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ABSTRACT

Work on the design, fabrication and testing of three broadband antennas is described. The antenna types are; 1) high gain constant beamwidth, 2) omnidirectional and 3) loaded conical helix. Impedance and pattern information are given for a broadband ridged horn designed for use as a feed in the constant beamwidth antenna system. Results are given on modeling tests with a pillbox designed to simulate the performance of the high gain reflector in one plane.

Work on two broadband omnidirectional antennas is described. The first is a cross plate antenna somewhat bulky physically but having promising impedance behavior. The second is a monopole with traps, the measured and calculated impedance behavior of which is excessive at several frequencies.

The results obtained in loading of bifilar helix antennas with cores and layers of ferrite powder material is described. A decrease in axial mode frequencies from 710 MHz (unloaded) to 550 MHz (loaded) regardless of layer thickness is obtained. The thinnest layer tested was 0.5" on a 4" diameter helix. Work on the loading of conical helices with capacitors and with ferrite materials is presented.

FOREWORD

This report was prepared by The University of Michigan Radiation Laboratory of the Department of Electrical Engineering under United States Army Electronics Command Contract No. DA 28-043 AMC-01263(E). The contract was initiated under United States Army Project No. 5A0-21101-A902-01-08, "Broadband Antenna Techniques Study". The work was administered under the direction of the Electronics Warfare Division, Advanced Techniques Branch at Fort Monmouth, New Jersey. Mr. Anthony DiGiacomo is the Project Manager and Mr. George Haber is the Contract Monitor.

The material reported herein represents the results of the preliminary investigation into techniques applicable to the design and development of broadband antennas.

The authors wish to acknowledge the contributions of E. Rupke, J. Brigham, A. Loudon, J. Bosel and W. Henry for the fabrication and testing of the antennas.

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I

INTRODUCTION

This contract is divided into three tasks; 1) broadband constant beamwidth high-gain antenna, 2) omnidirectional broadband antenna and 3) broadband loaded conical helix.

Under Task 1, a high-gain antenna is to be developed that covers the frequency range 1 - 10 GHz. The beamwidth is to vary less than 2 : 1 such that a relatively constant gain of 20 db above an isotropic source is achieved with a VSWR less than 3 : 1 with respect to a 50 ohm load. The investigation is to include a theoretical and experimental study of broadband, constant beamwidth, high-gain antennas. Electronic switching, electromechanical or mechanical motion to effect the constant beamwidth characteristics of the antenna are not to be considered. As a result, the constant beamwidth characteristics must be achieved employing antenna beam shaping techniques.

Under Task 2, a broadband omnidirectional antenna of the monopole or dipole configuration is to be developed which will be operational over the frequency range of 100 MHz to 1 GHz having a VSWR of less than 3 : 1 with respect to a 50 ohm load. It is desired that the configuration be as thin as possible and its overall length comparable to that of a half-wave dipole at the low end of the frequency band (100 MHz). The maximum diameter of the configuration is to be less than 20 " .

The objective of Task 3 is to design a circularly polarized antenna covering the frequencies of 50 MHz to 1.1 GHz, with a 2 : 1 reduction in size, and a maximum weight of 20 lbs. The antenna is to be a loaded conical helix. Various loading techniques are to be investigated including ferrites and dielectrics. The conical sections of the antenna may be truncated with the possibility that one may be set within the other. Cross-over networks which cause different sections to operate at different frequencies may be required.

II

BROADBAND CONSTANT BEAMWIDTH HIGH-GAIN ANTENNA

To effect the 10:1 bandwidth requirements of the high-gain antenna, a ridged horn design (Fig. 1) based on the work of Walton and Sundberg* was chosen. Figures 2 and 3 present the basic dimensions of the antenna. The details of the ridge configuration are presented in Fig. 2. To achieve the 10:1 bandwidth, the .053" gap is the most critical dimension in the design. Experimentally and theoretically, double ridged waveguides having the .053" gap along with the dimensions of Figs. 2 and 3 have exhibited a bandwidth greater than 10:1. Figures 4 and 5 provide further details as to the design of the waveguide and back cavity section of the horn. The design of the waveguide section was accomplished with the aid of Figure 6. The design of the ridges in the horn section of the feed was based on

$$Z=Z_{0\,\infty}e^{KX},\quad 0<X<\frac{\ell}{2}\quad \text{and}\quad Z=377+Z_{0\,\infty}(1-e^{K(\ell-X)}),\quad \frac{\ell}{2}<X<\ell\ ,$$

where ℓ is the overall length of the antenna and K is obtained by solving the two equations simultaneously at $X = \ell/2$. From the throat of the horn to its mid-point the ridge dimensions were determined by the first equation and for the remaining length of the horn a smooth curve was fitted to insure a gradual transition from the ridge impedance to the aperture impedance. Curve fitting was employed since the dimensions obtained from the second equation alone would require the ridge to extend into the radiating aperture of the horn. This was undesirable as it would disrupt the RF fields and produce undesirable far field patterns. The measured impedance behavior of a horn built according to the above design is presented in Figs. 7 and 8 and shows, over the 10:1 bandwidth, a standing wave ratio of 3 : 1 or less with respect to 50 ohms. Figure 9 is the plot of the VSWR vs frequency over the 10:1 band of the antenna.

Patterns have also been recorded in both the E- and H-plane for the ridged horn across the frequency band and typical results are shown in Figs. 10 - 15.

To obtain design information on the parabolic reflector to be used with the ridged horn a pillbox has been employed. Its aperture is 4' x 1" and it simulates the pattern from a segment of the full reflector, making it possible to determine the desired F/D ratio. Several feeds for the pillbox have been constructed which simulate the primary feed aperture size. The pillbox configuration was selected for the preliminary design because of its simple construction and low cost.

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*Walton, K. L. and V. C. Sundberg, "Broadband Ridged Horn Design," Microwave Journal, pp. 96-101 (March 1964).

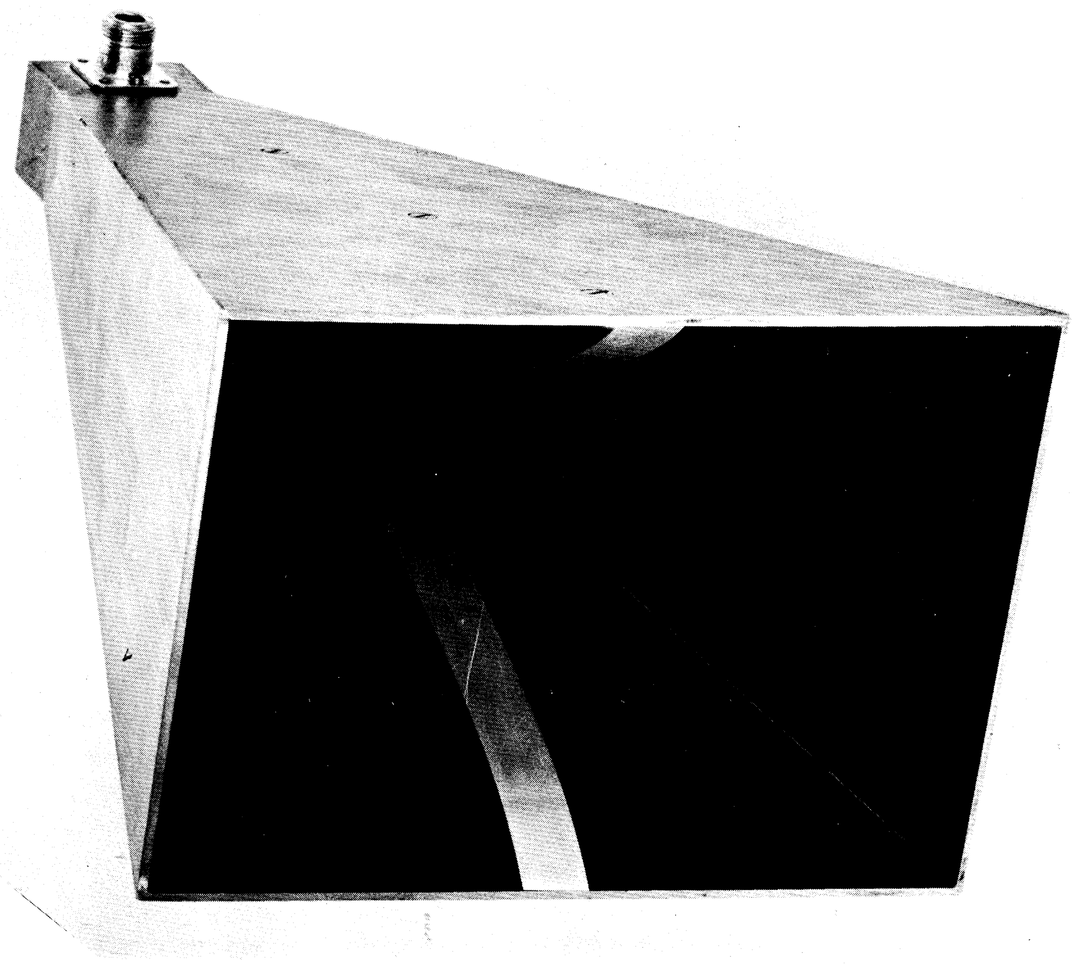

FIG. 1: RIDGED HORN FEED FOR BROADBAND HIGH GAIN ANTENNA

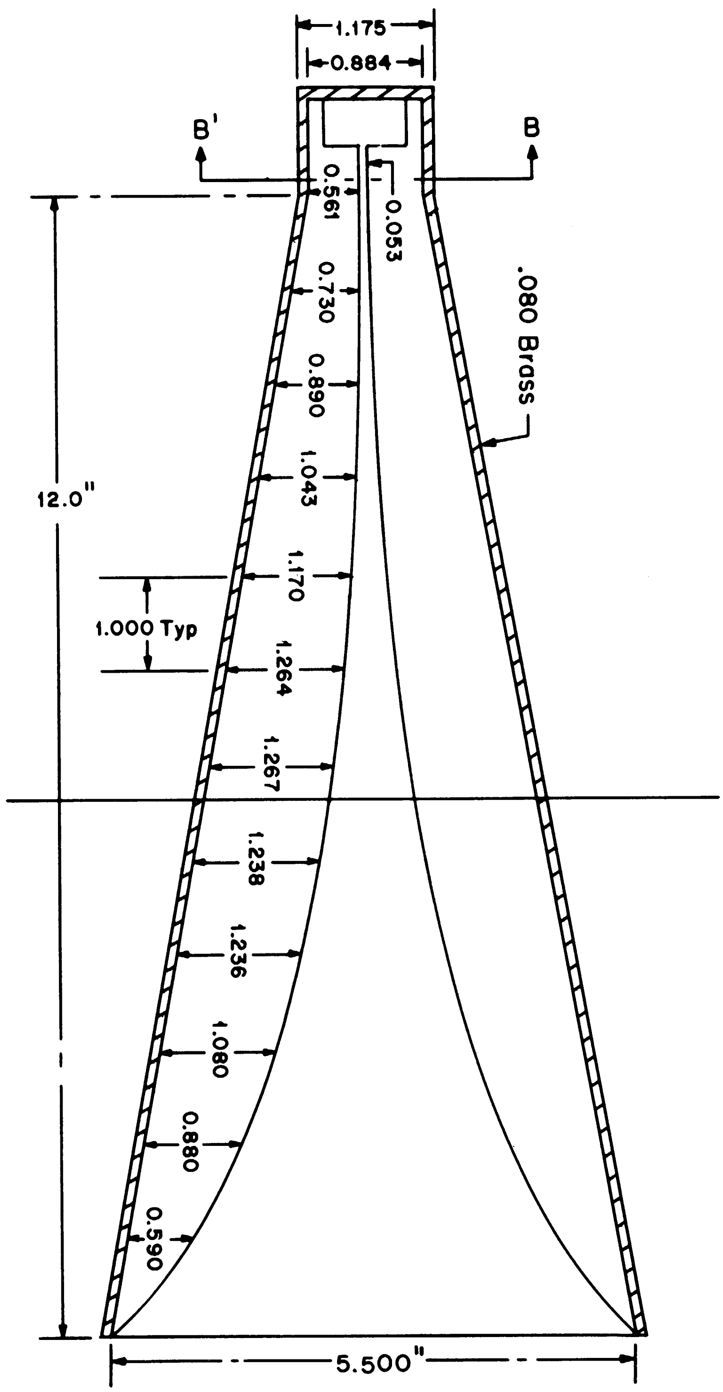

FIG. 2: HORN RIDGE DETAILS (Tolerances $\pm$ .003)

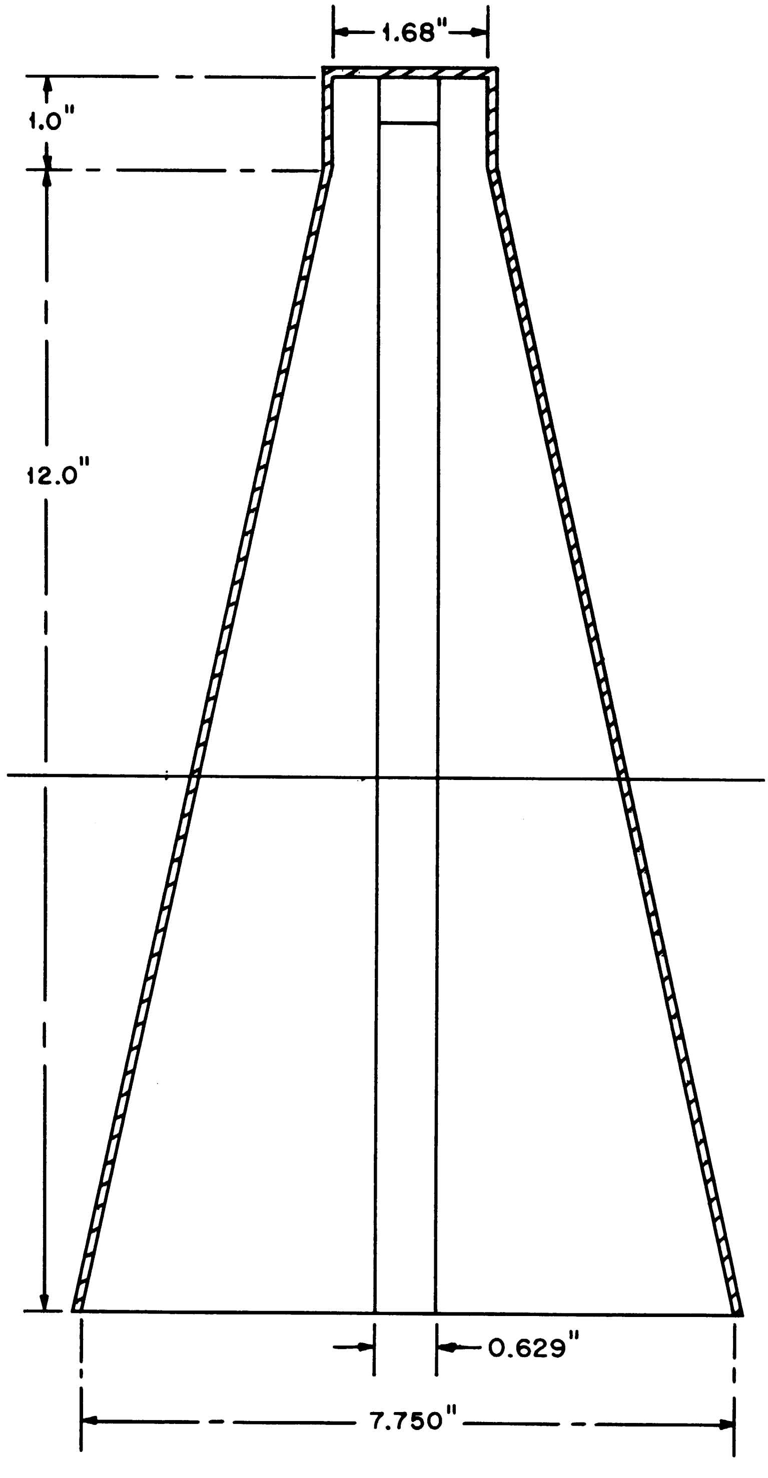

FIG. 3: RIDGED HORN

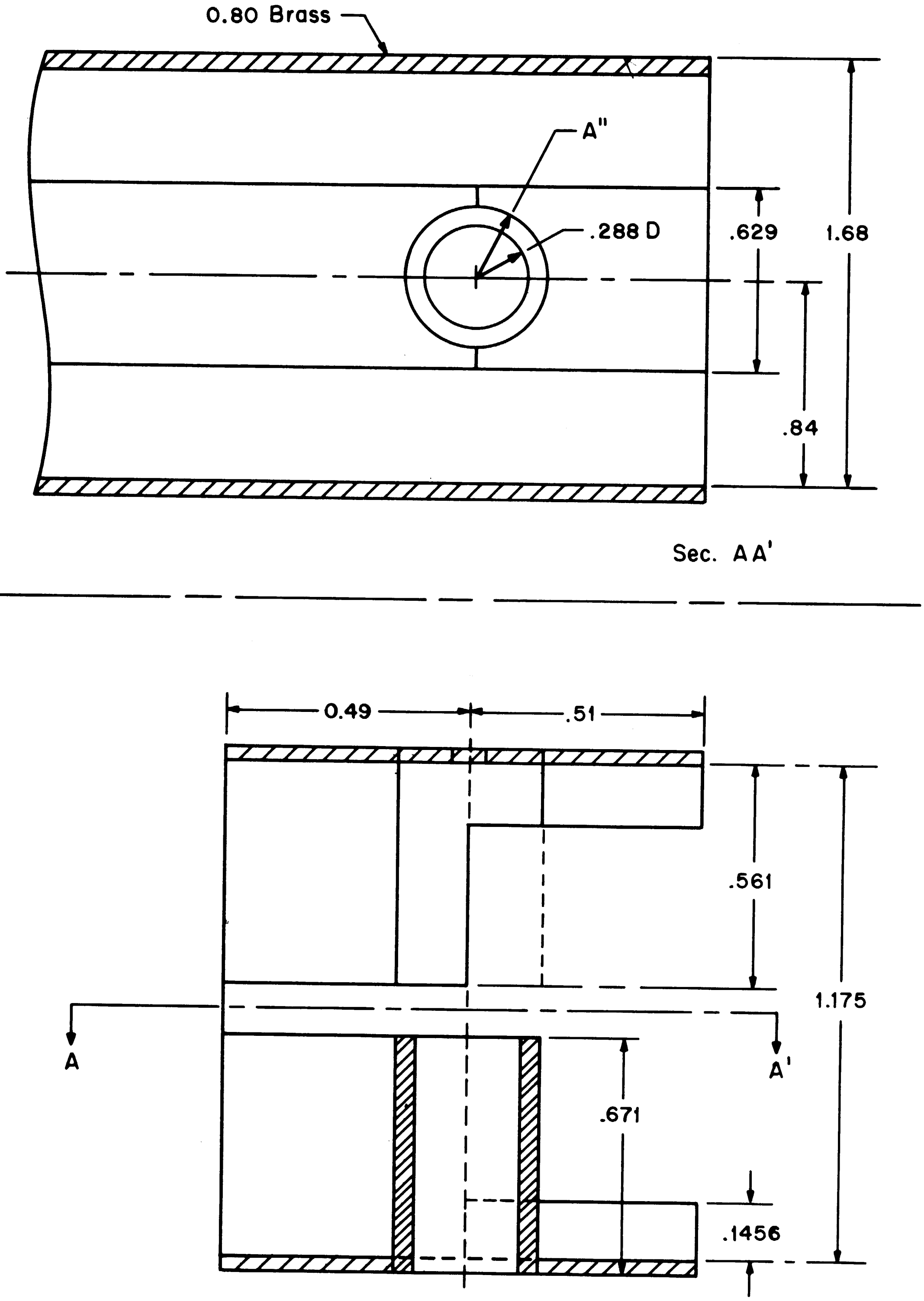

FIG. 4: BACK CAVITY OF RIDGED HORN

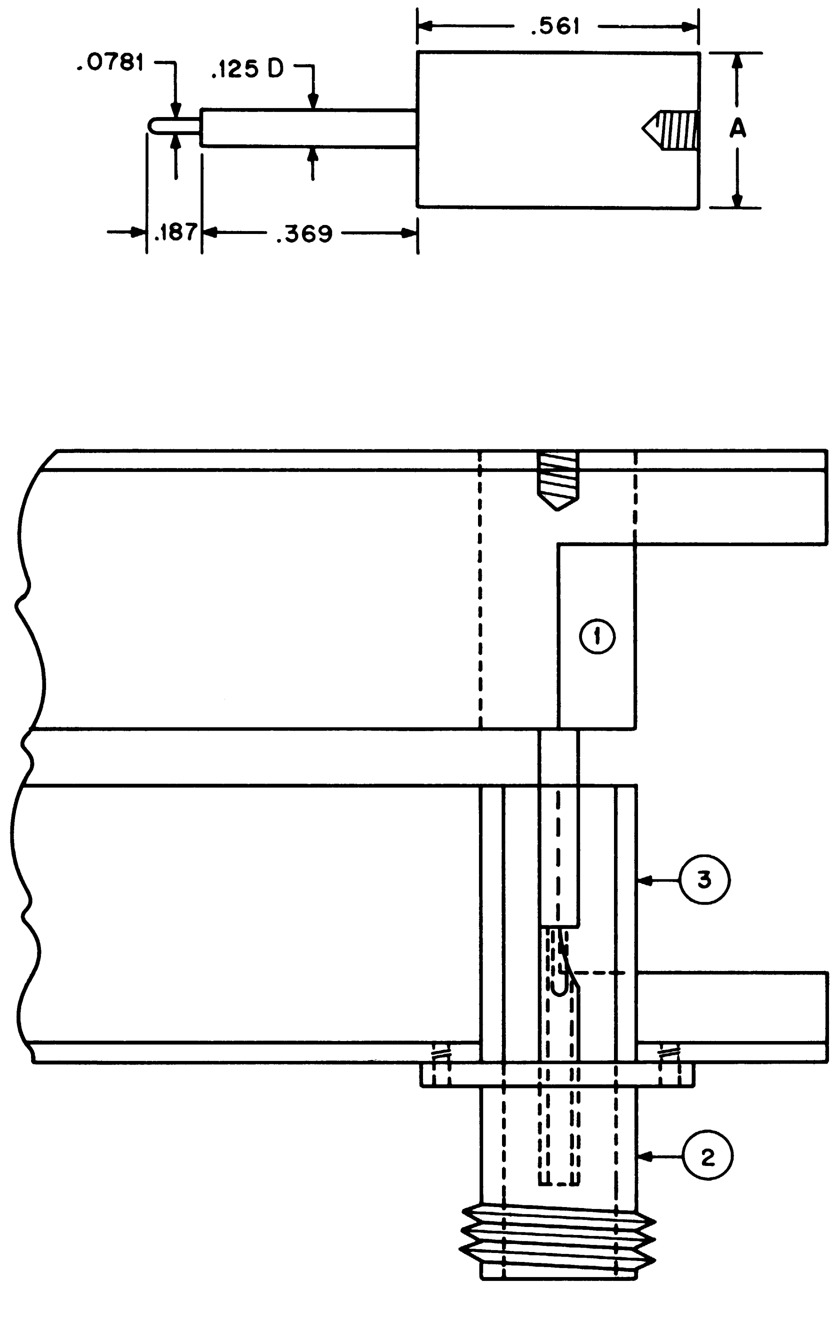

FIG. 5: BACK CAVITY PROBE

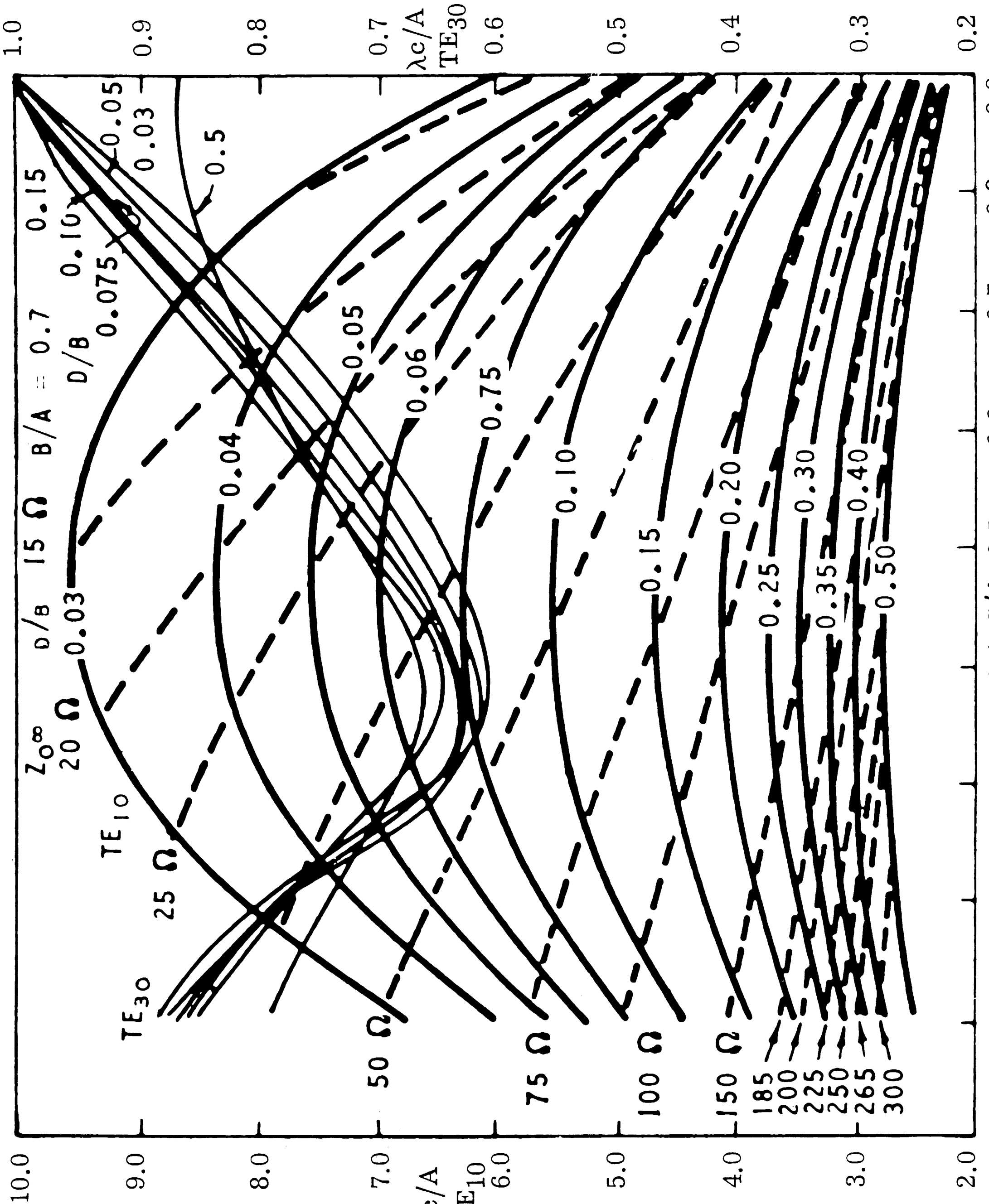

FIG. 6: TE_{10} AND TE_{30} MODE CUTOFF WAVELENGTHS IN DOUBLY-RIDGED WAVEGUIDE. $Z_{0\infty}$ IS ALSO SHOWN B/A = 0.7 (Walton and Sundberg, 1964).

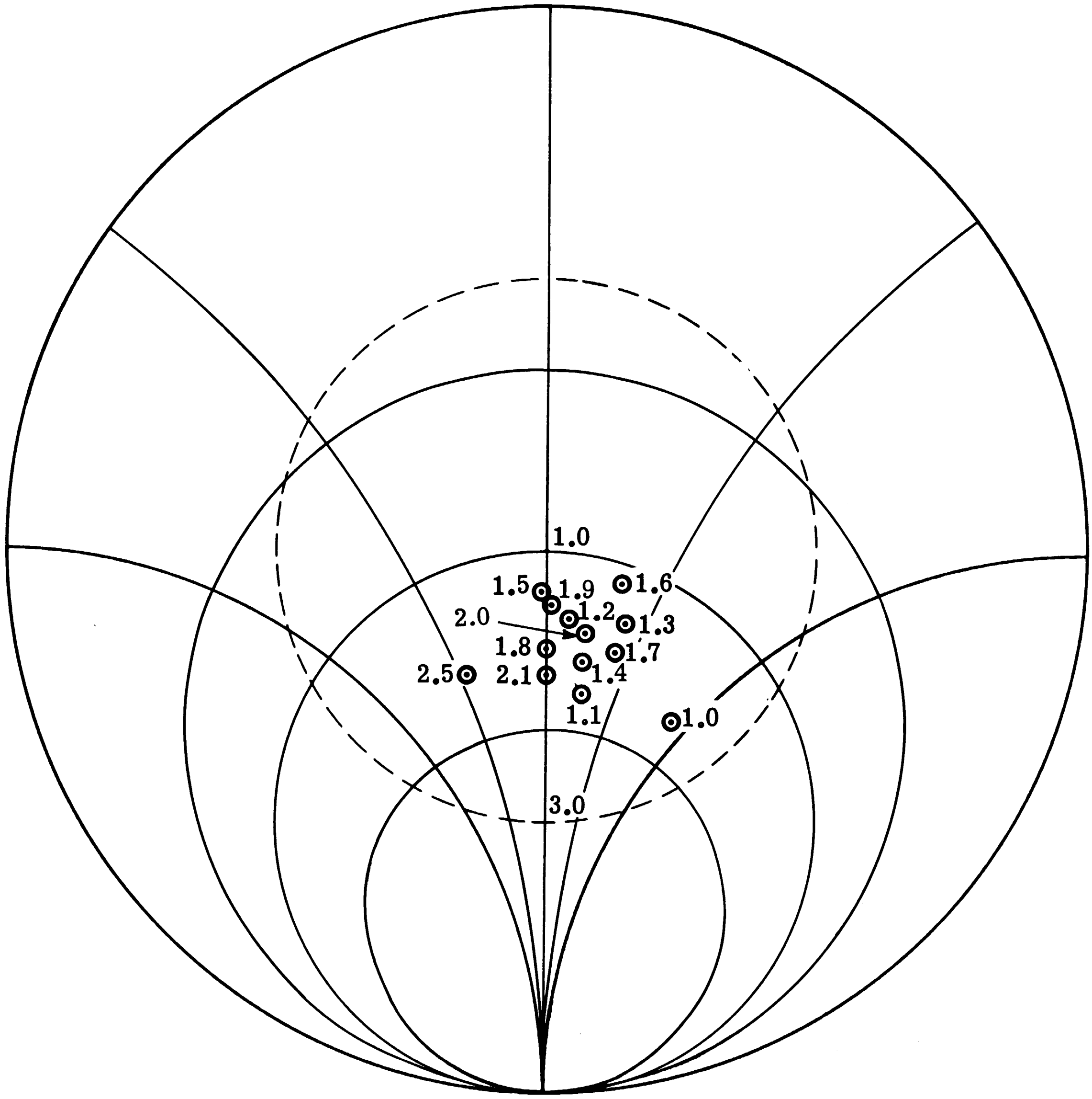

FIG. 7: RIDGED HORN IMPEDANCE, 1 - 2 GHz

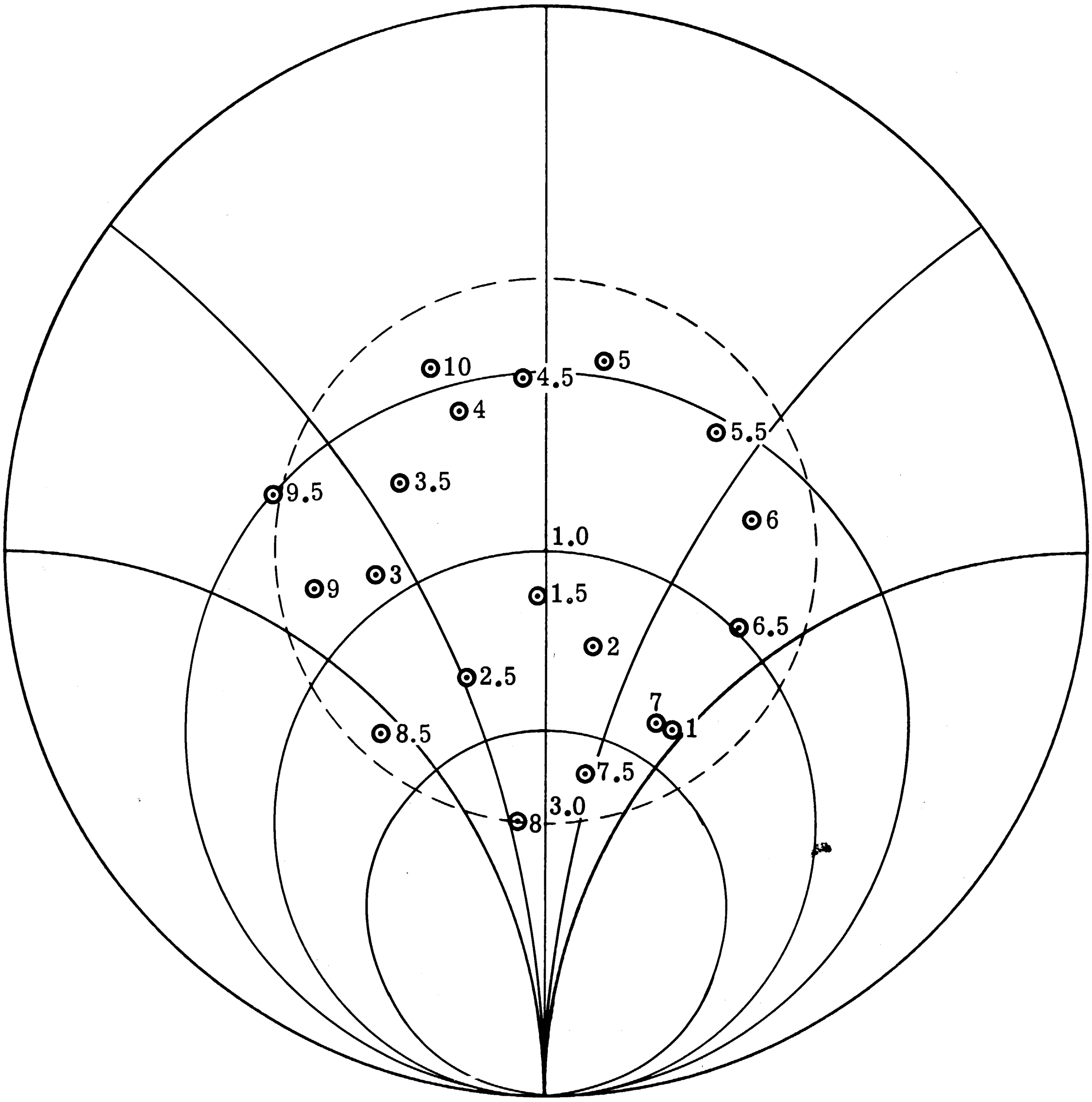

FIG. 8: RIDGED HORN IMPEDANCE, 1 - 10 GHz

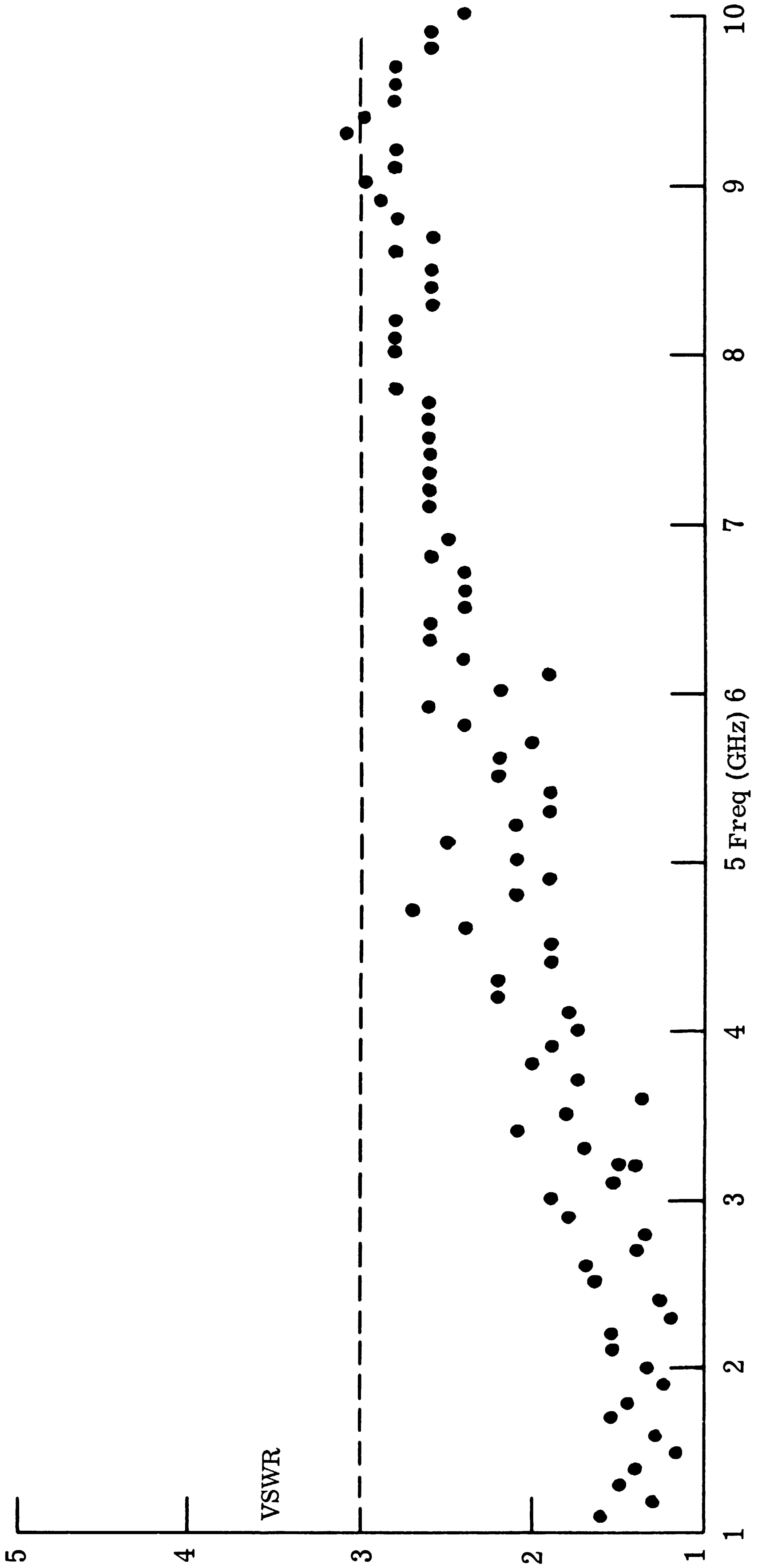

FIG. 9: 10 : 1 RIDGED HORN VSWR VS FREQUENCY

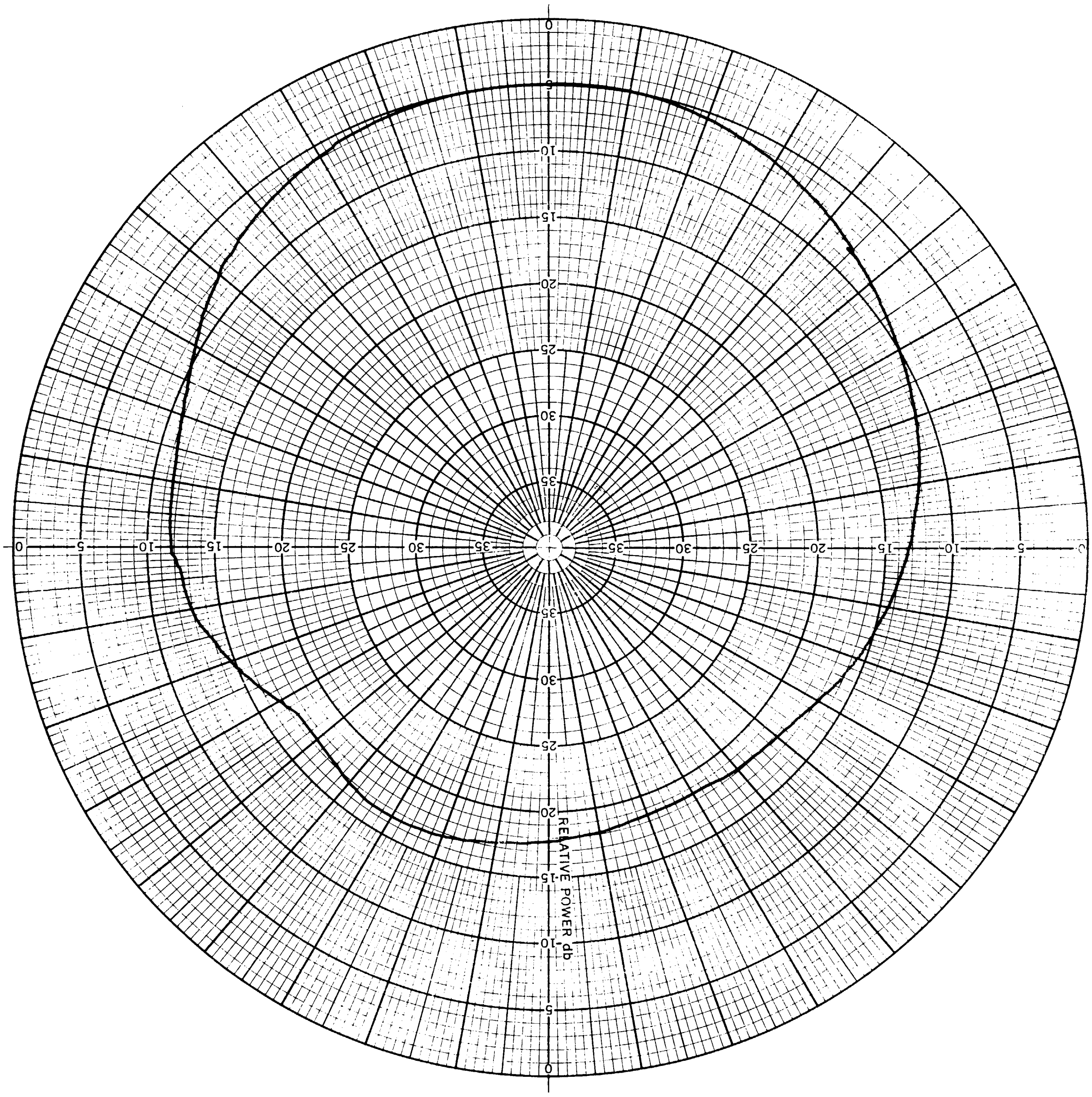

FIG. 10: E-PLANE PATTERN FOR RIDGED HORN, F=1.0 GHz.

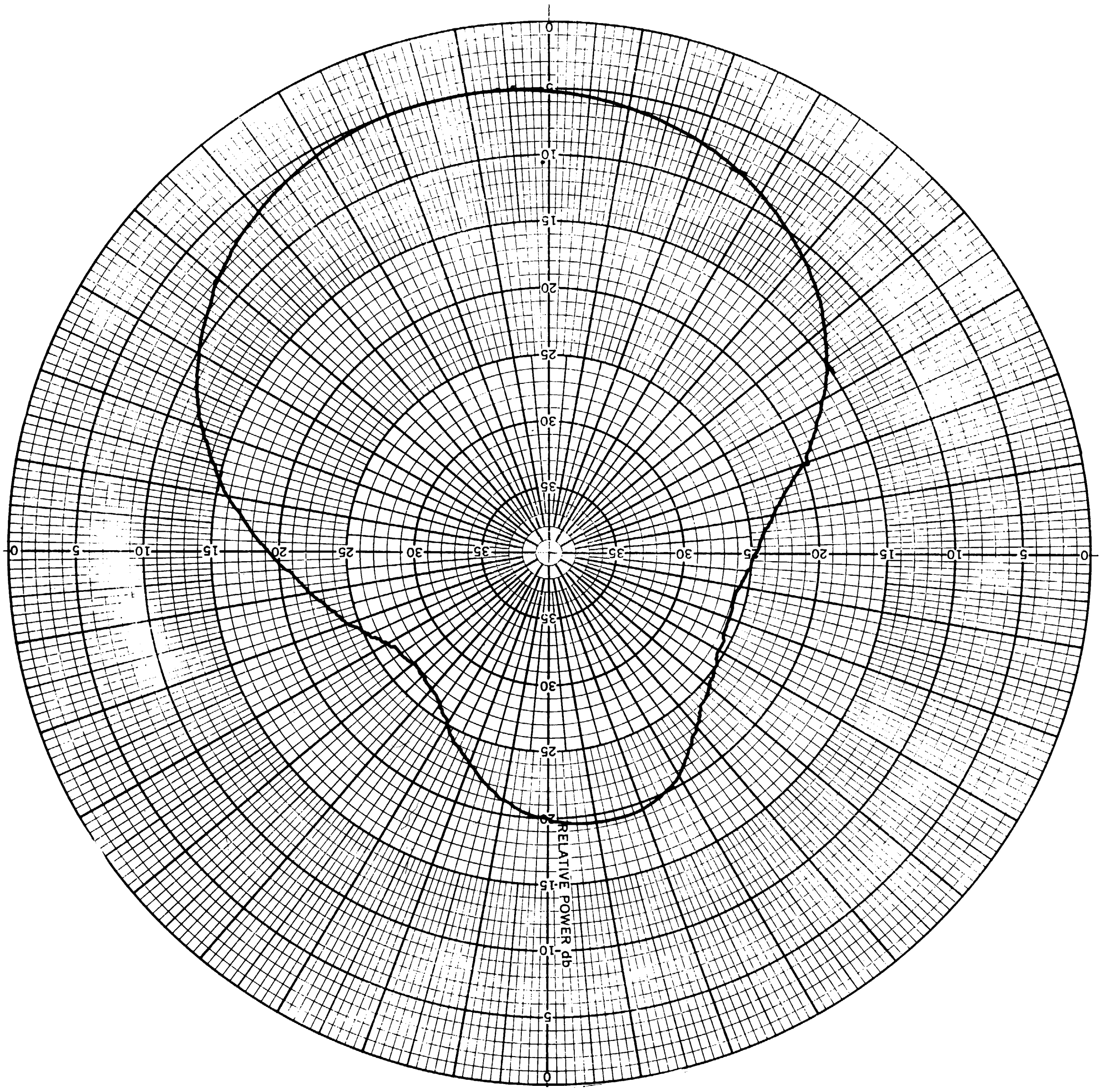

FIG. 11: H-PLANE PATTERN FOR RIDGED HORN, F=1.0 GHz.

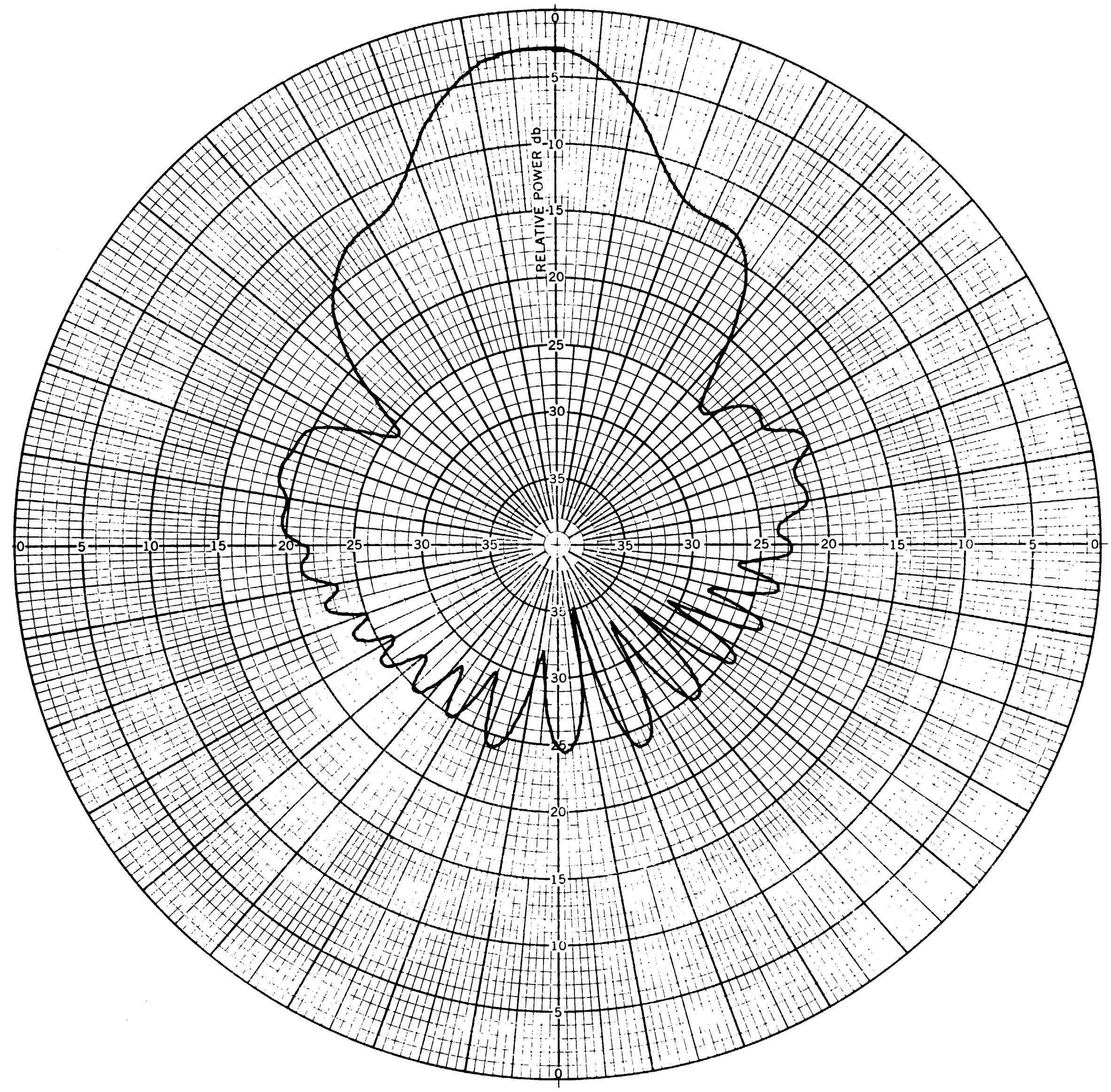

FIG. 12: E-PLANE PATTERN FOR RIDGED HORN, F = 5.3 GHz.

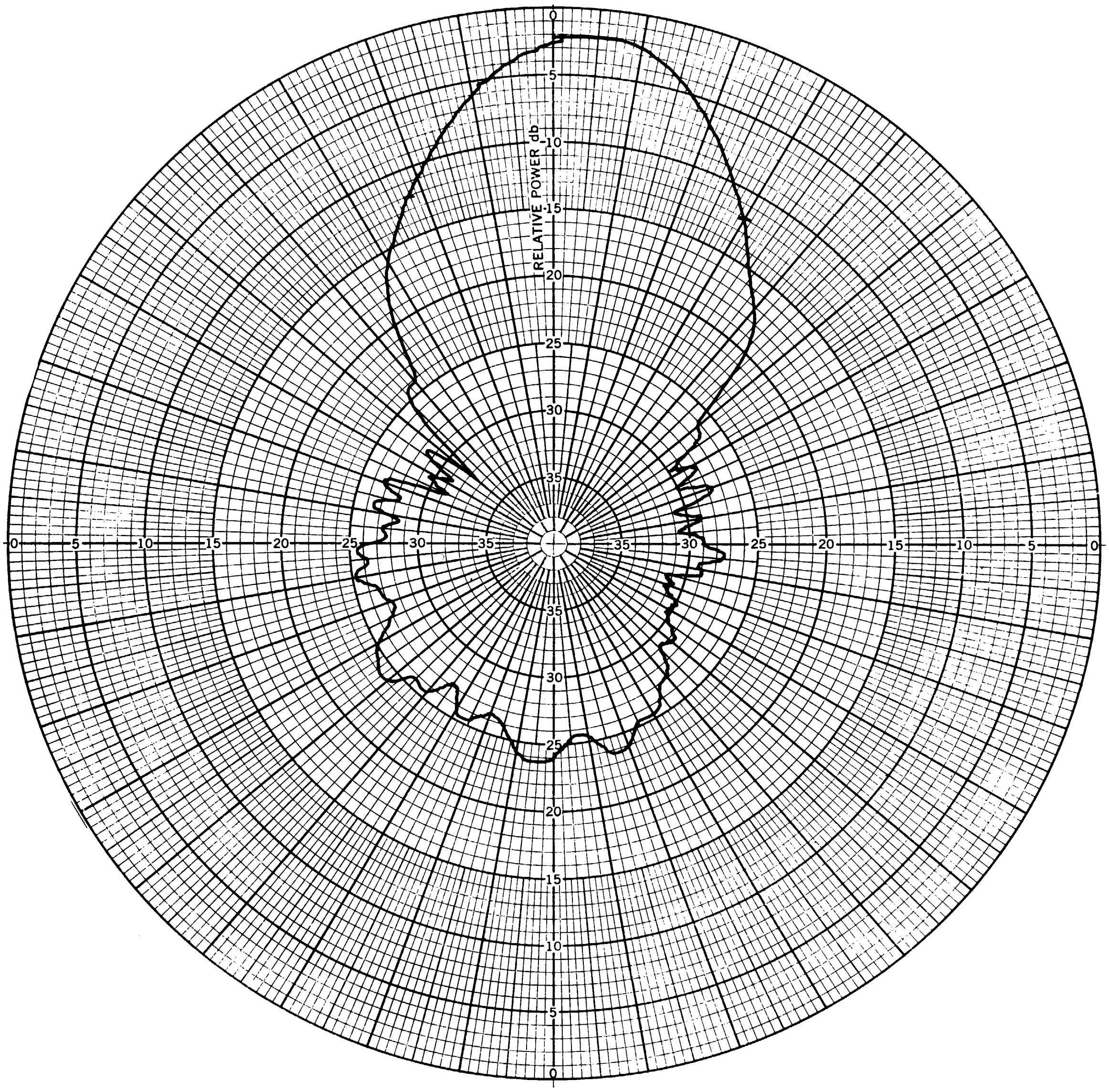

FIG. 13: H-PLANE PATTERN FOR RIDGED HORN, F=5.3 GHz.

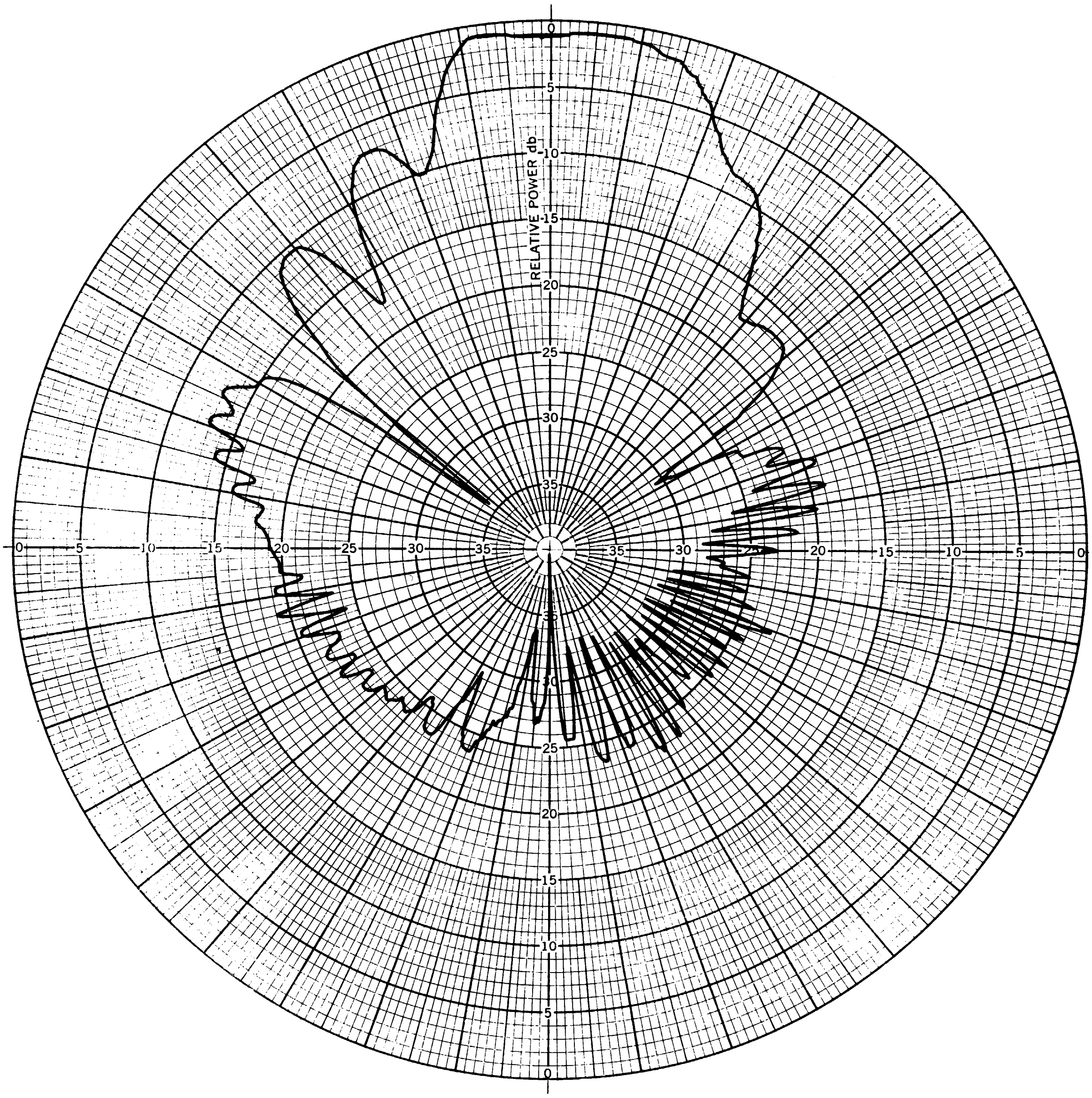

FIG. 14: E-PLANE PATTERN FOR RIDGED HORN, F = 10.0 GHz.

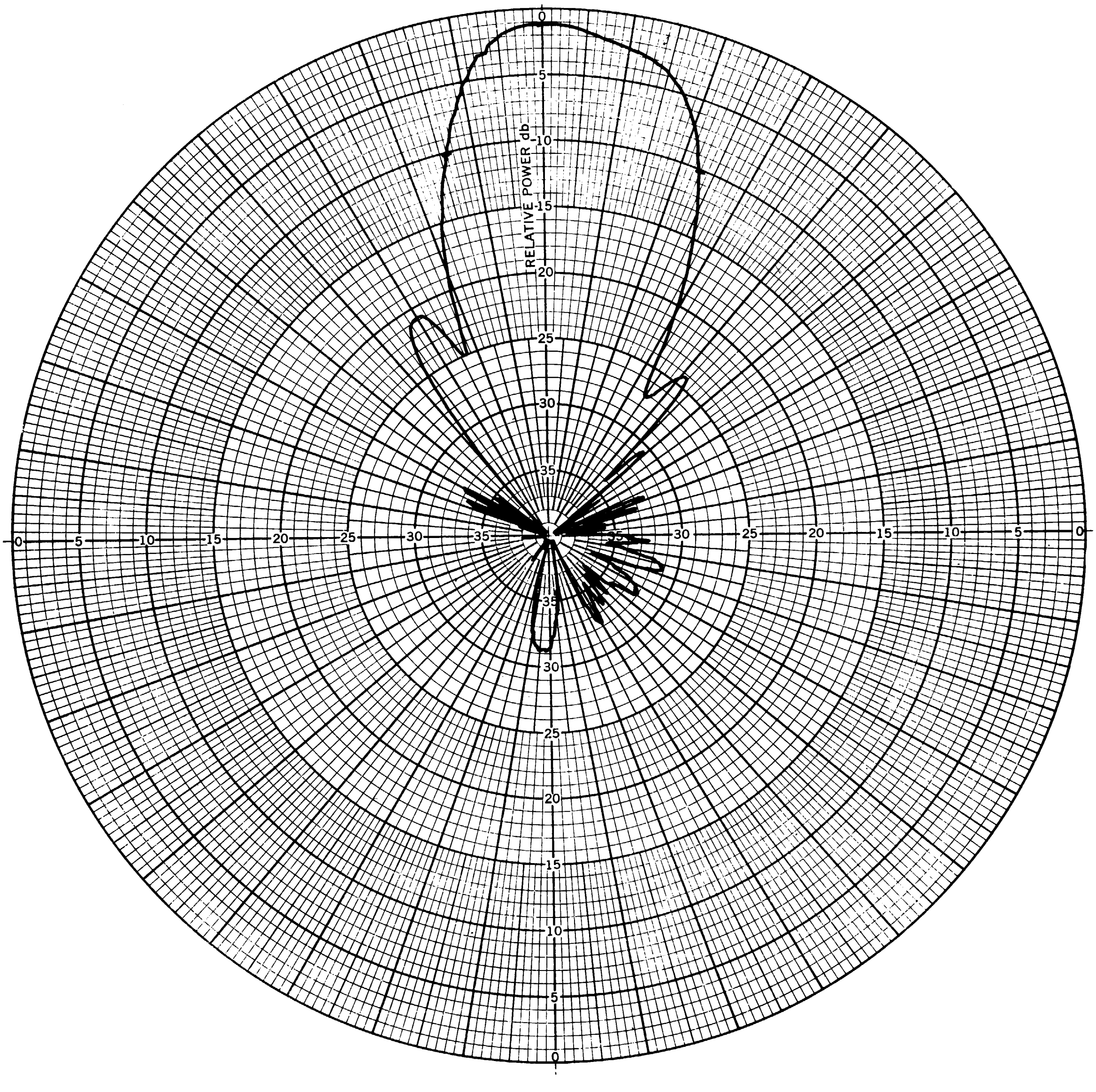

FIG 15: H-PLANE PATTERN FOR RIDGED HORN, F = 10.0 GHz

Early pillbox patterns displayed high back and side lobes. The back lobes have been lowered and the pattern greatly improved by attaching flanges to the radiating aperture. These flanges tend to reduce the excited field believed to be the result of currents induced by the long wave effects on the surface of the pillbox (Fig. 16). Repositioning the primary feed reduced the sidelobes and improved the pattern.

A set of primary feed patterns obtained from the pillbox feeds has been used to aid in a further analysis of the required aperture dimensions. An angle equal in degrees to the measured 10 db width of the feed was placed over a sketch of the pillbox, the vertex being on the focal point. The section of the pillbox reflector subtended by the sides of the angle was used to determine the effective radiating aperture for that particular horn or frequency. The effective aperture was then used to calculate the beamwidth of the far field pattern; the results are shown in Table I. Good agreement has been obtained between experimental and calculated beamwidths.

By using the above procedure to predict the half-power beamwidth of the field reflector from the ridged horn pattern data, it has been shown that the original design of the horn has too small an aperture. This is not too surprising and is not considered to be a problem. The original horn has served its purpose in showing the feasibility of obtaining satisfactory impedance performance over a 10 : 1 bandwidth.

TABLE I: MEASURED AND CALCULATED PILLBOX BANDWIDTHS

Frequency (GHz)	Feed Beamwidth 10 db (deg.)	Calculated Beamwidth 3 db (deg.)	Measured Beamwidth 3 db (deg.)
1.20	126	15.1	13
2.50	82	10	7
5.3	43	10	7
10.0	27	8.25	6

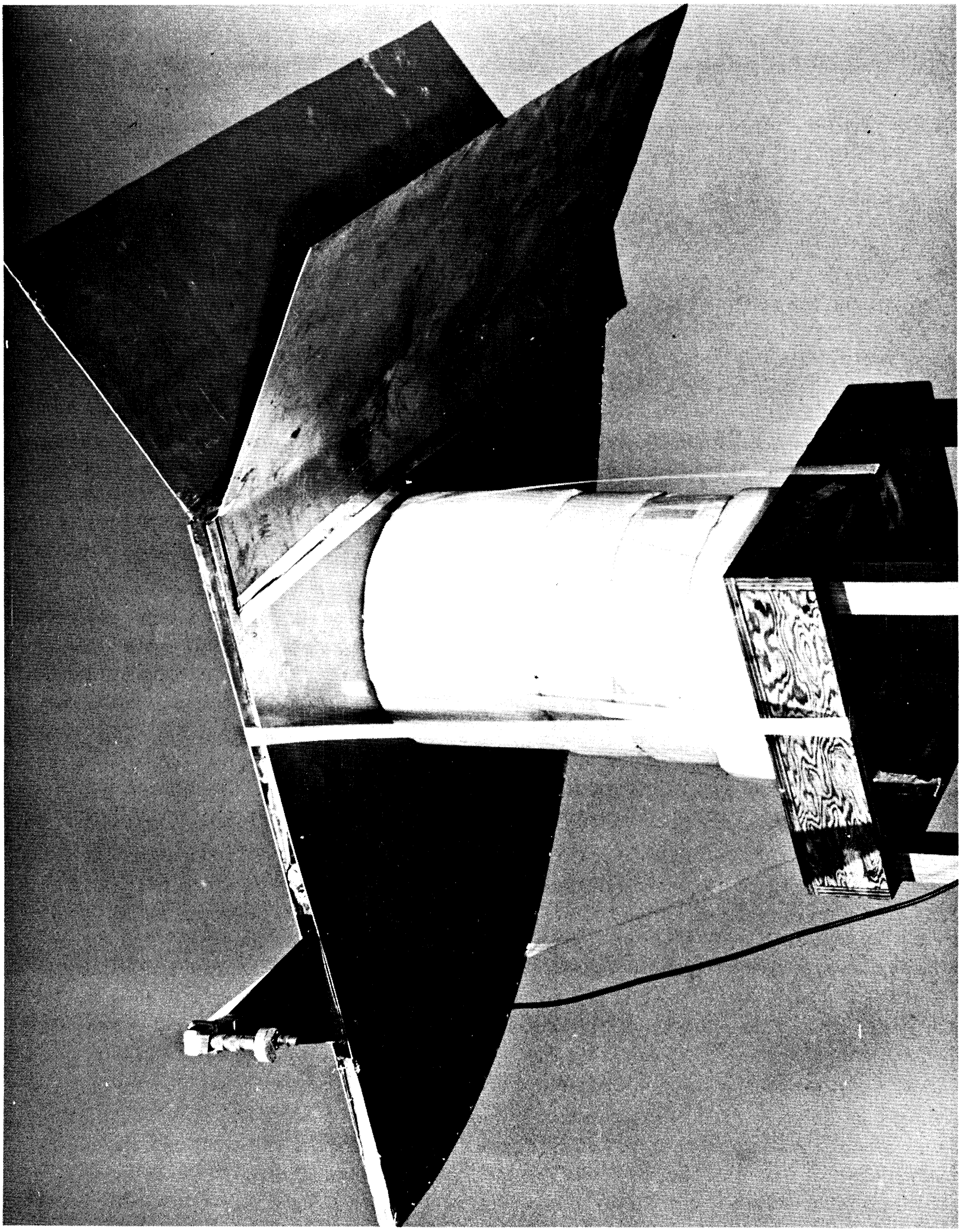

FIG. 16: PILLBOX WITH FLANGES

III

OMNIDIRECTIONAL BROADBAND ANTENNA

Several antenna types have been considered for this problem and two models have been assembled and tested. A major portion of the time spent on this task has been devoted to experimental data, however, some theoretical work has been performed and will be discussed below.

The first antenna to be considered was the cross plate configuration shown in Fig. 17. This antenna was first reported on by Lamberty * and consists of four plates attached to a central mounting post. The theoretical analysis of this antenna configuration would be difficult and therefore no attempt has been made to do this. However, it is believed that at the lower frequencies the currents induced on the plates cause the entire antenna to radiate much as a fat monopole. At the higher frequencies the spacing between the plate and the ground plane functions as a slot antenna. The VSWR of the present configuration has been measured and found to be constant (approx. 4: 1) across the band as shown in Fig. 18.

During an early visit by the Contract Monitor, we learned that there was considerable interest in the development of a slim dipole configuration to satisfy the broadband omnidirectional requirements. A configuration which has been in use for a number of years by radio amateurs appears to meet this requirement. This antenna employes traps to effect the broadband widths of interest to the radio operator, however, the amateurs are generally interested in operating the antenna in a relatively narrow frequency range within specified bands. That is, the antenna may function over a 10 : 1 band of frequencies but will be operative at only four distinct frequencies.

A simple trap monopole configuration (Fig. 20) was designed and fabricated. This antenna consists of a quarter-wave monopole (3") and three traps (3", 6" and 12"). The first section is a fat $\lambda/4$ monopole at 1 GHz. The remaining three sections are traps tuned for the fundamental, second and third sub-harmonics of the monopole.

The traps were of a coaxial configuration (center conductor O. D. = 0.500" and outer conductor I. D. =2. 63") shorted at one end and designed to be a quarter wavelength long at the trap fundamental, second and third sub-harmonics. The impedance appears as an open circuit $1/4\lambda$ from the shorted section so that the antenna section below is not affected by the remaining traps. The measured impedance of the antenna is shown in Fig. 21. The VSWR is plotted in Fig. 22 and is seen to be oscillating with high peak values.

* Lamberty, B. J., "A Class of Low Gain Broadband Antennas, " IRE Wescon Conv. Record, Pt. 1, 251-259 (1958).

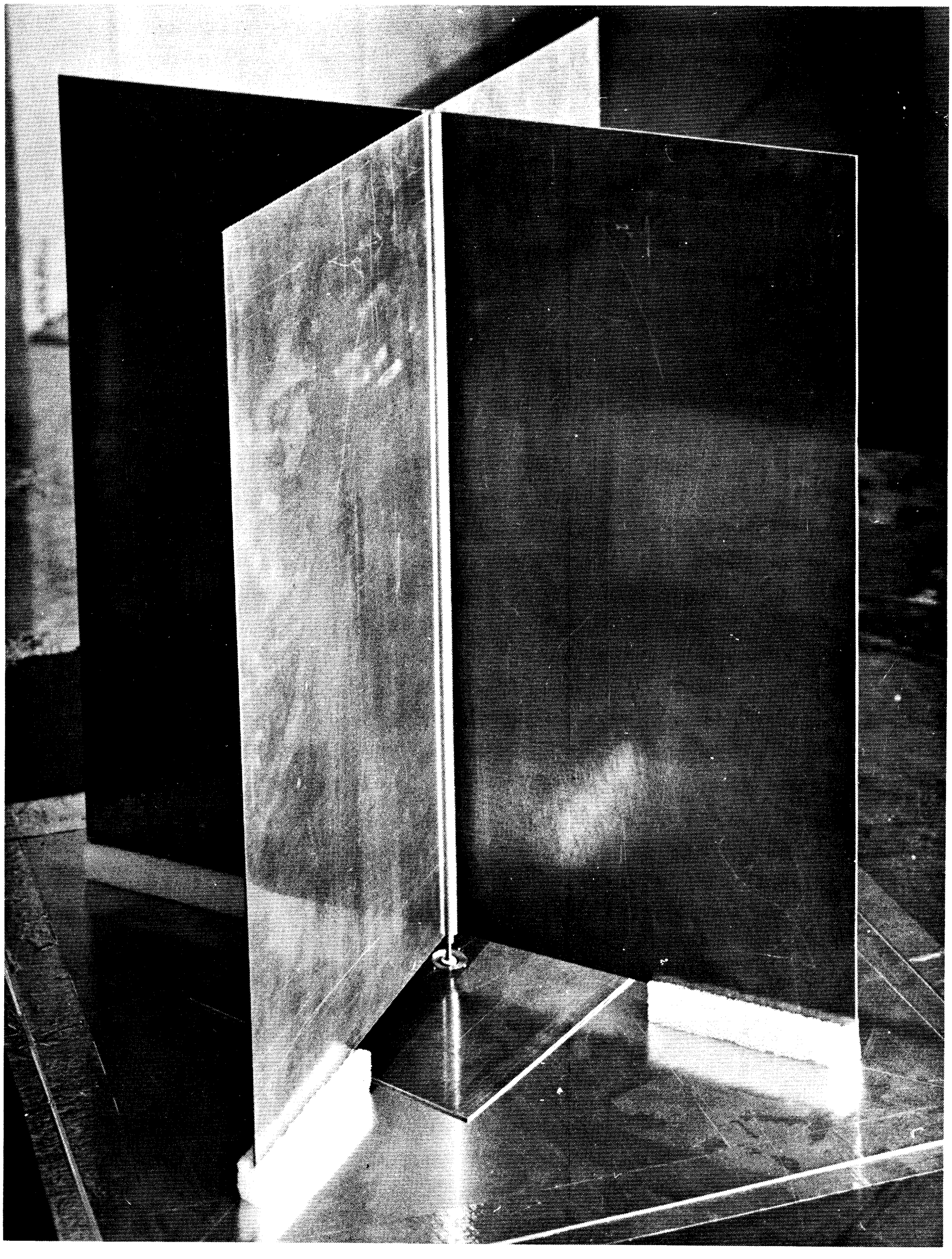

FIG. 17: CROSSED PLATE ANTENNA (20"x20"x20")

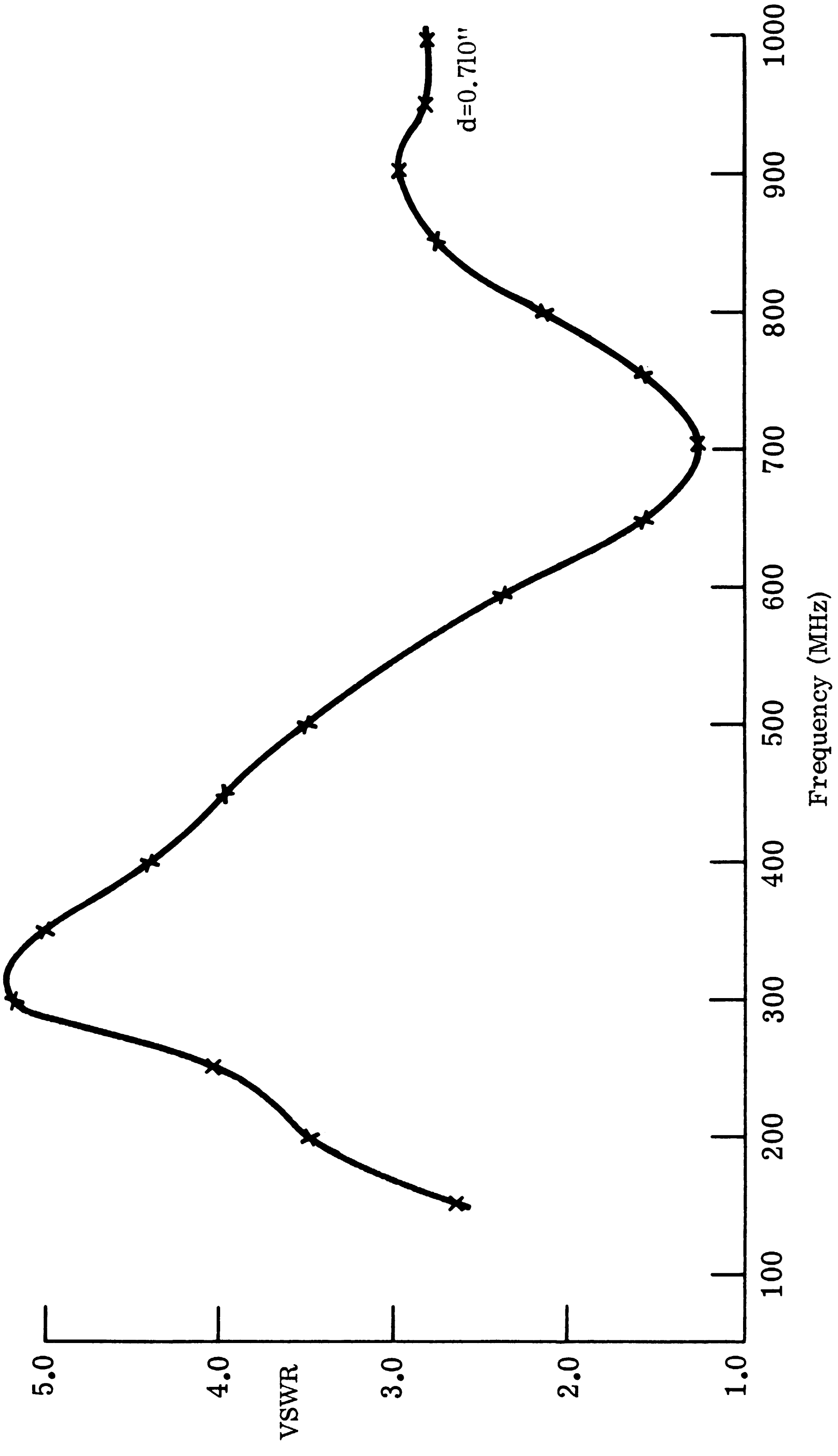

FIG 18: CROSSED PLATES VSWR VS FREQUENCY

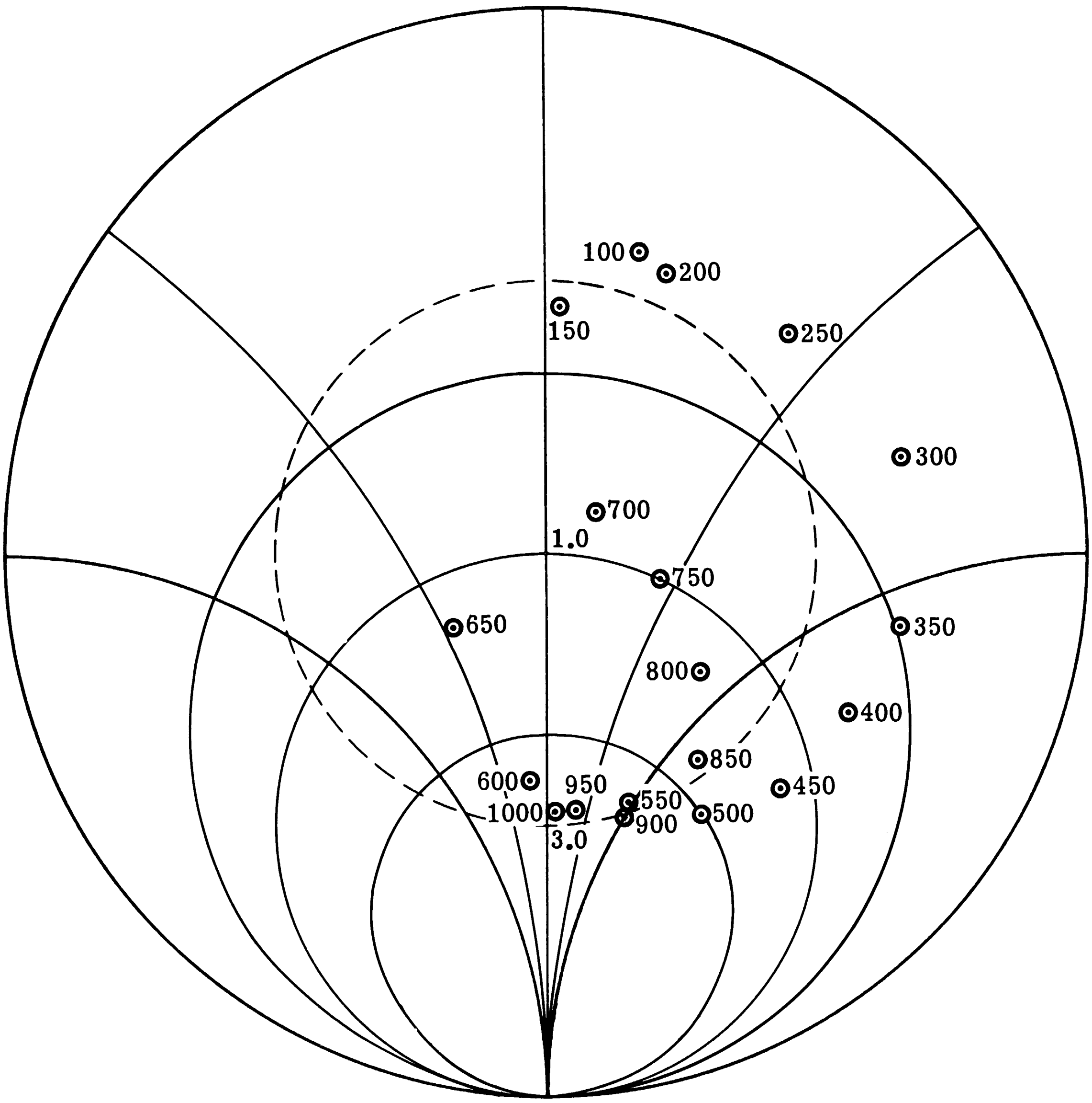

FIG. 19: CROSSED PLATE IMPEDANCE MEASURED AT BASE OF ANTENNA, 100 - 1000 MHz.

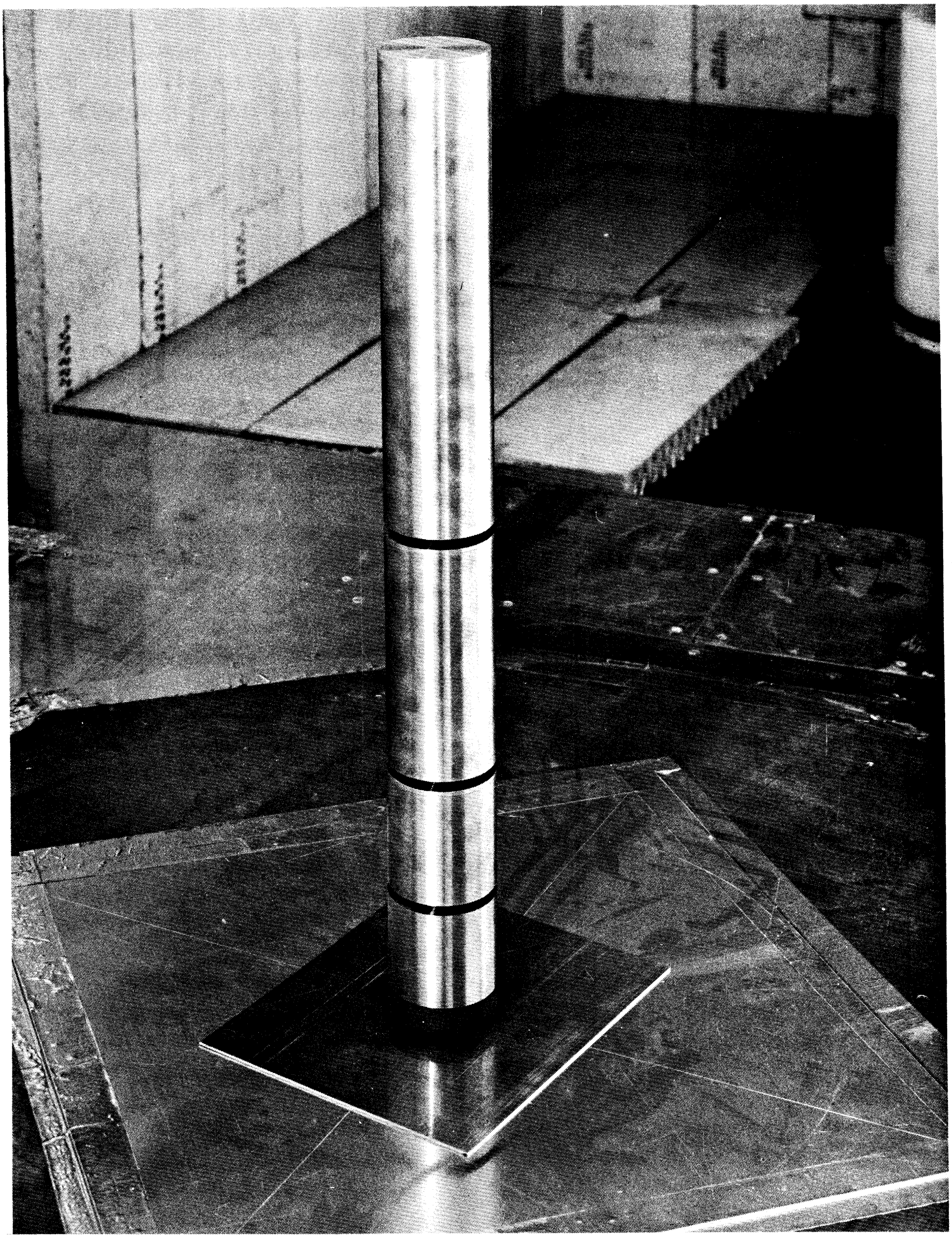

FIG. 20: TRAP MONOPOLE ANTENNA (2.75" x 24.41").

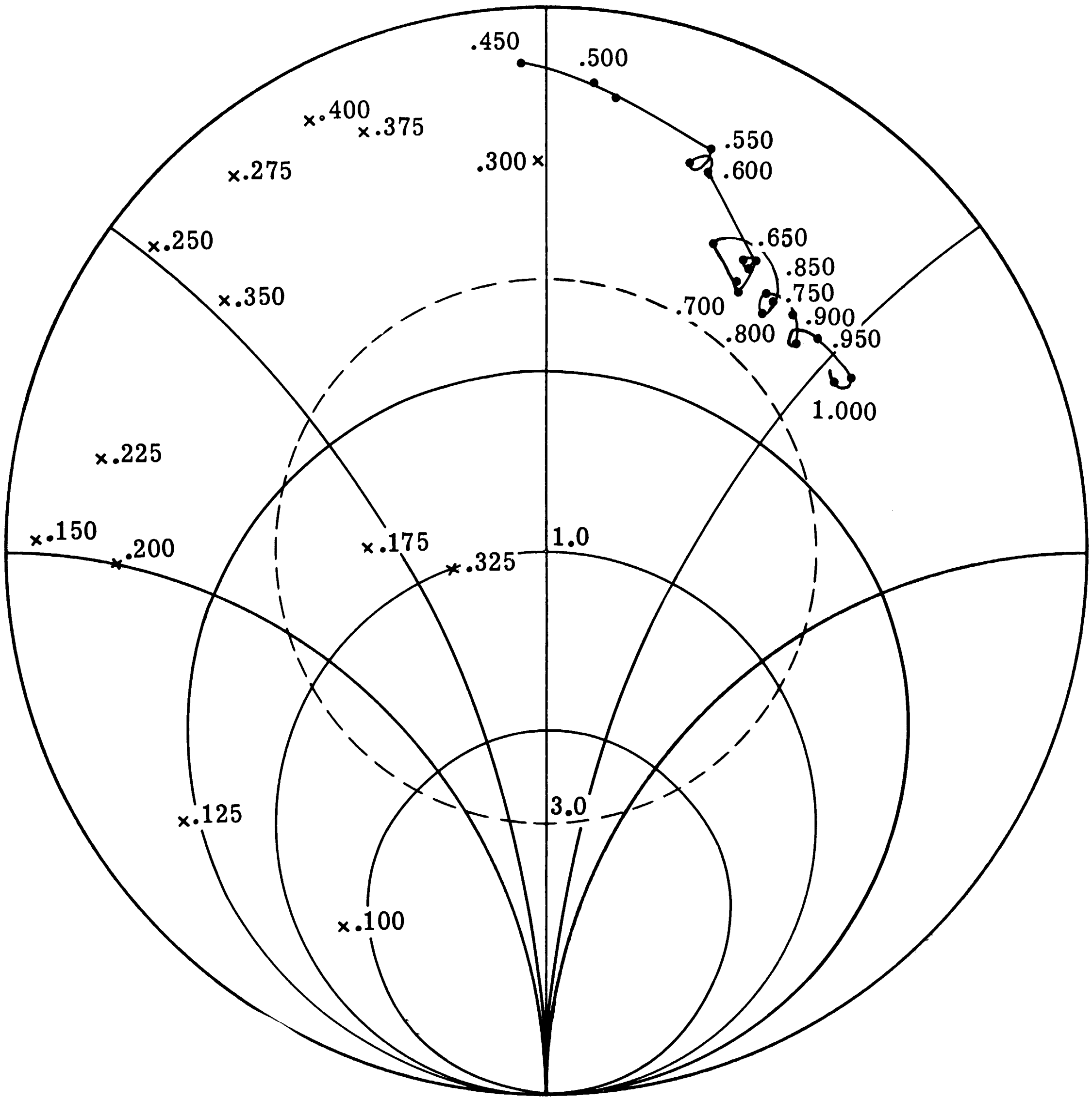

FIG. 21: TRAP MONOPOLE IMPEDANCE MEASURED AT BASE OF ANTENNA, 100 - 1000 MHz.

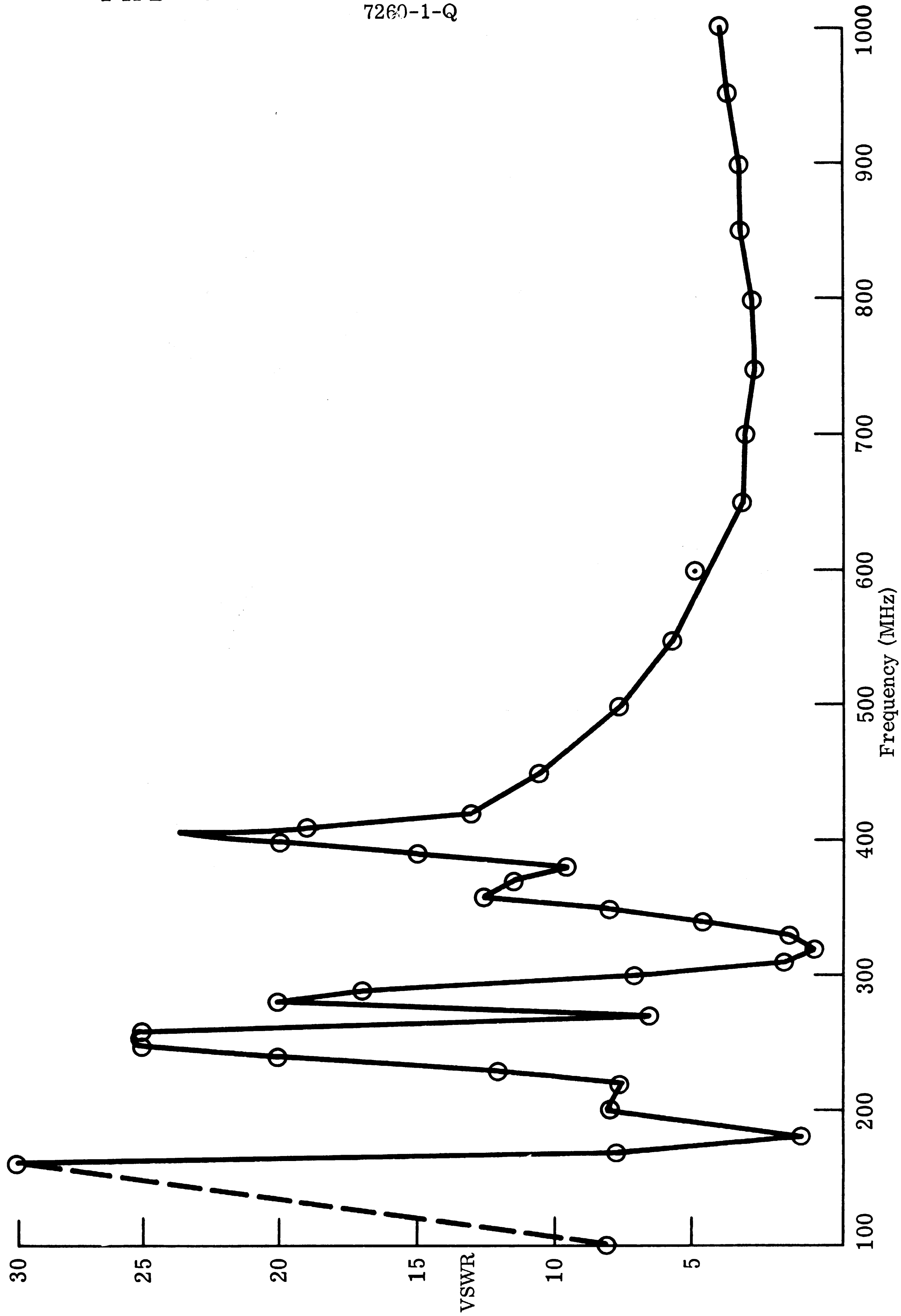

FIG. 22: TRAP MONOPOLE VSWR VS FREQUENCY

To understand the effect of the traps and explain the poor behavior, the monopole was analyzed with a single trap. This analysis is performed in three stages. First, for the 1000 - 700 MHz range, the antenna is considered to be a simple monopole with the trap above providing reactive loading. From 700 - 350 MHz the trap is considered to be a poor radiator as well as a reactive load. For this case the total antenna length includes the trap's electrical length. For the 350 - 100 MHz range the entire antenna is assumed to be radiating such that the total structure appears to be approximately $1/4\ \lambda$ long. Here the trap is considered to be a part of the antenna and provides no loading.

Due to large diameter-to-length ratio of the monopole section, theoretical impedance data was not readily available. Experimental data was used for the high frequency analysis. The trap reactance was added to the experimental monopole impedance as though the two impedances were in series.

The normalized experimentally determined impedance for the fat monopole is $0.29+j\,0.335$. The impedance of the trap is given by

$$Z_t = Z_o \frac{Z_r + Z_o \tanh d}{Z_o + Z_r \tanh d} \quad . \tag{1}$$

Since Z_r is a short circuit this reduces to

$$Z_t = Z_o \frac{0 + Z_o \tanh d}{Z_o + 0 \tanh d} = Z_o \tanh d \quad . \tag{2}$$

and Z_o=99.5 ohms. The hyperbolic function may be expanded in the form

$$\tanh(\alpha + j\beta) = \frac{\sinh(\alpha + j\beta)}{\cosh(\alpha + j\beta)} = \frac{\sinh\alpha\cos\beta + j\cosh\alpha\sin\beta}{\cosh\alpha\cos\beta + j\sinh\alpha\sin\beta} \quad . \tag{3}$$

Assuming the transmission line is lossless, this reduces to

$$\tanh(j\beta) = \frac{j\sin\beta}{\cos\beta} = j\tan\beta \tag{4}$$

at 800 MHz the trap impedance (referred to the antenna base) becomes

$$\frac{Z_t}{Z_o} = j\tan(\beta d) = -j\,Z\,(0.59) \tag{5}$$

where

$\beta = 2\pi/\lambda$,
d = distance from short to antenna base ; 5.95"
λ = 14.35"

Adding this trap impedance (5) to the monopole impedance gives

$$\frac{Z_r}{Z_o} = 0.29 + j\,0.34 - j\,0.59 = 0.29 - j\,0.25 \qquad (6)$$

A VSWR of 3.7:1 is obtained as compared to a measured VSWR of 3.65:1.

In the second case (700 - 350 MHz) the monopole plus twice the trap length gives a length-to-diameter ratio of normal value making it possible to use the experimental data of Brown and Woodward*. Here it is felt that the trap behaves like a slow wave structure. The trap reactance was then added to the Brown and Woodward impedance data.

At 400 MHz the electrical length of the antenna is

$$L = \frac{3.00''\text{ (monopole only)}+2\text{x}2.95''\text{ (trap length)}}{29.5''\text{ (free space wavelength at 400 MHz)}} = 0.305\,\lambda \qquad (7)$$

$$= 108.5 \text{ electrical degrees.}$$

From Figs. 5 and 6 of Brown and Woodward, the impedance for a monopole (with no flare) is 220+j 110 or 4.2+j 2.2 normalized to 50 ohms. At this frequency the trap impedance may be obtained from their Fig. 5;

$$\frac{Z_t}{Z_o} = +j\,3.18 \quad . \qquad (8)$$

The total impedance is then:

$$\frac{Z_r}{Z_o} = 4.20 + j\,2.20 + j\,3.18 = 4.20 + j\,5.38 \qquad (9)$$

which has a VSWR of 11 as compared to the measured VSWR of 15.

The third area (350 - 100 MHz) uses only the electrical length of the radiating structure with no trap loading. Again the trap was treated as a slow wave structure.

---

*Brown, G. H and O. M. Woodward, Jr., "Experimentally Determined Radiation Characteristics of Conical and Triangular Antennas," RCA Review, p. 430 (Dec. 1952)

For this frequency range we may neglect the trap impedance due to the strong radiation of the trap structure. The electrical length for 200 MHz is

$$L = \frac{3.00+2x2.95}{59.04''} = .151\lambda = 54.3 \text{ electrical degrees} \tag{10}$$

From the data of Brown and Woodward the impedance of a monopole whose electrical length is 54.3^o is found to be 18-j90 or 0.36-j1.8 when normalized to 50 ohms. This gives a VSWR of 12.0 as compared to the measured VSWR of 14.

Figure 23 shows the experimental and calculated VSWR as a function of frequency for the monopole and one trap. Correction for cable loss of the experimental data is not included; it is believed that this correction will bring the two curves into much closer agreement.

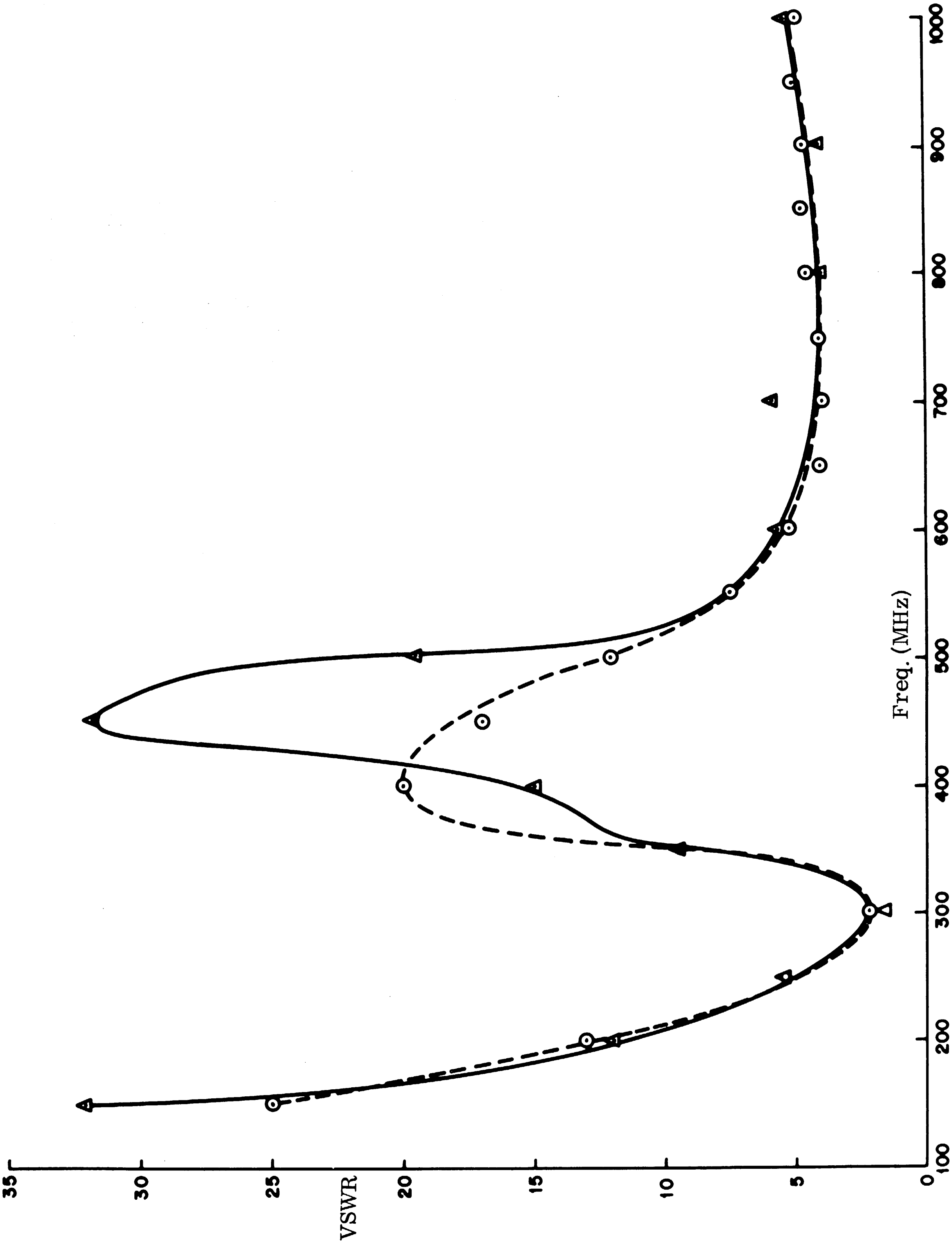

FIG. 23: EXPERIMENTAL AND THEORETICAL VSWR FOR MONOPOLE AND ONE TRAP
CALCULATED △ EXPERIMENTAL ⊙

IV

BROADBAND LOADED CONICAL HELIX

In the study of the loaded conical helix antenna during this report period attention has been given to a number of related problems.

4.1 Ferrite-loaded Helices

The ferrite-loaded helix is simpler to evaluate both theoretically (because it is a periodic structure) and experimentally because far field patterns of a helix are a more sensitive indication of the antenna parameters than those of a conical helix.

Recently published work in the USSR on the effect of solid loaded helical antennas has been helpful in planning experiments. Some theoretical work in the literature* indicates that for a solid dielectric loaded helix antenna, a reduction factor in antenna size for axial mode radiation is

$$\frac{\text{size before}}{\text{size after}} = \sqrt{\frac{\epsilon+1}{2}} \quad . \tag{11}$$

Another major question to be answered is the relationship between the thickness of material layer and the size reduction factor. Experiments were performed to ascertain the effect of material thickness and permeability on the loading of axial mode helical antennas as well as to verify the reduction equation (11). Figure 24 shows the relative effectiveness of both full core and 0.5" internal radial layer of ferrite powder on the backfire radiation frequency of a 4" diameter bifilar helical antenna. It is clear that a substantial reduction in size is obtained with this ferrite powder having properties of $\mu = 2.2$ and $\epsilon = 3.8$. In addition it appears that a 0.5" thick loading (.023λ at the center frequency, 550 MHz) of the helical antenna causes substantially the same size reduction as full core loading. This is a significant result in the feasibility of material loading of a large antenna with a maximum weight reduction and efficiency.

4.2 Solid Ferrite Loaded Conical Helix.

In previous work, personnel of the Radiation Laboratory have demonstrated an approximate 2:1 reduction in size with the application of ferrite layers of powder to conical helix antennas. Powdered ferrite of a given composition has lower permeability and permittivity than solid ferrite of the same composition. Therefore, it was desired to test material with a higher μ and ϵ to ascertain whether further

*Shestopalov, V. P., Bulgakov, A. A. and Bulgakov, B M., "Theoretical and Experimental Investigations of Helix-Dielectric Aerials," Radio Eng. and Elect. Phys., 6, No. 7, 159-172 (1961).

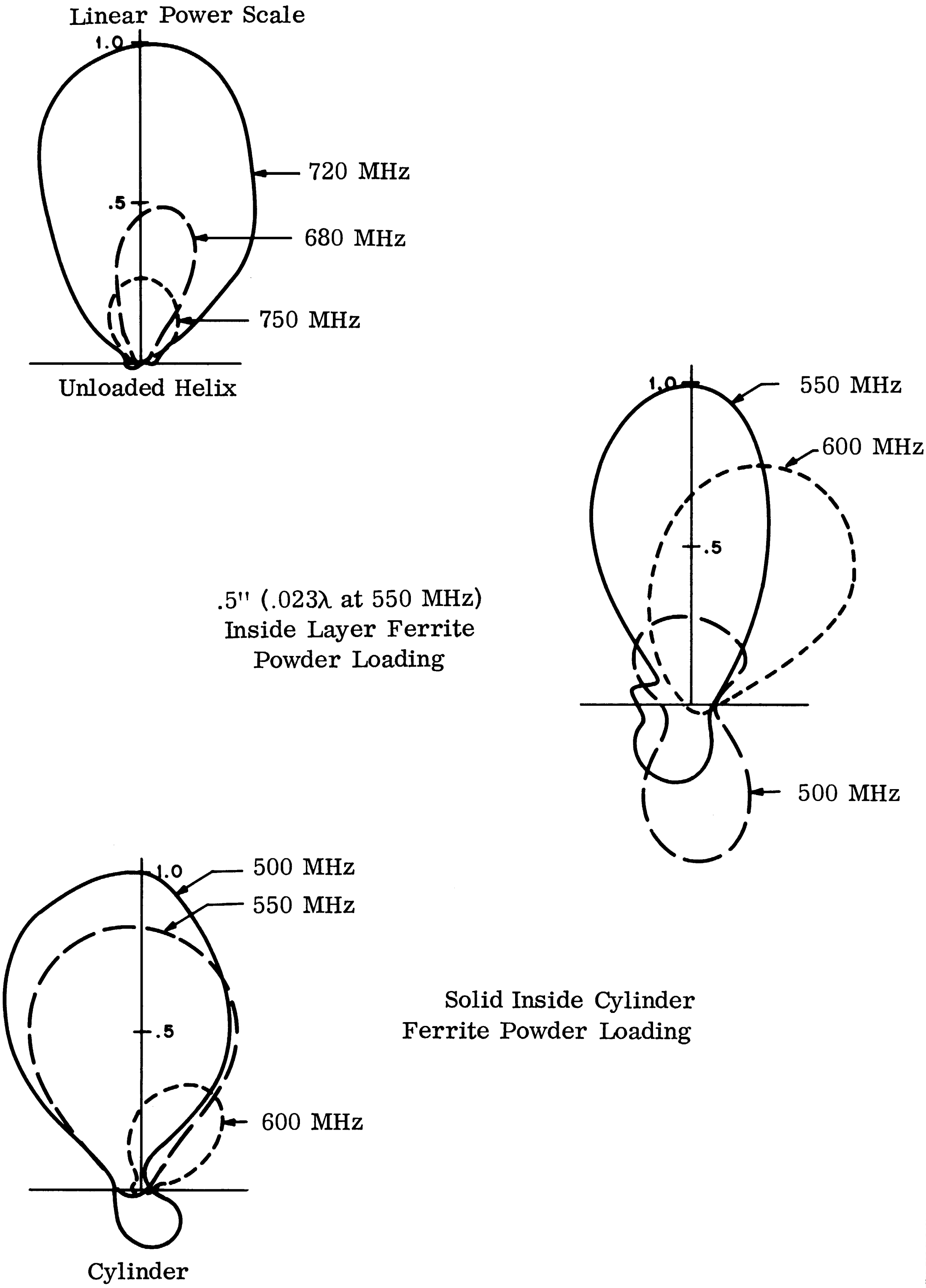

FIG. 24: EFFECT OF FERRITE POWDER LOADING OF 4" DIA. BIFILAR HELIX.

reduction in size was possible. Since low loss solid ferrite bars with rather high μ and ϵ (μ = approx. 7 and ϵ = approx. 12) were available, a small conical helix (base diameter = 5") was loaded with the solid ferrite bars by taping them inside the windings and in contact with the windings. Preliminary tests of this antenna showed a reduction of more than 2 : 1 in size according to VSWR characteristics. However, pattern measurements show a large reduction in the desired single forward lobe. The infinite feed balun used in this conical helix also failed to perform satisfactorily due to large fields at the base of the antenna. This indicated the absence of good backfire radiation from a small active region of the antenna. Further work on this loaded antenna is anticipated.

4.3 Discrete Capacitor Loading

An attempt was made to obtain slow wave modes through the use of discrete capacitors or inductors in place of continuous solid material loading with the spacing between the loading elements, small compared to a wavelength. Such loading would require very little weight or extra space and if successful could be extended to still lower frequency antennas in the future. In this experiment capacitors at every eighth of a turn were soldered between arms of a bifilar conical helix designed to operate down to 100 MHz without loading.

The theoretical calculation of the capacitors began with the rather drastic assumption that the current wave down the two arms of a bifilar helix is similar to the well known two wire transmission line. This assumption ignores the coupling between different turns of the helix and appears to be verified by the fact that the current wave velocity along the helix wires is close to the velocity of light which is the characteristic velocity of the TEM wave on a two wire transmission line.

The two wire transmission formulas commonly known may then be used for finding capacitance and velocity of propagation of a loaded transmission line. If D and d are the distance between wires and wire diameter, respectively, then capacitance per meter of line is given by

$$c = \frac{24.1}{\log_{10} \frac{D\sqrt{2}}{d}} \ \mu\mu f \quad . \tag{12}$$

The velocity of propagation is given by

$$v = \frac{1}{\sqrt{\ell c}} \ (\ell = \text{inductance/meter}) \quad . \tag{13}$$

The velocity of propagation (and thus wavelength) of the current wave around the helix must be reduced by approximately 2 : 1, which reduces the helix diameter by 2 : 1 and maintains an axial mode with the helix circumference equal to one wavelength on the transmission line. We may do this by increasing either the distributed series inductance or parallel capacitance by 4 : 1, by equation (13). Using (12) for the capacitance, and a ratio D/d equal to 1000 representative of our thin wire helix, we find the capacitance per meter of unloaded line is approximately $3\mu\mu$ f/meter. The logarithm makes the capacitance very insensitive to the exact D/d chosen. Thus the loaded line capacitance should be 4:1 greater, or $12\mu\mu$ f/meter, making the additional capacitance to be added equal to approximately $9\mu\mu$ f/meter.

If discrete capacitors are used to increase the distributed capacitance, they must appear as continuous loading and thus must be spaced much closer than 1 λ apart. A capacitor every $\lambda/8$ was chosen. For a wavelength of 300 MHz, the wavelength (and radiation zone circumference) is one meter. Approximately $1.1\mu\mu$f must be added each $\lambda/8$ in order to achieve $9\mu\mu$ f/meter added capacitance. This value will change with conical helix position. Actually, $.5\mu\mu$ f and $1\mu\mu$ f capacitors were used; soldered to opposite arms across the center of the helix.

The experimental results of the capacitor loaded helix were negative. Although the unloaded 'square' conical helix worked well down to 100 MHz (Fig. 25), the impedance of the loaded antenna dropped so low that it could not be fed. The long wires for connection of the capacitors probably reflect the traveling wave. In addition, the transmission line assumption is questionable, although not disproven.

Since series inductor loading would eliminate the long connector lines, and might capture more of the surrounding fields, this method is being considered as an alternate discrete loading, although continuous material loading looks more promising at present.

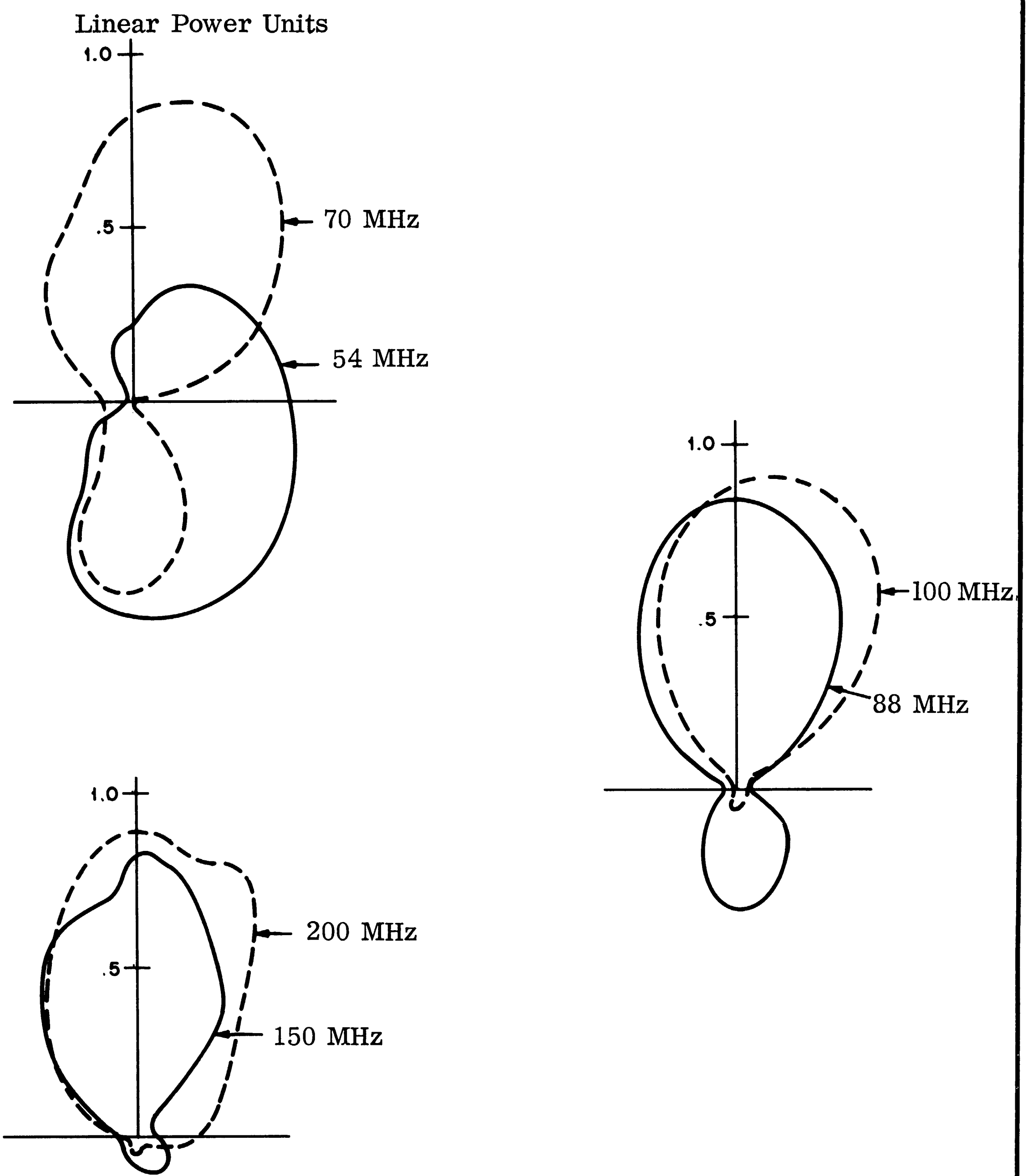

FIG. 25: ANTENNA PATTERNS FOR LARGE UNLOADED CONICAL HELIX 5' DIA. BASE.

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13. ABSTRACT

Work on the design, fabrication and testing of three broadband antennas is described. The antennas are (1) high gain constant beamwidth, (3) omni-directional and (3) loaded conical helix. Impedance and pattern information are given for a broadband ridged horn designed for use as a feed in the constant beamwidth antenna system. Results are given on modeling tests with a pillbox designed to simulate the performance of the high gain reflector in one plane.

Work on two broadband omni-directional antennas is described. The first is a crossed plate antenna somewhat bulky physically but having promising impedance behavior. The second is a monopole with traps, measured and calculated impedance behavior of which is excessive at several frequencies.

The results obtained in loading of bifilar helix antennas with cores and layers of ferrite powder material is described. A decrease in axial mode frequencies from 710 MHz unloaded to 550 MHz loaded regardless of layer thickness is obtained. The thinnest layer tested was 0.5" on a 4" diameter helix. Work on the loading of conical helices with capacitors and with ferrite materials is presented.

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