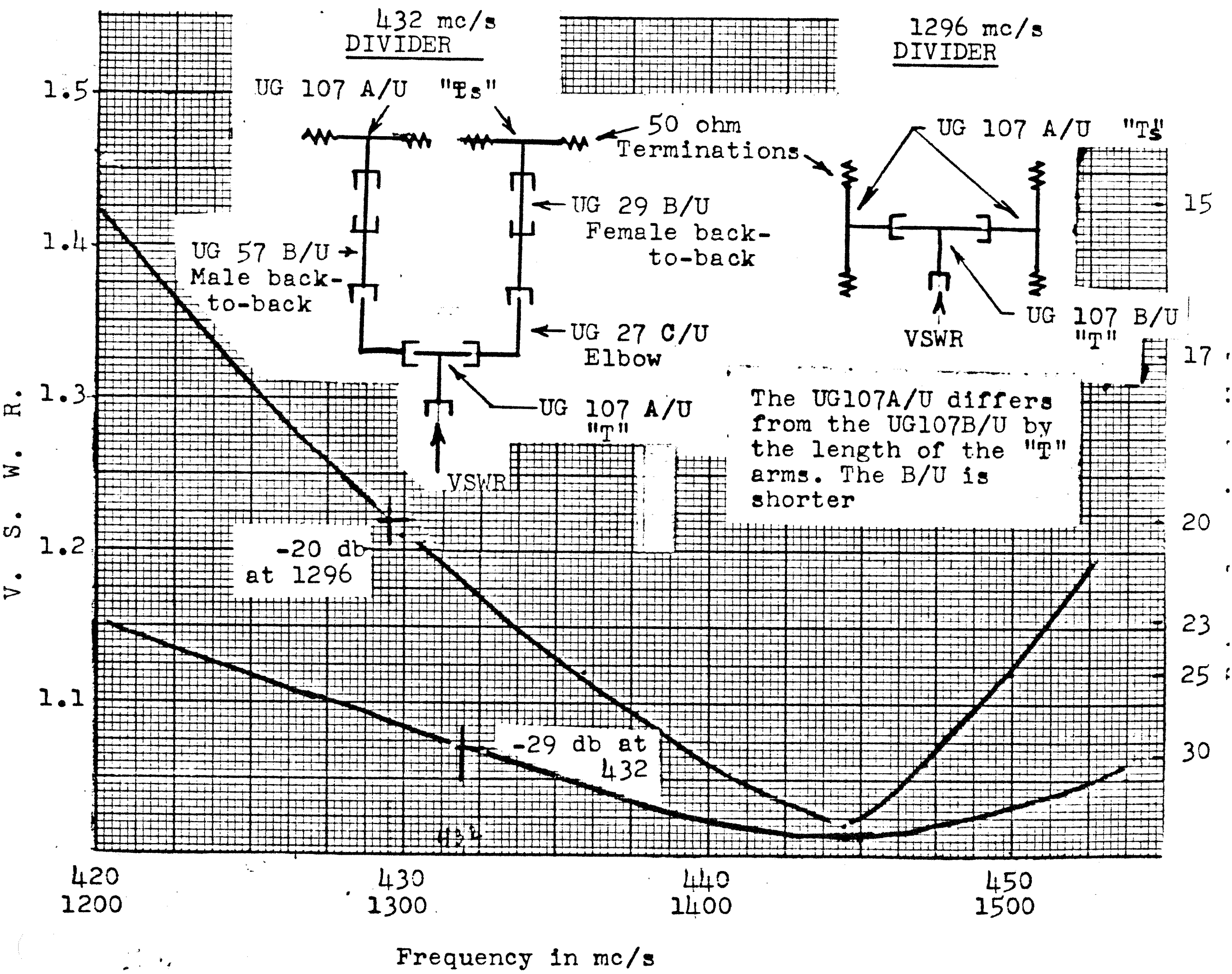
From: The Crawford Hill VHF Club   Date: December, 1971

Subject: UHF POWER DIVIDERS

Presented in this report are two 50 ohm 4-way power dividers of the reactive type composed entirely of standard type-N fittings. One of the dividers is for 432 mc/s and the other for 1296 mc/s. These power dividers are particularly useful for feeding 4 identical antennas in an in-phase array.

Below are shown the design of each divider and the required type-N fittings for each. The curves are the measured VSWR at the common port for each divider. Note that each divider has a natural best match frequency which is slightly higher than the desired operating frequency.

The 432 mc/s divider was developed by W2CCY and the 1296 mc/s divider by W2CQH.



TECHNICAL REPORT # 11

From: The Crawford Hill VHF Club
Date: Aug 1972 (revised Sept 1986)

Subject: Use of Solar Noise in EME System Evaluation

The purpose of this lengthy report is to show how Sun noise measurements may be used to obtain good estimates of the following EME station system parameters:

1. Receiving System Performance Factor, Geff/Tsys.
2. Antenna directivity from which effective gain is estimated.
3. System operating temperature from which the receiver input noise figure (temperature) can be estimated.

The Sun noise measurement methods require no laboratory equipment other than a baseband a-c volt, current or power meter, but do require patience, care and understanding of measurement procedures.
Interpretation and use of measured results must also be with care and understanding of limitations and accuracy.

1. A Receiving System Performance Factor, Geff/Tsys

The limit of weak signal reception in EME transmission is determined by the effective gain of the antenna, Geff, and the system operating temperature, Tsys, as well as detection methods, signal modulation, Moon reflection characteristics and the EME path. However, all other factors being equal, Geff and Tsys will be the limiting factors.
A measure of practical single side band (no image response) receiving system performance is the ratio of effective antenna gain to system operating temperature, Geff/Tsys.
A very simple and readily available method for measuring Geff/Tsys is by means of Sun noise. It is the purpose of this report to convince the EME enthusiast that Geff/Tsys is of more practical significance than knowledge of Geff and Tsys separately, and to show how it may be measured and used in system evaluation.
While the engineering purist would like to know each system parameter, Geff, Tsys, feedline losses, noise figures, etc., all of these factors are lumped into the Sun noise measurement of Geff/Tsys. Any improvement in receiving system performance is related directly to this ratio regardless of whether it is the result of improved preamplifier NF, higher antenna gain, lower feedline losses, corrected impedance match, etc. Any adjustment or system change that results in a greater Sun noise ratio will improve the receiving system performance. It is therefore evident that Sun noise measurements may be used for on-line adjustment of the receiving system. Indeed the most critical

adjustment may be the interface between low noise preamplifier and the antenna. The impedance interactions at this interface may be of sufficient magnitude as to demand that adjustments be made as a complete system for peak performance.
The complete receiving system may therefore be optimized without the need for laboratory equipment and with the antenna in a physical attitude of practical significance for EME communication. Note the word optimized. It is not readily possible to determine which component is faulty in the system by Sun noise ratio measurement alone. However, when an experimenter is faced with inadequate measuring tools, or none at all, Solar noise measurements become a valuable tool. Used with understanding, care and patience, the EME receiving system can be optimized to its full potential. Furthermore, Solar noise measurement can always be used as a receiving system performance check to determine if anything has changed or deteriorated.

Theory

In this section derivation of the mathematical relations between Sun noise and system parameters at UHF will be presented.
The Sun at 1300 mc/s and above appears as a disc of nearly uniform brightness equal in size to its optical disc which subtends an angle of approximately 1/2 degree viewed from the Earth. It is therefore not a true point source radiator, but may be considered as one for most practical size amateur EME antennas at 1296 mc/s, i.e., antennas with maximum aperture dimension less than 15 meters (20 wavelengths).
At lower frequencies, 144 mc/s for example, the Sun appears as a slightly larger source extending about 1.5 diameters beyond its optical disc size and with a temperature distribution that peaks towards the center of the disc.
Radiation from the Sun is random in amplitude, phase and polarization and is smoothly distributed over any practical bandwidth. Random polarization means that receiver output noise will be the same in average amplitude regardless of the orientation of a linearly polarized receiving antenna or sense of circularly polarized receiving antenna aimed at the Sun.

Figure 1, presents the distribution of Solar noise radiation over a wide portion of the radio spectrum. The radiation level is presented as a total flux density arriving at the surface of the Earth, in watts per square meter of area, per cycles of bandwidth; and, is for a quiet Sun. The flux level should always be higher than the curve, but during low Sun spot activity periods (1985-1986) levels close to the curve will be observed. Fluctuations can occur however due to Solar flares.

In the United States, a Solar flux density data service is available through the National Bureau of Standards World Data Center. Telephone 1 - 303 - 497 - 6223 and ask for Solar flux in Solar Units for the date of interest. A Solar Flux Unit converts to Solar Flux density by multiplying it by 10E-22. Data is available for at least several years from the current.

date. Ask for data for the frequency range of interest and one day before and after the date of interest to determine relative stability. Also ask for any notes regarding flares or Solar flux disturbances on the date of interest. The data will be given for a specific measurement frequency, 2800 mc/s or 1440 mc/s for example, and should be noted. This data point is then plotted on Figure 1, and an equal displacement above (or below) the curve given to the flux level at the EME band of interest. This will give a good value for the Solar flux density on the date and frequency of your measurement.

An antenna with a physical area (or effective capture area), A, in square meters and a single polarization, be it linear or circular, will intercept Solar noise power of

$$P\ sun = \frac{\Psi}{2} A \quad \text{watts/cycle of bandwidth.}$$

Where Ψ is the Solar flux density extrapolated from NBS measured value, using Figure 1 for the desired frequency. The factor of 1/2 is necessary because only half the total flux is available for a single polarization.
The actual power delivered to the antenna output terminals is always less than the power collected by the physical aperture by an antenna efficiency factor η. The antenna efficiency factor is intimitely related to the type of antenna, its design and construction. In general it is composed of a small amount of heat losses but mainly by the way in which the antenna output port is coupled to the collecting aperture. A paraboloidal antenna, for example, has an efficiency which is largely due to the way in which the feed illuminates the reflector to achieve optimum gain. (See Report # 5 for calculation of this factor.)
The useful Sun noise power available at the antenna terminals (port) may be written as

$$P\ sun\text{-}ant = \frac{\Psi}{2} A \eta \ .$$

In terms of effective antenna gain

$$G\ eff = \frac{4 \pi A \eta}{\lambda^2},$$

(Antenna Gain in this Report and in all other Reports in this collection is given with reference to an isotropic (Gain = 1) antenna.)

$$P\ sun\text{-}ant = \frac{\Psi}{2} \frac{G\ eff\ \lambda^2}{4 \pi} .$$

Since noise power is related to temperature through Boltzmann's constant (in the radio spectrum), k = 1.38 x 10E-23 watts/cycle-deg K (the computer notation of 10E-23 is used here for printing clarity, meaning 10 with an exponent of -23),

$$k\,T_{sun\text{-}ant} = P_{sun\text{-}ant} \,,$$

or

$$T_{sun\text{-}ant} = \frac{\psi}{2}\;\frac{G_{eff}\,\lambda^2}{4\pi k}\,. \qquad (1)$$

The above equation gives the electrical temperature of the antenna at its output port due to radiation from <u>only</u> the Sun (antenna pointed at the Sun). This expression is correct for antenna beamwidths greater than about 1 degree. See the section on Gain by Beamwidth Measurement, in this report.
It is important that Sun measurements are not made at low elevation pointing angles because of ground and atmospheric effects on the noise level. It is recommended that an elevation angle greater than 20 degrees be employed.
When the antenna is pointed away from the Sun (several beamwidths and at a high elevation angle away from the Galactic plane) the noise power at the antenna port will drop to some low value which is principally due to far sidelobes (rear lobes or leakage through a reflector) directed toward the warm Earth, and also I R losses in the antenna itself and background Universe noise. See Report # 5 for evaluation of antenna noise for a paraboliodal reflector antenna system. The Universe background temperature is generally accepted as 3.5 degrees K in modern times.
Finally, when a receiver (preamplifier, etc.) is connected to the antenna output port, an additional noise temperature appears there due to the overall receiving system noise figure (temperature). Since all of these noise powers are uncorrelated, they may be summed arithmetically, referenced at the antenna port for convenience.

The sum of receiver noise plus all other antenna noise when the antenna is pointed away from the Sun is called system noise and is characterized by a system temperature, Tsys, also referred to as the system operating temperature. (See Report # 3)
When the antenna with receiver connected is pointed at the Sun, the additional Solar noise produces a total noise temperature at the antenna output port of Tsun-ant + Tsys. The ratio of noise temperatures "ON" and "OFF" the Sun will be

$$R = \frac{T_{sun\text{-}ant} + T_{sys}}{T_{sys}}\,. \qquad (2)$$

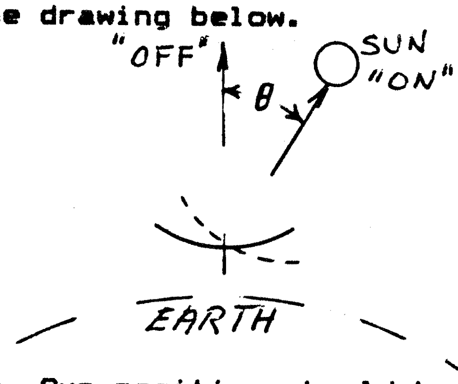
The ratio R may be measured directly at the receiving system output. This may be done with a suitable baseband power meter or voltmeter. Most VOMs have an a-c voltage range low enough to measure receiver output noise directly and added post detection filtering can be used to minimize the random fluctuations for more accurate readings.
Substituting equation (1) into equation (2) and solving for the ratio Geff/Tsys gives

$$\frac{G_{eff}}{T_{sys}} = \left(\frac{4\pi}{\frac{\Psi}{2k}\lambda^2}\right)(R-1) \tag{3}$$

The first term is just Tsun-ant and is a constant for a given wavelength and Solar flux, while the second term is the measured output noise ratio "ON" to "OFF" the Sun expressed in real number form, not decibels, minus one. Thus the accuracy of Geff/Tsys has been determined within the accuracy of the Solar flux, known at the time of measurement.

<u>Sun Noise Measurement of Geff/Tsys</u>

The measurement procedure should be obvious from the theory discussion. However. some details should be elaborated on with the aid of the drawing below.



1. The Sun position should be high in elevation to avoid ground noise and excess atmospheric noise.
2. The angle, θ , of pointing "OFF" the Sun should be several beamwidths or until no further decrease in receiver output noise is noted.
3. Point the antenna beam accurately at the Sun by maximizing the receiver output noise. Remember that the Earth is rotating so that the position of the Sun is constantly changing at a rate of approximately 15 degrees per hour. In some cases it may be advantageous to 'lead' the position of the Sun, and allow the Sun to drift through the peak of the beam. Take several observations to be sure that the beam peak is traversed.
4. A damped noise meter will be more accurate but requires more care in antenna pointing.
5. The output noise meter must be calibrated in voltage,

current, power or directly in decibels. The relations between these quantities are

$$\text{Ratio(R) in decibels} = 20 \log \frac{V1}{V2} = 20 \log \frac{I1}{I2} = 10 \log \frac{P1}{P2}$$

The use of a calculator is desirable when converting decibels to real number ratios. Those unfamiliar with the use of the decibel scale are urged to consult the ARRL Handbook or other similar references. The ratio may also be obtained directly by simply dividing the measured output levels.

6. Operate the receiver with the AVG (AGC) "OFF", BFO (CW mode) "ON", and just sufficient gain to obtain a readable output level when pointed "OFF" the Sun. Front end noise MUST predominate the system.

7. A receiving system non-linearity (compression) check should be made if possible. However, in view of the limited Sun noise ratio expected (< -20 db) with reasonable size amateur antennas, compression should not ordinarily be a problem. Poor receiving system design by providing more gain in amplifier stages than necessary to just swamp succeeding stage noise can result in compression. Good linearity is represented by no compression at about 10 db above the average maximum expected noise level to be measured.

8. Digital type readout voltmeters may be used for noise measurements provided sufficient post detection filtering is employed to obtain at least two digit readout without jitter. A large non-polarized capacitor at the receiver output should suffice for filtering.

9. Take several measurements for confidence. If wide variations in measured values of R are obtained check for system instabilities, poor measurment procedures and also possible Solar flare activity. Take measurements over a period of several days and check with NBS Solar flux service. Discard measurements if Solar flux is unstable during the period you measured.

10. Check by listening to the receiver output to determine if other sources of interference are present to cause erroneaous readings. Interference from 'man made' sources such as automobile ignition, electric motors, computers, etc., and even signals (QRM) in the wide receiver pass band can produce very erroneous results.

Example and Results

To bring into focus the preceding discussion, a useful example will be presented here.

Assume that you have an antenna of the type described in Technical Report # 5 (Geff = 35 db) and a receiving system temperature of 300 deg. K. Then assume that a measurement of Sun

Using equation (3) and appropriate constant values for 1296 mc/s gives

$$\frac{Geff}{Tsys} = \left(\frac{4 \times 3.1416}{\dfrac{80 \times 10E-22}{2 \times 1.38 \times 10E-23} \times (0.23)^2} \right) \times (15.9 - 1)$$

Carrying out the arithmetic gives

$$\frac{Geff}{Tsys} = (0.82) \times (14.9) = 12.2 \text{ or } (10.86 \text{ db}).$$

The expected value of Geff of 3162.2 (+35 dB) and (your) assumed system noise temperature of 300 deg. K, gives an expected value for your system

$$\frac{Geff}{Tsys} = \frac{3162.2}{300} = 10.54 \text{ or } (10.22 \text{ dB}).$$

Since the expected value is less than the measured value, this indicates that either the assumed value of Tsys is a bit high or that the expected effective gain assumed is a bit low, or both. See the next section on Gain estimate from beamwidth measurement for additional diagnostic information.

How does the value of Geff/Tsys relate to receiving Moonbounce signals?
To answer this question again assume another example of interest. Assume that W7XYZ is transmitting on 1296 mc/s with an antenna gain of 40 db (Geff = 10000) and power into the feed port of his antenna of 200 watts (Pt = +53 dbm). The signal-to-noise ratio at <u>your</u> receiver output will be

$$\frac{S}{N} = \frac{Pt\ Gt\ L}{k\ B} \times \frac{Gr}{Tsys},$$

where the average path loss, L, between isotropic (Geff = 1) antennas for the complete EME path is -271 db at 1296 mc/s. See Technical Report # 3 for tabulation of path loss for other EME bands. Gr/Tsys is your system performance ratio computed using equation (3) and your measured Sun noise ratio, R. (Gr = Geff of your system) k is Boltzmann's constant from page 2 (k in dbm = -198.6) and B is the effective noise bandwidth of the receiving system. Typically for the ear-brain system, B is about 100 cycles (B = +20 db). The signal-to-noise equation expressed in decibels can be written

$$\frac{S}{N}\text{ db} = Pt + Gt + L - k - B + (Geff/Tsys) \quad \text{all in decibels}$$

$$= +53 +40 -271 +198.6 -20 +10.86$$

$$= +10.46 \text{ db} \quad \text{a good EME signal !}$$

Under the above assumed conditions, any one of the factors, except B, or a combination of them may be reduced by about 7 db to reach a S/N ratio of +3 db, a barely perceptible signal to detect by ear.

Moon Noise

A curiosity about receiver system performance might be the question of at what level of the receiving system performance factor, Geff/Tsys, will Moon noise be just discernably measurable?

One must first know the level of Moon noise flux at Earth for the frequency of interest and then apply equation (3) with R set at about 1.15 (1/2 db of Moon noise). Moon flux is also shown by Figure 1 and is derived for an apparent Moon temperature of 210 degrees Kelvin and varies as the square of the frequency. At 1296 mc/s the Moon flux is 8.6 x 10E-24, and a Geff/Tsys ratio of 19.7 db would be required to observe 1/2 db of Moon noise.

Another curiosity about Moon noise is that as the receiving antenna effective gain increases, its beamwidth decreases. Since the Moon subtends an angle of about 1/2 degree, observed from Earth at 1296 mc/s and higher frequencies, when the antenna -3db beamwidth decreases to 1/2 degree the antenna temperature due to the Moon noise flux will be one half the apparent Moon temperature. This temperature is very nearly 210 degrees Kelvin at UHF. Further decreasing the antenna beamwidth will gradually increase the antenna temperature towards a limiting value of 210 degrees Kelvin.

Since the antenna must be aimed at the Moon for EME communication, the excess Moon noise will eventually limit the operating system temperature to a value of 210 degrees plus the receiving system temperature. Under these conditions (very unlikely for amateur EME antennas to be that large at UHF) the system performance cannot be improved by reducing the receiver noise figure much below 1 db.

2. Antenna Gain by Sun Noise Beamwidth Measurement

Introductory Remarks

In this section the effective antenna gain, Geff, will be estimated from measurement of the -3db beamwidth of the antenna by means of Sun noise. Geff is the single most important antenna parameter when transmitting. Paradoxically, optimum receiving system performance, Geff/Tsys, and maximum Geff are not generally arrived at by adjusting an antenna for maximum gain. (See Technical Report # 5 for a discussion of these factors with a parabolic reflector antenna). This effect becomes increasingly important as lower receiving system (lower noise preamplifiers) temperatures and lower antenna temperature are achieved, and lower Galactic noise bands are employed.

High gain antennas required for successful EME communication are difficult to measure by the usual slant range methods (along the ground) because of the length of unobstructed range required, ground reflections, the need for a signal source and antenna, and the need for a calibrated standard gain antenna for comparison gain measurement.
In general, direct antenna gain measurements are impossible for most backyard installations and Sun noise may be the only method available.
Several factors make this method attractive. The signal source (the Sun) is well beyond far-field requirements, for high elevation angles there can be virtually no problems with ground reflections or path obstructions, no fancy calibrated test equipment is required and finally, accuracy is limited by careful measurement. While the final estimate of effective gain may not be accurate by other standards, it does give diagnostic information which is valuable in determining where improvements should be made.

Theory

The effective gain of a high-gain pencil-beam large antenna of the type usually employed in EME communication can be considered to be the product of two basic factors, efficiency and directivity.

$$\text{Effective Gain (Geff)} = \eta_L D$$

The efficiency factor includes all power unavailable to the aperture. This can include heat losses associated with r-f current on lossy materials used in construction as well as spillover losses and leakage through the surface in reflector antennas.

Directivity is a measure of the way in which the available r-f energy is concentrated in a particular direction. The concentration of r-f energy into a small volume (the pencil beam) is directivity. In general for microwave antennas such as parabolic reflector antennas which are carefully constructed, the heat loss efficiency factor is close to unity, typically 0.9 or higher (not including spillover losses). The directivity factor is related to the antenna by the way in which r-f energy is launched, principally by the way the r-f energy is distributed over an area called the collecting area or aperture.
It is indeed possible to have an antenna with high directivity but low effective gain simply because the heat losses are excessive. It is also possible to have an antenna with a large collecting aperture area but low effective gain because of poor launching efficiency (excessive spillover).
Concentrating energy into a pencil beam implies a narrow radiation beam characteristic. The higher the directivity, the narrower the pencil beam.
A direct measure of directivity for a high gain antenna with a circular aperture and circularly symmetric pencil beam radiation is to determine the radiation characteristic -3db beamwidth(s), in degrees, which can then be related to directivity by the formula

$$\text{Directivity (D)} = \frac{33707}{(\text{BW})^2} \qquad (4)$$

Where BW is the measured beamwidth in degrees from the -3db level on one side of the beam to the -3db point on the opposite side of the beam. This formula can be extended to include radiation beams that are not circular but instead slightly elliptical, by measuring two beamwidths in the planes of maximum and minimum beamwidths and using their product in the denominator of equation (4). Measurement of two beamwidths in planes that are at right angles to each other can be a formidible task for a large EME antenna design.
In general, many factors including efficiency, favor large antennas which produce circularly symmetric pencil shaped beams.

Beamwidth Measurement Method
The remainder of this section will detail measurement methods and in particular corrections for systems with limited Sun noise amplitude range capability.

Before detailing the beamwidth measurement it is necessary to digress into the subject of measurement corrections with small signal-to-noise ratios, in this case the signal is Sun Noise. If the receiver system noise output is designated Nsys and the excess noise from the Sun as Nsun then the ratio of output noise

$$A = \frac{Nsun + Nsys}{Nsys} . \qquad (5)$$

The measurement process is exactly the same as in Section 1 of this Report, with all conditions of linearity, gain stability, etc., still invoked.
If the system noise were very small or the Sun noise very high then the S/N would be high and no correction would be needed. When the S/N ratio is not very high, as will be typical of many EME systems, then the system noise will partially mask the Sun noise making a correction necessary.
To correctly measure the -3db level of only the Sun noise in the presence of system noise, the correct measurement level may be derived by taking the ratio of both cases, maximum Sun noise and 1/2 maximum Sun noise in terms of the measurable quantity A.

$$\text{Change in output noise level for - 3db below Max} = \frac{\dfrac{Nsun/2 + Nsys}{Nsys}}{\dfrac{Nsun + Nsys}{Nsys}}$$

In terms of A this is identically,

$$\text{Change in output noise level for - 3db below Maximum} = \frac{1}{2}\left(\frac{A + 1}{A}\right). \qquad (6)$$

This equation is plotted in decibel form in Figure 3 for convenience. The measured change will always be less than the 'true' - 3db level because of the masking effect of system noise. In order to make use of this correction, one must first measure the Sun noise ratio A aimed directly "ON" and "OFF" the Sun and then move the antenna beam from "ON" the Sun (maximum Sun noise) until the corrected level is reached which corresponds with the 'true' - 3db beamwidth level.
A caution is in order here about measuring when the Sun position is in the Galactic plane area. Avoid beamwidth measurements under these circumstances, especially at the lower EME frequency bands. The severity of this problem is illustrated by the plot of Galactic noise v.s. frequency, shown by Figure 2.

To illustrate and clarify the measurement process, assume a measured value for A of 12 dB (15.9, from the previous example), then from the curve of Figure 3, a measured output level change from maximum to - 2.75 db corresponds to the half power point on the beam.

The first step in the measurement of the half-power beamwidth is to locate the Sun and point the antenna beam as accurately as possible by observing the noise output meter connected to the receiver. Remember that the Sun will appear to be moving from East to West, due to the rotation of the Earth, at a constant angular rate of very nearly 15 degrees per hour. An antenna with a beamwidth of say 3 degrees to the half power points would, if held stationary, pass from maximum noise level to the -3db point in 6 minutes.
Next measure the Sun noise ratio, A, by swinging the antenna beam far off the Sun (and out of the Galactic plane) to observe system noise only.
After computing A by direct division of the "ON" and "OFF" the Sun noise output levels, determine from Figure 3 or Equation (6) what the correct reading should be for the - 3db level. A decibel calibrated meter scale can be very helpful. Gain instability (drift) in the receiving system can be very troublesome in the course of these measurements and should be corrected before proceeding.
Finally, a good procedure to actually determine the angular displacement between the -3db points of the antenna beam when the angular readout system is not too accurate or suffers from backlash, is to use the Sun drift method. Point the antenna beam ahead of the expected position of the Sun and lock its position. Measure the time in minutes or fraction of minutes it takes the Sun to drift through the corrected -3db point on one side of the beam to the corrected -3db point on the other side. Multiply this by 15/60 to give the half-power beamwidth in degrees. Check carefully when the Sun drifts through the maximum of the beam that the level is the same as it was during the maximization of Sun noise when obtaining the Sun ratio A. If it does not reach the same peak, the Sun track and your antenna beam peak do not coincide or the receiver has changed gain level. Repeat this procedure until the levels match to within about 0.1 db. Take an average of the closest readings for confidence and use this value of half-power beamwidth to compute directivity from Equation (4).

This method of determining antenna directivity takes into account all factors of effective gain except unavailable power losses. For parabolic reflectors, only aperture distribution efficiency, which includes reflector surface irregularities, is included in directivity, but spillover and leakage through the reflector mesh as well as I R losses anywhere in the antenna are considered 'heat'loss because that is power unavailable to the aperture radiation. In order to estimate the effective gain of a parabolic reflector antenna it will be necessary to know more about the feed illumination characteristics (see Tech. Report #5).
Yagi arrays on the other hand, have heat losses mainly due to only the corporate distribution harness and the high r-f currents in the narrowband linear elements. In this case a careful assessment of feed harness losses is essential to estimating effective gain. The use of the lowest possible feed harness

losses and parasitic arrays of relatively wide bandwidth (low-Q) have a double benefit in both increasing effective gain and reducing system temperature by decreasing antenna noise due to losses.

Numerical Example of Gain Estimate

A typical example of the procedure and numerical values involved in making a gain estimate is included here for illustrative purposes.

The example will be for the 20 foot paraboloidal reflector (f/d = 0.56) shown by Report #5 complete with dual-mode horn feed operated in the 1296 mc/s EME band.

A value of Sun noise ratio A (Equation 5) was measured and found to be 12 dB. Note that the actual value of Solar flux is not needed for this measurement procedure, nor is the receiving system noise temperature, however the stability (constancy) of flux is important during the measurements as well as the stability of the receiving system.

For the value A = 12 dB, the required half power measurement level is -2.75 dB, from Figure 3.

Using the Sun drift method it was found that it takes 12.88 minutes to go from the half power point on one side of the antenna beam to the other side. The resulting measured beamwidth in degrees is then 12.88 x 15 / 60 = 3.22.

With this value for BW in Equation (4), the directivity computes to be 3250.9 or 35.1 dB. In order to compare this value of directivity with the maximum theoretical area gain, we must add the loss factors for illumination distribution and surface errors. From Report # 5, the theoretical computed efficiency factor for illumination distribution only is 0.5 dB, for the dual-mode feed and a reflector with f/D = 0.56. In this example the surface error correction is assumed to be 0.1 dB (an optimistic assumption). Adding these to the directivity by beamwidth measurement gives the total effective antenna gain, Geff = 35.1 + 0.6 = 35.7 dB. And finally comparing this with the theoretical maximum aperture area gain for a 20 foot aperture at 1296 mc/s, ($Go = (4\pi Ao)/\lambda^2$), results in a total efficiency factor for the complete antenna system of 35.7 - 38.3 = -2.6 dB. This total efficiency factor is very reasonable considering that a typical antenna of this type has an actual measured maximum efficiency of 1.8 dB below aperture area gain in practice. (See Figure 4 of Report #5)

A total efficiency factor in this case of -2.6 dB means 55 % with 45 % of the feed input power unavailable for effective gain radiation, a typical situation for prime focus fed reflector antennas.

If your beamwidth measurements and comparison with maximum theoretical aperture gain with a parabolic reflector antenna are smaller than -1.8 dB, your antenna is remarkably good or more likely the measurements are in error and should be checked carefully. If an efficiency of less than 50%, i.e. - 3dB or greater is obtained, then the antenna may be suspect and should be checked for surface accuracy, leakage and excessive heat losses, as well as feed antenna radiation (excessive spillover).

3. System Temperature Estimate

Having determined an estimated value for the antenna effective gain, Geff, in Section 2, it is a simple matter to rearrange Equation (3) to obtain Tsys.

$$Tsys = \frac{Geff}{R - 1} \times \frac{\psi \lambda^2}{8 \pi k}$$

Which is simply $\frac{Tsun\text{-}ant}{R - 1}$.

Since in this particular example, R and A are the same measured quantities, the system temperature which is now very dependent on the value of Sun flux at the time of measurement, (assumed here to be 80 Solar Units) is

$$Tsys = \frac{3745.4}{15.9 - 1} \times \frac{80E{-}22 \times (0.23)^2}{8 \times 3.1416 \times 1.38E{-}23}$$

$$= 307 \text{ degrees K} \quad !!$$

In general, the value of R and A may not be the same if the measurements are made at different times. A value of Solar flux must be available at the time when R is measured.
The result obtained here indicates that the assumed antenna gain was a bit low. This contrived example is not necessarily indicative of practical measured results! Error in measurements can easily be of the order 0.5 dB and in particular, the beamwidth measurement is a sensitive parameter and should be done several times for confidence.

Final Remarks

These methods of estimating system parameters, i.e., Geff, Tsys, are useful in ferreting out deficiencies in a system where performance appears to be below expectations. While the accuracy of such methods may not be of laboratory quality, the availability and simplicity of Sun Noise measurements make them good tools for the EME enthusiast.

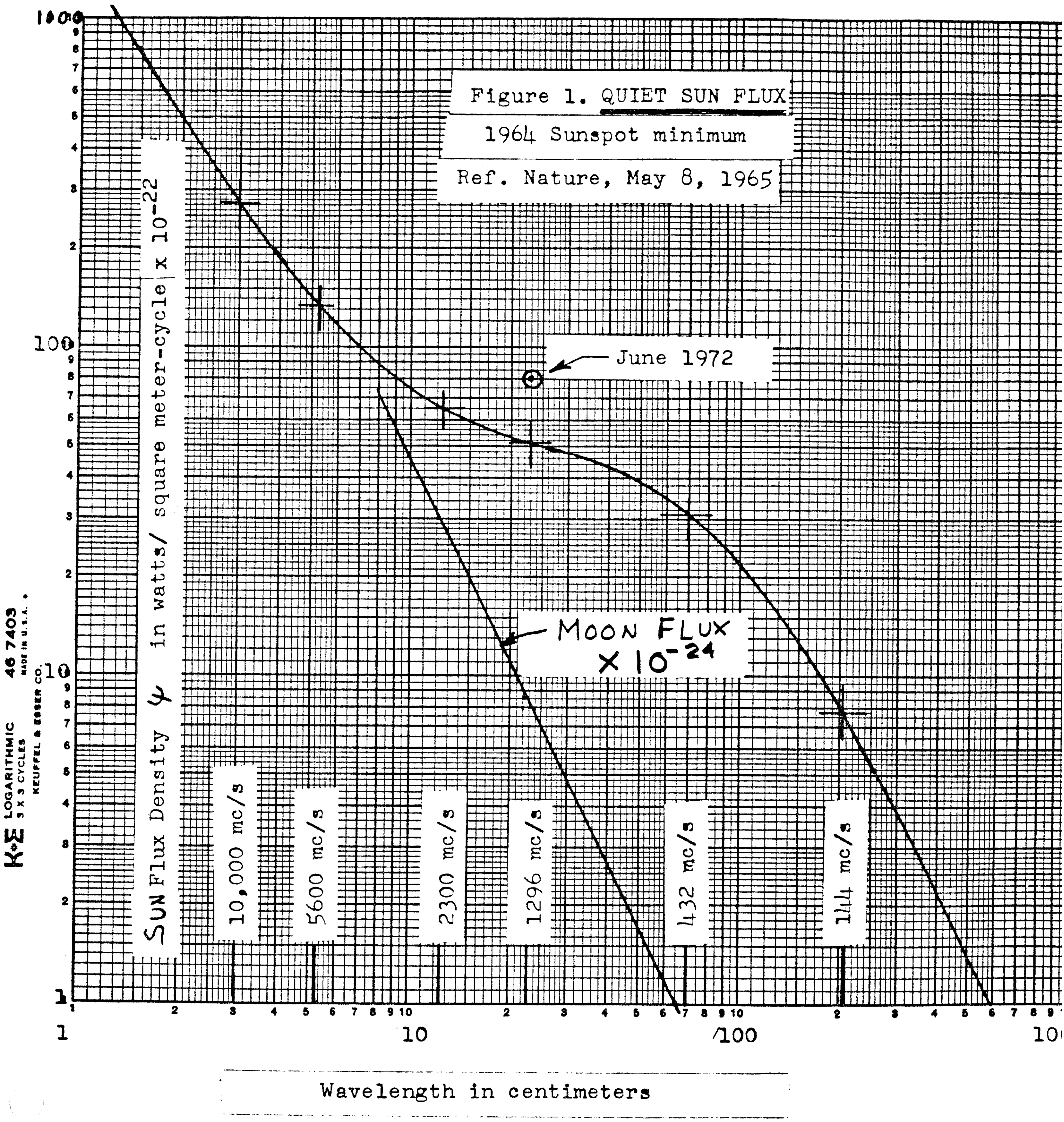


Figure 1. QUIET SUN FLUX
1964 Sunspot minimum
Ref. Nature, May 8, 1965
June 1972
MOON FLUX × 10⁻²⁴
SUN Flux Density ψ in watts/ square meter-cycle x 10⁻²²
10,000 mc/s
5600 mc/s
2300 mc/s
1296 mc/s
432 mc/s
144 mc/s
1000
100
10
1
1
10
100
Wavelength in centimeters
K+E LOGARITHMIC 3 X 3 CYCLES KEUFFEL & ESSER CO. 46 7403 MADE IN U.S.A.

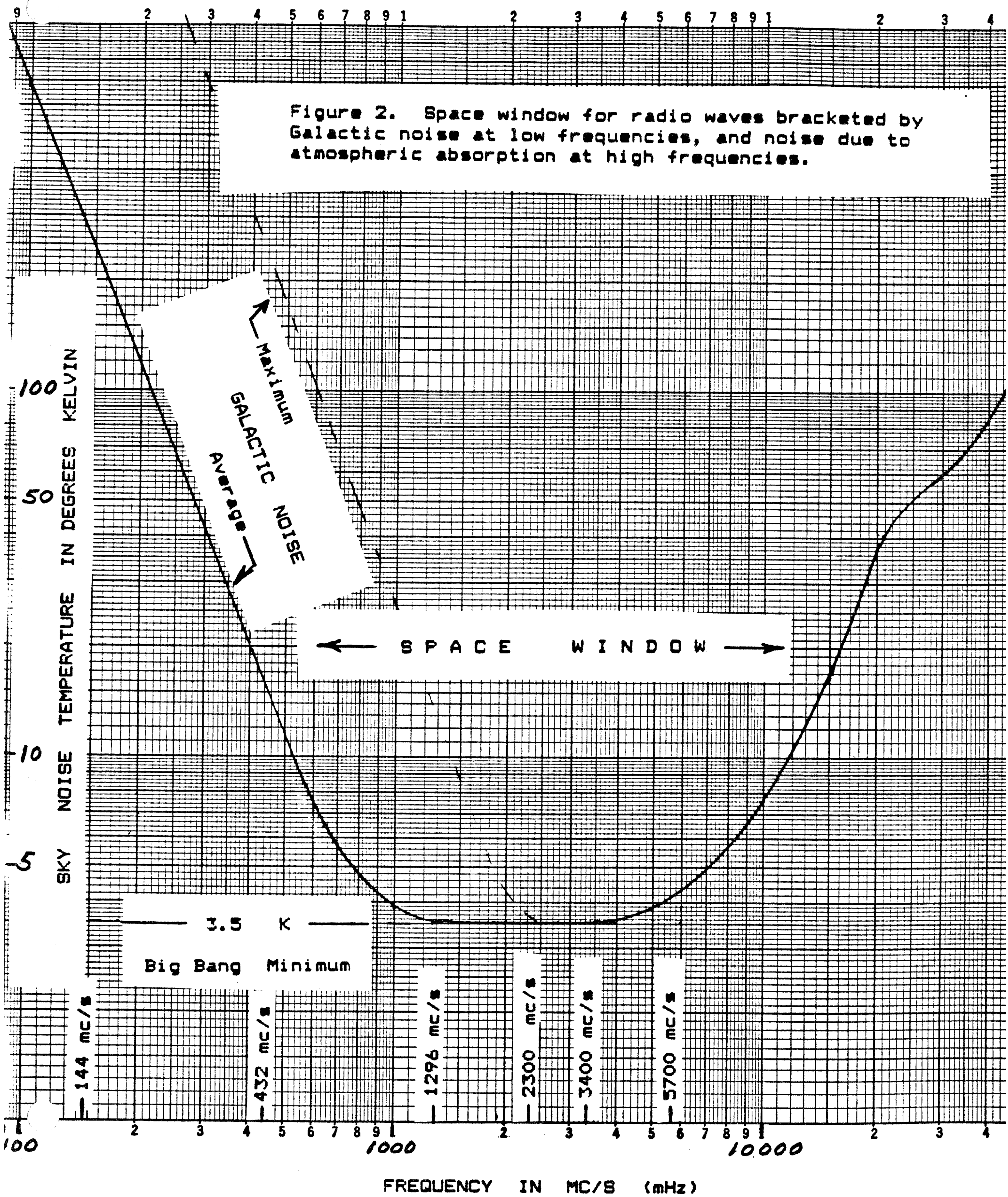


Figure 2. Space window for radio waves bracketed by Galactic noise at low frequencies, and noise due to atmospheric absorption at high frequencies.
Maximum
GALACTIC NOISE
Average
SPACE WINDOW
3.5 K
Big Bang Minimum
144 mc/s
432 mc/s
1296 mc/s
2300 mc/s
3400 mc/s
5700 mc/s
SKY NOISE TEMPERATURE IN DEGREES KELVIN
100
50
10
5
100
1000
10000
FREQUENCY IN MC/S (mHz)

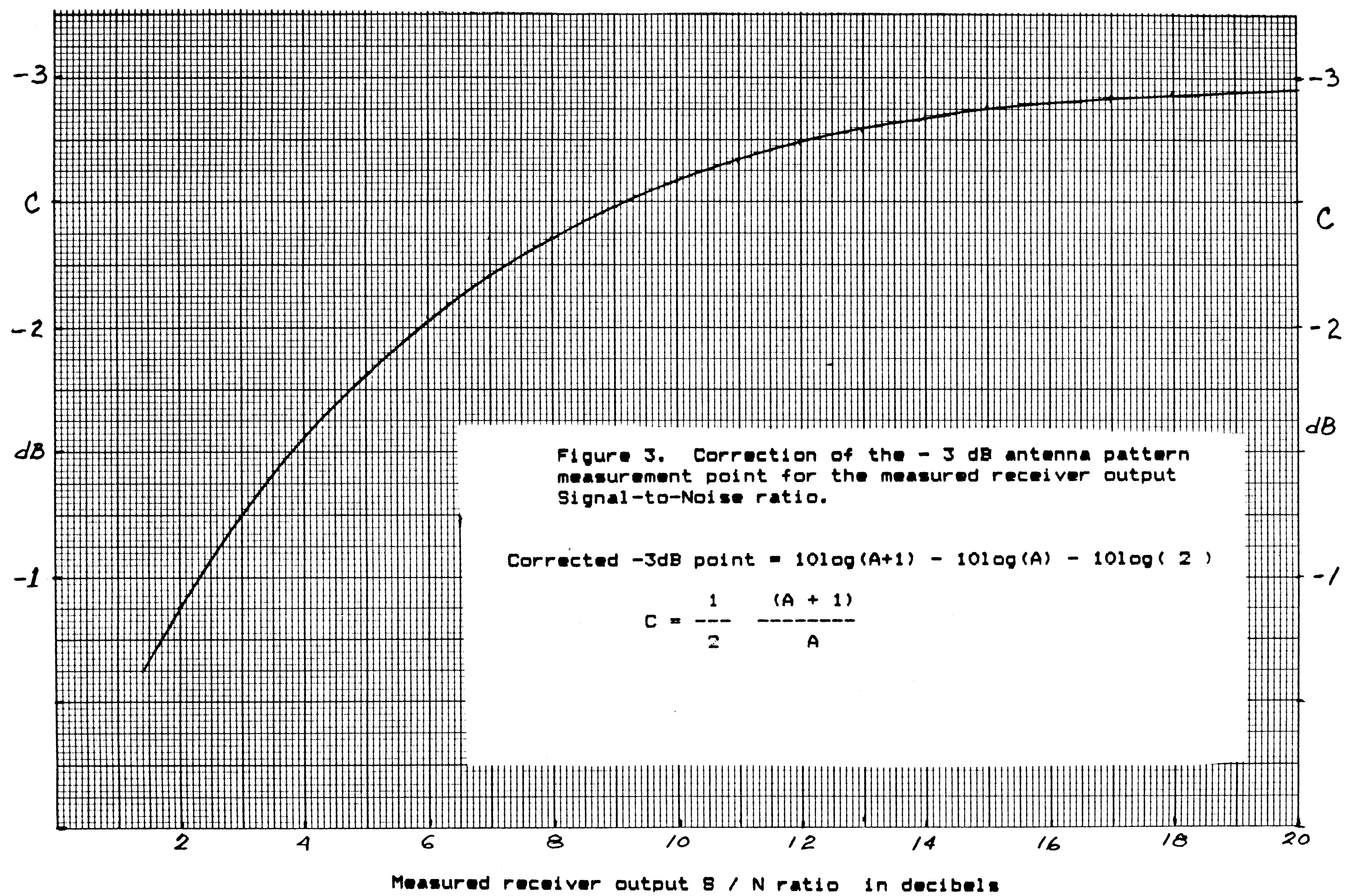


Figure 3. Correction of the - 3 dB antenna pattern measurement point for the measured receiver output Signal-to-Noise ratio.

Corrected -3dB point = 10log(A+1) - 10log(A) - 10log(2)

$$C = \frac{1}{2} \frac{(A + 1)}{A}$$

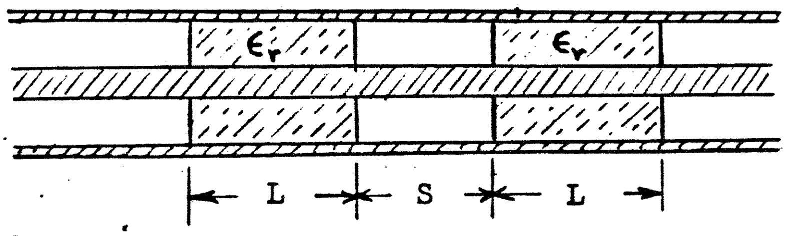
TECHNICAL REPORT # 12

From: The Crawford Hill VHF Club, W2NFA December 1972

SLUG TUNER FOR 1296 MHz

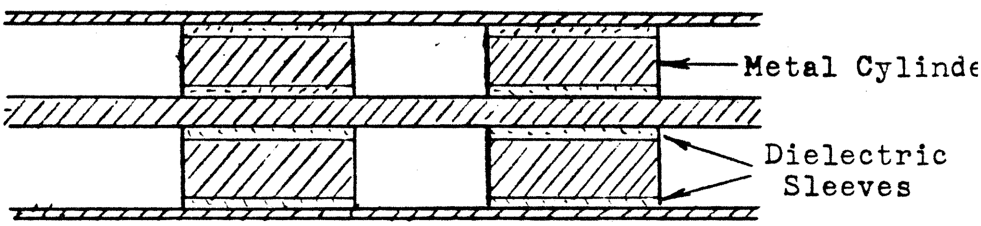
This report describes a low-loss double slug tuner which can be used for impedance matching over a limited range of SWR in coaxial transmission lines at one frequency in the UHF region. This double slug tuner introduces two electrically quarter wavelength line sections into the coaxial line in the form of dielectric slugs or insulated metallic slugs as shown below.

TYPE-A



$$L = \frac{\lambda_o}{4\sqrt{\epsilon_r}} \qquad \lambda_o = \frac{3 \times 10^{10}}{freq_{Hz}} \text{ cm.}$$

TYPE-B



Adjustment of impedance transformation is accomplished by sliding the slugs along the line. This is done through a single narrow longitudinal slot in the outer conductor wall. In a preferred arrangement, the two slugs may be moved simultaneously as a pair. For this case, movement of the pair with constant spacing between slugs affects the phase of the mismatch while movement of the slugs equally in opposite directions affects the magnitude of the mismatch. In general the slugs may be moved individually in an empirical manner until the desired degree of match is achieved. The maximum value of SWR that may be tuned out is equal to the square of the dielectric constant of the slug.

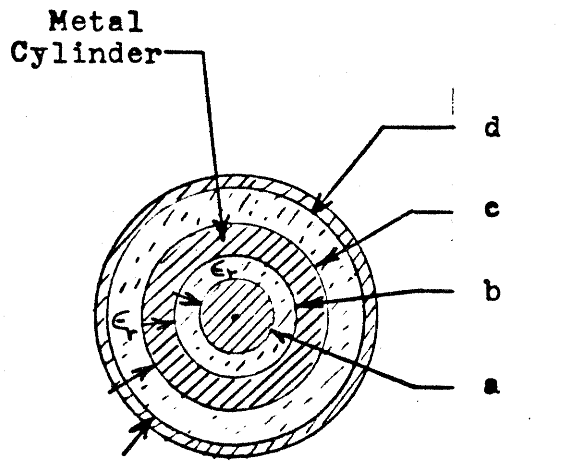
The property of the double slug tuner which ranks it superior to other type tuners is that there are <u>no</u> sliding metallic contacts which carry r-f currents. The presence of the dielectric inside the line changes the characteristic impedance of the line in the region of the dielectric by a factor of $1/\sqrt{\epsilon_r}$ where ϵ_r is the dielectric constant of the material used. For a low loss tuner, the material must have low loss at the frequency of interest. In the UHF region, Teflon (ϵ_r=2.1) is the most readily available low loss material which is also easily machined. Some ceramics and other plastics such as polystyrene may be used for low powers but Teflon is recommended at powers exceeding a few watts.

The use of any external impedance matching device interposed in a transmission line or between circuit and line should be regarded as a "crutch". In some cases however, the tuner may be a necessary convenience. For example, in matching a high power r-f amplifier to a transmission line, it may be inconvenient to have an adjustable matching device directly in the amplifier circuit. It should be pointed out philosophically that if a large SWR exists on a line, it is <u>not</u> good practice to use a tuner in the line but rather to minimize the SWR initially and then if necessary mop up the residual SWR with a low loss tuner of the type described here.

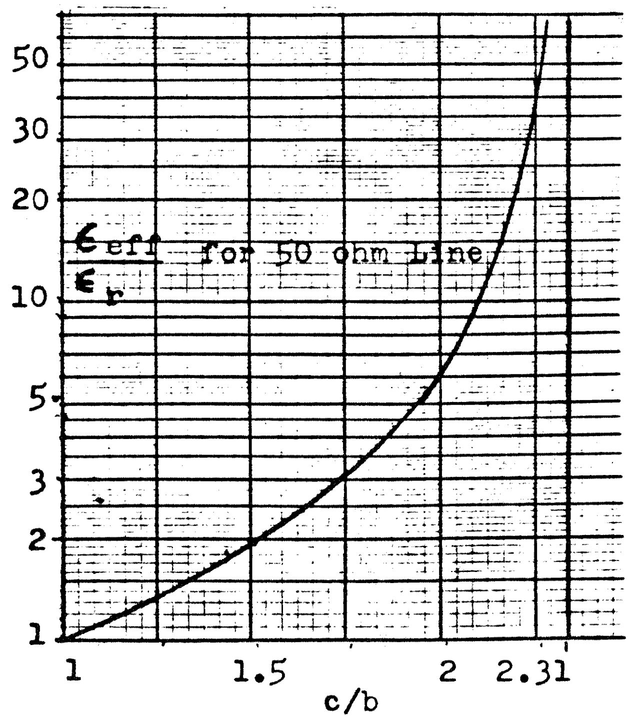
Two designs are shown on the previous page. Type-A uses simple dielectric slugs and is easier to fabricate but has a limited SWR tuning range. Where the SWR may be higher than type-A can handle, type-B is suggested in which the slugs are part metal and part dielectric. In this case, thin Teflon sheet may be used to separate the metal slug from both coaxial conductors. The thickness of the Teflon sheet is determined by the amount of power and magnitude of SWR in the line. For typical cases where powers exceed a few watts, the Teflon should be greater than .010 inches thick.

A cross section drawing of the type-B compound slug is shown below together with formulas to compute the effective dielectric constant and hence the maximum SWR which can be handled. Note that the length, $L=\lambda_0/4\sqrt{\epsilon_r}$ for the type-B slug is determined by the ϵ_r of the insulating dielectric and <u>not</u> the ϵ_{eff} of the composite slug.

$$\epsilon_{eff} = \epsilon_r \left[\frac{\log_{10} \frac{d}{a}}{\log_{10} \frac{db}{ac}} \right]^2$$



For 50 ohm air line d/a=2.31
" 75 " " " d/a=3.46



Construction

Construction of a double slug tuner first requires selection of suitable material for the coaxial line commensurate with the desired impedance. For a 50 ohm line, the ratio of the I.D. of the outer conductor to the O.D. of the inner conductor should be as near to 2.31 as possible. Of the commonly available material, ½ inch O.D. copper tubing used in household plumbing has an I.D. of 0.43 inch. For the inner conductor, 3/16 inch (0.1875 inch) diameter rod or tubing results in a ratio of diameters of 0.43/0.1875 or 2.29 which is adequate. The copper tube is of the rigid variety. Although aluminum may be substituted for the outer conductor, the inner conductor should be of copper or at least brass (silver plated if possible) since the surface current density is greatest on the inner conductor.

Figure 1 shows all of the dimensional details of a 50 ohm double Teflon slug tuner for 1296 MHz. Slotting of the tubing can be done in the home workshop on a bench or radial arm saw equipped with a suitable thin metal cutting blade. It is advisable to dowel the inside of the tube first and then clamp the tube securely before sawing. After all burrs are removed from the slot, both ends of the tube should be carefully squared (faced off). Next, the locking nuts of the cable connectors should be bored out to be an easy fit over the ½ inch tube. The nuts may now be soldered on the tube as shown. The use of aluminum for the outer conductor may present a problem with soldering and should be considered when selecting materials.

The center conductor should be cut ½ inch longer than the outer conductor to allow for ¼ inch extensions on either end which are cut down in diameter to fit inside the connector pin. The slight taper at each end should be machined after the pins are properly registered and soldered in place. The taper is not critical and is done to avoid an impedance "bump" and to avoid sharp corner corona. The use of tubing for the center conductor requires that the ends be fitted with short pieces of solid rod to permit soldering of the pins and tapering of the ends as described. Since standard cable connectors do not have captive pins, the panel receptacle (UG58A/U or equivalent) which does have a captive pin may be preferred as an alternate on one end in some cases.

The Teflon slugs should be machined carefully to be a snug slide fit on the center conductor and an easy clear fit to the outer wall. The slugs are made 1.6 inches long since this is an electrical quarter wavelength at 1296 MHz for a dielectric constant of 2.1. Adjustment of the slugs for tuning is accomplished by means of a thin wooden or plastic blade through the slot. Radiation leakage through the slot should be negligible but do not attempt "hot" adjustment with a metal blade device as coupling will result. Should the very small amount of radiation leakage be objectionable in cases where the tuner is to be included on a permanent or semi-permanent basis, the tube may be wrapped with copper or aluminum foil and taped securely. Weatherproofing may also be accomplished with ordinary plastic electrical tape.

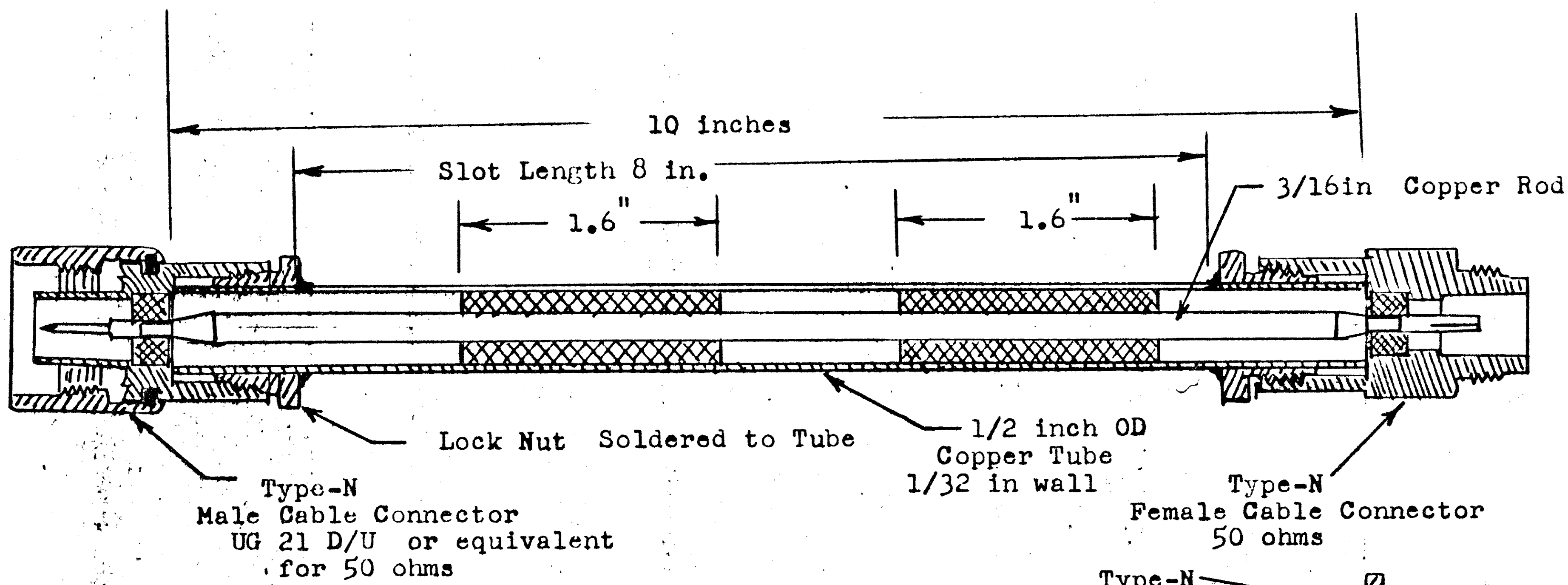
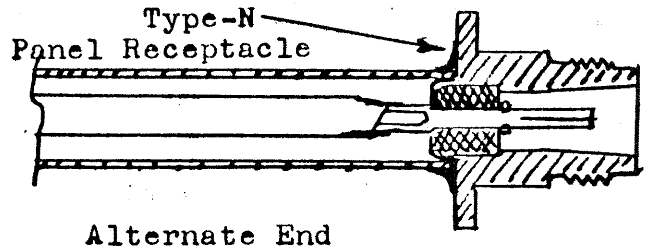


Figure 1. Teflon Slug Tuner
for 1296 mc/s, 50 ohms



Alternate End
Connector Arrangement

TECHNICAL REPORT # 13

From: The Crawford Hill VHF Club
Date: December 1972 (rewritten December 1986)

Subject: A Kilowatt Power Amplifier for 1296 mc/s

Introduction

This report is provided as both a modification procedure and as a guide for designing a parallel operated multi-tube power amplifier for UHF. The modification procedure is specifically for a World War II U.S. military surplus pulsed amplifier designated as part of a UPX-4 IFF system. Original units are very scarce but are still available at an exhorbetent price*. The UPX-4 was originally used in an L-band (900-1100 mc/s) pulsed plate modulated transponder ground station and employed 6-2C39 planar triodes in parallel in a grounded grid configuration.
The modified version of the UPX-4 described in this report retains only the modified anode cavity and grounded grid configuration. Some of the physical parts were also used as source material. This somewhat drastic modification was felt necessary to bring the tuning and impedance matching into the 1296 mc/s range. The original silver plated brass casting and machined parts bring the needed physical stability and low loss to the anode cavity, a necessity at UHF.
Although the original unmodified unit and the original modified version (December 1972 Report #13) employed air cooling, This report describes a water cooled version which improves anode tuning stability, a problem in air cooled versions which cannot cool the tube anode internal structure as efficiently. Some of the original drawings using air cooling have been included at the end of this Report for historical reasons.
In more recent times, OZ9CR with excellent foundry and machine shop facilities, had been duplicating the modified design. An estimate of 30 such units were released to amateur radio EME enthusiasts and are highly recommended for this service, if you can obtain one. OZ9CR no longer offers this service.
Recommended tube types for this amplifier are the 3CX100A5 or 7289 (1986 pricing $29 new). Some used tubes can still be found in surplus outlets. The glass sealed 2C39 is not recommended for this high power amplifier.

It is more efficient in terms of output power to operate a multi-tube power amplifier with all tubes in parallel coupled to a common anode resonator than to operate multiple amplifiers through a corporate combining network. Even though the latter method provides a 'graceful' degradation of output in the event of tube failure, the additional losses and complication of combining networks and the need for individual amplifier adjustment still favor the multi-tube design.

Radio Research Instrument Co. Inc., 2 Lake Ave., Ext., Danbury, Connecticut 06811. Tel: 203 792 6666 (Soon to move location)

The grounded grid configuration, though only conditionally stable in operation, is virtually dictated by the planar triode tube type. And finally, because of transit time effects in vacuum tubes, the planar triode with closely spaced elements is virtually the only type tube which will operate efficiently at 1296 mc/s.

Although the design shown in this report uses 6 tubes, amplifiers of this type have been built with as few as 4 tubes and as many as 16 tubes operated in parallel, successfully.* Plate circuit efficiency of up to 50 % has been obtained while operating the tubes above recommended conditions with output power for 6 tubes of 500 watts easily obtained, and up to 800 watts output reported. Operation of these tubes under stressed conditions can only be obtained by special precautions, with special attention to heat removal from the tube elements, anode and grid. Water cooling of the anodes is recommended (see Report # 6) and generous air flow past the grid ring is essential. Other precautions to observe are; never exceed a maximum cathode current of 300 ma. for the 3CX100A5/7289 type tube, provide an efficient r-f tank circuit with good impedance matching to the output load, provide secure and tight finger stock socketing, provide good r-f by-passing especially at the anode circuit, and use good UHF practice in construction of r-f circuits. This means well fitted parts assembled with a minimum of solder and removal of all excess solder from areas of high r-f current.

UPX-4 Modifications

The modifications presented here require the use of some machine tools and a fair knowledge of machine shop practices, none of which are beyond the capability of a devoted UHF-EME enthusiast.

Anode Circuit

Modification of the anode circuit requires the removal of the adjustable tuning capacitor at the center of the cavity, the fixed output coupling loop and output jack. The output line hole should be bored out to be a snug fit with a short piece of 1/2 inch copper tube which will include a coupling loop. The boring operation may be done by first drilling out the hole with a 1/2 inch drill and then increasing the inside diameter slightly with an expansion reamer until the copper tube fits snugly but can be rotated and slid in and out. Details of the output link and line will be given later.

Measurements on the original anode cavity circuit show that there is a current maximum exactly at the center of the cavity at 1296 mc/s. For this reason, it is necessary to extend the cavity mode resonance to 3/4 wavelengths by means of a coaxial extension stub from the center of the cavity. Tuning of the anode cavity is accomplished by dielectric capacitive loading of the 'hot' end of this stub.

"Annular Circuits for High-Power Multiple Tube, Radio-Frequency Generators at Very -High Frequencies and Ultra-High Frequencies", D. H. Priest, Proceedings of the IRE, May 1950, pages 515-520.

Figure 1 shows a cross sectional view of the UPX-4, modified. Note that the extension coaxial stub outer sleeve is at the anode voltage (+1500v. d-c) and should be provided with a safety shroud as shown by Figure 1. This shroud or cover also encloses all the tube water jackets and interconnect hoses as well as providing an r-f shield for the anode by-pass capacitor.

The extension center conductor of the coaxial stub is shown as a machined mushroom shaped piece. The size and shape are not critical and can be dimensioned directly from the drawing. The reason and necessity for the wide mushroom shaped bottom is that high r-f currents are flowing in the region where the stub is flared out. The flaring provides a larger surface area for the currents to flow thus reducing losses. The flaring should extend almost to the grid finger stock and a champhered edge should be machined at the outer bottom edge where good electrical contact with the cavity bottom wall is essential. The champhered edge should 'bite' into the cavity wall. The top of the stub should be rounded to minimize r-f field breakdown. In general, all sharp edges in r-f circuits at UHF should be removed or smoothed. The piece is secured to the bottom of the cavity by one or more flat head machine screws. The piece should be machined from copper with a fine finish or from brass, finished and silver plated.

The outer sleeve of the coaxial stub extension, which is a piece of 3/4 inch copper pipe, must be soldered to the anode tube socket plate. Soldering near existing finger stock must be done with extreme care in order not to anneal (remove the temper in) the hard copper finger stock, which will soften the material resulting in poor finger stock connection to the tubes. A technique which can be used for this soldering operation is to fit the pieces together snugly with solder paste in the joint. Then place the assembly on a 'hot plate', raising the temperature to a little less than the melting point of soft solder. Next apply heat to the top of the sleeve with a torch until the solder applied to the joint begins to flow. Do not overheat with the torch or 'hot' plate, and remove all heat as soon as the solder has flowed evenly. Avoid using rosin core solder if possible as the clean-up is more difficult. Rosin core solder residue may be cleaned away with alcohol and a brush while the piece is still hot. Do this in a well ventilated area as alcohol fumes are harmful to your health.

Solid, uncored, solder may be obtained from plumbing supply houses along with a suitable flux. A very good but highly corrosive flux for brass and copper is WETOIL, which is basically a zinc ammonium-stannous chloride solution.

Liquid flux may be applied to the work area with a small brush while the solder is still in a liquid state. Short brush strokes will produce excellent wetting and distribution of solder without destroying the brush bristles. Spattering of the sorrosive flux will occur, and it is advisable to shield clothing and your eyes as well as any other delicate apparatus within range of the spattering.

Tuning of the anode cavity is accomplished by means of a solid rod of Teflon* which has been machined to a slide fit inside the outer coaxial extension piece and with a hole bored in the Teflon rod to fit easily over the center conductor of the stub extension (see Figure 1). For convenience, the Teflon rod is coupled to a screw mechanism which allows movement of the rod into and out of the stub extension, thus effecting a capacitive loading by changing the material at the 'hot' end of the stub from air to a dielectric (Teflon, relative dielectric constant of about 2.1). The tuning range of this arrangement is about +/- 50 mc/s at 1296 mc/s.

In the original UPX-4 amplifier, the anode by-pass capacitor used 4 pieces of sheet Teflon. In the modified design, only one sheet of Teflon, 0.01 inch thick, need be used since the peak d-c plus r-f voltage is far less than in the original pulsed design.

Special care should be taken when installing the Teflon sheet to remove all foreign particles and chips from the parts. As a general rule, when the amplifier is completely modified, it should be thoroughly scrubbed with detergent and water using a brush. Then a careful and thorough inspection should be made for residue particles, sharp edges and foreign material. If any are found they must be removed and the scrubbing should be repeated until all parts are immaculately clean and free of any particles which can easily puncture the soft Teflon sheet and also cause poor fitting of parts.

In some cases it may be advisable to clean the 7289 tubes if they are discolored or have marks on the ceramic that might have been made by some conductive material, such as pencil lead. The ceramic may be cleaned in a similar manner with household scouring powder solution and carefully dried. Silvered parts may be bright cleaned with a material called TARNEX, available in houseware stores for cleaning silverware.

The original air cooling plenum which has no purpose with water cooling of the anodes, is extended above the amplifier to enclose all the water jackets for personal safety, and with a rigid top, to form a stable plate to mount the anode tuner mechanism, as suggested by Figure 1. Metal is suggested for this extension with close fitted holes for the water tubes and anode voltage lead (coax). This enclosure will then act as an r-f shield for any leakage from the anode by-pass capacitor.

<u>Anode Water Jackets for the 3CX100A5/7289 Type Tubes</u>

Water cooling of the 3CX100A5/7289 can be done with a home made water jacket of the type shown by Figure 2. The tube normally comes with air cooling fins attached which must first be removed. Most tubes have an aluminum air-cooling fin assembly secured to the pinch-off tube at the anode by means of two hex head locking screws. Other versions have a screw on cooling fin assembly which may require some machine work or heat to remove. Once the air cooling fin has been removed, the water jacket is inserted into the anode cup and secured with a water proof sealant. A

One such Teflon source is: ESTOK Palstics Company, Inc., 434 Whitehead Road, Trenton, New Jersey, tel: 609-586-4336 variety of such sealants are available as bathtub sealers, gasket material and caulking compounds. This simple method of attachment is adequate because there is little excess pressure from the water pump, there is no need for electrical bonding to the water jacket, and the jacket may be easily removed for tube replacement. The jacket is made from easily obtained and inexpensive copper water pipe fittings which are prepared and soldered together as an assembly, as shown in Figure 2.*

If silicone bathtub sealer is used, the water jacket may be removed by heating the assembly until the sealer softens (below the melting point of solder).

Figure 1 also shows a hold-down device to insure that the tube is kept seated in its finger stock socket. These clamps, one for each tube, should be made of non-conductive material, and their exact design is not critical as long as a small amount of pressure is exerted at the top of the water jackets.

N6CA suggests operating the entire amplifier in an inverted position with the tubes at the bottom. In the event of water leaks, no damage will be done to the amplifier circuit.

An important consideration is electrolysis in which small leakage current through the water supply from the anode voltage will erode away metal in and around the tube anode cup. As a precautionary measure, it is suggested here that the vacuum exhaust pinch-off tube extending away from the center of the anode cup, be coated with a sealant which is both water proof and heat resistant. One such material is Permatex found in auto supply stores, or any of the other gasket forming liquids. A generous coating only over the end of the pinch-off tube which has been squeezed is necessary. Do not coat the anode block or threaded shank where maximum heat transfer is needed.

<u>The Water Cooling System</u>

The tube anode water jackets are connected together in series fashion, and to the water pump and reservoir with clear flexible plastic hose. Series connection of water jackets is simple and gives adequate cooling with little differential temperature between tubes due to the high heat capacity of water and the flow rate. If the r-f driver amplifier is also water cooled, its tubes may be added in series also, simplifying cooling of all tubes with a common water supply and pump.

Since the tube anodes are at high d-c potential, the water supply should be adequately grounded for personal safety. In the system

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* A set of 6 water jackets of this type may be obtained from ANATECH, P.O. Box 65, Colts Neck, New Jersey 0 7 7 2 2.

See 432 and Above News by K2UYH for:

1. Manometer sensor switch - June 1986, Vol. 14, No. 7
2. Alternate water jacket construction - July 1986, Vol. 14, No.8
3. Water pump and heat exchanger - February 1986, Vol. 14, No. 2

Source - Herbach & Rademan Inc. 401 E. Erie Ave., Philadelphia, PA 19134, tel: 215-426-1708

Evaporator Coil Assembly - Part No. TM 23K722

Submersible water pump - Part No. TM22K945 shown in Figure 2, a short section of metal tube is inserted in the supply a minimum of one foot from the tube anodes and grounded to the system H. V. supply ground.

Leakage in the distilled water may be monitored by placing a low range (0 - 100 uamp) d-c meter in series with this ground. This monitor is very important since the distilled water can become contaminated and the resulting electrolysis at the anode pinch-off tube can eventually result in loss of a tube. Should the leakage current exceed about 50 microamperes with a 1500 volts supply, the system should be cleaned and a new supply of distilled water installed.

Never operate the tubes without the water flowing through the jackets, and also air across the grid ring and finger stock sockets. Provide some means of monitoring the water and air flow and an automatic sensor which will disable the anode supply in the event of loss of water or air flow. Such a sensor can be made as shown in Figure 2, which is a simple pressure sensor that includes water level electrodes. These electrodes rely on the water conductivity to complete a circuit, therefore a high gain d-c amplifier must be used to sense the high resistance of the the distilled water. This sensing circuit may also be used to monitor the contamination of the water with a suitable meter or LED indicator at the amplifier output. Some experimenting will be required with the tubing length to accomodate your particular circulating system pressure.

A similar air pressure sensing device may be made with a captive small conductive water reservoir as shown by Figure 2. The OP amplifier may be omitted if the fluid conductivity will operate a small low field current relay directly.

The distilled water supply should be several gallons in capacity, or optionally a smaller reservoir and a small heat exchanging radiator with fan assembly to cool the water, may be used. Distilled water is readily available from any Super Market.

The reservoir should be non-metallic and covered to avoid airborne contaminants.

The circulating water pump need not have a capacity more than a gallon or two per minute. Such pumps are found in acquarium supply stores and are usually submersible.

The heat capacity of water is rather large. Fundamentally it takes 1 calorie of heat energy to raise 1 gram of water by 1 degree celsius (centigrade). Converted into more useful terms, 500 watts of power dissipation into a gallon of water for 10 minutes will raise the temperature of that gallon of water by about 20 degrees celsius or 68 degrees Fahrenheit. In the system described in this report the entire water supply is operated at ambient room temperature without insulation and will dissipate heat to the surrounding air, thus increasing the cooling capacity somewhat further.

The circulating system shown in Figure 2 includes an optional heat exchange radiator which should be included if experience indicates that your operating practices drives the water temperature up by more than about 50 degrees celsius.

When starting up a water cooled system, care should be taken that air pockets in the hoses should be cleared. With clear plastic flexible tube, air pockets may be observed visually and cleared by pinching the tube where pockets appear.

Cathode Circuit Modification

This cathode circuit modification is aimed at using as much of the original amplifier material as possible. An alternate modification, which will be described later, discards the entire original cathode assembly. Both modifications are essentially the same electrically and perform equally well. The electrical cathode circuit consists of a single high-Q resonator to which the tube cathodes are magnetically coupled in a non-resonant manner. This arrangement was first suggested by K4QIF and has proven to perform better than others which have been tried.

The lower portion of Figure 1 shows the completed modification in which the original cathode casting has been machined. Figure 4 shows the basic modification to the original cathode casting which consists of cutting off the casting at the indicated parting line. The cut surface should be faced off on a lathe for accuracy. The parting line location is determined by having the electrical length of the Teflon loaded cathode coaxial by-pass capacitor be an electrical quarter wavelength at 1296 mc/s., or 1.6 inches long. This dimension dictates the height of the modified cathode chamber.

The new cathode chamber is made from a very short piece of 5 inch diameter (1/8 inch wall) copper or brass tube 1.10 inches high, which is permanently soldered to the cathode casting and carefully machined (lapped on a flat surface or turned on a lathe) to be a tight fit with the bottom outside wall of the anode cavity. Since the cathode chamber wall shape and diameter are not critical, an alternative is a strap of brass or copper which may be formed into a cylinder or hexagon approximately 5 inches in diameter.

The tapered hole at the center of the casting is filled with a piece of the original tapered coaxial center conductor which is truncated and soldered in as shown in the lower portion of Figure 1. A short piece of 1.0 inch copper tube is pressed into the central hole of the truncated cone section and soldered in place. This is the cathode resonator and is equipped with finger stock and a sliding copper cylinder which provides capacitive loading and resonance adjustment. A non-metallic plunger handle and guide ring complete the resonator construction. Capacitive loading of the resonator will be high since the resonator is shorter than a quarter wavelength. For this reason, small changes in plunger setting will greatly effect the resonant frequency. The builder may wish to consider a vernier arrangement for tuning (either mechanical or electrical, see Figure 6).

The original heater-cathode tube stem assemblies are used as-is but extended further into the cathode chamber than in the original design. The fibre annular ring used to retain the

heater-cathode stems is also used in its original role. A minimum of three holes should be drilled and tapped into the casting to provide an anchor for the retaining ring. Make sure that the registration of the retaining ring and holes in the casting is good before drilling the anchor holes. Six threaded studs were used as anchors in the original design.
In addition, one 1/2 inch hole must be bored through the casting and chamber wall exactly between two tubes for installation of the cathode r-f input line. This will be a non-adjustable line assembly as shown by Figure 3.

Alternate Cathode Circuit Modification

A variation of the cathode circuit is shown by Figure 6, where virtually all fabrication is from new material. The major difference is in the cathode by-pass capacitors. These are non-resonant capacitors, and because of the large surface area used, capacitance greater than 100 pf can be obtained with reasonable dielectric material thickness (0.01 inch Teflon sheet).
This variation also shows a vernier screw method to obtain smooth cathode resonator tuning. The captive non-metallic shaft screws in and out of the solid metallic moving piece of the resonator.
If the finger stock bears tightly on the moving cylinder and the lead screw is lubricated, no guide is necessary to keep the cylinder from rotating.
The heater-cathode socket assemblies in this design are made from pieces of tubing, rod and finger stock as shown.* Good physical alinement of this assembly is essential with the grid and anode finger rings for stress free tube socketing.

Input and Output Coupling

Figure 3 shows details of both the input and output coupling coaxial lines. Though shown on the same side of the cavity, it may be advantageous to locate the input and output lines diametrically opposite to minimize possible coupling and for physical ease in adjustment and assembly. The output coupling is by means of a magnetic loop located between two of the 6 tubes in the anode cavity. The thin copper tube (1/2 inch O.D. water pipe with 0.04 inch wall thickness) and a 3/16 inch diameter inner conductor rod form very nearly a 50 ohm line. This section of coaxial air line is made as close to a half wavelength so that the characteristic impedance is not critical. The coupling loop is made from a strap of copper and very carefully soldered or screwed to the 3/16 inch center conductor. If soldered, a longitudinal slot cut in the rod is preferred. The strap may be soldered to the inside of the outer tube by first tinning the area of the inner wall and then fitting the strap carefully in

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Finger stock material and or prepared socket finger assemblies may be obtained from Instrument Specialties, P. O. Box A, Delaware Water Gap, PA, ~~18237~~ 18327. Tel: 717 424 8510. The cathode and heater part numbers are 97-420A and 97-280A respectively. Anode and grid rings are also available as part Nos. 97-70A and 97-74A respectively. Ask for a catalog.

place and sweating it together. Attachment of the coupling loop to the coaxial line section is a critical operation and should be done carefully and with a minimumn of solder in order to keep losses down.
A preferred way to fabricate the coupling loop is to use an extension of the outer copper wall pipe, bending it around to form the coupling loop with only a soldered connection to the center conductor of the coaxial line. In this way, no soldering to the outer coaxial wall is needed.
If an ordinary N-type cable connector is used for this line construction, use one with a captive pin arrangement since there is no provisions for supporting or registering the center conductor. A panel connector may be substituted which has a secure center pin. (See Report # 12 for more details on assembling rigid coaxial lines to connectors)

The coaxial line section with magnetic loop is inserted into the anode cavity through the original output coupling hole which has been bored out to be a snug fit with the copper pipe O. D. The depth of the loop should be adjusted as shown by Figure 3 and a mechanical stop in the form of an outer section of plastic or metal tube fitted to prevent the loop from being pushed any further into the cavity. Coupling adjustment is by means of rotating the loop, and a locking screw should be installed to secure the position of rotation. Maximum coupling will occur when the plane of the loop is parallel with the axis of the tubes.
To obtain best impedance match to a 50 ohm load, the double slug dielectric tuner will be necessary (Report # 12). Output loading adjustment must be done with the amplifier operating at full anode voltage and r-f drive. Both rotation of the coupling loop and adjustment of the dielectric slugs should be done carefully with an appropriate output instantaneous power measuring device.

Input
The input coupling method is also shown by Figure 3 and consists of a semi-permanent tap on the cathode resonator. Details of the coaxial line section are the same as for the output coaxial line except that the cathode input line is permanently soldered into the cathode chamber wall as shown. The actual tap on the resonator is made with a piece of wide finger stock or a piece of beryllium copper which is soldered to the center conductor of the coaxial line and presses tightly against the resonator stub. The position of the tap is approximately at the middle of the stub. However, it may be necessary to move the tap point to obtain best match from a 50 ohm source. The use of a double slug dielectric tuner to adjust the impedance match is desirable but not necessary (see tune-up procedure).

Grid Air Cooling
Although water cooling of the anodes is ample for this service, the grid structures must also be well cooled. It is also the

hardest part to cool since the actual heating effect is from within where the physical grid (a screen mesh) which sustains the heat must cool by conduction through the grid connection ring and finger stock. In this design, grid cooling is accomplished by passing a large flow of air between the anode cavity and the cathode chamber, thus forcing the air to flow past the grid ring and finger stock slits. Since the cathode chamber can be made essentially air tight (except for the grid finger stock) it is necessary to attach the air flow supply to the cathode chamber. Vacuum or pressure may be applied capable of one or two pounds per square inch above or below ambient pressure. Adequate air holes should be provided between each tube in the anode cavity side wall. An air pressure sensing device is recommended to disable the anode voltage to both the amplifier and driver in the event of air flow failure.

A two stage blower may be required since fans and squirrel cage blowers generally will not develop sufficient pressure.

Biasing and Metering

A suggested wiring diagram is shown by Figure 5. Besides water cooling of the tube anodes, this revised report includes a noteworthy change in the cathode biasing method.

It has been shown that a well regulated cathode bias supply will improve the performance of any grounded grid power amplifier, especially this UHF design.* The new wiring diagram includes two versions of this improved biasing scheme. One method is continuously variable while the other more simple method has fixed step adjustment.

Metering of cathode current is desirable for safety reasons and more meaningful of the individual tube performance. A single d-c milliampere meter in the anode supply, suitably enclosed for safety with a high-voltage type fuse may be used to determine the actual anode power input. This double metering of anode supply and tube cathodes can also be used to determine the grid current with fair accuracy.

The anode supply should be capable of delivering an ampere at between 1200 and 1500 volts d-c.

Control interlocks are recommended for water and air flow to protect the tubes.

Decoupling of power supply leads and metering leads is mandatory with this and any other grounded grid power amplifier. Although the UHF circuitry may be suitably by-passed and decoupled, it is not necessarily so at lower frequencies. Parasitic oscillations can occur at any frequency where the anode and cathode external circuitry impedances are appropriate. A grounded grid amplifier is only conditionally stable. Dress of leads, shielding and stray coupling can cause such parasitic oscillations which are usually indicated by erratic behavior of the amplifier and the presence of r-f outside of the UHF enclosure. Loss of efficiency

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For example, see an excellent design single-tube 1296 mc/s amplifier by "Chip" Angle, N6CA. QST March-April, 1985. Contains much valuable information and advice on UHF power amplifier design and construction.

at 1296 mc/s and the presence of sidebands far removed from the carrier frequency also indicate spurious oscillations which can even exist down to very low radio frequencies in the hundreds of kilocycles.

In order to minimize the possibility of the amplifier becoming a high power oscillator as well as UHF amplifier, resistive decoupling can be used in all external leads close to the UHF circuitry. Approximately 50 ohms of non-inductive resistance which will dissipate about 20 watts can be used in the anode lead and at the tube connection plate with a slight reduction in anode voltage. This resistance also serves a dual purpose in limiting the maximum peak current to the tubes in the event of internal flashover. This type of protection is recommended by tube manufacturers.

Dressing and shielding of leads may also be requires in special cases. For this design it is recommended that the + H.V. line be a piece of RG 8/U coaxial cable with the shield grounded. which has a breakdown level of over 4000 volts .

W A R N I N G
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High level radio frequency fields (non-ionizing radiation), especially in the UHF region of the spectrum, ARE harmful to health ! ! !

The amplifier described in this report is capable of delivering as much r-f energy as most microwave cooking ovens, and should be treated with equal or more respect.

The principle damage to human tissue occurs as a direct result of heating the tissue. As with microwave cooking (all modern microwave ovens use a magnetron operating at 2450 mc/s +/- 25 mc/s), heating occurs throughout the tissue mass and is not confined to the surfaces as with radiant heat cooking. Irreversible internal damage can be done to tissues if sufficient energy is present for a period of time. The eyes, gonads and internal organs are especially vulnerable in that order. The eyes are particularly vulnerable because of their inability to dissipate internal heat. If you can "feel" the effect of r-f heating with your fingers, the level is already far in excess of acceptable limits and should be discontinued directly.

In testing high power amplifiers, take care to see that the r-f fields are confined to circuits, lines and loads which are well shielded. R-f by-pass capacitors may be especially subject to leakage. In this amplifier design, the anode by-pass must be fitted tightly to minimize leakage. Other potential sources of r-f leakage are poorly fitted or poorly secured coaxial connectors, open microstrip circuitry, poorly fitted shield compartments, and braided outer conductor coaxial cable (RG 8 at 1296 mc/s leaks, as do all single braid flexible coaxial cables !).

A special precaution is indicated in testing with antennas and feeds where the ultimate purpose is radiation at high r-f power for EME communication. Never, never allow your body to come near

the mouth of a feed horn antenna or any small aperture antenna. Dipoles, groundplanes, short verticals and small yagis are especially dangerous since the radiation is not confined to a small angular sector.
A generally acceptable r-f energy density level in the past has been 1 milliwatt per square centimeter. In recent times however, this level has been raised to about 10 mw/sq-cm. This level is for long-time constant-exposure periods. Much higher levels can be tolerated for very short time periods.
Energy density levels can be estimated with good accuracy for a radiating antenna by knowing the power delivered to the antenna, the absolute gain (dBi) of the antenna in the direction of observation, and the distance to the observing point.

$$P_{mw/cm^2} = \left(\frac{Po}{4}\right) * (Gain(O)) * \left(\frac{1\ cm}{R\ cm}\right)^2$$

As an example suppose that you have a dual mode feed horn of the type described in Technical Reports #5 or #9 which has an on-axis gain of about 10. If your transmitter delivers 450 watts of r-f at 1296 mc/s to this feed horn, then the radiation density 1 meters directly in front of the horn is:

$$P_{mw/cm^2} = \frac{450{,}000\ mw * 10}{4 * (100cm)^2} = 36\ milliwatts/cm^2$$

which is an unacceptable exposure level for periods of more than a fraction of a minute.
High level leakage may be detected by holding a small neon lamp close to the suspected leak. Use extreme care around high voltage by-pass capacitors for personal safety. Small neon bulbs will glow violet in color when excited by UHF, and more orange in color at H.F. The field voltage to strike a neon tube is already in excess of harmful levels. Use this method for quick checks of leakage in and around questionable circuit areas.
In the matter of r-f radiation, an ounce of prevention is surely worth a megaton of cure, since there is no accurate knowledge of the effects of even low level radiation on the human organism.
Do not depend on regulations set by government agencies for your health and safety. Avoid exposure as much as possible.

Tune-up Procedure

Initial tune-up of this amplifier, as with any high power amplifier, should be done carefully and with caution. Hazardous voltages and r-f fields levels are involved. The high voltage supply should be initially equipped with a variable auto-transformer at its input in order to slowly increase the anode voltage while checking for faults in wiring, Teflon sheet by-pass breakdown and tube balance.
Tube testing for balance may be done directly in the assembled

amplifier without cooling it the test period is less than a few seconds. Temporarily replacing the regulated cathode bias supply with a 100 ohm 5 watt resistor, one tube at a time may be tested for cathode emission with a 1 KV anode supply. Typically, a useful 7289 tube with 6.3 volts heater supply will draw an anode current of 60 to 100 ma., depending on cathode activity. Tubes which draw less than 50 ma. are useful tubes but not recommended for this amplifier. Cathode current balance between tubes should be in the range +/- 10% at full operating conditions. Tubes which draw excessive or erratic current may be gassy and should not be used. Save these tubes for dimensional jigs in future projects.

With a set of six reasonably balanced tubes installed, all cooling systems in operation, cathode bias supplies connected, a high power r-f load connected to the amplifier output through a suitable power indicating device, and an r-f driver capable of up to 50 watts adjustable output, the first step will be to tune the cathode resonator. With some r-f drive applied and no anode voltage, adjust the cathode resonance tuner for a maximum cathode current through the tubes. Make the adjustment as quickly as possible and reduce the drive if the cathode current exceeds about 10 or 20 ma. per tube. Without anode voltage, this cathode current is all grid current and can damage the tube if allowed to persist or exceed these levels for even a few seconds. If resonance cannot be achieved or is at an extreme of the tuning mechanism, it may be necessary to change the cathode input tap. It may also be desirable to use an r-f tuner of the type described in Report # 12 to trim up and optimize the cathode match.
Once cathode resonance is achieved, the regulated cathode bias circuits should be connected and set for maximum bias voltage. Then full anode voltage may be applied with no r-f drive. The bias voltage setting should now be adjusted so that each tube draws a static current of about 70 ma. If no erratic behavior is noticed, r-f drive may be applied at low level, increasing until the anode current is about 90 ma. per tube. The anode tuning should now be adjusted quickly to find anode cavity resonance, noted by a 'dip' in total anode current and some indication of r-f output power. The output link coupling should be adjusted to an angle for the loop of 45 degrees to the axes of the tubes.
Do not allow the amplifier to run for more than a few minutes continuously during preliminary tune-up, at least until confidence is gained that all functions are operating properly. Any erratic behavior should be a sign of some problem. Proceed with caution until full power levels can be reached.
If all indications of operation are satisfactory at reduced levels, increase the r-f drive. When the tubes are drawing between 100 and 150 ma. of cathode current each, start trimming up the output coupling match and anode cavity tuning to achieve maximum r-f output. Output coupling match involves rotation of the magnetic coupling loop by incremental amounts and sliding of the dielectric slug tuner elements by small amounts.
Always obtain a shallow "dip" in cathode (better still anode)

current with each matching adjustment. Note whether the anode current dip corresponds with maximum r-f output. If they do not correspond, there may be regenerative coupling between anode and cathode circuits, faulty anode by-passing or excessive losses in the anode cavity caused by poor construction or even over coupling to the load. Some misalinement between maximum r-f and minimum anode current can be tolerated. Favor the anode tuning which gives more r-f output.

If your H.V. anode supply can deliver more than 1 ampere of current at 1500 volts, then increased r-f drive will increase anode d-c input power and r-f output, and should be in compliance with the amateur regulations which apply to your license limitations. Maximum cathode current to any of the 6 tubes should not ever exceed 300 ma.

<u>New Construction</u>

Since the UPX-4 is in limited supply and the OZ9CR clone no longer available, the reader may wish to try his hand at building a similar amplifier from materials available. This section then is <u>not</u> a cookbook step-by-step receipe, but simply guidelines and suggestions to be observed in laying out and building a similar 6-tube design.

A radically different approach will be outlined that virtually eliminates the need for machine tools, lathe, milling machine, etc., and makes use of readily available materials and simple hand tools for construction. A drill press would be desirable for accuracy and convenience.

The anode cavity is the most difficult area of construction. For six tubes, the cavity shown by Figure 1 may be used as a guide. The important dimensions are the inside diameter and inside height. Fortunately with this design, if the cavity is not made accurately resonant at 1296 mc/s, resonance can be reestablished by changing the length of the coaxial extension tuning stub.

It is important that the tubes be located so that the anode-grid gap (inside the tube) be near a voltage maximum in the cavity resonant mode. It is equally important that the cavity walls be made structurally strong and highly conductive so that flexing of the walls, which will change the tuning, do not occur, and r-f losses are minimized.

The radical construction approach, suggested here, is to use sheet copper, available as roof flashing material in any building supply house, to line the walls of the cavity and all other r-f active surfaces. By making the inside of the cavity <u>hexagonal</u> instead of round permits lining the cavity with only small slits in the side walls where r-f fields are zero and r-f currents are in a radial direction, not interrupted by the slits. The copper sheet is bonded to rigid back-up support members and plates with Epoxy cement.

Figure 7 illustrates the salient points of this construction. Suitable material to laminate the copper onto may be hard wood. Oak or maple parkay floor tiles a quarter inch thick are available at building supply houses and are remarkable stiff. An Epoxy (2-part dispenser) cement suitable for copper water pipe assembly is recommended. Epoxy cements will decompose at elevated temperatures approaching the melting point of solder, therefore care and quickness are important when soldering finger stock and other copper pipe pieces to the laminated copper sheet. Use jigs, blocks and guide pins where necessary to properly aline the precut copper sheet pieces in position when bonding with Epoxy, as the cement when first applied is extremely slippery. Pressure must be used for laminating during the setting and curing period of the Epoxy. Solid back-up blocks must be used with sufficient pressure to ensure a flat copper surface. The Epoxy should be spread evenly and thinly over the surface in the manner of any surface adhesive. A small home made serrated trowel can be very helpful. Read the manufacturers directions and precautions when using any Epoxy cement.

Copper sheet should be selected for maximum flat area devoid of bumps, bents and scars. Never try to flatten the sheet by hammering to remove even the smallest bump as the process will cause stresses in the material which will prevent it from lying flat and bonding properly. Buy extra material if necessary to obtain sufficient good surface area for all inside and mating surfaces, especially in the area of the anode by-pass capacitor.

Keep in mind also that the height of the anode cavity is dictated mainly by the tube structure and finger stock used. Therefore, one of the first items to collect is suitable finger stock socketing materials. In particular, the grid ring finger stock must have sufficient height to reach both copper sheet laminates, or extra shim stock must be used to bridge the gap.

The cross sectional view in Figure 7 illustrates some of the suggested use of hard wood as structural members. In particular note that the anode cavity is entirely lined with copper except for the anode by-pass capacitor gap. The Teflon dielectric by-pass capacitor is deliberately made to have an electrical length in the radial direction of the cavity, of a quarter wavelength. Careful thought and drawing must be given to the copper sheet layout for the anode cavity to ensure that it will not have excessive gaps along joints. A paper model is highly recommended before the final copper design is cut.

Cutting copper sheet with hand shears can be used but will result in uneven edges with stresses. Cutting any thin sheet accurately is best done by clamping the sheet between blocks of hardwood as stiffening agents and as a guide for cutting with a fine tooth coping saw.

All copper surfaces should be cleaned with emery cloth (or fine sandpaper) until a bright copper finish is obtained and then washed with strong detergent and thoroughtly rinsed and dried.

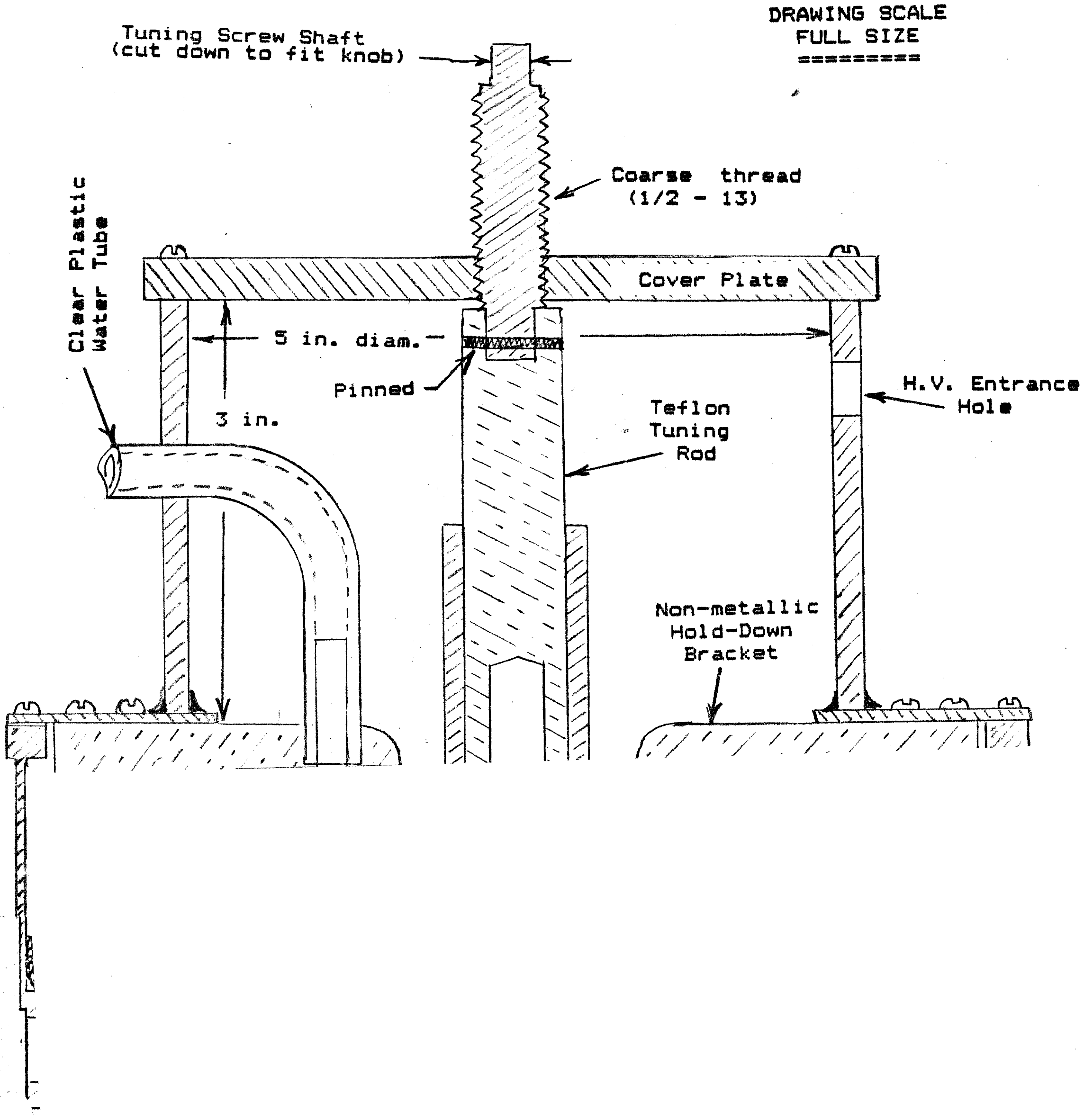
Flat surfaces once laminated, especially in the area of the anode by-pass capacitor should be given extra attention. If the plate surfaces are reasonably flat, they may be further polished by hand lapping. This requires a very flat and solid surface to which a new sheet of emery cloth or sandpaper is secured. The part to be lapped is then rubbed against the cutting surface in a "figure-8" pattern motion (similar to quartz crystal grinding)

until all minor irregularities are removed.
Once again, the final height of the cavity must be estimated from the finger stock socketing of the 7289 tube. Use the upper lip of the anode cup as a stop against the anode finger stock ring to set the tube registration in the cavity. The grid finger stock ring should be registered at the bottom of the tube grid ring assembly, not against the annular grid ring. Faulty contact will be made if the grid annular ring is allowed to seat on top of the finger stock.
Holes may be bored in the laminated copper-wood members with a hole saw, if a suitable size can be found, or with a fly cutting tool mounted in a drill press, with suitable safety precautions.
Cut the copper sheet on each side before cutting out the wood.
Soldering of finger stock and other copper tubes to the laminate copper should be done by first tinning all edges, and then assemblying the pieces and applying heat from a large soldering iron on the <u>inside</u> of the piece. In this way solder can be confined to a very small area of the seam joint. Excess solder should be scraped away from all surfaces that are not in intimate contact. Solder should appear as a very fine line between the pieces joined.
The output coupling assembly may be similar to Figure 3, however since there is only a thin copper wall for the cavity at the point where the output coax line enters, it will be necessary to provide a copper pipe sleeve soldered carefully to the inner wall and located as shown by Figure 7. The sleeve may be a "straight" copper pipe coupling from plumbing supply. It will be best to achieve a good snug fit at the end of the sleeve next to the cavity wall. The home made half wavelength section of coaxial output line can be made from 1/2 inch copper tubing, similar to the line used in the UPX-4 modification.
Finally, air holes must be drilled into the anode cavity edgewise to permit free flow of air intended to cool the grid ring. These holes should not be larger in diameter than 3/16 inch and located at the corners of the hexagon, as shown by Figure 7. Coupling through these air escape holes should be negligible owing to their size. If any leakage occurs due to unfortunate UHF impedance at the outside of the structure, the leakage may be eliminated by soldering in short pieces of thin wall brass tubing at each hole. The tubes should barely extend into the cavity, just far enough to solder.
Other parts of construction may be similar to the UPX-4 modifications. Cathode resonator assembly and water jackets are shown by Figures 6 and 2, respectively, but adapted to the new suggested design as shown by Figure 7.

While a larger number of 7289s may be used in a similar design, recent information on proper biasing of these tubes indicates that the 6 tube amplifier can easily achieve the U.S. legal power limit for amateur service.
This section of the report has been written as a suggested method of building a high power amplifier for 1296 mc/s with minimum cost, and using hand tools. All the details have not been worked out, and it is advisable to think through the design and all

steps of construction before starting to build. Such an amplifier has not actually been built to date, however, it is presented with reasonable confidence that it can be achieved. The singular disadvantage to this method is the detail work which will require more time and care to accomplish.

Figure 1. Modified UPX-4 for 1296 mc/s.
Water and air cooling are used.



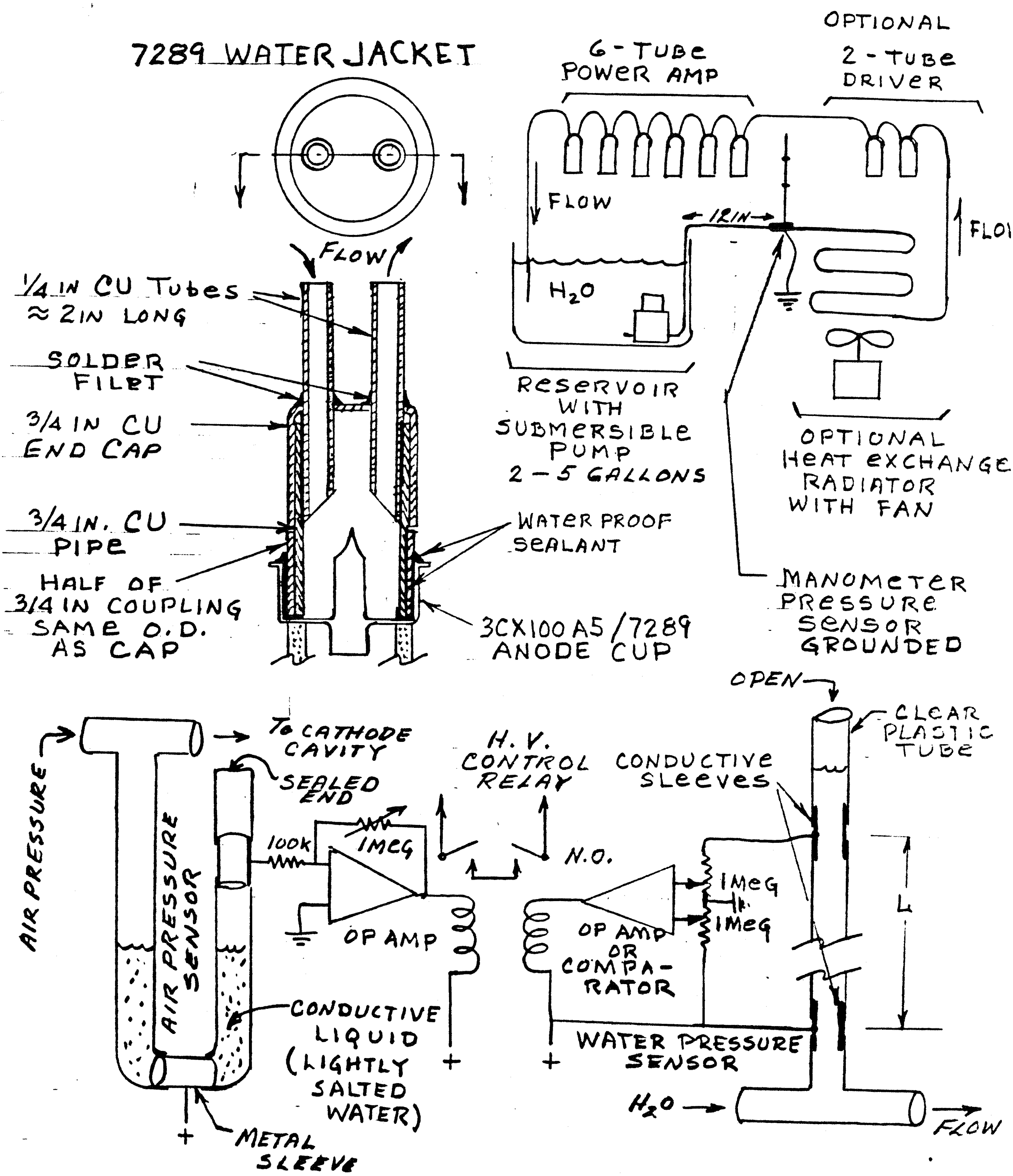
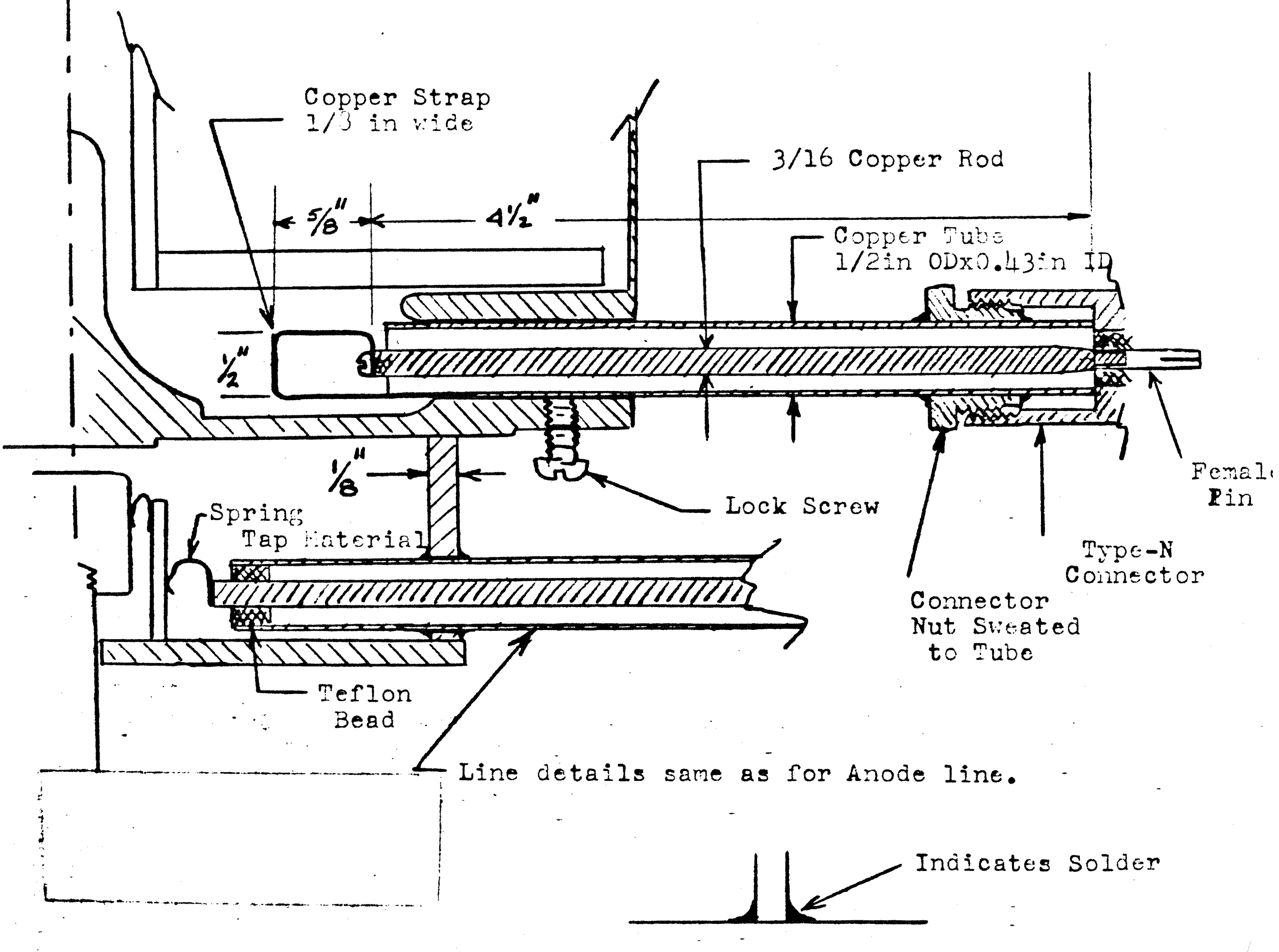


Figure 2. Water cooling system components and with air and water pressure sensors. The 7289 water jackets are made from readily available copper water pipe parts, soldered into a unit and cemented to the anode cup with a waterproof sealant. The pressure sensors are optional for automatic protection.

Figure 3
Input-Output coupling methods for the modified UPX-4 1296 mc/s Power Amplifier. The coaxial lines may be located between any pair of tubes in the amplifier. Locating the input and output lines on diametrically opposite sides of the amplifier may be desirable for convenience and minimum external feedback.

The line sections are made about one half wavelength long so that if the characteristic impedance of the line section is in error by a moderate amount its effect will not introduce additional mismatch.



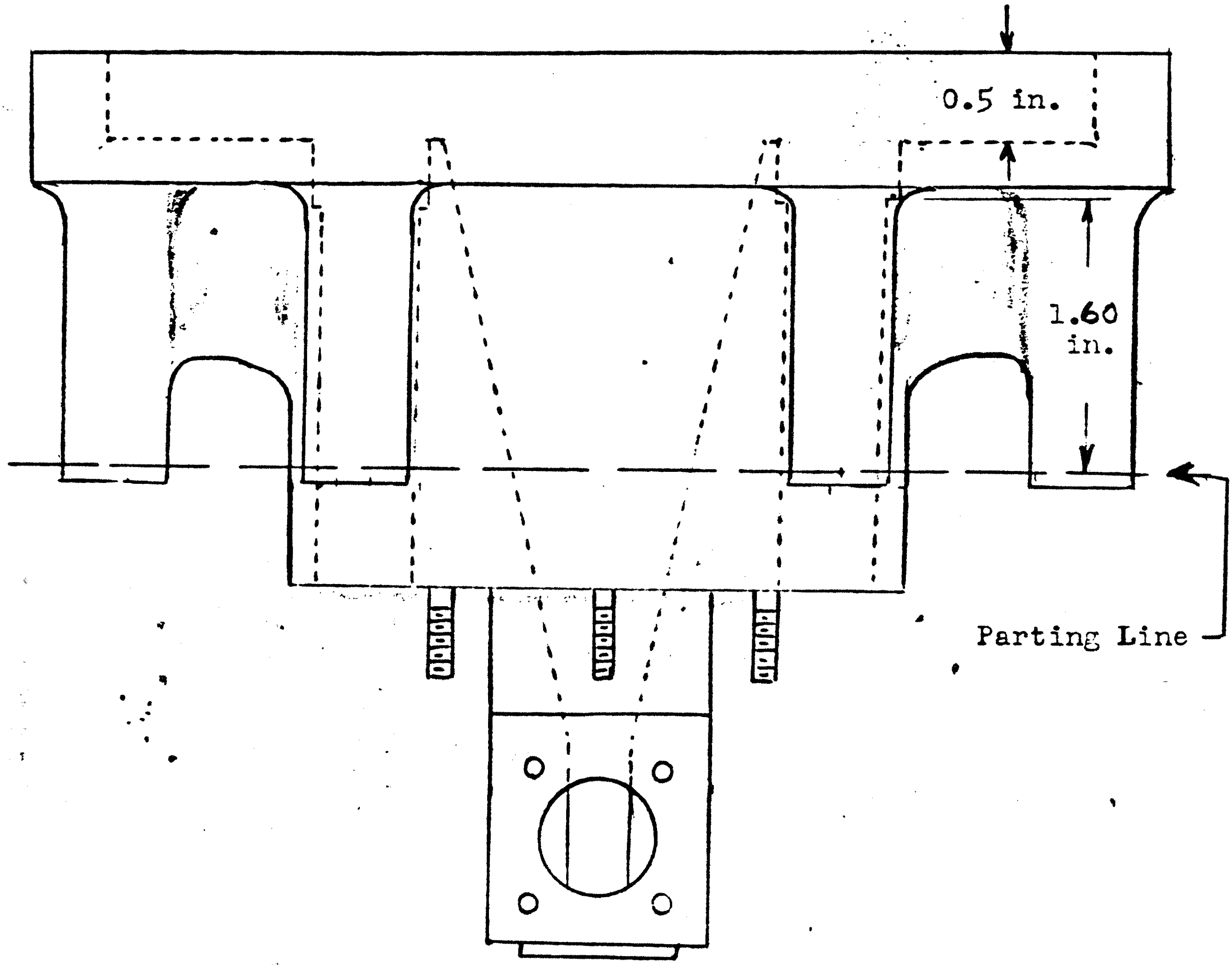


Figure 4. Unmodified cathode casting, stripped, showing parting line and important dimensions.

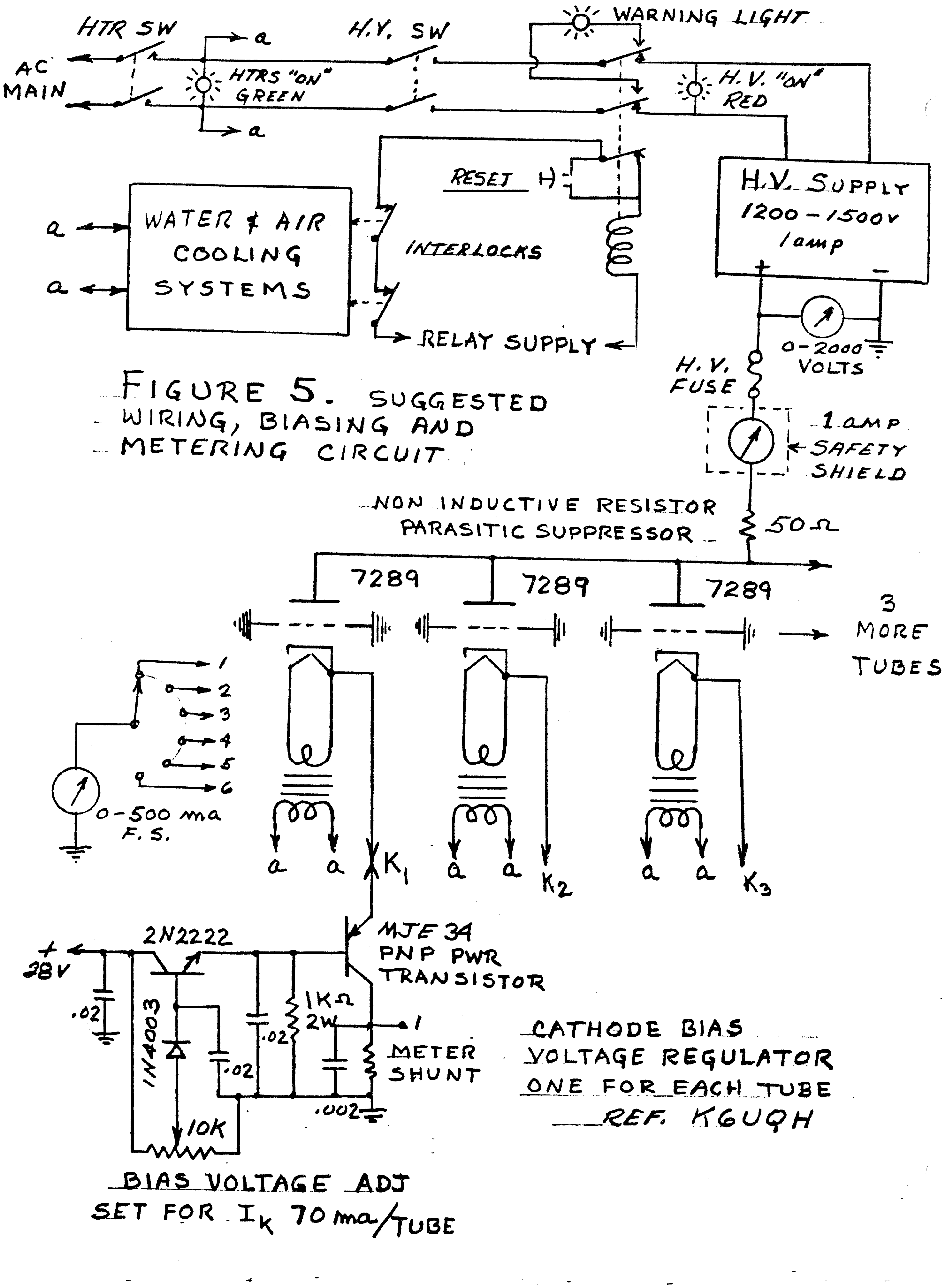


FIGURE 5. SUGGESTED WIRING, BIASING AND METERING CIRCUIT

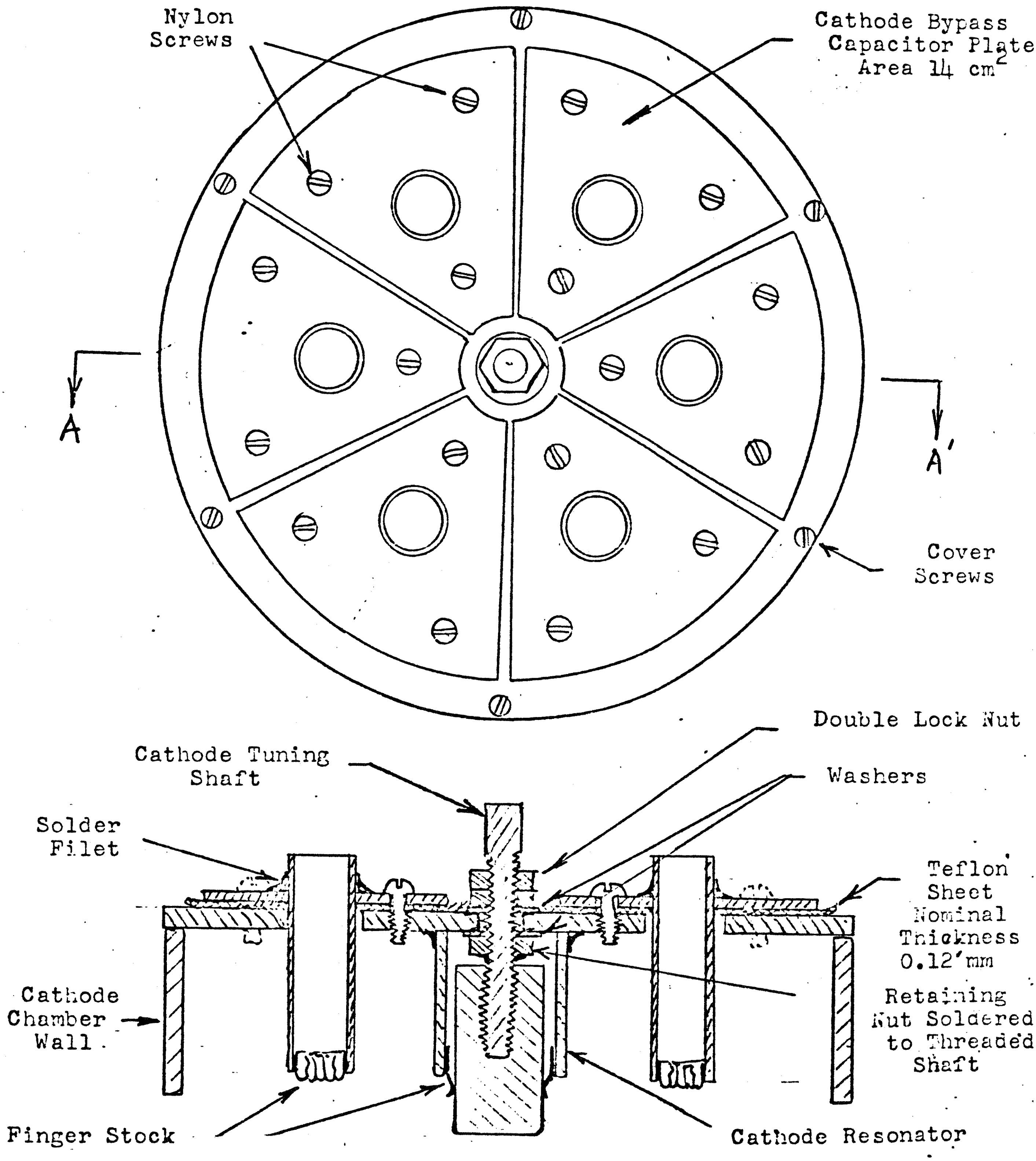


FIGURE 6 Cathode bypass capacitor details. Also a method of vernier tuning of the cathode resonator.

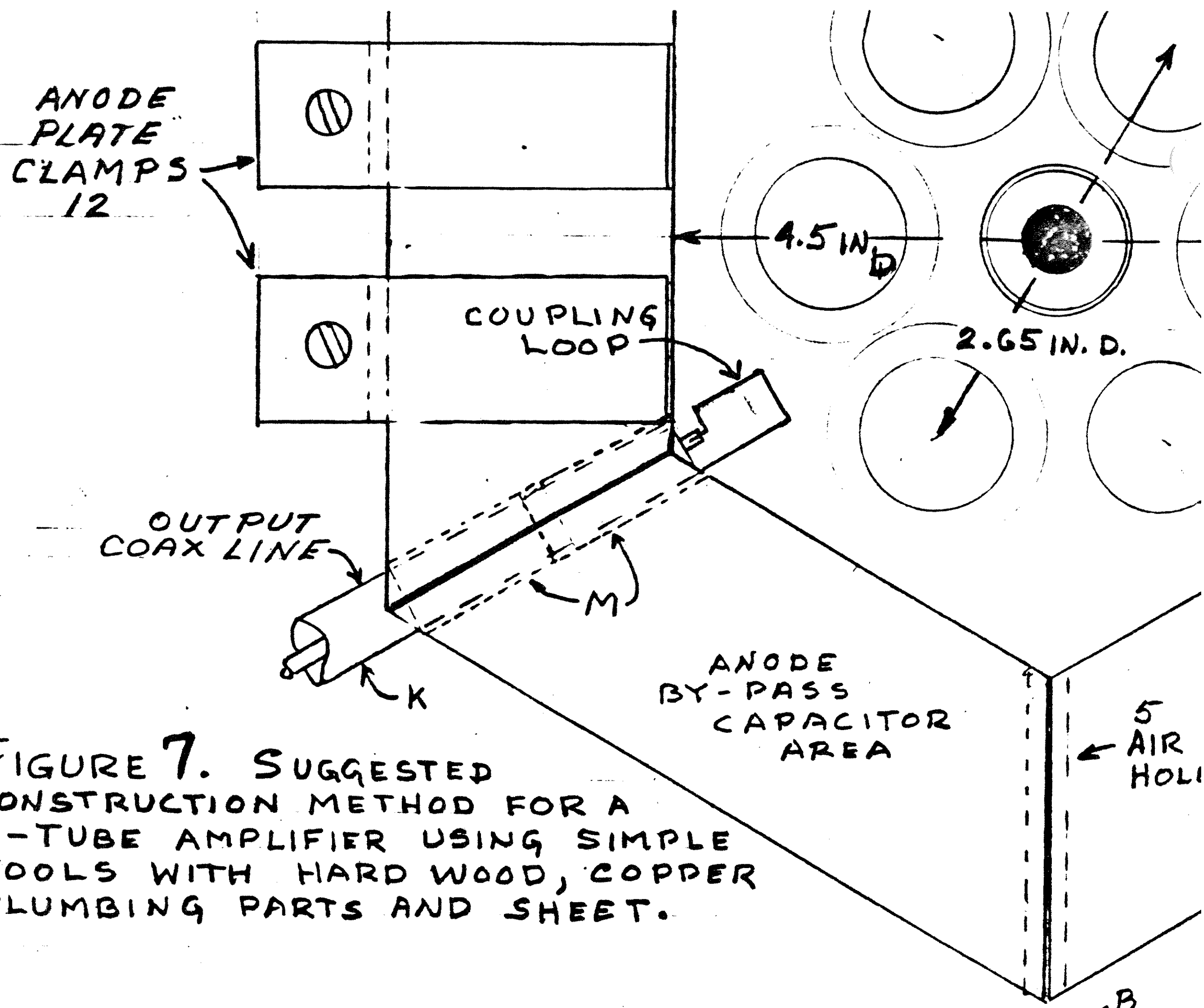
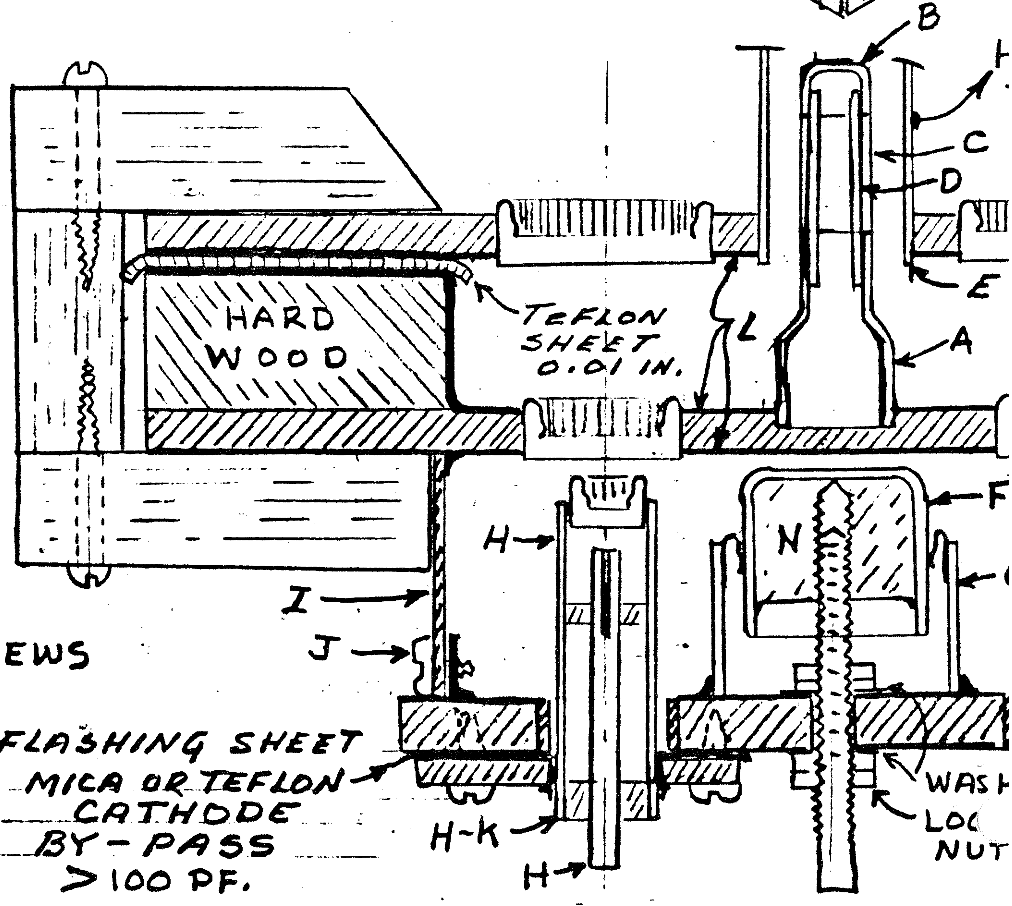
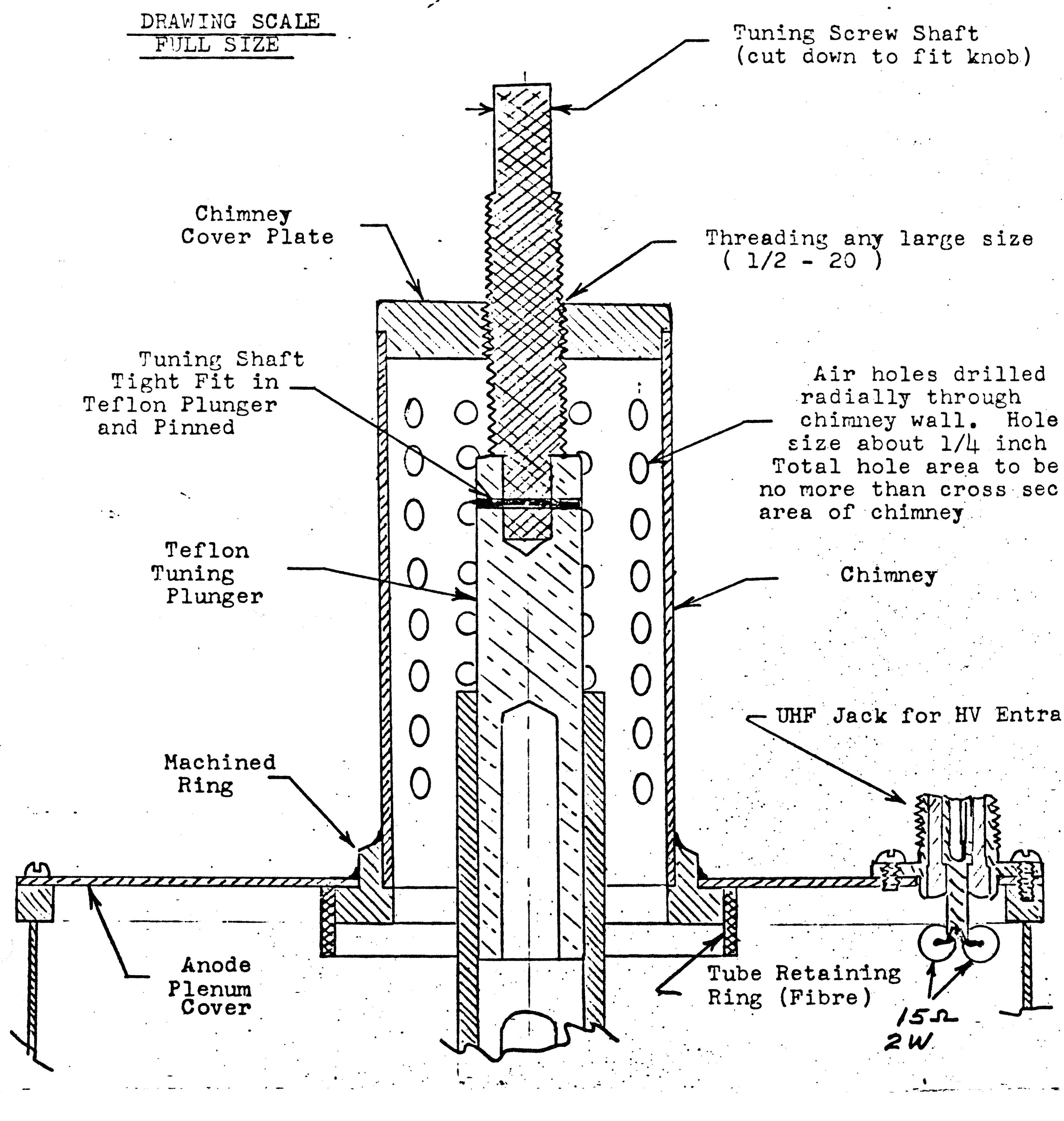


FIGURE 7. SUGGESTED CONSTRUCTION METHOD FOR A 6-TUBE AMPLIFIER USING SIMPLE TOOLS WITH HARD WOOD, COPPER PLUMBING PARTS AND SHEET.

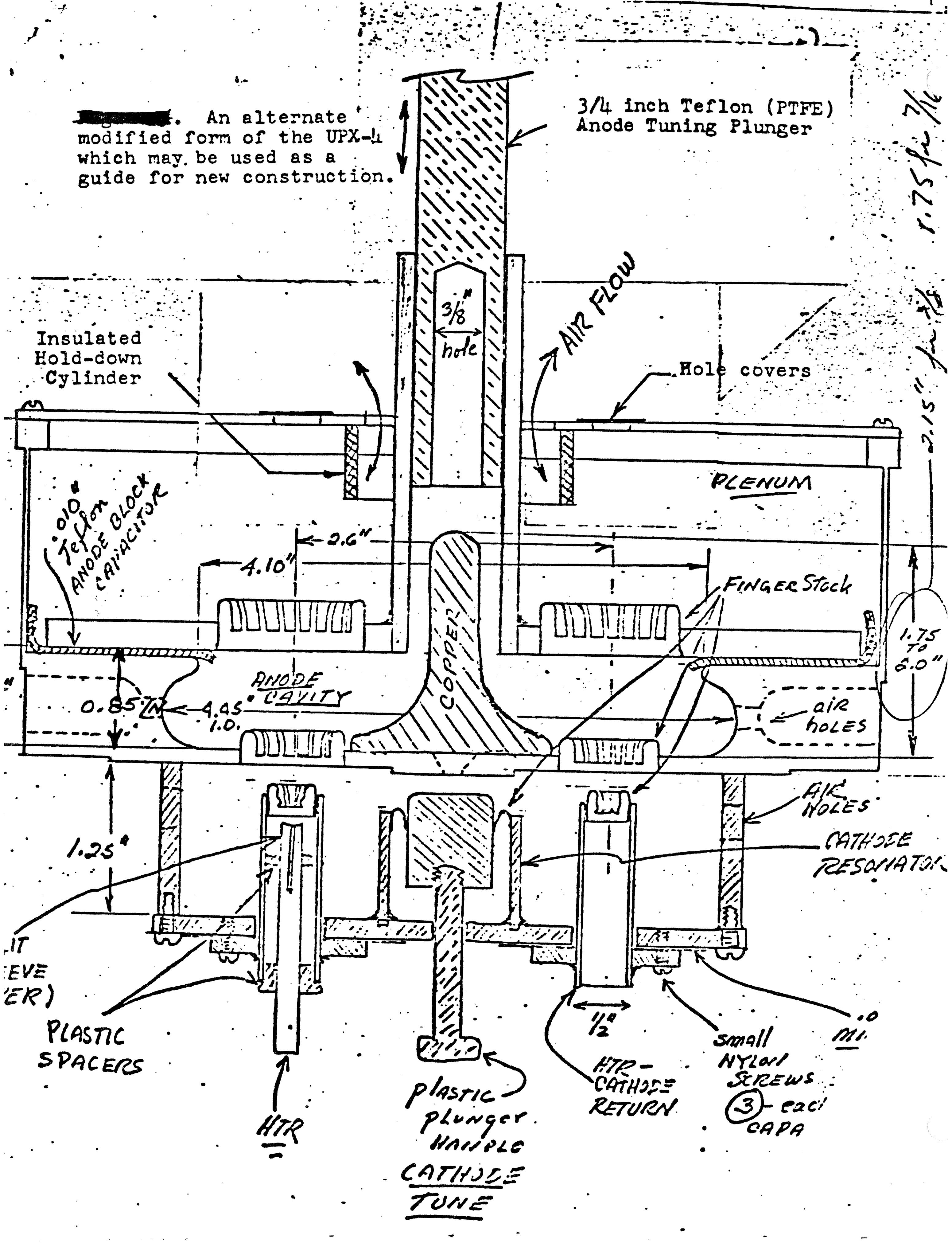


LEGEND
COPPER PARTS
A - 1/2→5/16 REDUCER
B - 5/16 END CAP
C - 5/16 SLEEVE
D - 5/16 NIPPLE
E - 3/4 PIPE
F - 3/4 END CAP
G - 1 IN. SLEEVE
H - 1/2 PIPE
I - 1/16 PC BOARD
J - SELF TAP SCREWS
K - 1/2 PIPE
L - 0.02 OR LESS FLASHING SHEET
M - 1/2 SLEEVE
N - PLASTIC

[redacted]. Modifications to the UPX-4 1296 mc/s power amplifier described in Tech Report #13 which eliminate d-c and r-f hazards as well as improve anode tuning control.



[redacted]. An alternate modified form of the UPX-4 which may be used as a guide for new construction.



W2IMU
12-86

WATER JACKET FOR 7289 TUBES

Here's a simple and inexpensive water jacket design which can be used with any of the 3CX100A5/7289 type tubes. All parts are readily available copper water pipe fittings. The 3/4 inch sleeve O.D. is an easy fit into the anode cup of the tube. The parts are cut, drilled and cleaned with sandpaper or emery cloth, and sweated together. A simple clamp jig or bench vise may be used to hold the assembly together by the 1/4 inch stems while soldering with a torch.

Before attaching the jacket to the tube cup, the most vulnerable part of the tube to electrolysis, the pinch-off stem should be coated with a hard-drying water-proof heat-resistant material such as gasket forming liquids found in auto supply stores. Even acrylics such as nail polish would be adequate. Apply several coats especially around the pinched edge at the top. Do not coat the anode stem, or threaded shank if your tubes have a threaded stem.

To attach the jacket to the tube, coat the lower outside area of the jacket with a layer of bath tub sealer. Insert the jacket into the cup while twisting to get better distribution of the sealer, and finish by wiping the lip of the anode cup with your finger to obtain a smooth filet with no voids. Place the assembly in an upright position, and allow overnight curing.

This jacket arrangement will be water tight and more than adequate in strength for the low pressure water system needed.

Electrical bonding between tube anode and water jacket is unimportant here.

The jackets may be removed from dead tubes and reused by simply heating the assembly slowly with a torch until the sealer softens and the jacket separates from the tube. The heat required should not melt solder in the jacket.

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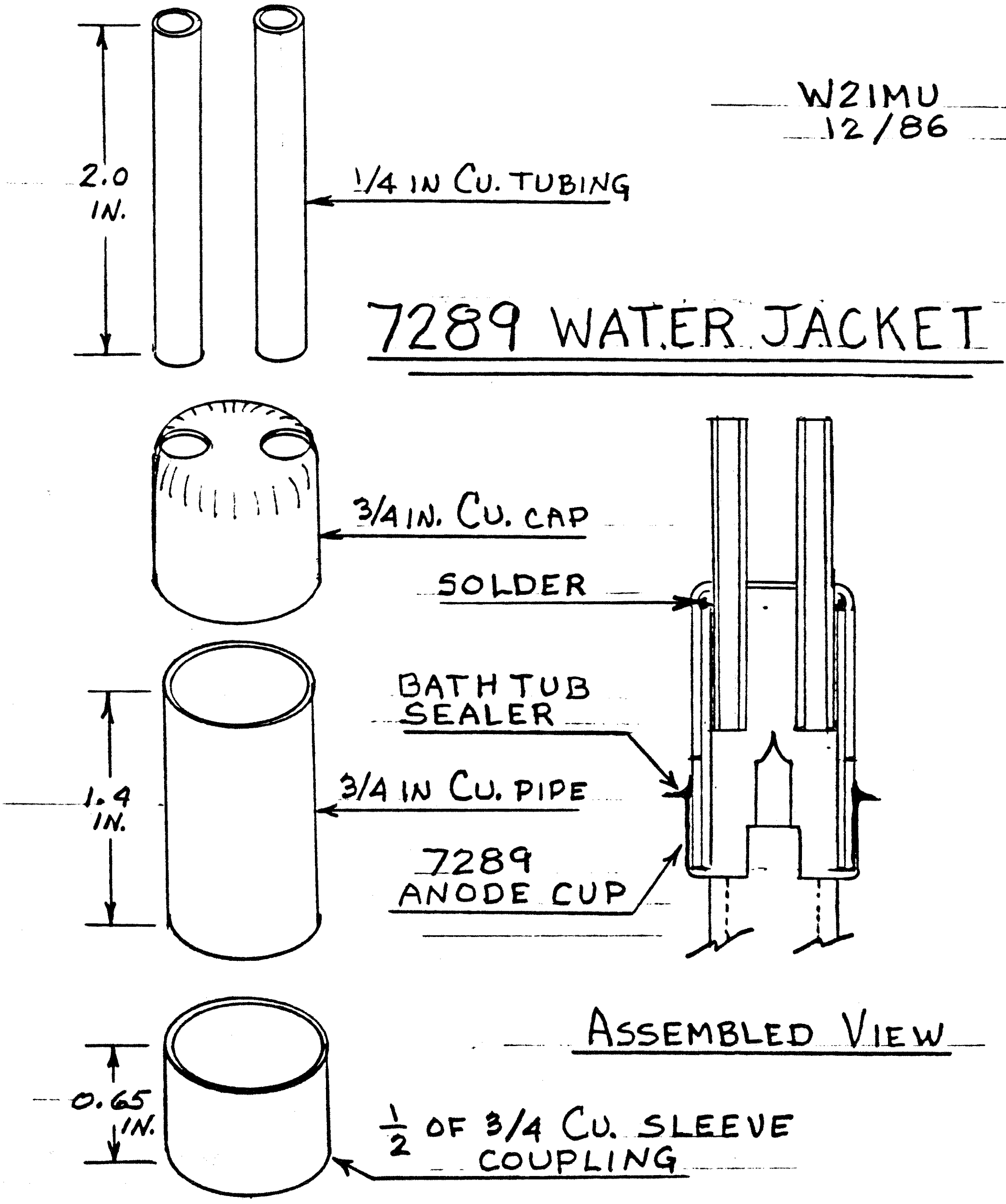
7289 WATER JACKET



EXPLODED VIEW

TECHNICAL REPORT # 14

From: The Crawford Hill VHF Club, W2NFA

Date: August 1973

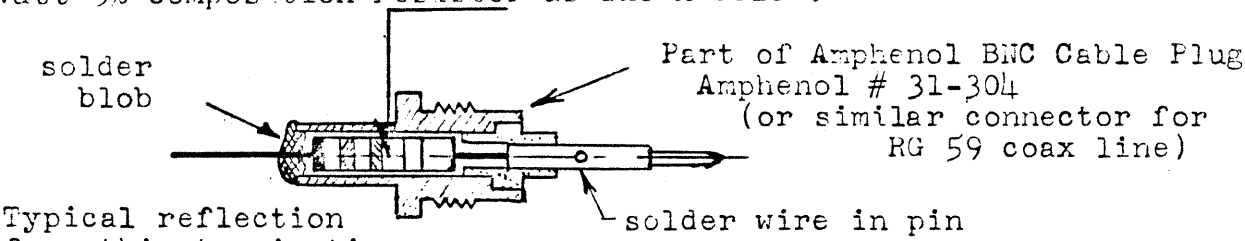
Subject: A HIGH POWER DIRECTIONAL COUPLER AND POWER MONITOR

A necessary adjunct in the EME station is a high power in-line power monitor which may also be used as a directional coupler to measure reflected power from the antenna. Two 50 ohm designs are presented in this report which are relatively easy to construct from readily available materials. The basic design is for a nominal coupling coefficient of -40 db which is useful for measuring high power levels with low power measuring equipment. For example 500 watts of available power will produce 0.05 watts in the -40 db coupled port. The coupling coefficient may be altered in this design by as much as ±20 db without violating the basic design.

The coupler consists of two 50 ohm unbalanced transmission lines weakly coupled. One line called the main line can handle hundreds of watts of power while the other line called the coupled line is only required to handle less than one watt. The couplers are built as 4-port networks with all 4 ports available. This means that for proper use external termination and detector must be used. The reason for not including the termination and detector on the coupled line is to facilitate accurate calibration at low power levels and to increase its usefulness as a measuring tool.

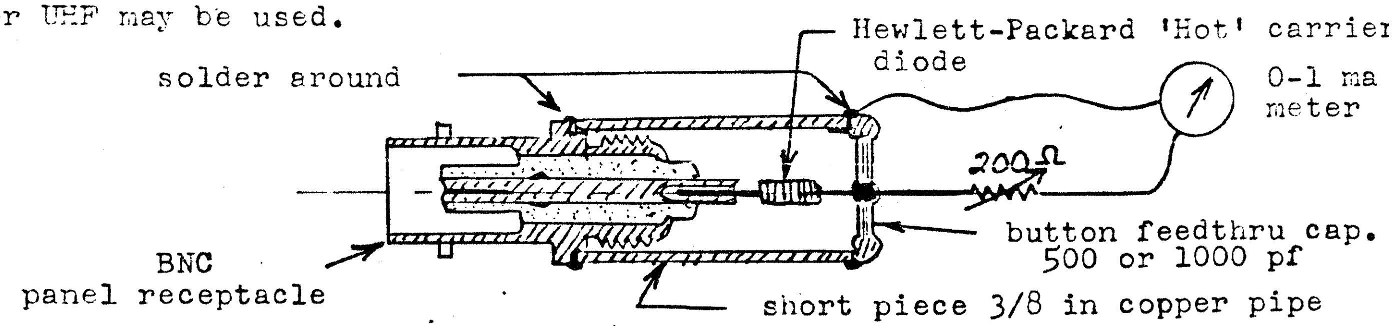
Other uses include measuring reflected power on feed lines, feeds and other devices and also to monitor the reflection coefficient while adjusting a device such as an antenna for best 50 ohm impedance match. These measurements can be performed at very low power levels using a well shielded receiver as the detector. A shielded r-f source is also advisable especially when adjusting antennas as direct r-f leakage can cause considerable error in readings.

A suitable low power termination for the 4th or null port of the coupler can be made from a specific BNC cable connector and a 47 ohm 1/2 watt 5% composition resistor as shown below.



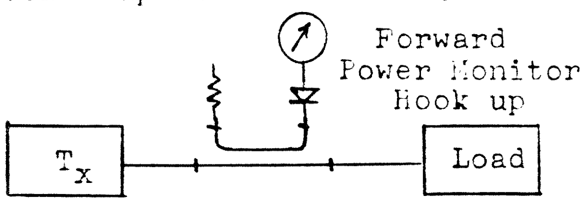
Typical reflection from this termination is -25 db at 1296 mc/s

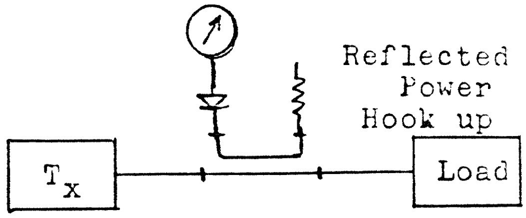
A suitable crystal detector for high power monitor may also be constructed as shown below, however any available detector suitable for UHF may be used.



Using the Coupler

As a power monitor the coupler may be installed at the output of a power amplifier with extremely little loss. This arrangement will permit continual monitor of output power. The coupler is hooked up as shown below.





To check reflected power, simply reverse the connections to the coupled line. In the designs described later, one coupler has a rotatable coupling loop while the other may be built with two coupled lines for continuous monitoring of power flow in both directions. The same hook ups apply to low power measurements where a receiver is substituted for the crystal detector.

Calibration

This coupler is not self calibrating or are the dimensions accurate enough to predetermine the coupling coefficient. In lieu of calibrated laboratory equipment properly used, the EME enthusiast can perform the coupling coefficient calibration in a variety of ways. The accuracy of these ways depends largely on how carefully the measurements are performed and what confidence can be attached to the equipment used. All methods must depend on a calibrated attenuator (±0.5 db accuracy) whether it be at UHF, IF or AF. Accurate attenuation measurement over a wide dynamic range (40 db in this case) are subject to non-linearities (saturation limiting) in the equipment as well as impedance mismatch in the circuits at either end of the calibrated attenuator. The experimenter is advised to acquaint himself with these problems of measurement before attempting accurate calibration.

One method which uses readily available equipment will be described. Most V-O-M meters are equipped with a decibel scale for AF measurement. This method will use the V-O-M as a calibrated decibel indicator in lieu of a calibrated attenuator. Particular attention must be given to saturation of equipment with this method since the station receiver will be used. The receiver must be operated in a linear mode, i.e. AVC-OFF, BFO-ON, Wide bandwidth, AF and RF gain settings for minimum saturation. A shielded low power source at 1296 mc/s is required which includes a crude adjustable output level, a variable link or adjustable capacitor output coupling will suffice.

Since the V-O-M scale is accurate only over a 10 db range it will be necessary to cover the 40 db range in four 10 db steps. In this way the receiver is only required to be linear over a 10 db range.

The calibration procedure then goes as follows:

1. Connect an adjustable low level signal source (1296 mc/s) to the coupler main line.

2. Connect receiver to other end of main line with some matched attenuator ahead of the receiver input. Twenty feet of RG58 cable will suffice as a matched attenuator.

3. Turn OFF the source and adjust the receiver gain controls to give noise output reading of -10 db on V-O-M scale. Impedance matching may be required to get optimum transfer from receiver to V-O-M.

4. Turn ON source and set its level so receiver output is near 0 db on V-O-M scale. This assures that S/N is high enough not to cause error from noise.

5. Next reduce receiver RF gain to give output reading of -10 db on V-O-M. This is the initial reference point for calibration.

6. Note very carefully the receiver gain control settings and also how sensitive the RF gain control rotation is on output level. If it is very hard to re-set the gain for output level within ±0.5db then it will be necessary to remove the small knob and replace with a large knob preferably having an indexed scale. Or mark the knob so that resetting can be done within ±0.5 db limits. If the gain controls are erratic replace or abandon receiver. It is also wise to let the equipment run at the reference level for a few minutes to see how steady the output level remains. If the level wanders more than ±0.5 db, more drastic stabilization may be required of the source, receiver or both before measurements can be made. Frequency as well as amplitude stability are very important.

7. Starting from the reference gain setting and output reading of -10 db, increase the source level until the receiver output is at 0 db.

8. Turn down RF gain to give output of -10 db

9. Increase source level until receiver output is again 0 db.

10. Repeat steps 8 and 9, 4 times which should increase the signal source level by exactly 40 db ± small errors.

11. After the last increase of source level to give receiver output level of 0 db, disconnect receiver and cable attenuator together from the main line and connect cable to coupled line port. Put good termination on main line where receiver was connected and also another termination on the other end of the coupled line. This latter termination should be kept as a part of the coupler and only this termination used for subsequent measurements with the coupler.

12. With these connections completed the receiver gain control should now be re-set exactly to the reference position. The receiver output may be above or below the 0 db reading.

13. Now carefully adjust the coupling by moving the coupled line closer for more coupling or farther away from the main line for less coupling until the 0 db output reading is again obtained.

Repeat the entire procedure several times for confidence as well as practice in the procedure. The results each time should check within ±0.5 db. Finally, clamp the coupled line or set it securely so that subsequent checks will always give -40db ±0.5 db.

The coupler is now calibrated and may be used for absolute power measurement in conjunction with calibrated low power measuring equipment. If the expected level at the coupled port is higher than the limit of the low power meter, the coupling may be reduced and recalibrated accordingly. If the coupler is only used for VSWR (reflection) measurements, the calibration procedure is not required. Construction as per the drawings will give a coupling coefficient of -40 ± 5 db, which is quite adequate for reflection measurements.

Strip Line Design

This design employs flat material in an easily built strip-line configuration. Figure 1 shows all dimensional details. The main line section is in symmetric strip-line with center conductor a copper strip 0.055 inches thick, nominally 1/16 inch sheet material. The length of the line section shown as 3.5 inches is actually an electrical half-wave long at 1296 mc/s, foreshortened by end loading due to the transition from strip to coax.

The coupled line section is built directly on the BNC receptacles which are screwed into the brass side wall (3/8-32 thread). The mounting holes for this side wall should not be drilled, if calibration is desired, until after the coupling has been set. The 45° mitered corners in the coupled line are characteristic of strip-line technology to preserve impedance around a 90° bend.

The bars which comprise the end walls where the N-type receptacles are located should be assembled first and then top and bottom plates fitted. The main line holes should be drilled undersized (3/8 in) and then taper reamed from the outside until the N-receptacle shoulder fits snugly in to the reamed hole. Retaining screws may now be installed.

Top and bottom plates may be made of copper clad board (copper on inside) if the coupler is not subjected to large physical forces. Squeezing the plates can deform the surfaces enough to cause significant error in coupler calibration (1 db or more.).

Aluminum may be substituted for the brass, in which case all parts should be aluminum to avoid electrolysis.

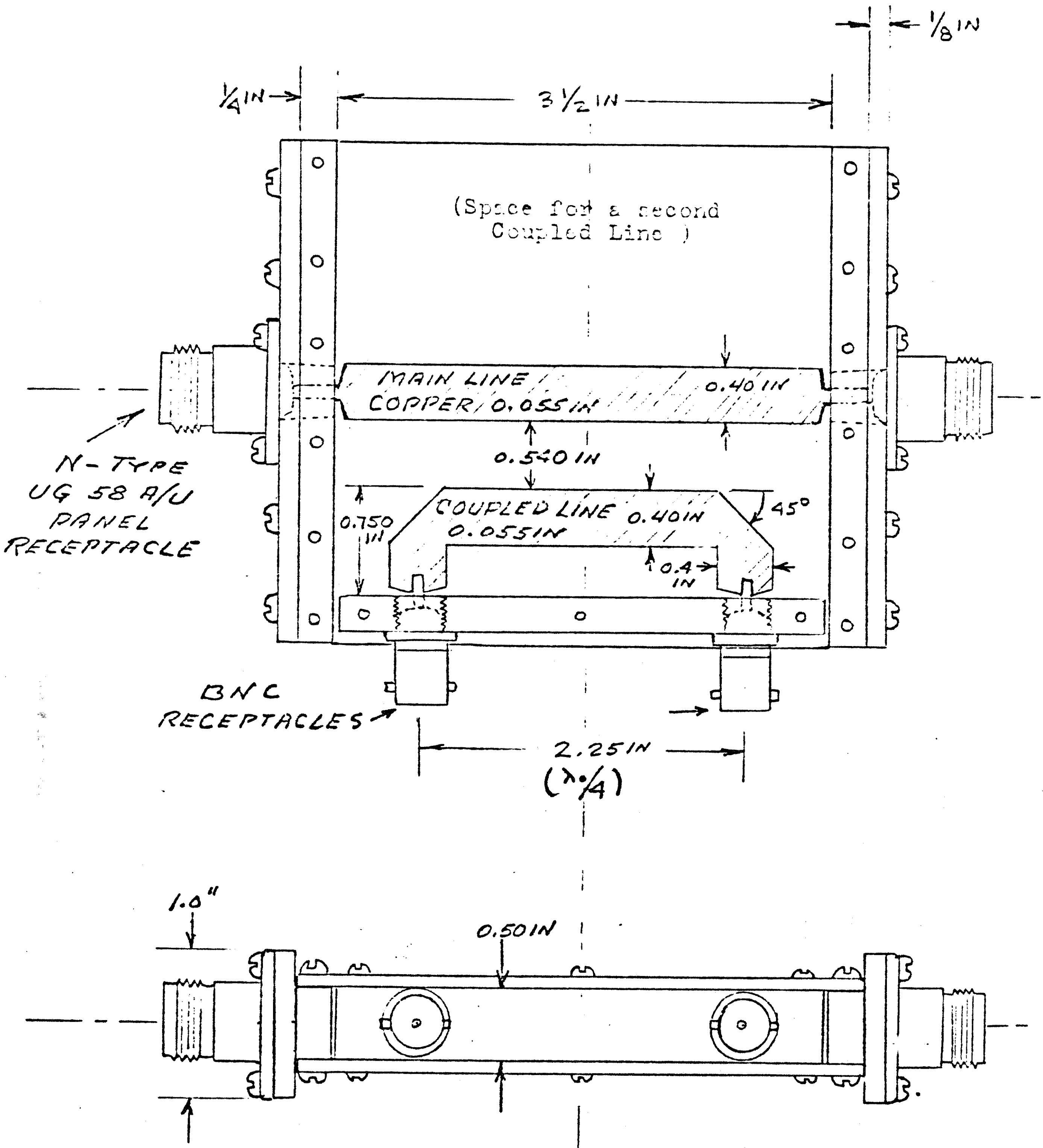
Coax Design

This design shown by Figure 2 is constructed of standard readily available copper water pipe and a 'T' fitting. These are available in most discount department stores, hardware stores and plumbing houses. The N-type connectors are actually panel receptacles (UG 58 A/U) which have been carefully machined down to be a press fit into the copper pipe. The center jack is removed from one of the connectors to allow assembly of the coupler. Simply drive it out with a flat ended rod and hammer in the direction toward the soldered end. The jack is soldered to the main line center conductor separately so that the coupler may be assembled starting from the opposite end.

The main line section is essentially 50 ohm. The Teflon bead at the center of the 'T' is to compensate for the inductive discontinuity of the 'T' arm missing wall. The dielectric bead adds shunt capacitance to tune out the inductive reactance.

The coupled line is actually a short piece of RG 58/U fitted inside of the short soft copper tubes. The coupled region is a very short exposed section of the inner conductor and dielectric. The outer vinyl jacket is stripped away and the coupled section is exposed by carefully pressing the woven wire outer conductor back away from the coupled region without breaking the wires. This may be done after the BNC connectors and line are installed. The coupled assembly is built into a short section of copper pipe with the two bent tubes sweated (soldered) inside the pipe diametrically opposite each other. Install the RG58/U last to avoid overheating of the polyethelene. The BNC connectos may be any of a variety which preserve the impedance integrity (50 ohms). The plane of the coupled line should be parallel with the main line for correct operation as a directional coupler.

The main line assembly does not require soldering if the parts fit together snugly.



HIGH POWER DIRECTIONAL COUPLER 50 ohms

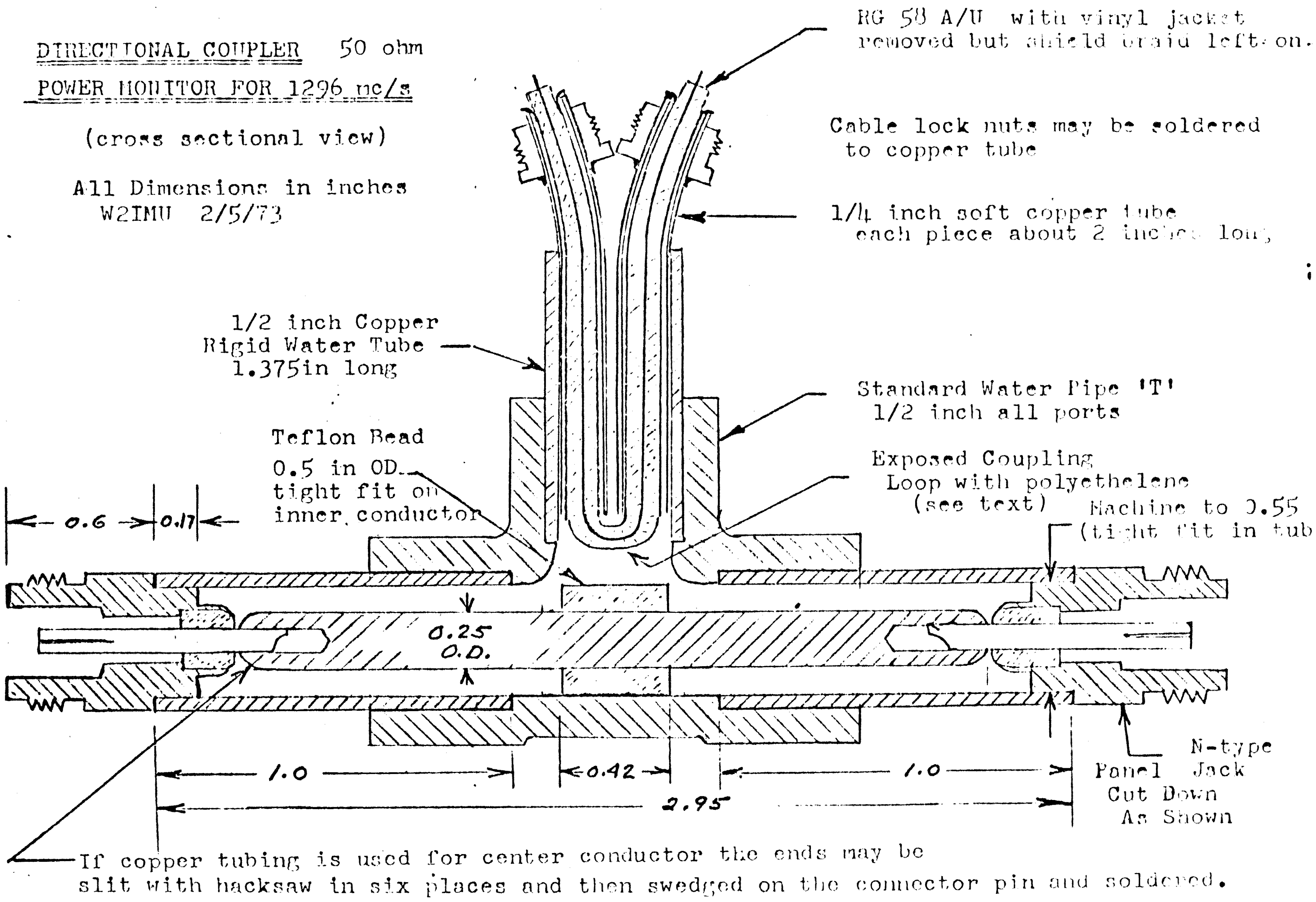
POWER MONITOR FOR 1296 mc/s

DIRECTIONAL COUPLER 50 ohm

POWER MONITOR FOR 1296 mc/s

(cross sectional view)

All Dimensions in inches
W2IMU 2/5/73



TECHNICAL REPORT # 15

From: The Crawford Hill VHF Club
Date: November 1986

Subject: <u>Spin Casting a Paraboloidal Reflector</u>

Introduction

The fabrication of an accurate paraboloidal reflector depends primarily on its required size and application. For EME, the operating environment must also be a considered factor. Wind loading and ice loading are important factors which effect the size of the reflector and type of surface material used. Large reflectors, 10 feet (3 meters) or more in diameter almost always requires some form of trussed rib arrangement for high strength-to-weight ratio. The addition of circular members will enhance the surface accuracy.
Smaller reflectors can be made using a reverse mold technique in which a reverse mold is constructed of some stable material which can be worked into a paraboloidal shape with a template and pivot axle. Such molds can be made of cement or plaster but require surface finishing to achieve accuracy. The actual working reflector is then molded over the surface in the form of a thin layer of Fibreglas or other durable material, with a suitable parting agent coating the mold. Some form of rib or honeycomb back-up structure is usually required for strength. The inner surface (reflector surface) must then be coated with a conductive material such as Silver paint or aluminum foil. This method can result in good surface accuracy if the mold has been prepared accurately.

This report describes an alternative mathod of achieving a very high accuracy paraboloidal reflector with very high surface accuracy. The method is called <u>spin casting</u> and has been around for a long time. While primarily used to fabricate small reflectors (even optical reflectors), it has been used to make reflectors up to 14 feet in diameter !
Surface accuracy considerations have been discussed in Report # 5 and should be consulted to realize that it is a factor which is difficult to accurately determine on a self fabricated reflector, and that it is a serious problem in fabricating high efficiency antennas. This problem becomes accute as higher frequency bands are used both for amateur EME and terrestrial microwave communications.

Spin Casting

The basic physics of spin casting a paraboloid is that when a container of liquid is rotated about its vertical axis, gravity and centrifugal forces will act on the liquid in such a way as to form a perfect paraboloid at the surface of the liquid **. This remarkable property can be employed to form a high accuracy

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** Roberson, J.A. and Crowe, C.T., <u>Engineering Fluid Mechanics</u>, Houghton Mifflin Co., Boston, 1975, p. 119.

paraboloidal reflector by using a material which is initially liquid, but will harden in a reasonable period of time. One such material is acrylic casting resin available in any hobby shop or art supply house.

The relationship between the spinning speed and the parabolic shape is simply

$$\text{focal length in feet} = 1468.14/(\text{spin r.p.m.})^2,$$

regardless of the material used as long as it is a homogenous fluid during the spin forming period.

The spin speed must be held very constant during the time it takes the casting material to harden. The container may be any convenient circular shape, however in the interest of saving expensive casting resin and reducing weight, the container might be in the form of a crude paraboloid. In fact, this spin casting method can be readily used to restore the surface accuracy of a damaged dish.

The spinning mechanism should be vibration free and have some inertia to help keep the spinning smooth and constant. A spin speed monitor which can be used to control the drive motor is desirable although not essential.

Typically rotational speeds of less than 100 rpm will be required. For example, a 45 rpm old record player turntable will produce a focal length of 0.73 feet. To obtain a paraboloid with f/d of 0.5 will require a container about 18 inches in diameter.

When the spin casting is complete, the surface may be coated with a conductive material to obtain a radio reflector. Coating can be with Silver paint (sprayed on but very expensive), or copper or aluminum foil cemented to the surface with any good adhesive. Ordinary household aluminum cooking foil applied with a contact cement is entirely adequate and may be hand rolled to take out wrinkles, and slit to remove air pockets.

If the initial attempt at spin casting produces an unsatisfactory surface, for whatever reason, the entire operation may be repeated with a thin layer of the same material. It is usually desirable to start the spin forming process with the material at the bottom of the container (bowl) but not necessary. The material may be slowly poured onto the container surface from center to edge as long as the material is fluid during this process. Experience will dictate how fast to introduce material on the surface.

Although the acrylic casting resin (with hardner) is a very suitable material, it is expensive and is also exothermic. This means that if a large quantity is mixed with the hardening agent, allowance should be made for heat dissipation. Spread the material as soon as it is mixed thoroughly, and spread leftover material over a wide area to avoid possible fire and or explosion due to rapid heating.

An alternative material is non-shrinking plaster also available at art supply houses. Primarily used for statue molding, this material can be used for spin casting by making a slurry that has the consistency of heavy cream. It sets relatively slowly (about 5 to 10 minutes for a noticeable set) and may require up to an

hour or more to completely harden.
For a large paraboloid, plaster will be heavy and will have to be sealed to prevent weathering. It may also be advisable to place reinforcing rods or screening material within the casting to add strength.

Additional Remarks

Spin casting of reflectors for antennas finds its chief application at microwave frequencies of perhaps 4000 mc/s and higher where surface accuracy requirements are stringent and cannot be readily realized by other methods of home fabrication. Commercial spun dishes (not spin casting) can have good accuracy but are limited by the mold used to spin-form the maleable material into shape.
For amateur application, spin casting can be used for reflectors in the 2 to 4 foot diameter range at X-band (10,000 mc/s) and higher to achieve excellent surface accuracy and thus highest aperture efficiency from parabolic reflector antennas.
Larger diameter reflectors will be heavy and of course be of solid surface with the full impact of wind loading. Larger spin cast reflectors should be fabricated as a thin shell, perhaps 1/8 to 1/4 inch thick, and backed up with a trussed rib and hub assembly for strength.
Fine optical surfaces have been formed by spin casting but special precautions must be taken to obtain the required surface accuracy. These include a very stable spin mechanism, completely free of vibration. The casting resin must be carefully prepared to avoid contaminants and excessive air bubbles. Spinning must be done in a clean-air environment free of dust, lint and low in water vapor content. The reflecting surface may be coated with fine optical silvering used in telescope reflector surfacing.
Optical reflectors of a foot or so in diameter can be used at optical and infrared wavelengths to give fantastic gain. At these extremely short wavelengths, this size antenna will produce a virtually collimated beam over many miles and can provide long line-of-sight reliable transmission paths with very low transmitter power requirements. Such paths are ideal for high security links but easily wiped out by heavy weather conditions, fog, rain, sleet, snow, etc.

Date: January 1974
From: The Crawford Hill VHF Club, W2NFA

Subject: Libration Fading on the EME Path

One of the most troublesome aspects of receiving a moonbounce signal besides the enormous path loss and Faraday rotation fading, is libration fading. This report will deal only with libration fading its cause, effects and possible measures to minimize it.

Libration fading of an EME signal is characterized in general as a fluttery rapid irregular fading not unlike that observed in tropospheric scatter propagation. Fading can be very deep, 20 db or more, and the maximum fading rate will depend on the operating frequency. At 1296 mc/s the maximum fading rate is about 10 cycles per second and scales directly with frequency.

On a weak CW moonbounce signal, libration fading gives the impression of a randomly keyed signal. In fact on very slow CW telegraphy, the effect is as though the keying is being done at a much faster speed. On very weak signals only the peaks of libration fading are heard.in the form of occasional short bursts or 'pings'.

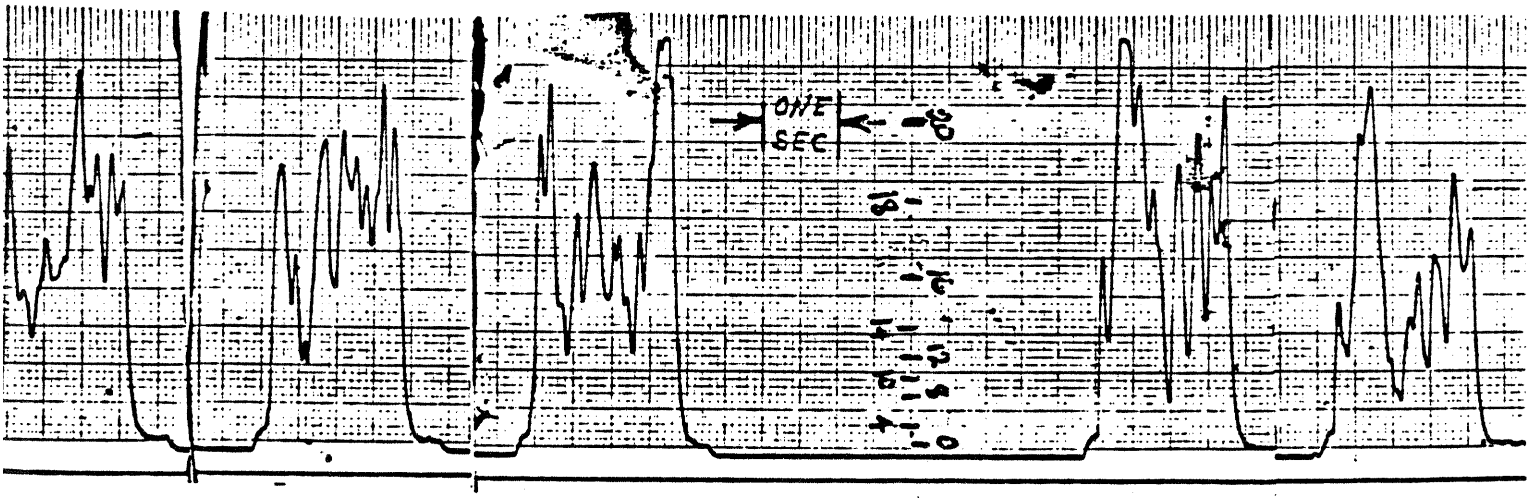


Figure 1. Moon echoes received at W2NFA July 26, 1973, 1630GMT. Antenna gain 44 db, transmitting power 400 watts and system temp. 400°K.

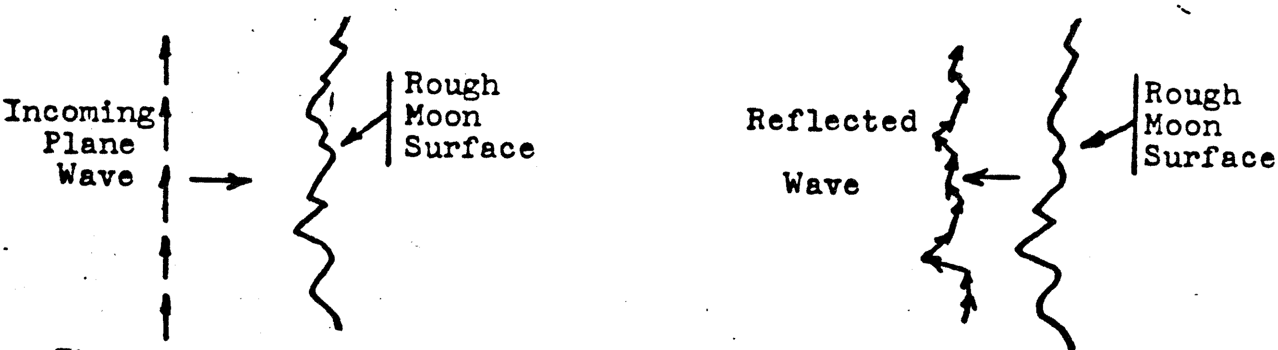
Figure 1 above shows samples of a typical EME echo signal at 1296 mc/s. These recordings made at W2NFA show the wild fading characteristics with sufficient S/N to record the deep fades. Circular polarization was used to eliminate Faraday fading, thus these recordings are of libration fading only. The recording bandwidth was limited to about 40 Hz to minimize the higher sideband frequency components of libration fading which persist but are much smaller in amplitude. For those who would like a better statistical description, libration fading is Raleigh distributed.

An interesting and useful aspect of libration fading is due to the mechanism which causes it. In the recordings shown by Figure 1, the average signal return level computed from pathloss and mean reflection coefficient of the Moon is at about the +15 db S/N level.

It is clear that enhancement of echoes far in excess of this average level are observed. This point should be kept clearly in mind when attempting to obtain echoes or receive EME signals with marginal equipment. The probability of hearing an occasional peak is quite good since occasional enhancements as much as 10 db is possible. Under these conditions however, the amount of useful information which can be copied will be near zero. The enthusiastic newcomer to EME communications will be stymied by this effect since he knows that he can hear the signal strong enough on peaks to copy but can't make any sense out of what he tries to copy.

What causes libration fading? Very simply it is due to multipath scattering of the radio waves from the very large (2000 mile diameter) and rough Moon surface combined with the relative motion between Earth and Moon called librations.

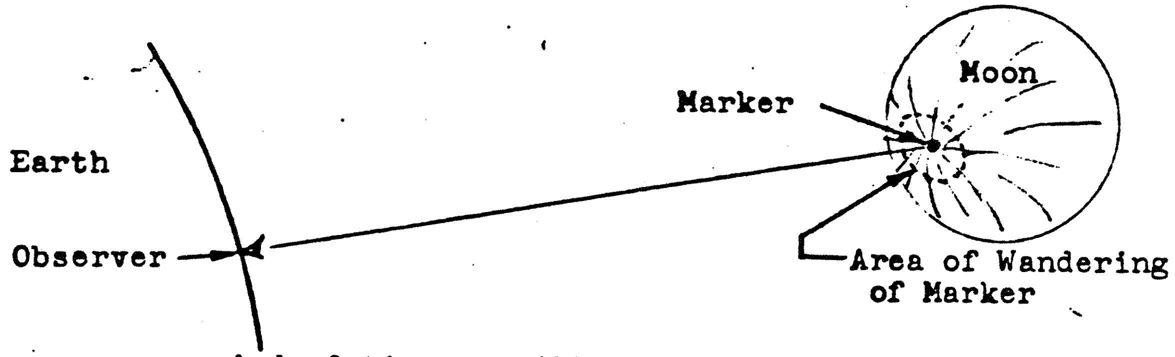
To understand these effects, consider first that the Earth and Moon are stationary (no libration) and that a plane wave front arrives at the Moon from your Earth bound station as shown below on the left.



The reflected wave in the right hand drawing consists of many scattered contributions from the rough Moon surface. It is perhaps easier to visualize the process as if the scattering were from many small individual flat mirrors on the Moon which reflect small portions (amplitudes) of the incident wave energy in different directions (paths) and with different path lengths (phase). Those paths directed toward Earth arrive at your antenna and appear as a collection of small wave fronts (field vectors) of various amplitudes and phases. The vector summation of all these coherent (same frequency) returned waves (and there are a near infinite array of them) takes place at the feed of your antenna (i.e. the collecting point in an antenna system). The level of the final summation as measured by a receiver can of course have any value from zero to some maximum. Remember now that we assumed the Earth and Moon are stationary which means that the final summation of these multipath signal returns from the Moon will be one fixed value. The condition of relative motion between Earth and Moon being zero is a rare event about which we will discuss later in this report.

Consider now that the Earth and Moon are moving relative to each other (as they are in nature) so that the incident radio wave "sees" a slightly different surface of the Moon from moment to moment. Since the Moon surface is very irregular, ~~so~~ the reflected wave will be equally irregular changing in amplitude and phase from moment to moment. The resultant continuous summation of the varying multipath signals at your antenna feed point produce the effect called libration fading of the Moon reflected signal.

The term libration is used to describe small perturbations in the movement of astro bodies. Earth libration consists mainly of its diurnal rotation while Moon libration consists mainly of its 28 day rotation which appears as a very slight rocking motion with respect to an observer on Earth. This rocking motion can be visualized by # considering placing a marker on the surface of the Moon at the center of the Moon disc which is the point closest to the observer, as shown below.



Then over a period of time we will observe that this marker wanders around within a small area. All this means is that the surface of the Moon as seen from the Earth is not quite fixed but changes slightly as different areas of the periphery are exposed due to this rocking motion. Moon libration is very slow (of the order of 10^{-7} radians per sec.) and can be determined with some difficulty from published Moon ephemeris tables.#

Although the libration motions are very small and slow, the large surface area of the Moon having so many scattering points (small areas) means that even these slight geometric movements can alter the total summation of the returned multipath echo by a significantly large amount. Since the librations of Earth and Moon are calculable, it is only logical to ask if there ever occurs a time when the total libration is zero or near zero? The answer is yes, and it has been observed and experimentally verified on radar echoes that minimum fading rate, not depth of fades, is coincident with minimum total libration. Calculation of minimum total libration is at best tedious and can only be done successfully by means of a digital computer. It is a problem in extrapolation of rates of change in coordinate motion and in small differences of large numbers.

At W2NFA, several libration fading minima on echoes have been observed, recorded samples of which are shown ~~below,~~ Figure 2. Comparison with Figure 1 shows the lack of rapid fading indicating that libration has slowed down considerably. Note that the fades are still quite the same in depth and enhancement. In general libration fading on echoes will be most severe at Moon zenith and will have minimas only in the regions near Moon rise or set, elevation angles of roughly 20 to 30 degrees but not with daily regularity. And most important of all, the minima are cusp like in occurance and are easily distinguishable for brief periods of approximately one half hour or less.

The American Ephemeris and Nautical Almanac, published annually by the U.S. Govt. Printing Office and obtainable from the Superintendent of Documents, Washington, D.C. 20402; or,
The Astronomical Ephemeris, published by Her Majesty's Stationery Office, 49 High Holburn, London W.C.1

These effects are for echo recordings and it is interesting to contemplate the effect for a given point-to-point path on the Earth which is of more immediate interest. This more useful case involves determining the simultaneous minima for both station locations on echoes. A problem which is more complicated and as yet has not been considered. It is conceivable that the point-to-point libration fade minima may rarely occur in general except for specific paths such as two stations located on the same meridian. Much computation for an extended period and variety of paths has to be undertaken before a clear picture of what to expect can be formulated.

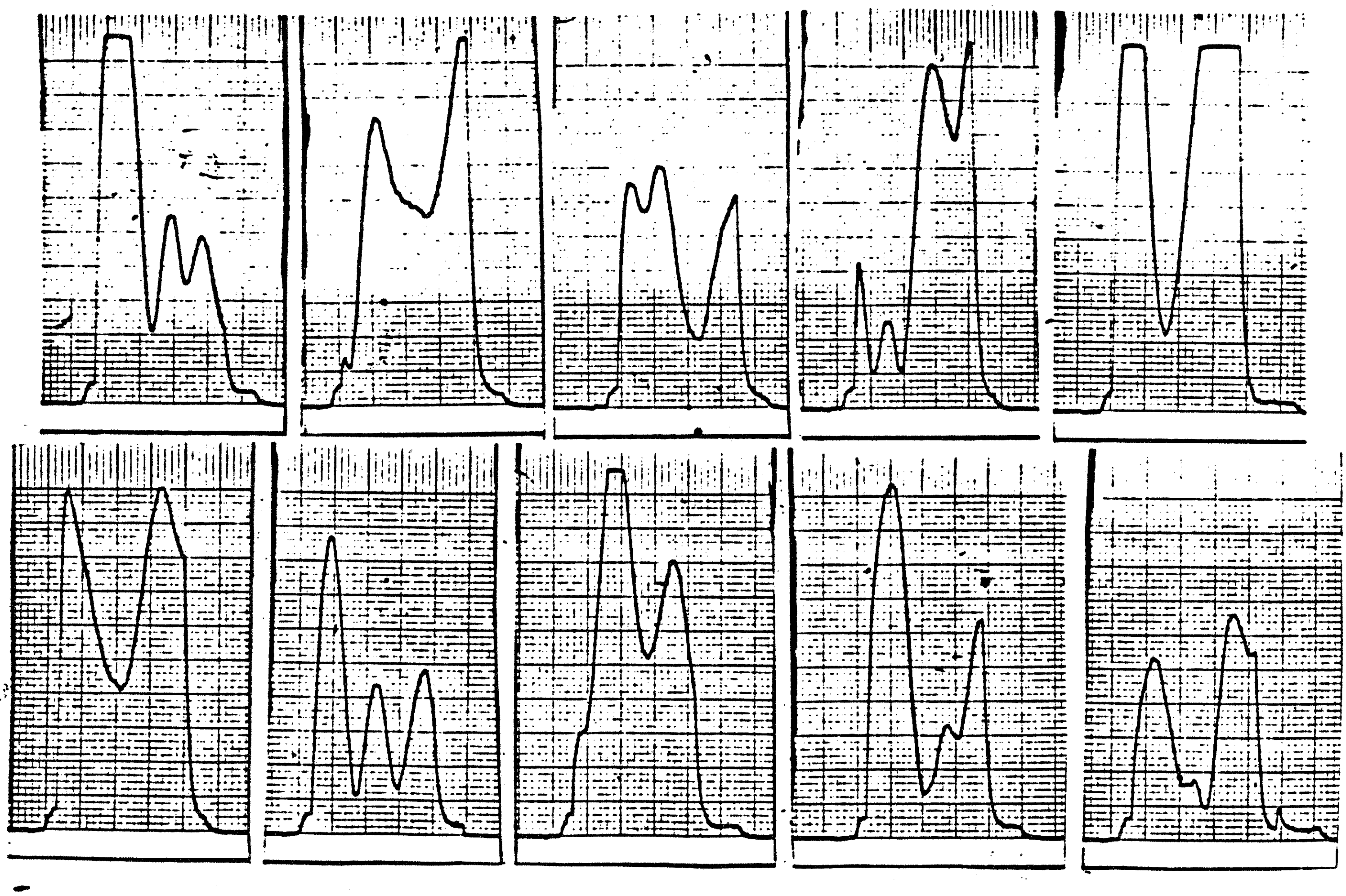
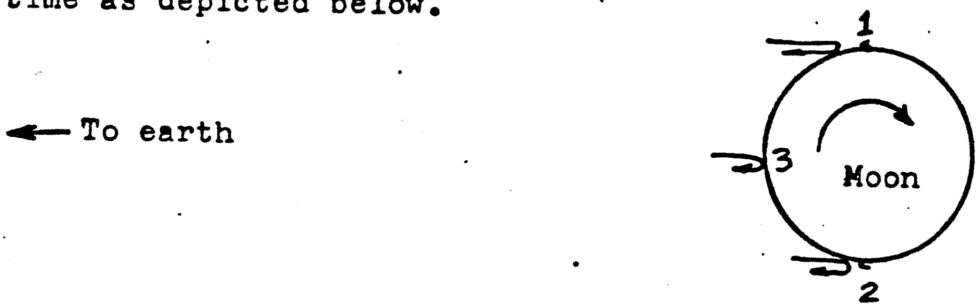


Figure 2. Echo samples of low libration taken at W2NFA at 1296 mc/s on July 28, 1973 1130 GMT, 30° elevation, 88° azimuth. Note much lower fading rate than in Figure 1 but equally deep fades.
The time scale and level calibration are the same as for Figure 1.

Further Considerations of Moon Libration

The simple concept of multipath scattering from the Moon must be extended for completeness to include libration spreading. The main component of libration motion is due to the earth's rotation and occurs when the moon is in its zenith position (directly over your station meridian). At this time the observer on earth is moving with considerable tangential velocity which may be translated to simple rotational motion of the moon.

Total libration motion of the moon and earth may therefore be considered as simple rotation of the moon over a small interval of time as depicted below.



Reflections occurring towards the limbs of the Moon designated above as points 1 and 2 have their radial distance to the earth changing in opposite directions due to the rotation of the Moon. Reflections from point 1 are increasing in distance to the earth while reflections from 2 are decreasing. From well known rules for Doppler frequency shift it is clear that the radio wave reflected from point 1 will be decreasing in carrier frequency while those from point 2 will be increasing in frequency. Reflections from the center of the Moon disc, point 3, will not be shifted in frequency because the radial distance to the earth is not changing.

Since reflections are occurring over essentially half the Moon surface (the half facing the earth), it is obvious that each reflection point (multipath scattering) will have a doppler shift of its own according to the change in distance to the earth with libration rotation. The grand total effect summed at your antenna terminals is a spreading or smearing of the incident CW signal into a carrier surrounded by symmetric sidebands which extend out to a maximum of

$$F = \pm \frac{2 L_t r_m}{\lambda} \quad \text{cycles per second.}$$

Here λ is the carrier wavelength, r_m is the moon radius in the same units as λ, (r_m = 2000 s. miles or about 10,000,000 feet) and L_t is the total maximum libration rate which is about 12×10^{-7} radians/sec. For a carrier frequency of 1296 mc/s, the maximum extent of the sidebands will be about ±20 cycles. At 144 mc/s it is about 2 cycles. The total energy in these Doppler spreading sidebands is small in the VHF-UHF range but increases at higher frequencies.

It should be made quite clear at this point that libration spreading caused by Doppler shift of the many reflections over the surface of the Moon is a small disturbance of the radio signal. The mean or average Doppler shift in carrier frequency caused by the rate of change in average range distance between earth and Moon is not considered in this report but can amount to as much as ± 4000 cycles/sec. at 1296 near Moonrise (+) and Moonset (-). This is the more familiar Doppler shift in frequency of a Moonbounce signal and sounds like a slowly drifting oscillator.

An overview of Moon reflection distribution with respect to frequency is useful to complete the libration fading and doppler spreading effects. When looking at a full Moon optically we see a bright uniformly illuminated disc. At much lower frequencies than light waves the picture is somewhat different. At 25,000 mc/s, the disc appears to be brighter at the center falling off in brightness by only 5 or 6 db at the edge. As we procede lower in frequency the central brightness area becomes smaller and the intensity falls off more rapidly towards the edge. At 1296 mc/s, the central area of most energy reflection (bright spot) comprises less than 1/3 the total disc area and the reflected energy falls off by 30 db or more towards the edge. At 144 mc/s the bright central area is even smaller and the edge of the Moon disc is all but invisible. Curiously, the average reflection coefficient of the Moon remains virtually constant over the entire VHF-UHF range at about 6.5%.

This distribution with frequency of the reflection characteristics of the Moon has been experimentally verified and is important to keep in mind as we consider further aspects of libration fading and doppler spreading.

Coherence Bandwidth

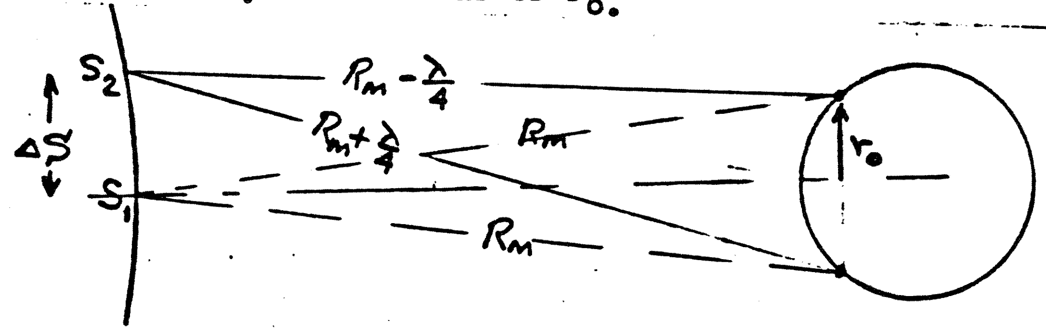
The total effect of libration fading and doppler spreading can be summed up in the following hypothetical experiment. Suppose that we transmit an AM signal to the Moon. ~~And~~ Suppose further that the carrier is modulated by a single tone whose frequency can be varied from about 100 Hz to 10,000 Hz. For low modulation frequencies the return signal can be demodulated and, except for the fading, will demodulate normally and be the same as the original signal. As the modulation frequency is increased, ~~we notice a curious result.~~ the demodulated signal ~~is~~ becomes distorted and appears to have 'selective fading'. The fact is that the two sidebands of the returned signal (an AM signal consists of a carrier with symmetric sidebands, two sidebands for a single tone modulation) have undergone different phase and amplitude changes, i.e. they are becoming incoherent.

The coherence bandwidth can then be defined as the frequency separation at which some degree of acceptable coherence still exists. At 1296 mc/s, the AM sidebands start becoming incoherent for a modulation frequency of a few kilohertz. This effect essentially scales inversely with frequency so that at 144 mc/s the sidebands will be coherent out to about 10 kHz. For these reasons AM and FM and other forms of modulation requiring coherent double sidebands are not recommended for voice or wideband moonbounce communications, especially at frequencies above about 1000 mc/s. And for the same reasons, SSBSC is recommended for all voice communication via the Moon .

Space Coherence

One possible method of reducing libration fading is by means of diversity reception where two spaced antennas are used and the receiver outputs are combined on a power basis. To implement such a system, the first consideration is how far apart ~~must~~ the antennas must be spaced so that the signals are totally uncorrelated. By power combining of these incoherent signals, a substantial reduction in deep fades will be achieved resulting in a more constant average signal level.

An estimate of the spacing of the two antennas can be made by a simple geometric analysis of the radio wave path lengths on a radar basis. The geometry is shown below for a single pair of reflection points at an arbitrary disc radius of r_o.



This geometry permits a simple formula to be derived which gives the distance separation between S_1 and S_2 such that the radio signals are in phase at S_1 and out of phase by 180^o at S_2. The formula is

$$\Delta S = \frac{\lambda \; R_m}{4 \; r_o}$$

, where R_m, the distance to the Moon, is approximately 250,000 statute miles, λ is the operating wavelength, and r_o is the disc radius in the same units as R_m. r_o may be any value from zero to a maximum of 1000 miles. It is obvious then that the space coherence distance ΔS can have a value of infinity down to some minimum where r_o = 1000 miles. Recalling the discussion of energy distribution reflected from the Moon, it is possible to make a reasonable estimate for r_o such that a reflection point in a region of high energy reflection is chosen. At 1296 mc/s for example, a value for r_o would be 100 miles , for which ΔS computes to be about 470 feet!

What about the energy from within the 100 mile radius? It is large and indicates that a much greater spacing is demanded. The choice of r_o = 100 miles was made partly in consideration of the following analysis.

Inspection of Figure 1 indicates that fading rates of about 3 Hz are easily observed along with lower frequency components. At Moon zenith, the major contribution of libration is the rotation of the Earth. A short calculation will reveal that an observer standing at one point near the Earth's equator will actually be moving with a tangential velocity of about 1500 feet/second. Since the coherence distance, ΔS, computed above is only half a space cycle it becomes clear that for reflectors at r_o = 100 miles, the fading rate as observed on Earth will be of the order 3Hz!! This means that for a diversity antenna spacing of 470 feet all those fading components above 3Hz will be uncorrelated. Those below 3 Hz <u>will</u> be correlated and hence cannot be reduced at this spacing. Since these quantities are linearly related a spacing of about 1000 feet will be required to reduce fading rates in the region of 1 Hz by diversity reception.

It should be kept in mind that each antenna must be large enough to receive a signal with reasonable S/N since the addition of the second antenna will not substantially increase the average power level but will only minimize the deep fading. If one attempts to add the signals coherently from the antennas, the effective gain of the system will double (if both antennas have the same gain) but the fading ~~will~~ will be essentially the same as with a single antenna.

Another method of minimizing libration fading is to utilize the coherence bandwidth property in the form of a frequency diversity system. In its simplest form, one would transmit two signals spaced in frequency by greater than the coherence bandwidth and of nearly equal power through one antenna. To instrument this type of signal a double sideband suppressed carrier , DSBSC, type signal can be produced using a balanced modulator even at high power levels. The modulation frquency should be at least 5 to 10 kc/s at 1296 mc/s. Although the sidebands will be coherent in this type DSBSC signal ~~however~~, this property is not required.

An appropriate receiving system would be the usual front end converter down to some IF where selectivity can be achieved. At this point in the receiving system the signal is split and filtered. The two signals are separately detected and the outputs are combined on a simple rms basis through a resistive network. For CW operation the final beat note output of the detectors should be adjusted for the same pitch.

Frequency diversity is of course much easier to implement than antenna space diversity; however, ~~[illegible]~~ neither system has been employed in amateur moonbounce work to this authors knowledge.

Polarization Scattering

Another effect of scattering from the rough surface of the Moon is depolarization. This effect, which occurs for all polarizations including circular and linear, simply means that some of the incident radio wave energy in one polarization is scattered into an orthogonal or cross polarized wave energy. Specifically, if the EME transmitting antenna is linearly polarized and oriented say vertically, most of the reflected EME energy will be vertically polarized (neglecting Faraday rotation of course) but a small amount of reflected energy will be horizontally polarized (cross polarized).

At 1296 mc/s the average cross polarized energy will be about 15 db lower than the main or parallel polarized energy. At lower frequencies scattering is less and a smaller amount of energy will appear in the cross polarized component. The ratio of these levels (cross polarized to parallel polarized) varies considerably with time also. It has been observed on 432 mc/s EME (K2UYH) with linear polarization that the orientation (rotation) of the dipole feed in a parabolic reflector antenna system can be used to maximize signals while at other times the orientation makes little difference. Limited experience at W2NFA on 1296 mc/s EME with circular polarization indicates a difference of 10 to 15 db between right and left circular polarization on echoes and other signals which were known to be circularly polarized.

When depolarization is small, of the order -10 db or less, the loss in signal level for the desired polarization is small. However, when considerable depolarization occurs (such as observed on 432 mc/s EME) the loss in signal level can be as much as 3 db !

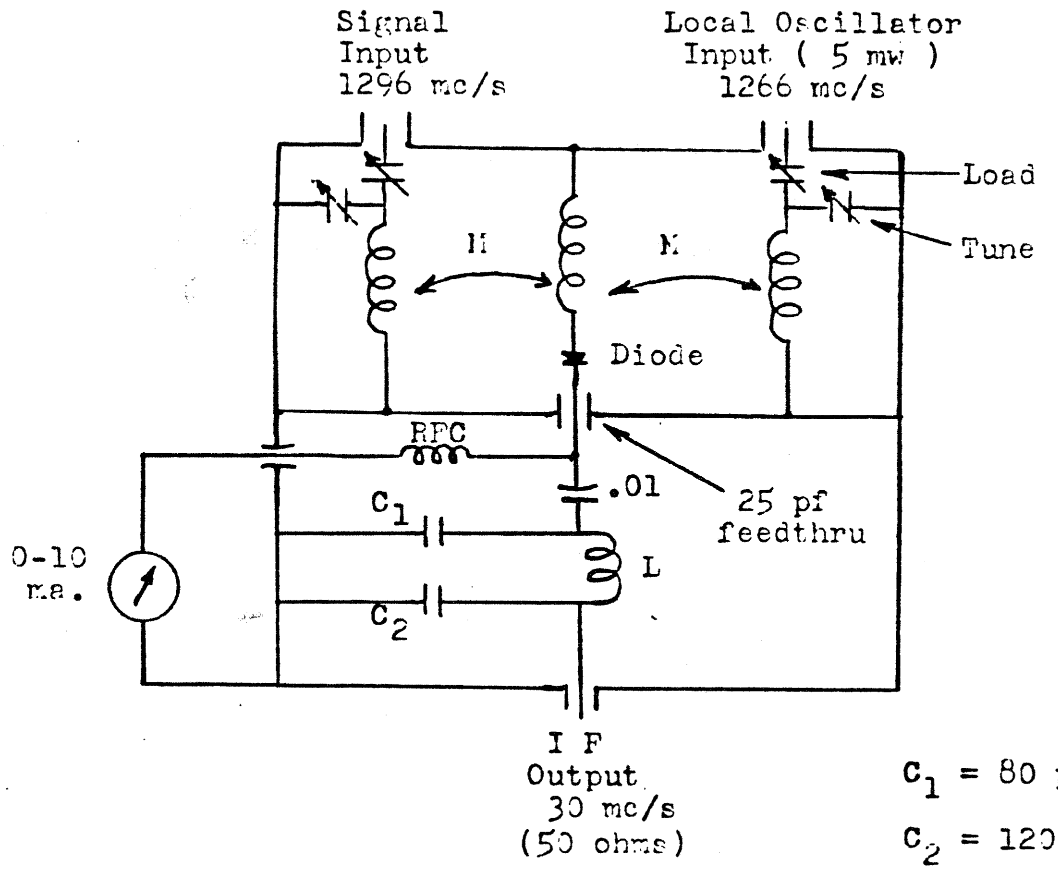
TECHNICAL REPORT NR 17

Date: Sept. 1974
From: The Crawford Hill VHF Club, W2NFA
Subject: A Low Noise Converter for 1296 mc/s

This report describes a low noise frequency converter for 1296 mc/s to 30 mc/s IF using a single inexpensive Hewlett-Packard "hot carrier" diode based on a design by R. E. Fisher, W2CQH *. The circuit is narrow band and uses single pole interdigital resonators for separation of signal and local oscillator voltages to insure SSB performance. This design differs from that of Fisher's by substituting adjustable capacitive coupling at the signal and L.O. inputs instead of the fixed inductive coupling. The local oscillator power source is not included in this design.

This report is intended for those experimenters who would like more information and variation on an excellent converter design. All are urged to consult the reference.

The converter to be described in this report is shown schematically below.



C_1 = 80 pf - $C_{feedthru}$

C_2 = 120 pf

L = 0.45 uhy

Q = 3

* QST January 1974

The π impedance matching network at the mixer output (C_1, C_2 and L) should be located at the mixer. Unlike the Fisher design, the IF output impedance for minimum NF was found to be about 200 ohms at the diode and the IF preamplifier is separated from the mixer at a 50 ohm interface impedance level. The impedance transformer which includes the 25 pf feedthru by pass capacitor as part of the network has an impedance ratio of about 4;1 and is not critical.

The working IF may be moved to 28 mc/s for convenience but should not be moved much below this frequency because of the selectivity of the r-f resonators. When this circuit is optimized for best SSB low noise performance the image rejection is about 16 to 20 db and the L.O. signal is rejected at the signal port by a similar amount. If these rejection levels are objectionable or there is evidence of strong interference at the image frequency (60 mc/s below the signal frequency in this case) an interdigital filter of the type described by Fisher * may be connected to the signal port. Keep in mind that the L.O. and image frequencies should be in the stop band of the filter.

The L.O. level must be sufficient to drive the rectified diode current to about 4 ma for optimum performance. This is not critical but must be over 1 ma for satisfactory performance. The L.O. level may be adjusted by means of the capacitive coupling. This coupling should be measurably less than at the signal port to assure a higher-Q in the L.O. resonator which results in little signal energy loss in the L.O. circuitry. For this reason, the available L.O. power should be at least 5 mw. The diode d-c return path, r-f choke and meter should have low resistance. If the meter is removed a strap should replace it.

The r-f resonators and diode coupling construction are exactly as in the Fisher design which essentially limits the resonator maximum loaded Q. This design assumes that the pumped diode r-f impedance is nearly 50 ohms, an assumption which is very reasonable judging by the mixer circuit performance.

Construction

Details and dimensions of the mixer circuit construction are shown by Figure 1. The broad walls of the enclosure may be made of 1/16 inch sheet brass or copper. Copper clad printed circuit board is also acceptable. The resonators are made from 3/8 inch O.D. hard drawn copper tube while the coupling post is a brass rod 1/2 inch O.D. (outside diameter - not critical) and tapered as shown to minimize stray capacitance where the diode is connected. The grounded ends of the resonators and coupling post may be bolted securely as shown or alternatively soldered carefully in place. Experience has indicated that the highest Q for these resonators is obtained by carefully mating the resonator to the brass bar before drawing tight or soldering. Soldering is not required if the inside chamfered edge shown in the detail on Figure 1 is used.

The adjustable coupling capacitors are small metal discs which pass through the mounting hole of the BNC jack and are soldered to the jack pin. Capacitor adjustment is by means of the threaded barrel of the UG 625 B/U jack in the 3/8-32 tapped hole.

Resonator tuning is by means of the strip tab soldered to the broad wall and pressed toward the resonator by the insulated screw.

* QST March 1968

Unlike the Fisher design, the narrow sides of the resonator enclosure must be covered to prevent leakage radiation. Strips of P.C. board may be secured to the brass bars on either side (not shown by drawings). The corners need not be electrically tight since resonator field current is along the axial direction and no leakage will occur through the axial slits.

Mounting of the diode which is physically small and has wire leads is detailed by Figure 2. Two methods of r-f bypass are shown. The diode wire lead is first soldered into the bypass plate or button capacitor with zero lead length. Do this quickly with a fairly hot small tipped iron so as not to damage the diode. Diode polarity is not important to the operation of this circuit. After the bypass capacitor is secured in its mounting place, the other end of the diode may be tack soldered to the soft thin copper tab at the end of the coupling post. This copper tab provides not only a low inductive connection to the diode but also a very important soft mechanical mount to prevent strain and diode breakage. It would be wise to solder the diode in at the very last.

Capacitor C_2 may be a small ceramic trimmer in order to optimize the match.

If it is desirable to include the IF preamp with the mixer as in the Fisher design, the space occupied by the matching network may be expanded to include the preamp. A single π network matching 200 to about 1500 ohms will now be required.*

As in all UHF construction care, cleanliness and good surface joints are essential to success. Since the brass bars and sidewalls are part of the resonator construction it is recommended that the bars be lapped clean and flat on all surfaces especially the narrow sides which join to the broad walls. Lapping is conveniently done by taping a sheet of medium fine emory cloth or fine sandpaper to a flat surface. The brass bars may be hand held between square wooden blocks for accuracy and then rubbed in a figure '8' pattern until the entire surface has been worn down evenly. The broad wall material should also be checked for flatness especially where the brass bars are joined. All drilling burrs must be removed to insure good joints.

One of the broad wall plates may be permanently soldered to the bars the other should be removable.

Tune-Up

Initially a suitable local oscillator source should be connected to the L.O. port and the L.O. resonator tuned for resonance as indicated by maximum rectified diode current. Then adjust the coupling for about 4 ma of diode current and trim the resonator tuning.

Next a suitable IF preamplifier and linear noise measuring receiver are connected to the IF output. If automatic noise measuring equipment is available the tune-up consists of first resonating the signal circuit for maximum response at 1296 mc/s and then making small adjustments of input coupling and IF output tuning (C_2) for minimum NF.

* A. R. R. L. Handbook, 1967 page 49.

Without automatic measuring gear, the procedure is essentially the same but more tedious since the noise source must be turned ON and OFF after each adjustment. It is not necessary to have a calibrated noise source for the tune-up, just a stable one. The object is to adjust for maximum difference in receiver output noise from ON to OFF of the noise source.

If no measuring gear is available, set the coupling capacitor gaps as per the dimensions in Figure 1 and adjust the signal resonator at 1296 mc/s using a weak signal from your exciter or from a nearby station on 1296 mc/s. The use of on-the-air testing signals should be used as a last resort since the uncertainty of signal level over a period of time may be considerable.

It is recommended that a 50 ohm 10 db pad be connected to the converter input during initial signal tune-up adjustments to insure that the signal source impedance is close to 50 ohms. Do not use a pad for noise measurements but do know the noise source impedance, which should be near 50 ohms.

If this converter is to be used directly with an antenna whose feedline is not well matched or of unknown match, some trimming of the signal loading and tuning may be required. Similarly with a preamplifier of unknown output impedance trimming will be required.

Some Theory

Conversion of radio frequencies is accomplished by mixing two or more signals in a non-linear circuit element. A receiving down converter usually employs a non-linear resistive element (a semi-conductor diode) which translates a high rf signal down to some lower frequency called the intermediate frequency, IF. To accomplish the frequency translation it is necessary to introduce a second RF signal (local oscillator) above or below the signal frequency by an amount equal to the IF. Ideally the non linear element should behave as a lossless switch driven by the L.O. signal. In practice diodes used in UHF mixers can approach ideal characteristics when driven with sufficient power at the L.O. frequency and used in an appropriate circuit.

In a receiving down conterter the frequency translation from RF to IF should be accomplished with good efficiency (low conversion loss) and the mixing element should introduce a minimum of extra noise. Mixers designed around good quality Schottky barrier ('hot carrier') type diodes can be characterized by a noise factor equal to the noise factor of the IF amplifier, F_{IF}, multiplied by the conversion loss, L, expressed as a number greater than 1.

$$F_{mixer} = F_{if}\, L$$

In decibel form, $F_{mixer} = F_{IF}(db) + L\ (db)$.

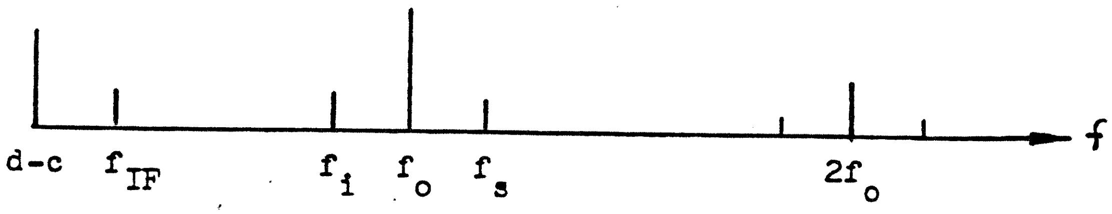
For example, if the IF amplifier noise figure is 1.5 db and the mixer conversion loss is 6 db, then the combined mixer plus IF amplifier noise figure is 7.5 db. The equivalent temperature of the mixer is then the usual formula

$$T_{mixer} = (F_{mixer} - 1)\ 290^{\circ}K$$

It is evident that the overall NF may be decreased by using as low an IF noise figure as possible and or reducing the conversion loss. A practical limit on the IF noise figure is about 1 db in the HF and VHF region using bipolar transistors. Lower IF noise figures can be achieved but with considerable more difficulty. Current FETs and MOSFETs are around 1.5 db noise figure in the same region.

Conversion loss can be considered separately. In a mixer the local oscillator power level must be much greater than the signal level in order to affect an efficient linear amplitude conversion. Unfortunately the large local oscillator level impressed on a non-linear element produces harmonics of the local oscillator frequency. In addition energy converted from the signal frequency to the IF will modulate the local oscillator and its harmonics and appear as sidebands on either side of the L.O. and its harmonics. The sideband associated with the L.O. fundamental is called the image frequency.

A spectral diagram of frequencies which appear across the diode terminals is shown below.



It is intuitively obvious that if energy is dissipated at all the sidebands, less will be available at the IF. It is therefore desirable to reactively terminate all sidebands except the signal frequency and the IF. A rigorous analysis indicates that the image frequency, f_i, is most important and should be terminated by an open circuit effectively across the diode terminals. This process is sometimes referred to as image enhancement or image recovery in a mixer circuit.

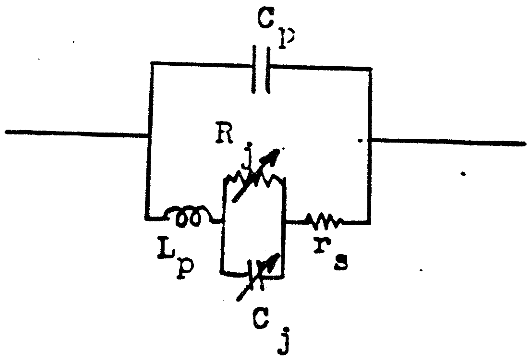
Next in importance are the sidebands on either side of the second harmonic of the local oscillator frequency, $2f_o \pm f_{IF}$. Although it is desirable to open circuit these sidebands, it is an impractical circuit problem and therefore more expedient to suppress the second harmonic of the local oscillator by means of a short circuit effectively across the diode terminals at $2f_o$. Treatment of higher harmonics depends on the diode quality in generating higher harmonics. In general an open circuit at the image frequency and a short circuit at $2f_o$ accounts for most of the available improvement in conversion loss. Little improvement will be achieved by treating higher harmonics of the local oscillator.

Diode quality can be described in terms of the ratio of its cut-off frequency to the operating frequency. The cut-off frequency is defined in the usual way

$$f_{co} = \frac{1}{2\pi r_s C_j},$$

where r_s is the bulk spreading resistance of the diode and C_j is the average junction capacitance under full local oscillator drive. In general C_j will be very nearly equal to the small signal junction capacitance at zero bias, a value usually available from the manufacturers specifications. A value for r_s is also usually specified. Both can be measured with proper equipment. Analysis of the minimum conversion loss of a Schottky barrier diode mixer, neglecting circuit losses indicates that for a ratio of f_{co}/f_s, of 20, L_{min} is 3 db. And for a ratio of 100, L_{min} is 1.5 db. In practice these minima cannot be achieved and measurement data indicates that corresponding values for L should be increased by 1 db. In addition, the spreading resistance, r_s, should be as low as possible. If two diodes have nearly the same cut-off frequency, the one with the lower r_s will give slightly better performance.

An additional factor which must be considered is the series resonance of the diode junction capacitance and the inevitable inductance which takes the form of a contact wire or strap. This strap is also required to take up thermal stresses and is therefore usually folded or bent which increases its inductance. The equivalent citcuit of a Schottky barrier diode of the wire lead glass encapsulated type is shown below.



C_p is the package capacitance which is resonated at the image frequency to open circuit the diode terminals. L_p is the package inductance described above and does not include the wire leads of the diode. The equivalent circuit shown above is contained within a very tiny volume at the center of the glass bead which supports the wire leads. The series resonance considered here is for the circuit elements L_p and C_j. If this series resonance is near or on the second harmonic of the local oscillator all efforts to short circuit the diode will not reduce the second harmonic sideband losses. It is therefore a further requirement of the diode that its natural series resonance be well above the second harmonic of the local oscillator.

Typically for the Hewlett Packard glass bead diodes, L_p is about 2×10^{-9} henries and C_j under drive conditions will be about 1 to 1.2 $\times 10^{-12}$ farads. Series resonance will therefore lie in the region around 4 Gc which is adequate for 1296 mc/s operation but is marginal for 2300 mc/s operation. For this particular encapsulation it is therefore suggested that diodes with minimum C_j be chosen in order to increase series resonance above the local oscillator second harmonic. A better approach is to use a different diode encapsulation more suitable for higher frequency operation, such as the beam lead type.

Armed with these theoretical requirements it is instructive to analyze the mixer circuit designed by R. E. Fisher. The interdigital resonators are single pole narrow band filters tuned to the signal and local oscillator frequencies. The central rod is an inductively coupled non-resonant line terminated by the diode. The RF impedance of the diode is approximately 50 ohms at either signal or LO frequencies. At frequencies other than f_s and f_o the central rod can be considered as an isolated unterminated section of line whose characteristic impedance is about 50 ohms.

At f_{image} the rod is approximately a quarter wavelength long and together with the case capacitance of the diode a parallel resonance is affected across the diode terminals. This is more nearly valid when the image frequency is below the local oscillator frequency, as shown by the spectral diagram. (It is of course possible to operate a mixer with the image and signal frequencies reversed in designation.) The Q of this image resonance is very low because the RF resistance of the diode is in shunt with the circuit impedance. It is therefore unnecessary to obtain exact parallel resonance, only in so far as the parallel resonant impedance is at least ten times the diode RF resistance of about 50 ohms This condition is relatively easy to realize without exact knowledge of the resonant frequency of the rod plus diode case capacitance.

Furthermore, the length of the central rod is very nearly a halfwave at the second harmonic frequency of the local oscillator. Since the end of the rod opposite the diode is grounded, the halfwave resonance presents an effective short circuit across the diode terminals at $2f_o$. This resonance although relatively critical in tuning so that an impedance much lower than 50 ohms is effected at the diode, is of lesser importance in terms of improving the conversion loss of the mixer circuit.

It is evident therefore that the interdigital mixer design of Fisher fullfills the primary circuit requirements for low conversion loss. In a quantitative accounting using an H.P. 5082-2835 diode which has a cut-off frequency of about 20 Gc the optimized noisefigure of the mixer plus 1.5 db IF was 5.5 db. Which indicates a conversion loss of 4.0 db which is approximately 0.5 db higher than the theoretical lower limit of 3.5 db for the ratio of 20Gc/1.3Gc. A remarkable accomplishment for so simple a circuit!!!

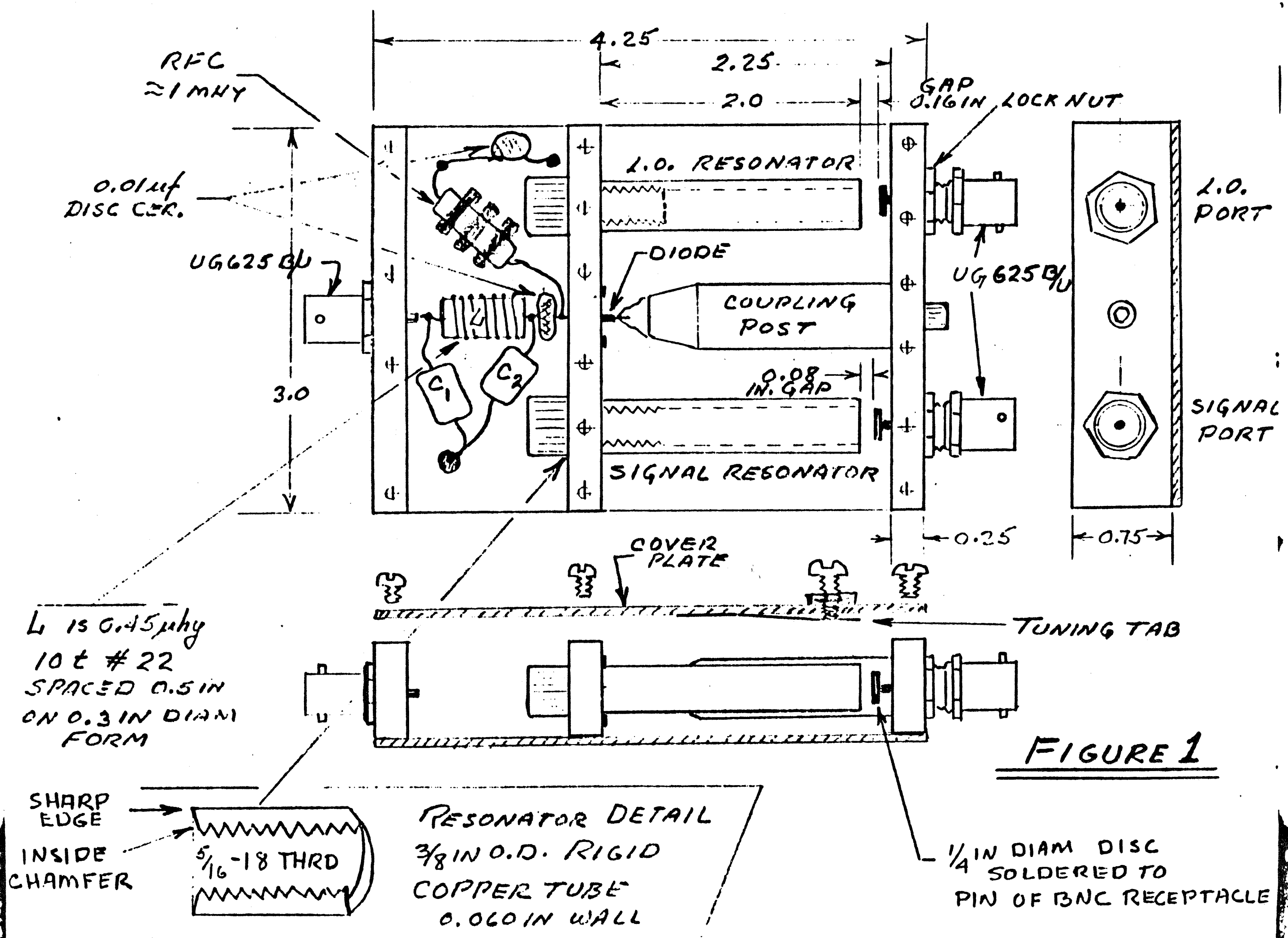
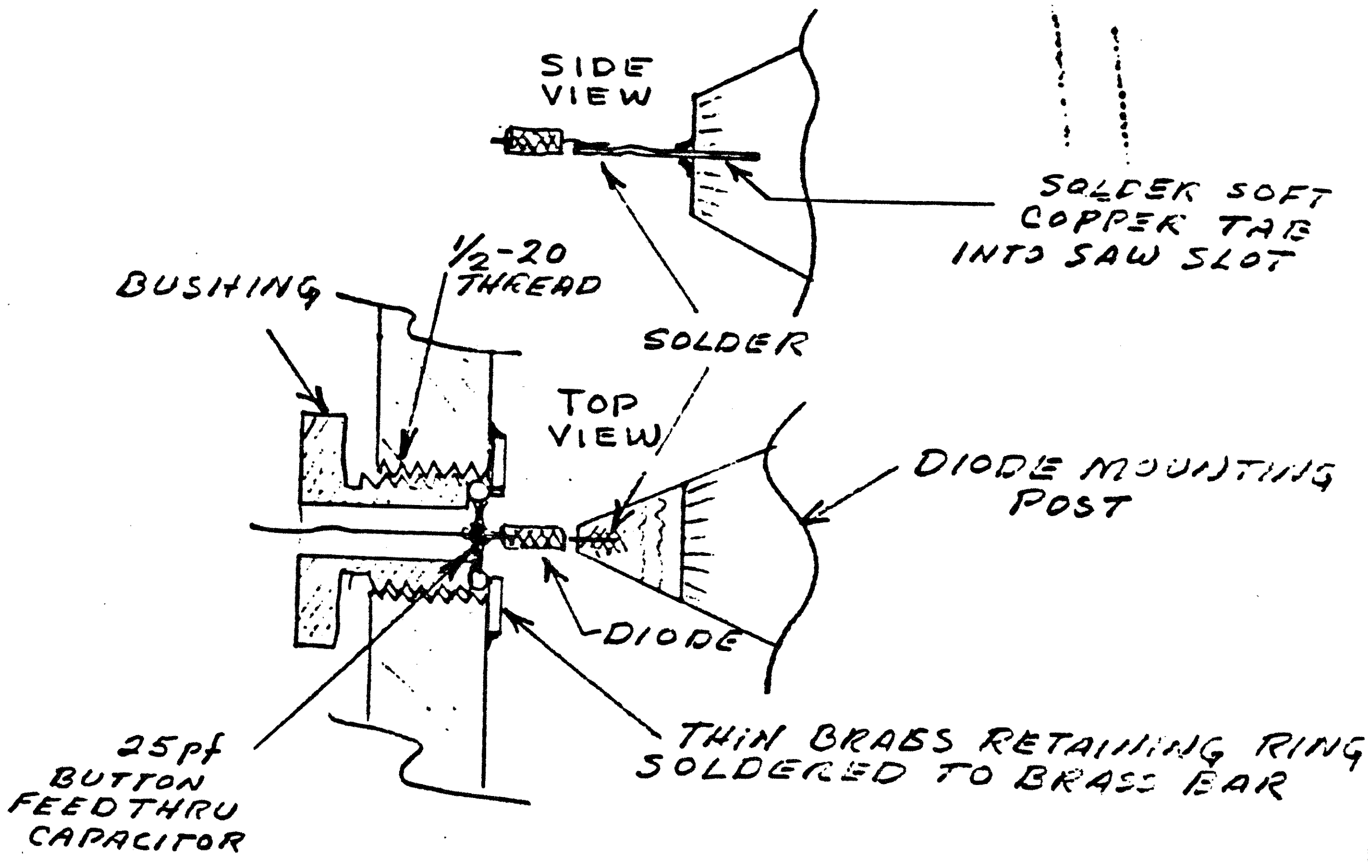

4.25
2.25
2.0
RFC
≈1 MHY
GAP
0.16 IN
LOCK NUT
0.01 µf
DISC CER.
L.O. RESONATOR
L.O.
PORT
UG625B/U
DIODE
UG625B/U
COUPLING
POST
3.0
0.08
IN. GAP
SIGNAL
PORT
SIGNAL RESONATOR
0.25
0.75
COVER
PLATE
TUNING TAB
L IS 0.45 µhy
10 t #22
SPACED 0.5 IN
ON 0.3 IN DIAM
FORM
SHARP
EDGE
INSIDE
CHAMFER
5/16-18 THRD
RESONATOR DETAIL
3/8 IN O.D. RIGID
COPPER TUBE
0.060 IN WALL
1/4 IN DIAM DISC
SOLDERED TO
PIN OF BNC RECEPTACLE

FIGURE 1



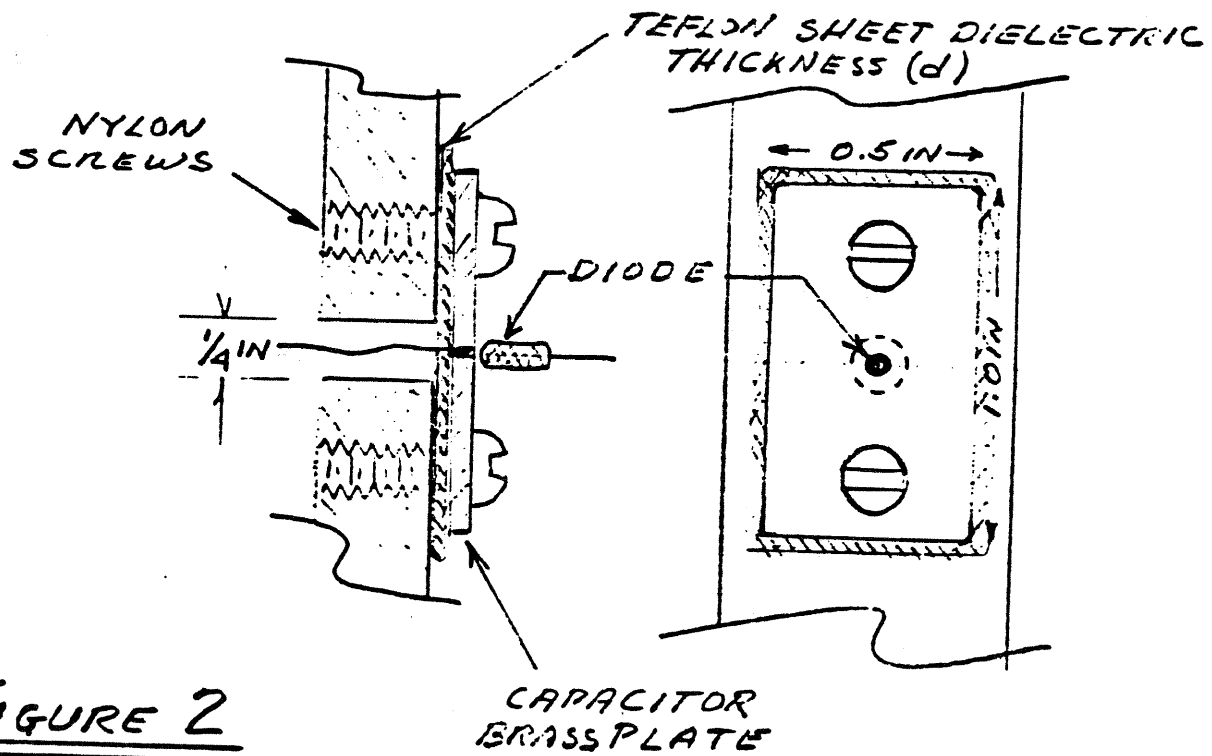


FIGURE 2

$$C_{pf} = 0.224 \frac{KA}{d} \approx 25 pf$$

$K = 2.0$ FOR TEFLON

TECHNICAL REPORT # 18

From: The Crawford Hill VHF Club W 2 N F A
P.O. Box 65 Colts Neck, N.J. 07722
Date: April 1980

Subject: An Off-Set Fed Parabolic Reflector Antenna for 1296 mc/s.

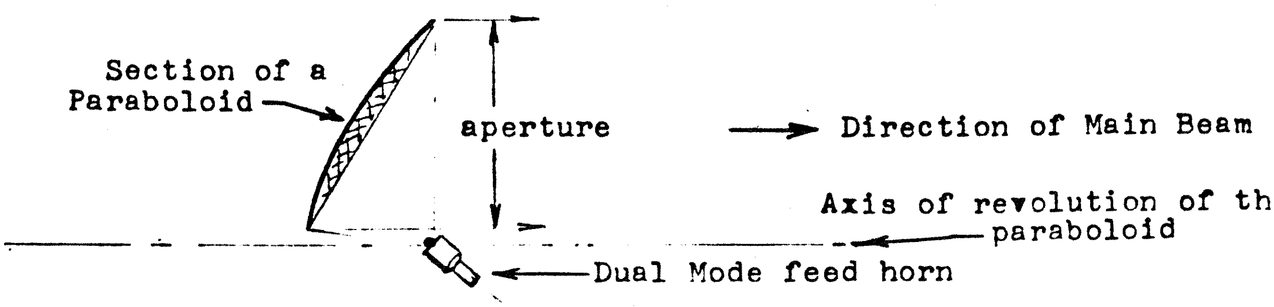
The parabolic reflector antenna design presented in Technical Report # 5 was a standard front fed 'dish' type antenna. This widely used antenna type provides high gain with good efficiency in the UHF and microwave frequencies. There are however several disadvantages to this standard design. (a) The prime focus feed is directly in front of and in the center of the electrical aperture. (b) The feed must be supported by struts which are also in front of the aperture. (c) With the large reflectors required for EME service, the feed is difficult to reach for adjustment while in service. The latter is a primary disadvantage not only because it is physically difficult to reach the feed but for EME service both the receiver low noise preamp and the last stage(s) of the power amplifier should be located at the feed to minimize feedline losses which can be very significant at 1296 mc/s. These equipments when mounted at the feed can increase the aperture blockage and scattering of energy from the center of the aperture where the energy density is maximum.

The implication of the above is that initial tune-up and subsequent maintenance and adjustments of the feed, preamp or PA to optimize the system in service will place the person doing the work also in the area of maximum energy. This results in exposure to high r-f fields on transmit and further blockage to the aperture area for either transmit or receive. In short, 'hot' adjustments of the feed will be hazardous and difficult to perform properly. The implication of (a) and (b) are decreased antenna gain by blockage and higher antenna noise by scattering.

While all these effects are relatively small (the order of 1 db typically) in terms of the antenna gain they are definitely detriments to high performance required in EME service. It is the express purpose of this report to acquaint the EME enthusiast with methods and techniques to obtain maximum performance with parabolic reflector antennas at UHF. To this end, the off-set fed parabolic reflector antenna is presented as an improvement over the front-fed parabolic reflector antenna.

The Off-Set Fed Parabolic Reflector Antenna

All of the above shortcomings of the front fed reflector antenna can be overcome by changing the geometry of the parabolic reflector to an off-set fed reflector as shown below in side view.



The term off-set refers to the section of the paraboloid off-set from the axis of the paraboloid. The feed horn aperture (phase center) remains at the focal point of the paraboloid but the feed is aimed so as to illuminate the off-set section of the paraboloid. In this report a design directly related to the high efficiency dual-mode feed horn is chosen as a desirable and compact arrangement. The geometric considerations presented later are specifically for this particular design.

With the off-set geometry the feed is at once out of the antenna aperture, the feed supports need not be in the antenna aperture at all, the feed is closer to the ground and most importantly, the feed may be loaded with equipments which may be adjusted without electrical intervention of the person doing the adjustments. The importance of being able to get to the feed and associated equipments for initial adjustments of focal position, impedance matching and optimizing preamp noise performance as well as maintenance cannot be stressed enough.

A more subtle advantage of the off-set feed is that since the feed is outside of the electrical aperture there will be no interaction between feed and reflector as there is with a front fed reflector. design and all of the initial optimizing adjustments will be more straightforward.

Elimination of reflections between feed and reflector is especially important when using circular polarization. In the front fed design circularly polarized energy transmitted in one sense of rotation is reflected back to the feed in the opposite sense. This means that the transmit-receive port isolation is limited to $G\lambda/4\pi f$,* where f is the focal length, G is the feed absolute gain (typically G = 10 for the dual-mode feed) and λ is the free space wavelength. A 20 ft diameter front fed reflector with f/D = 0.6 will have an isolation of about 26 db at 1296 mc/s. For 500 watts at the feed transmit port there will result in 1.25 watts at the receive port! In the off-set design the T-R isolation can be nearly perfect with any polarization.

It is characteristic of the off-set parabolic reflector assymetric geometry to radiate cross polarized energy with linear polarization**. The cross polarized radiation increases with shorter focal length and larger off-set angle. Cross polarized radiation is found not on axis of the main beam but in the 45° planes close to the main beam. This radiation is small but is completely lost energy in terms of antenna gain.

Another characteristic of the off-set reflector antenna is that with circularly polarized feed energy there results no cross polarized radiation but instead a slight misalinement of the main beam.

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* Radiation Lab. Series, MIT, Vol 12, page 440.

** T. S. Chu and R. H. Turrin, "Depolarization Properties of Off-set Reflector Antennas", IEEE, PGAP, Vol AP-21, #3, May 1973, pp.339-345.

The main beam misalinement is in the horizontal (i.e. transverse) plane and is to the left of center for right circular polarization for a transmitted wave receding from the antenna. The error in pointing of the main beam from left to right circular polarization is only a fraction of the beamwidth which results in a gain difference from left to right C.P. of only a fraction of a decibel. For example a 20 foot (3 meter) aperture off-set reflector antenna with off-set angle of 45° at 1296 mc/s (the design detailed in this report), the pointing angle will change 3/4° from right to left CP while the main beamwidth is of the order 4°. The resulting gain difference is about 0.1 db.

It is desirable to employ circular polarization for the EME path to eliminate Faraday rotation fading through the ionosphere. The off-set reflector antenna is suited to circularly polarized application. As mentioned previously the off-set feed virtually eliminates the coupling between transmit and receive ports via reflections from the reflector surface back into the feed. It now is highly desirable to consider the circularly polarized feed method suggested in Tech. Reports #1 and #2 which makes use of the C.P. dual mode feed presented in Reports #9. This scheme takes advantage of the isolation between opposite sense circularly polarized ports to eliminate the need for a T-R relay or switch. With this method if all EME stations transmit one sense C.P. and receive in the opposite sense then all stations will be compatible with each other and with their own echoes. The obvious advantages of this scheme are elimination of Faraday fading, no polarization tracking or searching, no T-R switch required in the antenna feed, complete polarization compatibility and elimination of cross polarized radiation which results in maximizing gain.

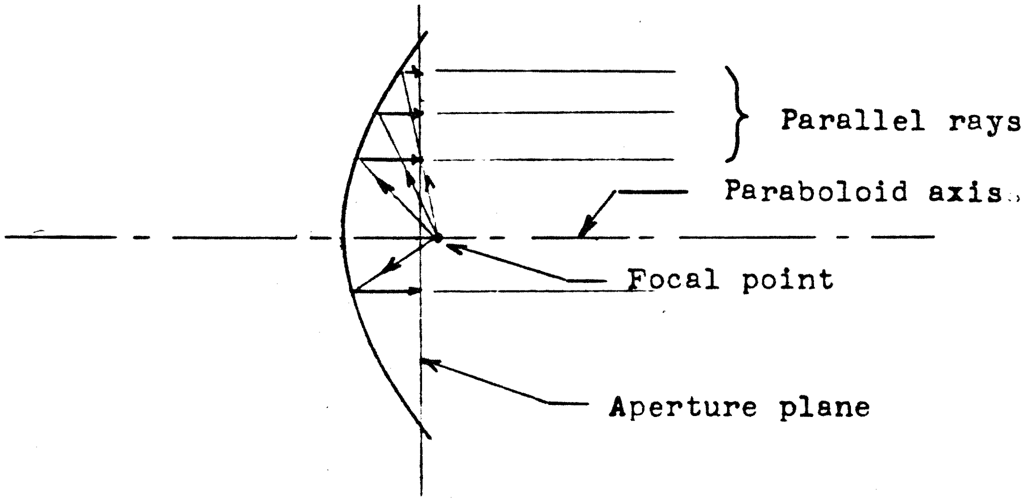
Another advantage of the off-set geometry is for multi-band operation where the feed package may be easily changed by virtue of its accessibility and non critical space requirement. That is the feed and equipments may be packaged in almost any size and shape since there is no blockage problem. It is conceivable to mount two complete feed packages on swinging booms to facilitate changing bands.

For strictly EME service with good foreground clearance in the direction of the Moon orbit, the lower edge of the reflector may be mounted near to the ground and may also be the location of the elevation axis. In this arrangement the feed will be easily accessible even for moderate elevation angles. For a given aperture size and with an off-set angle of 45° the feed will be 16% closer to the reflector in the off-set design compared with the front fed design.

If there is any disadvantage to the off-set reflector antenna design it is the different and somewhat awkward geometry. The remainder of this report will present geometric details and suggestions for construction of a specific off-set parabolic reflector antenna design.

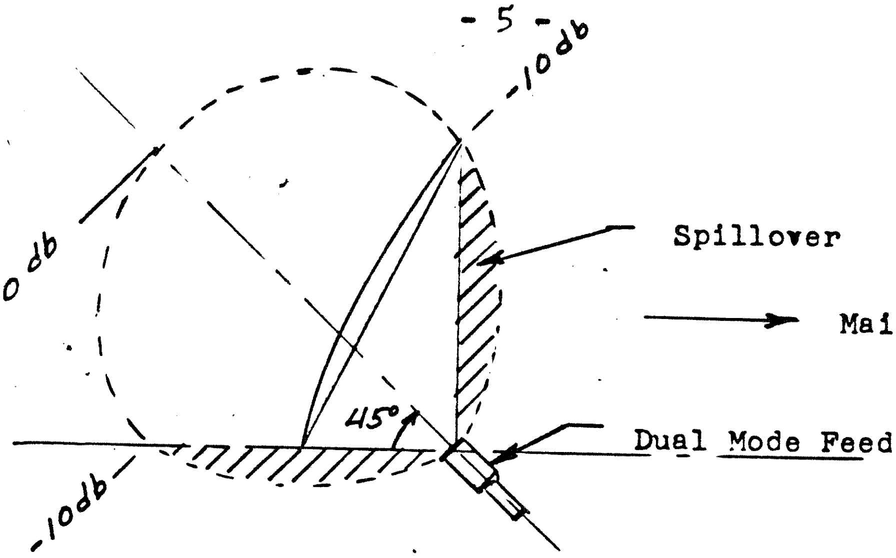
Off-Set Geometry

The geometry of the off-set reflector antenna may be developed from basic properties of a parabolic reflector. One property is that all rays of energy that originate at the focal point of a paraboloid will be reflected in a direction parallel with the axis of the paraboloid as shown below.



It is also a basic property of a parabolic reflector that the total length of each reflected ray from the focal point to the aperture plane is a constant length. This latter property means that energy at a fixed frequency originating at the focal point will be distributed over the aperture plane with uniform (constant) phase. These properties cause the beam forming characteristic of a parabolic reflector antenna much the same as a flashlight produces a beam of light when the bulb filament is located at the focal point of the small mirrored parabolic reflector in the flashlight.

Since all rays emanating from the focal point behave in the same way, we may choose to direct energy from the focal point to any region of the paraboloid which will then form a beam. In the design described in this report we choose a region just off the paraboloidal axis so that the feed antenna is outside of the region (aperture area) of the reflected rays. The term off-set therefore designates that the reflector surface region used is off-set from the paraboloidal axis but the feed horn antenna is still located at the focal point. The feed horn axis is however tilted to illuminate the chosen area of the reflector surface and the center of the feed horn aperture (phase center) is located at the focal point of the paraboloid. A cross sectional drawing of this geometry is shown below for the particular case of the dual mode feed. The dual mode feed is chosen not only because of its high illumination efficiency as pointed out in Report #5 but because it will also result in the most compact form of the off-set antenna design.



The constraint on feed illumination is the same as with the front fed case, that is for maximum gain the -10 db contour of the feed radiation pattern should fall at the edge of the reflector surface. The spillover will be the same in both cases and the gain between the two designs in terms of aperture efficiency will be the same. The difference in effective gain between the two cases is due to blockage of feed and struts which will be 0.5 db or more.

Since the feed radiation characteristic of the dual mode horn are circularly symmetric, the -10 db radiation surface contour may be represented by a cone whose apex is located at the focal point of the paraboloid and whose half cone angle is 45°. The geometry of the off-set fed parabolic antenna may therefore be simply described as the intersection of a cone and a paraboloid. Figure 1. shows various useful views of this geometry with simple formulae to calculate all necessary dimensions and angles. The views shown by the drawing are specifically for the dual mode feed with off-set angle of 45°.

Several useful properties of this geometry which may be readily proven are: (1) the projected area of illumination of the reflector in the direction of the main beam (parallel with axis of paraboloid) is a circular disc (the aperture area). (2) The edge or rim of the illuminated area on the reflector is elliptic and lies in a tilted plane. These characteristics make it easier to visualize the physical construction. Be assured therefore that even though the off-set geometry is awkward its physical properties are described by simple geometric figures.

In the aperture view, to the right in Figure 1, the dotted radial lines represent lines of the same parabolic shape. If a trussed rib type construction is employed these trussed ribs can all be made on the same template but of different lengths. All the ribs will converge at the vertex V at the bottom of the aperture where a support hub may be used as the main structural mount.

Construction Notes

The following notes are included merely as a suggestion guide for building an off-set reflector antenna including mount, strictly for EME use.

Since an EME antenna will be large, the basic considerations in construction will be selection of a suitable site and keeping the structure low to the ground to minimize structural wind loading problems. The site selection should allow minimum obstruction to the radio beam over as large a portion of the Moon orbit as possible.

Figure 2 is a sketch of a complete off-set fed reflector antenna and mount. The mount consists of a well anchored vertical post which serves as the azimuth pivot. It is located at the center of a circular track near ground level upon which the mount will turn on wheels.

The mount frame should be triangular with running wheels near the corners. This provides a simple three point support which will not teeter. The circular track may be of poured concrete with footing suitable to the local climate and just wide enough to support the total antenna load (estimate 1000 pounds for a 20 ft reflector). The wheels can be solid rubber type found in hardware stores as replacements for garden equipment and need not be very large in diameter. Several wheels may be stacked at each cornerof the frame for extra load bearing surface. Pneumatic type wheels might be satisfactory if inflated to high pressure to prevent tilting of the mount frame with unbalanced loads.

The central pivot should also provide for a thrust bearing to hold the frame down securely onto the track. A track diameter of no less than 15 feet is recommended for this size antenna (20 ft aperture The central pivot should be set plumb in concrete and the circular track leveled carefully. Since the antenna structure will turn slowly around the pivot, the actual bearing may be very crude.

Azimuth drive may be through friction drive directly to a support wheel. If the drive motor is coupled through a worm gearbox, braking will be automatic. Azimuth angular readout may also be taken conveniently from an idler wheel which runs on the track. By sizing the idler wheel diameter it may be used to drive a synchro directly for fine readout and through a suitable gearbox for coarse readout. A bench mark calibration of true North (or any other reference) will permit rapid check and realinement of the readout should it slip or accumulate error. A readout accuracy of 0.5 to 1.0 degree should be achieved.

Once the azimuth pivot, track and frame have been completed, construction of the reflector frame with elevation pivots can be started. The elevation pivots can also be relatively crude since rotation will always be slow.

The reflector frame should be constructed in the "stow" position, that is with the radio beam pointed straight up. This frame as well as the azimuth triangular frame may be fabricated of relatively heavy material to provide rigidity and strength for the actual reflector and feed support. Details of the reflector frame are shown approximately by Figure 2 with additional stiffaners and supports added where necessary. The depth of the frame from elevation pivot to reflector vertex need not be more than 3 feet for this size antenn a.

Since this design does not use counter weights to balance the elevation forces, the structure is inherently unbalanced except at one elevation angle. For this reason the elevation drive suggested is a system of dual lead screws or hydraulic jacks one adjacent to each pivot bearing as shown by the sketch. Also, because the antenna is near the ground it will be difficult to track the Moon to rise and set times due to obstructions in the radio beam. The range of elevation movement may therefore be limited for your site thus minimizing the requirements of the drive system. Auxiliary drives may be required to "stow" the antenna.

Construction of the reflector should begin with fabrication of an accurate reverse template and pivot assembly which will be the guide for placement of the trussed ribs and as final surface alinement tool. The template pivot should be secured to the reflector frame at exactly the vertex of the paraboloid. Actual construction of the trussed ribs and supports is left to the individuals imagination. Most of the support will be near the vertex and towards the center of the reflector. The rim need not provide much support in this design. Reflector construction using the materials and techniques described in an article by VK3ATN in Ham Radio Magazine, p 12, May '74. is recommended.

The feed support tower is added last and may be partially secured to the reflector by a few thin steel wires for added rigidity.

Surface material for the reflector should be chosen according to the highest frequency of operation. As a reminder, the surface material hole size should not exceed 0.1 wavelength at the highest frequency. The material need not be bonded together electrically at overlapping joints provided no joints are closer together than a wavelength at the lowest operating frequency and the overlaps are at least a quarter wavelength at the lowest operating frequency. Deviation of the surface from a true paraboloid should not exceed 0.05 wavelength at the highest frequency especially towards the center area of the reflector where the energy density is highest. If the rib construction is not too accurate, trimming up the surface is highly recommended.

Refer to Figure 1 for more details of the geometry and placement of the feed. The circular rings around the vertex shown in Figure 1 are added on top of the ribs and are the actual surface material support. These circular arcs may be made of lighter material an placed about a foot apart.

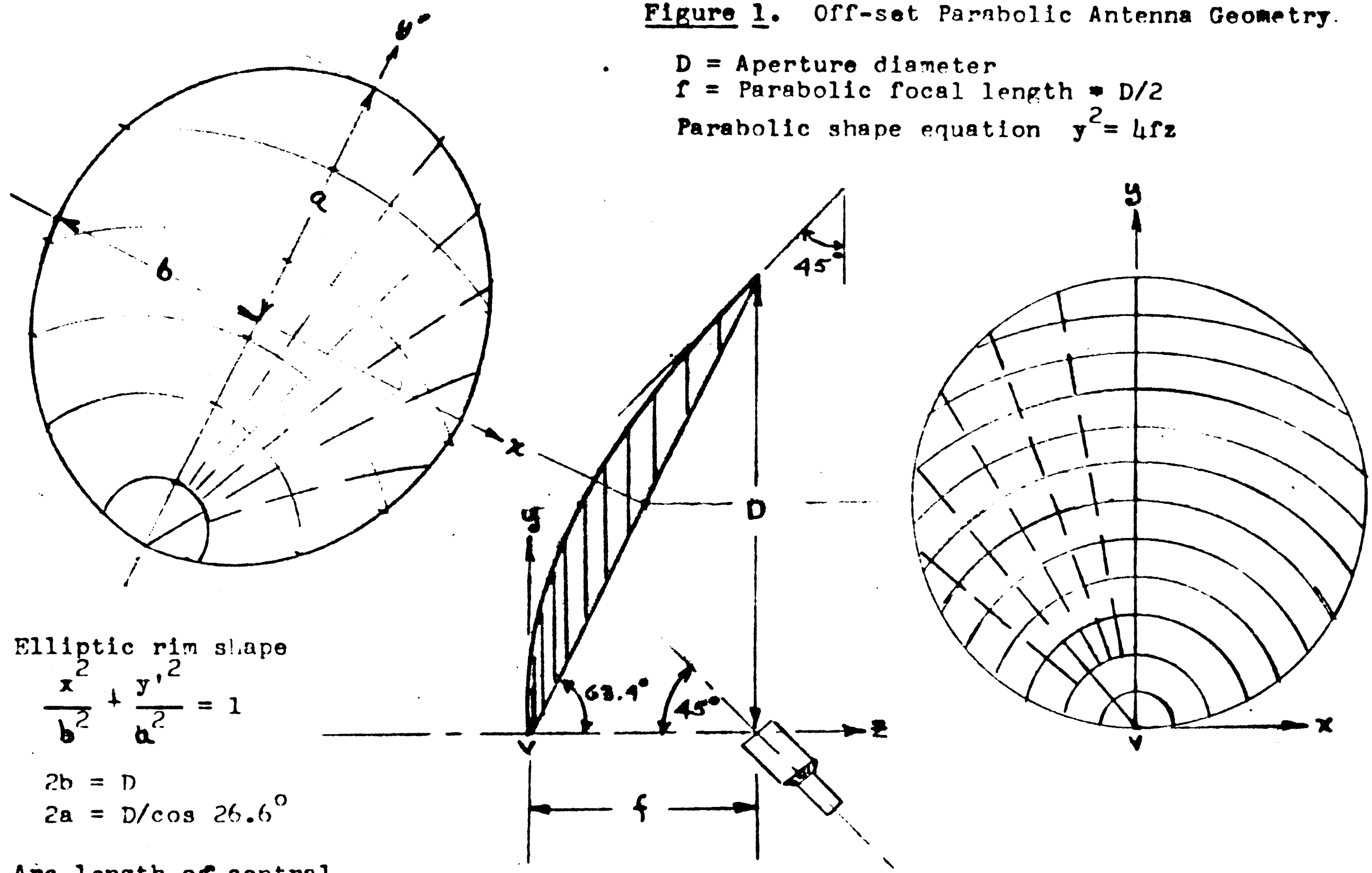
In the stow position the reflector frame should come to rest and be secured to the azimuth frame. This position provides the least wind loading. Provisions should be made to anchor the azimuth frame in severe weather.

A project of this magnitude should be given careful consideration especially with regard to expense, construction time, site location and family-neighborhood complications. In fact if your site is not acceptable it would probably be advisable to join forces with a nearby EME enthusiast who has an acceptable site. Remote control operation might be a solution for joint operation.

Since a reflector antenna may be used over a wide frequency range by changing feeds, it would be highly desirable to make the reflector surface as accurate as possible in anticipation of use above 432 mc/s, possibly 1296, 2400 and 4000 mc/s the satellite TV band!

Figure 1. Off-set Parabolic Antenna Geometry.

D = Aperture diameter
f = Parabolic focal length = D/2
Parabolic shape equation $y^2 = 4fz$



Elliptic rim shape

$$\frac{x^2}{b^2} + \frac{y'^2}{a^2} = 1$$

$2b = D$
$2a = D/\cos 26.6^{\circ}$

Arc length of central parabolic section
$S = 2.2956\ f$

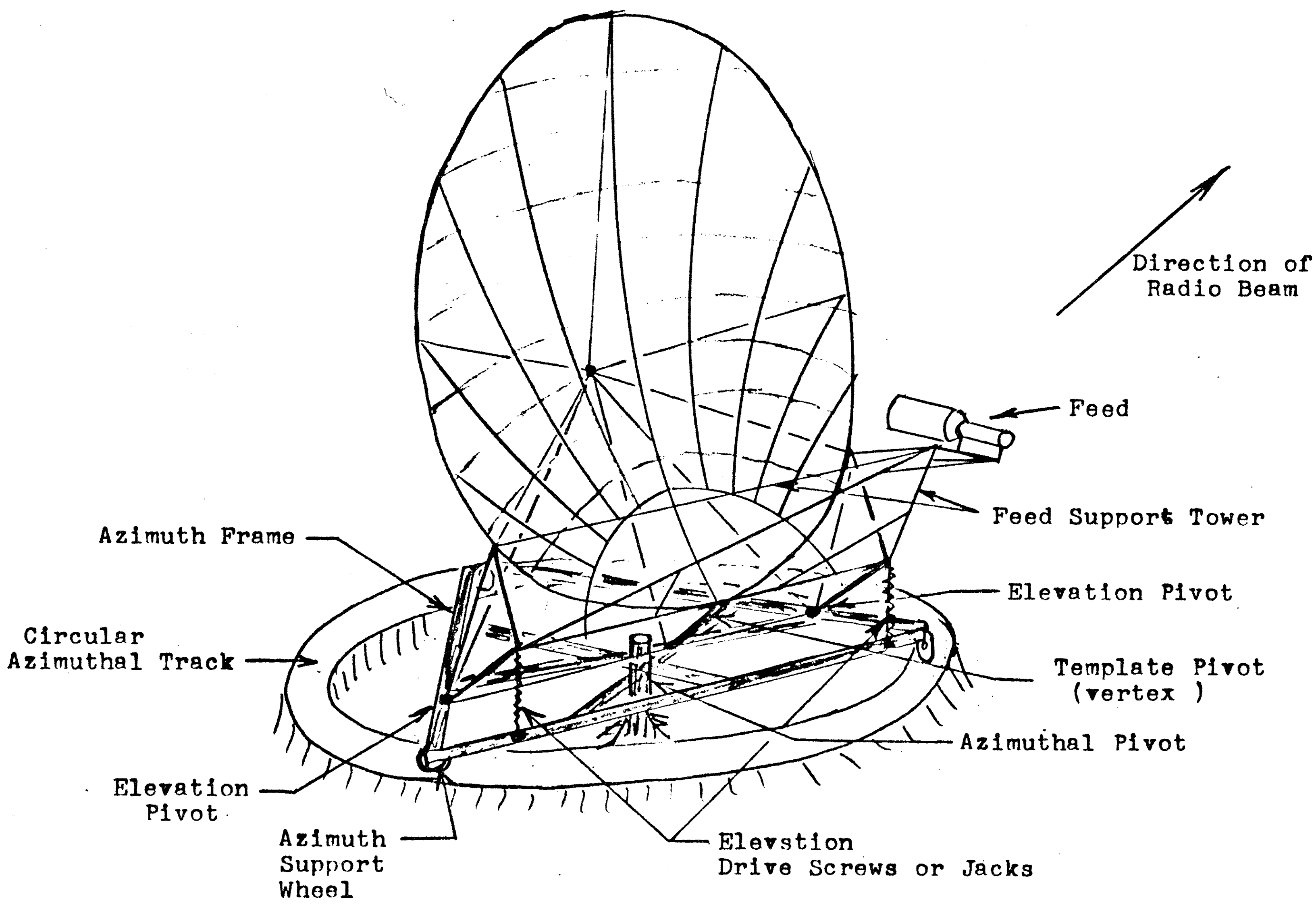


Figure 2. A sketch of an off-set fed parabolic reflector antenna for Moonbounce application.

Technical Report # 19 #19

From: The Crawford Hill VHF Club
Date: April 1982 (revised Oct. 1986)

Subject: A Six-port Quadrature Power Divider/Combiner

This report describes a relatively new and simple branch-line hybrid coupler which is a six-port planar network with four output ports. Each output port has equal amplitude, and phase sequence of 0, 90, 180 and 270 degrees. The network is shown schematically by Figure 1 and may be constructed in coax, stripline, or other type line sections. Particularly note that each line section is an electrical quarterwave in length, and that all line sections have the same characteristic impedance as the terminating and drive port impedances.

Alternatively, all line sections may be constructed with line sections whose characteristic impedance is not exactly the desired port impedances, and then quarter wave linear transformers used at output ports where necessary. This technique can be especially useful when constructing low loss couplers from air-line sections of available materials, such as copper tubing and wire.

An experimental model for 432 mc/s built with sections of RG214 (50 ohms, double shield, silver plated) had a measured insertion loss of 0.5 db and an input port isolation of over 30 db.

This coupler can be used in combining four power amplifiers for higher power generation, and in feeding a quad of dipoles such as the NBS feed antenna to obtain circular polarization. Figure 2 illustrates these applications in schematic form.

The power amplifier application, Figure 2a, permits combining 4 two-tube amplifiers of the type described in Report # 6 to achieve a kilowatt d-c input power amplifier at 1296 mc/s with little stress to the planar triodes. It is suggested that to minimize losses, at least the output combiner be constructed with air-line sections.

The antenna feed application is especially useful since it permits achieving isolated left/right circular polarization which is desirable in all EME communication to eliminate Faraday fading and also to eliminate high-power T/R switching. One form of this application is in feeding a dual linearly polarized NBS feed.

A basic NBS feed is two in-phase dipoles a quarter wavelength above a one wavelength square ground plane.

Figure 2b shows a quad of dipoles arranged for dual linear polarization. By feeding each dipole from a port of the six-port hybrid, the proper equal amplitude and phase relationship is obtained to radiate and receive left and right circular polarization. The planar network may be folded into a cube for convenience and 'dummy' line sections included to complete the cube and add physical strength as shown.

Note that pairs of parallel dipoles are fed at opposite sides of their baluns to preserve the necessary in-phase condition.

Reference: A. A. M. Saleh, "Planar Multiport Quadrature-Like Power Dividers/Combiners", IEEE Trans. on MTT Vol MTT-29, No. 4, April, 1981, pp 332-337.

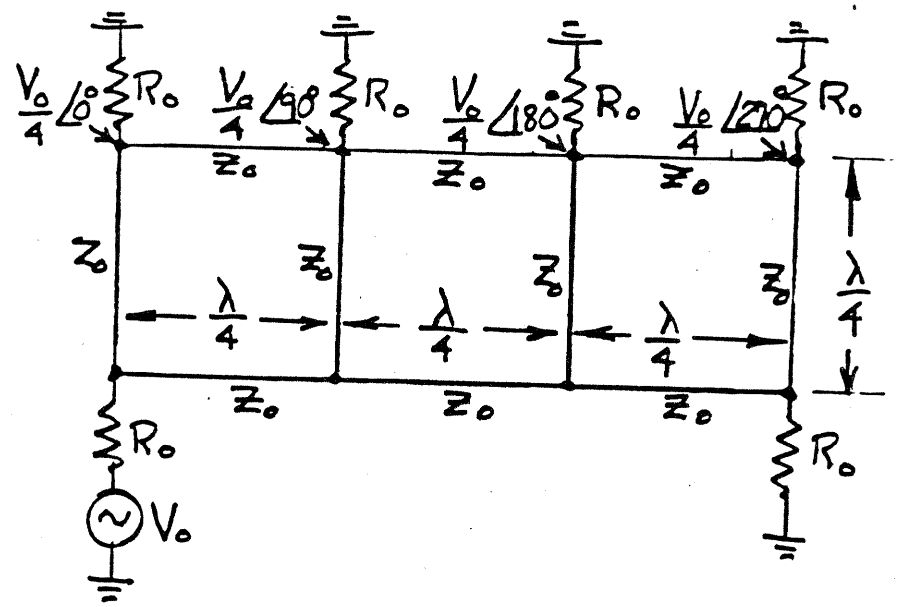
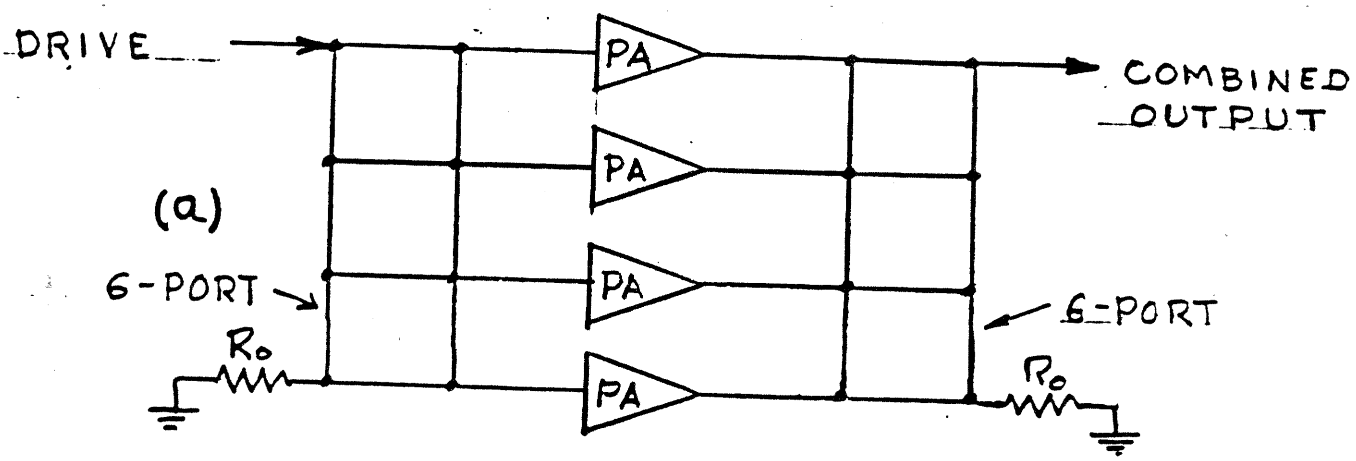


FIGURE 1. BASIC 6-PORT 90° HYBRID



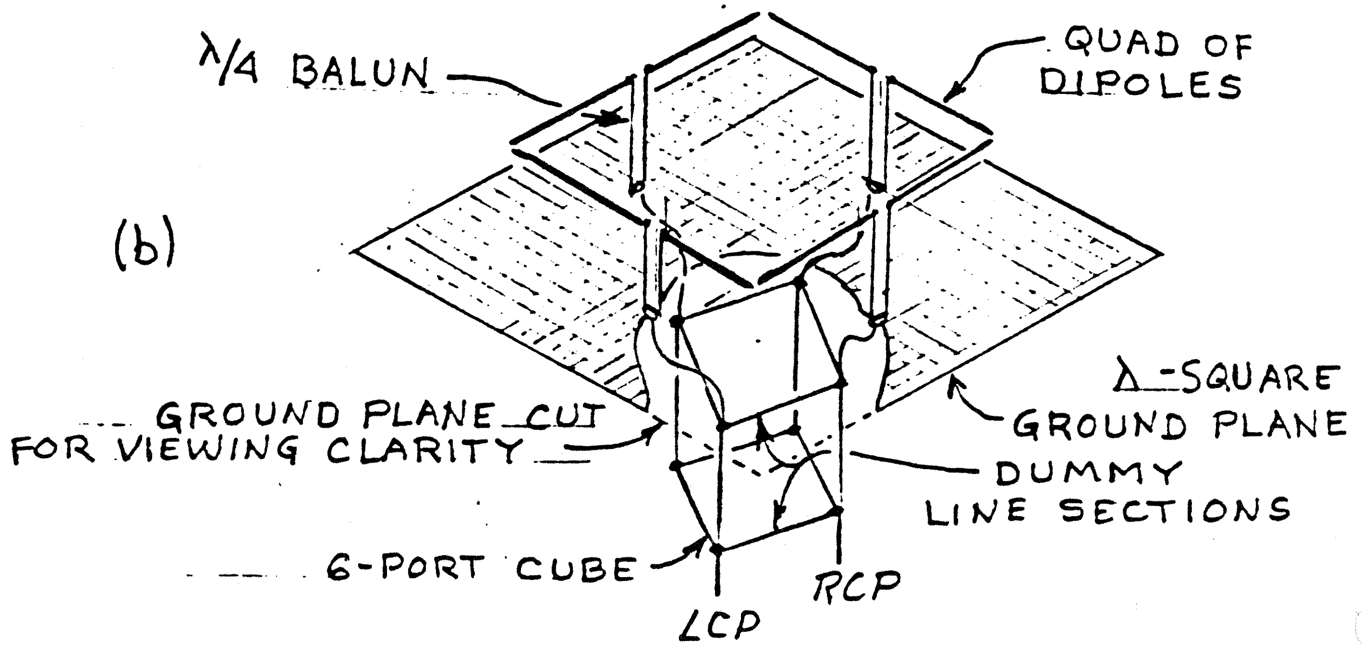


FIGURE 2. (a) POWER AMPLIFIER COMBINER,
(b) CIRCULARLY POLARIZED NBS FEED ANTENNA.

TECHNICAL REPORT # 20

From: The Crawford Hill VHF Club
Date: September 1986 (New)

Subject: <u>Method for Estimating Receiver Noise Temperature</u>

In an earlier version of Technical Report # 11, a method was described to determine the receiver noise temperature by comparison with a room temperature termination. This new Report will elaborate and attempt to eliminate some deficiencies in the method described earlier. Some of the text will be repeated here.

It should be expressly noted here that these measurement procedures should be employed mainly as diagnostic techniques in order to determine where improvements are needed in the EME system.

While a direct measurement of receiver NF using calibrated laboratory test equipment may seem like an accurate determination of the receiver (at least preamplifier) noise performance, it is not always indicative of exactly how well the preamplifier (receiver) is performing in your system. This may be largely due to the impedance presented to the preamplifier by your antenna as compared with what the impedance of the noise generator is when used to measure its NF. In general a good laboratory noise generator may 'look' like a broadband 50 ohm resistive source while your antenna may be approximately a 50 ohm (narrow band) source. Furthermore, since the preamplifier can be optimized (if adjustments are provided) when connected to a good 50 ohm source impedance, there is no guarantee that it will be optimized when connected to your antenna.

Finally, the input noise temperature (and possibly gain) of a low noise preamplifier is characteristically dependent mainly upon input source impedance, and possibly output load impedance.

For these reasons, reasonable precautions must be taken to optimize the preamplifier when connected in normal fashion to <u>your antenna</u>.

<u>Optimizing Receiver-Antenna Interface</u>

If there is any doubt of the impedance match to the antenna, the following procedure may be used to optimize the complete receiver-antenna interface connection.

This procedure will require a low noise impedance tuner of the type described in Report # 12; or, a preamplifier with input matching adjustments. Do not use a stub tuner or any tuner with sliding metal-to-metal contacts as these can be erratic and lossy when adjusted. The dielectric slug tuner described in Report # 12 is ideally suited because it is low loss and has a limited impedance transforming range. The rest of the procedure assumes that the preamplifier input is <u>not</u> adjustable.

Install this tuner between the antenna port and the preamplifier input. If you have a protective relay in front of the preamplifier in normal use, leave it in place and install the

tuner between the relay and antenna port.
As a general rule, never install a filter between the antenna port and the preamplifier unless intolerable interference exists. The unavoidable filter losses will degrade the system performance as well as result in extreme impedance match out-of-band.
Next a weak and well shielded signal source is needed to leak a small r-f signal into the antenna system. A good shielded source can be made from a battery operated crystal oscillator and harmonic generator all enclosed in a new paint can as a shielded enclosure. A coaxial port on the shield can be equipped with a small pick-up loop which can be rotated to adjust the output level. The output is then connected to a dipole or horn radiating device radiating directly into the antenna structure. For a parabolic reflector antenna system the weak signal source may be introduced at the vertex of the paraboloid with a small horn or dipole.
The signal source shielding is necessary to prevent stray radiation from reaching the receiver through any path other than directly into the antenna port feeding the preamplifier. A test for leakage should be made when the weak signal level is adjusted to about 10 db above the noise. Simply disconnect the preamplifier from the antenna and terminate the preamplifier with a well shielded resistive termination and check for signal at the receiver output. If a signal is heard, leakage is present and should be traced down and eliminated before proceeding.
With the weak signal feeding the antenna system, the preamplifier connected in normal fashion and the slug tuner installed, the optimization procedure is simply to measure the signal-to-noise ratio at the receiver output while adjusting the tuner for maximum signal-to-noise. The process may be made more convenient if the weak signal source is automatically keyed "ON" and "OFF" at a regular rate of about once a second (key the harmonic generator to avoid frequency "chirp"), and a level meter is wired to the receiver output and located at the tuner site.
If no improvement can be made, the interface impedance may be correct and the tuner may be removed, checking the S/N ratio again after removing the tuner. If an improvement is noted, even slightly, it will be necessary to either leave the tuner in the operating system or provide other adjustments to the preamplifier input circuit to facilitate matching without the tuner installed.
This completes the preamplifier-antenna interface optimization.
Don't be surprised if a very noticeable improvement was made, this only indicates the sensitivity of this interface in your set-up and the lack of proper initial adjustment to both preamplifier and antenna feed port match.
If the tuner slug positions indicate maximum mismatch (slugs separated by a quarter wavelength of air coax), then it would be advisable to readjust the antenna input impedance and repeat the noise matching procedure.
Do not use connectors and/or jumper cables of different impedance levels (50/70 ohm) anywhere in and around the antenna-preamplifier interface. Use only double shielded coaxial cable or <u>rigid line</u> for <u>short jumpers</u> at frequencies above 432 mc/s.

Receiver Noise Temperature Estimation

This measurement procedure is specifically tailored to an antenna system which uses a feed antenna and a focussing reflector.

A good measure of the receiver noise temperature alone may be made using the procedure of switching between the antenna feed and an ambient temperature termination. The procedure will be modified here to minimize the antenna noise effects and to assure that the ambient temperature termination is properly optimized with the preamplifier. This method is a 'hot-cold' load method with input impedance optimization.

The termination in this case will actually be a matched attenuator of over 30 db loss. The reason for using a matched attenuator instead of a direct termination is to permit a weak signal to be introduced into the receiving system through the attenuator for optimization (using a low loss tuner with the attenuator only). It is assumed that the antenna feed-receiver interface was optimized already.

To minimize feed antenna noise, the feed is removed intact from the reflector antenna and relocated at the vertex of the antenna and positioned to radiate maximum along the center line of the paraboloid. In this position and with the paraboloid in its stow position (rim horizontal) the paraboloid acts as a shield for the feed allowing it no access to warm Earth noise and aimed towards the zenith where the Universe temperature is very low at 1296 mc/s and higher bands. The procedure is illustrated by Figure 1 along with required switching and terminating devices. Normal feed support struts should be removed if possible.

The optimizing S/N procedure as described above may now be used with the terminating attenuator switched to the preamplifier. And, for accuracy, the feed-preamplifier interface may be reoptimized by introducing the weak signal over the rim of the dish. There may be some slight interaction when the feed is optimized with and without the active reflector because of reflections into the feed when normally installed. With circular polarization, this effect should be negligible.

Under the conditions illustrated by Figure 1, there should be a very noticeable change in receiver output noise when the preamplifier is switched from feed to attenuator (termination), weak signal turned "OFF", of course. The ratio of output noise change can be related to temperatures by the following formula:

$$\text{Noise Ratio (R)} = \frac{Tr + To}{Tr + Tfl + Tuniverse} > 1$$

Tr is the receiving system noise temperature. Tuniverse is the background Universe temperature which is of the order less than 10 degrees K at 1296 mc/s and higher frequency bands in the "space window". Tfl is the feed heat losses and will be assumed to be small, of the order of the Universe background temperature, or less.

Rearranging the equation to solve for Tr gives

$$T \text{ receiver system} = \frac{To - R\,Tu}{R - 1} \qquad (1)$$

where Tu is assumed to be of the order 20 degrees Kelvin maximum (background plus feed losses) and To is the ambient temperature of the matched attenuator measured with a thermometer. The conversion to the absolute temperature (Kelvin scale) is 273 + C. C is the temperature in degrees celcius.
As an example, if C = 17 deg. celcius, To is 290 degrees and with Tu = 20 and an expected receiver system noise-figure of 1 db (75 deg. K), then the expected value of R will be about 5.8 db.
If the procedure is carried out properly then the determination of the receiver noise temperature, Tr, will be dependent on how accurate To and Tu are known. Since Tu embodies the feed antenna loss temperature, Tfl, as well as the Universe background temperature, the determination is not truly complete.
When the feed temperature due to losses in the feed are more than the background temperature, lower values of the ratio R will be observed and Tr will be higher than expected by about the temperature of the feed due to losses. If this is the case in your measurements, then the feed should be suspect. Look for poor solder joints, cold solder joints, use of lossy materials in construction, faulty seams, etc. A well constructed feed made from good conductive material should not have more than 0.1 db loss (7.5 degrees K). An interconnecting cable and connectors may have as much or more loss than the feed itself and should be included as part of the antenna temperature if it is between the antenna port and the switching device.

It will be advantageous, when installing the feed back into the reflector antenna, to retain the receiver antenna-to-termination switch both as r-f overload protection, and as a simple check of system performance. When the system is in its normal EME receiving mode, a record of the noise level change when switching from termination to antenna should be made with the antenna aimed at a 'cold' place in the Universe. This level change should be used as a reference to check system performance from time to time.

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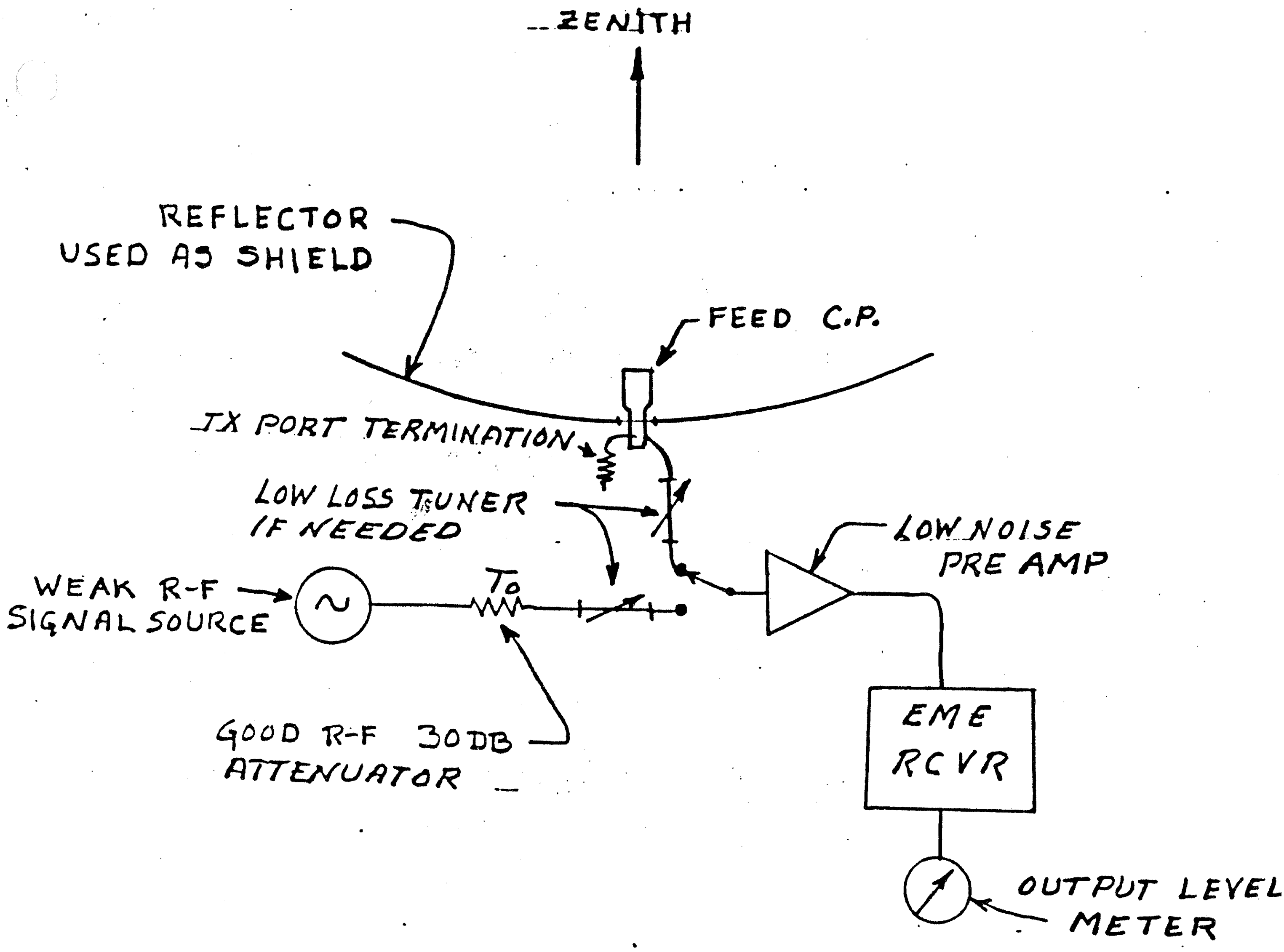
ZENITH
REFLECTOR
USED AS SHIELD
FEED C.P.
TX PORT TERMINATION
LOW LOSS TUNER
IF NEEDED
LOW NOISE
PRE AMP
WEAK R-F
SIGNAL SOURCE
To
GOOD R-F 30DB
ATTENUATOR
EME
RCVR
OUTPUT LEVEL
METER

FIGURE 1