STUDY AND INVESTIGATION OF UHF-VHF ANTENNAS

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Prepared by

J.A.M. Lyon, G.G. Rassweiler, N.G. Alexopoulos,
J.C. Parker, D.L. Smith, and P.R. Wu

Approved by J. A. M. Lyon
J.A.M. Lyon, Professor
Electrical Engineering

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O.E. Horton, Project Monitor

Air Force Avionics Laboratory, AVWE
Research and Technology Division, AFSC
Wright-Patterson Air Force Base, Ohio

FOREWORD

This report, 7848-4-Q, was prepared by The University of Michigan, Radiation Laboratory, Department of Electrical Engineering under the direction of Professor Ralph E. Hiatt and Professor John A. M. Lyon on Air Force Contract AF 33(615)-3609, under Task 627801 of Project 6278 "Study and Investigation of UHF-VHF Antennas (U)". The work was administered under the direction of the Air Force Avionics Laboratory, Wright-Patterson Air Force Base Ohio. The Task Engineer was Mr. Olin E. Horton, the Project Engineer Mr. E. M. Turner.

This report covers the period 1 November 1966 through 31 January 1967.

ACKNOWLEDGEMENTS

The experimental assistance of Mr. U. E. Gilreath and Mr. R. Carducci is gratefully acknowledged.

ABSTRACT

This report describes the accomplishment on each of four assigned tasks of this project for a three-month period. Under Task 1, a log conical spiral antenna having a diameter of 21 cm at the base and having a length of 65 cm along the axis is described. Various possible loading techniques are discussed which may meet the frequency bandwidth requirements for the antenna. One objective is to obtain a conductor which in itself is a slow wave structure.

Under Task 2, considerable progress has been made in obtaining an antenna which is a series of finger-like elements. This antenna is called an interdigital array. Effects of loading on this interdigital array are discussed. Under this same task, an array with ferrite filled slots capable of being controlled magnetically is also under consideration. During this particular report period, very little effort has been put on this type of array. However, in continuing Task 2, a major part of the effort will be placed on such an array using ferrite filled slots as elements.

Under Task 3, an analytical effort is described providing the most basic criteria for rod type antennas. The results of this analysis can be used for either ferrite rod or dielectric rod antennas. Experimental models of ferrite rod antennas are under construction.

Under Task 4, a survey has been made on various possibilities of reducing size and obtaining required bandwidth performance. Linear elements and combinations of linear elements have been considered in some detail.

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I

INTRODUCTION

During this report period progress has been maintained in each of the four tasks assigned under the contract. It will become apparent in reading the succeeding sections of this report that more effort has been spent on some tasks than on others. This has been occasioned to some extent by certain expediencies.

Task 1, dealing with a log conical spiral antenna, has been given a considerable amount of attention during the first year of this contract. In the last report, a detailed analysis was given of loading this type of antenna. In this report, emphasis is placed upon ferrite-loading of a coiled slow wave structure. This conductor is to be formed into the winding of a log conical spiral antenna. Attention has also been given to coating the winding of a conical helix with ferrite.

The work objective under Task 2 has been to study the feasibility of utilizing physically small ferrite loaded slot antennas at 300 MHz as elements in an antenna array system. The work on such an array would include the magnetic control of the phasing and amplitude of elements. This study has been carried out in earlier quarters and is presently being continued. However, in the quarter corresponding to this report, major attention has been given to the completion of work on the interdigital array. More specifically, the work on this array has emphasized the experimental effort of loading this array with ferrite to reduce its size. It is believed that the resulting interdigital array constitutes a worthwhile antenna of unusually small size even without loading. With ferrite loading, the size has been decreased still further. Experimental results for the loaded and unloaded cases are compared. Radiation patterns and VSWR data are included.

Task 3 requires an investigation of the design feasibility of an endfire ferrite rod antenna for use at 300 to 1000 MHz. An analytical study has been made showing the influence of various design parameters. This analysis has been briefly described and some of the results of the analysis have been given in the form of curves.

Under Task 4, consideration has been given to the basic requirements of antennas usable at frequencies as low as 30 MHz. These studies include the characteristics of single linear elements as well as the combination of a number of such elements. These preliminary studies have aimed at ascertaining the influence of changes in element configuration and loading upon radiation efficiency, radiation patterns, frequency bandwidth, size and weight. The study has included the use of multiple linear elements and their incorporation in compensating techniques whereby a radiating element is used at least partially as a tuning element. The studies under this task lay a good foundation for work during the next year.

A series of collateral studies have been made about suitable materials adequate for loading antenna elements. These studies were helpful in each of the assigned task areas. The report discusses the availability and desirability of certain materials for loading purposes.

II

FERRITE LOADED CONICAL SPIRALS

Emphasis has been placed on two techniques for obtaining a conical spiral antenna that is 21 cm in diameter at the base and 65 cm in height: 1) Coating the winding of the antenna with ferrite; and 2) Winding the antenna with a ferrite filled coiled slow wave structure.

Because of the construction difficulties associated with incorporating the Radiation Laboratory's supply of ferrites into a ferrite coated winding, a simple theoretical analysis of the problem was done. It indicated that a size reduction multiplicative factor of less than 0.50 can be obtained under favorable circumstances. However, this alone would not be sufficient to obtain the 0.365 reduction factor needed to build a conical helix covering 200 to 600 MHz. with the above dimensions.

Work on the ferrite filled helical slow wave structure or coiled winding has begun. However, only data for the unloaded antenna are reported.

2.1 Ferrite Coated Windings.

Because of the mechanical difficulties involved in trying to construct a helical or conical helical antenna with a ferrite coated winding out of the available supply of ferrite, a theoretical study was done of this problem to see how much help coating wires with ferrite would give in reducing the size of antennas.

The analysis is based on an observation as reported by Rassweiler (1966) who considers a helix or conical helix antenna as a two wire transmission line wrapped around a cone. Judging from his work it appears that this is a fair approximation for the loading effects of ferrite around each antenna wire. Of course, for the two wire transmission line model to be realistic, the ferrite coating of the wires cannot be too thick; otherwise, the coating is really closer to a layer of ferrite covering the surface of the windings.

Using a conventional approach to determine the capacitance and inductance per unit length of the two wire transmission line cross section depicted in Fig. 2-1, (for example, see Johnson, 1950, p. 82) it can readily be shown that the inductance per unit length, L, and the capacitance per unit length, C, are given by the formulas:

$$C = \frac{\pi}{\frac{1}{\epsilon}\ell n\frac{t}{d} + \frac{1}{\epsilon_o}\ell n\frac{2D}{t}}$$

$$L = \frac{\mu}{\pi}\ell n\frac{t}{d} + \frac{\mu_o}{\pi}\ell n\frac{2D}{t}$$

Thus, it follows that the phase velocity (and the group velocity, since the medium is assumed non-dispersive and the transmission line is assumed to be lossless) is given by the expression:

$$v_p = v_g = c\sqrt{\frac{\frac{1}{\epsilon_r}\ell n\frac{t}{d} + \ell n\frac{2D}{t}}{\mu_r\ell n\frac{t}{d} + \ell n\frac{2D}{t}}} \quad .$$

The above formula indicated a very interesting property of ferrite coating. For small t/d ratios, most of the reduction is produced by the relative permeability. The relative dielectric constant provides a greater influence at large t/d ratios.

This would mean that the coating would have to be tightly bound to the conductor. If there were a small gap between the coating and the conductor, the effectiveness of the coating would be greatly reduced.

Let us now look at two cases of a ferrite coating on the winding of a helix antenna. The first is illustrated in Fig. 2-1. In this case, the ferrite is assumed to be powdered EAF-2. The dimensions used correspond to a 4.8 cm diameter, bifilar, 14^{o} helix wound with no. 18 copper wire. The diameter corresponds to that needed at 600 MHz for a helix to radiate in the axial mode assuming a 0.365 reduction factor. (This reduction factor corresponds to the reduction needed to obtain a conical helix fitting the dimensions described in the technical guidelines.) The object of assuming the reduction factor was to see if, indeed, the reduction was possible with a thin coating of this ferrite.

Notice that the dashed line in the graph corresponds to the reduction factor expected with EAF-2 ferrite if the core of the helix were completely filled with the material. Note also the marking that indicates the ferrite diameter to wire diameter ratio that corresponds to the touching of the ferrite coating of successive turns. It is interesting to note that if the antenna were buried in a medium consisting of nothing but EAF-2 ferrite, the reduction factor should be 0.346.

The second case consists of a coating of Indiana General Q-3 Ferrite. (See Fig. 2-2.) The dimensions used correspond to a 14.4 cm diameter, 14^{o}, bifilar helix wound with 1/4 in. O.D. copper tubing. This particular helix diameter corresponds to that needed for axial radiation at 200 MHz, assuming a 0.365 reduction factor. The reduction factor was assumed in this case for the same reason as in the previous case.

The dashed line in the graph corresponds to the reduction factor obtainable with a full core loading of Q-3. Again the ferrite to wire diameter

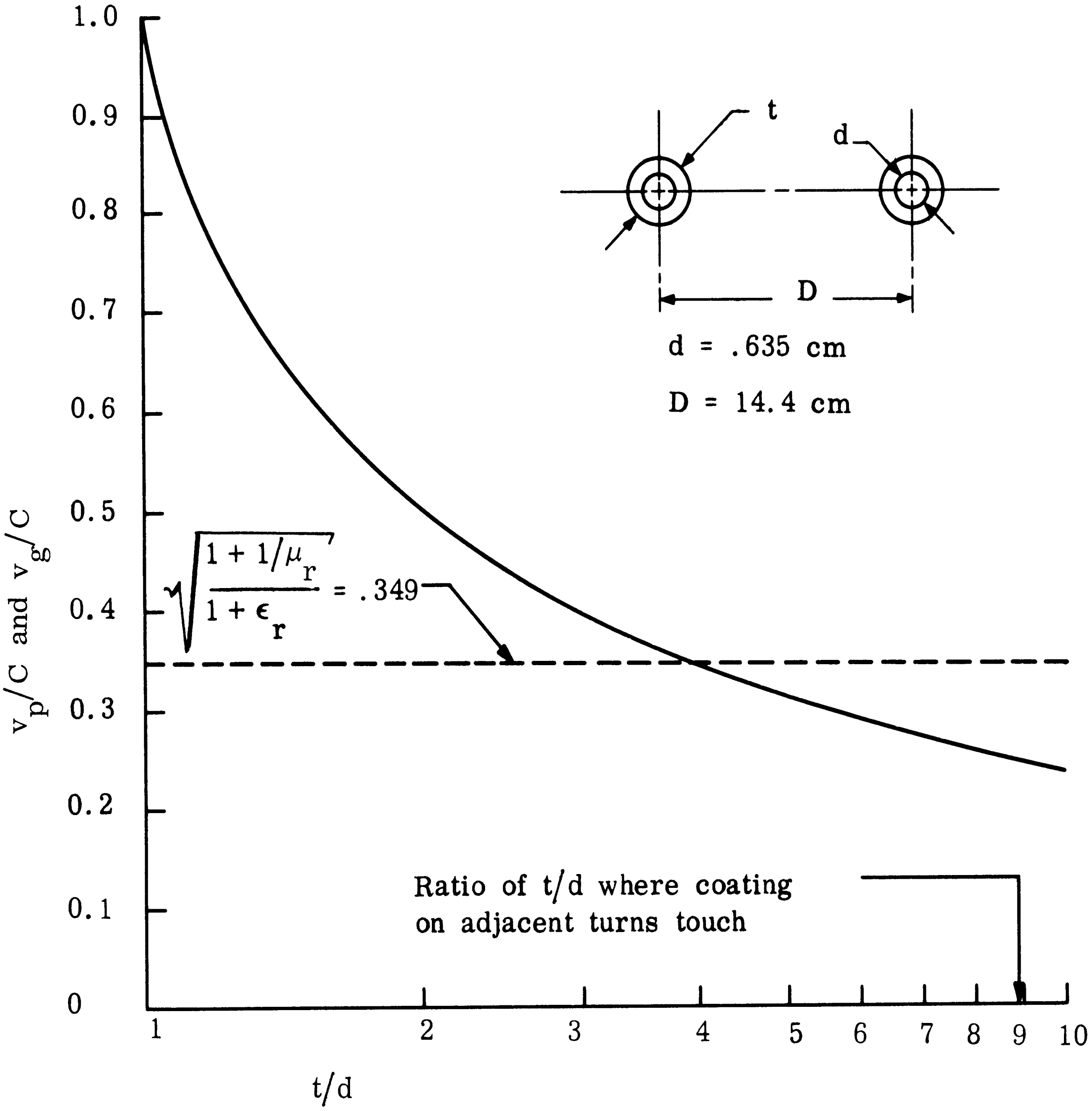


FIG. 2-1: PLOT OF THE REDUCTION FACTOR (v_g/C) VS. t/d FOR A TWO WIRE TRANSMISSION LINE LOADED WITH INDIANA GENERAL Q-3 FERRITE (μ_r = 14, ϵ_r = 7.8).

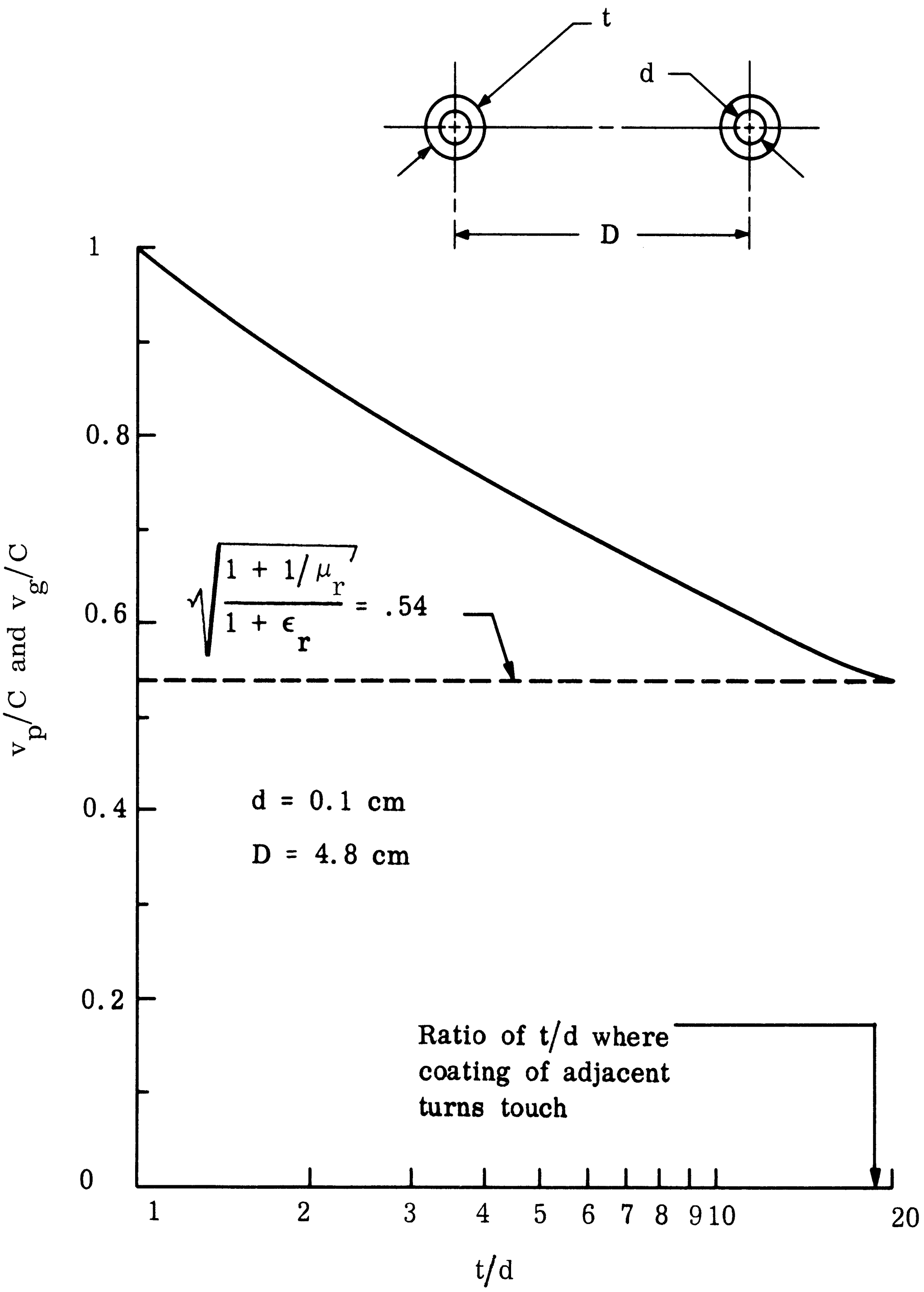


FIG. 2-2: PLOT OF THE REDUCTION FACTOR (v_g/C) VS. t/d FOR A TWO WIRE TRANSMISSION LINE LOADED WITH EAF-2 FERRITE (μ_r = 2.2, ϵ_r = 3.8).

where the coating of adjacent turns touch is indicated on the graph. If the antenna were buried in a medium consisting entirely of Q-3, the reduction factor should be 0.096.

For the Q-3 example, it appears that a ferrite coating diameter between 4 and 5 times the wire diameter should give the desired reduction. However, for thicknesses this great, the requirement that $t/D \ll 1$ is most certainly violated, and, indeed, the reduction factor may not be possible after all.

2.2 Ferrite Filled Coiled Windings

It is well known that a helix is a slow wave structure. It would seem that if a helical antenna were wound with a small coiled conductor, then the size of the antenna wound with such a coiled conductor could be reduced. This experiment has been tried by several workers, but unfortunately, no one has been completely satisfied with the results. (Turner, 1966.) The device acts as an antenna and appears to radiate quite well. However, the far field patterns do not resemble those of a regular helical antenna. Apparently, the interaction of a turn with other turns produces unusual results.

Since a magneto-dielectric material has a confining effect on electromagnetic fields, it was hoped that if the interior of a coiled conductor were filled with ferrite, the interaction might be reduced enough for the antenna with the coiled conductor to operate successfully.

Figure 2-3 shows a picture of a bifilar helix constructed for this experiment. The coiled conductor is wound on a 3/4 in. O. D. by 3/32 in. thick piece of Tygon tubing using no. 20 enameled copper wire with a pitch of 1/3 in. This coiled conductor should give a slowing factor of about 0.25 based on the data presented by Okubo (1965). The antenna itself has a pitch of 4 in. The antenna winding proper is wound on a 3-1/8 in. O.D. by 1/16 in. thick piece of NEMA Grade XXX paper phenolic tubing. The antenna is fed at the tip through a two wire shielded transmission line from a hybrid.

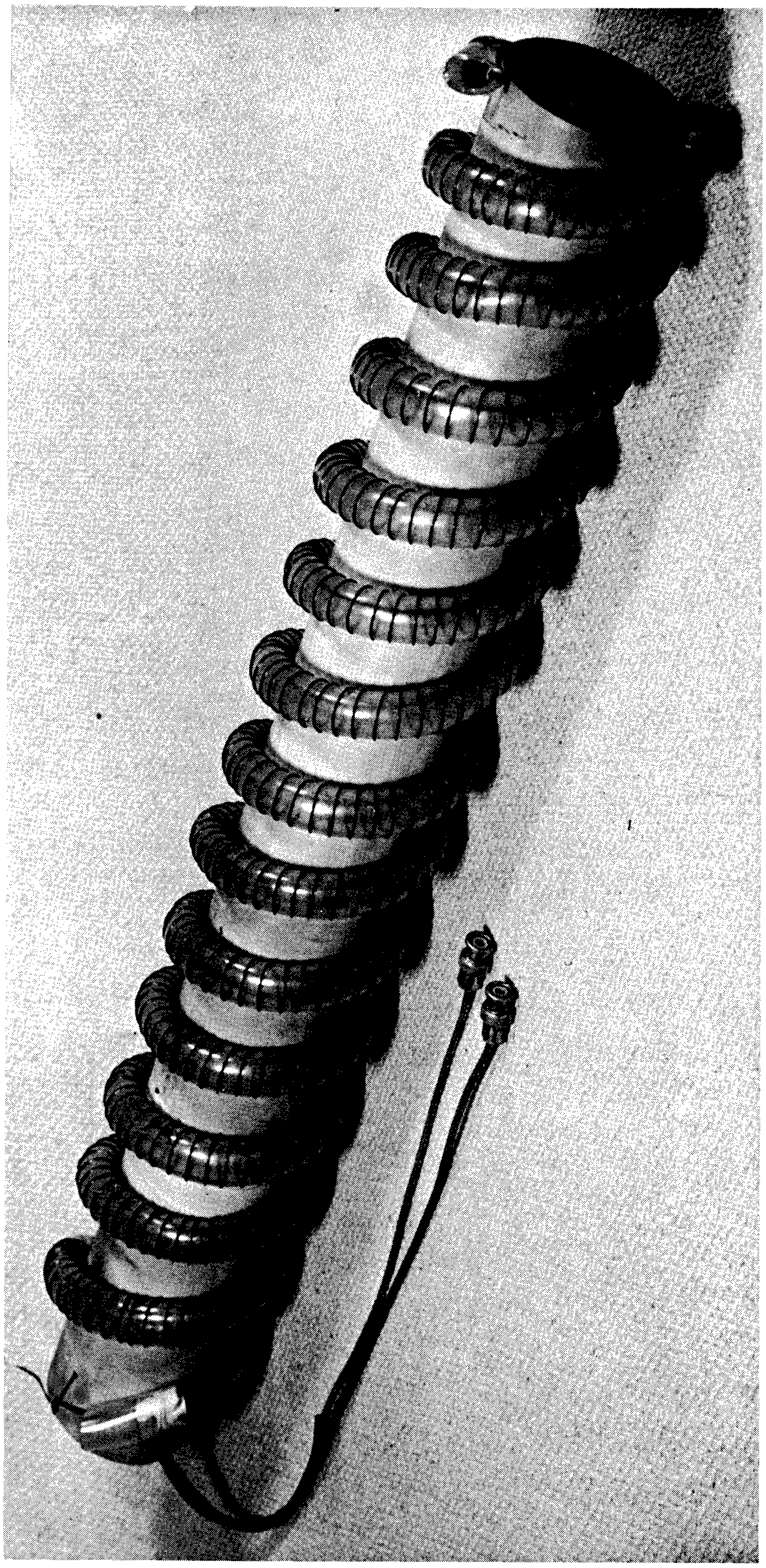


FIG. 2-3: ANTENNA 228, A 14^{o} BIFILAR HELIX WITH A HELICAL WINDING.

Figure 2-4 shows the far field patterns for this antenna. The center frequency of operation, if the antenna were wound with a simple wire on a cylindrical form of the same mean diameter as the antenna, is about 800 MHz.

Unfortunately, the plastic tubing used to construct the coiled conductor started to decompose before the antenna could be loaded with ferrite. Rather than risk contaminating the irreplaceable EAF-2 ferrite, it was decided to postpone loaded tests on the antenna until it could be rebuilt using a replacement tube.

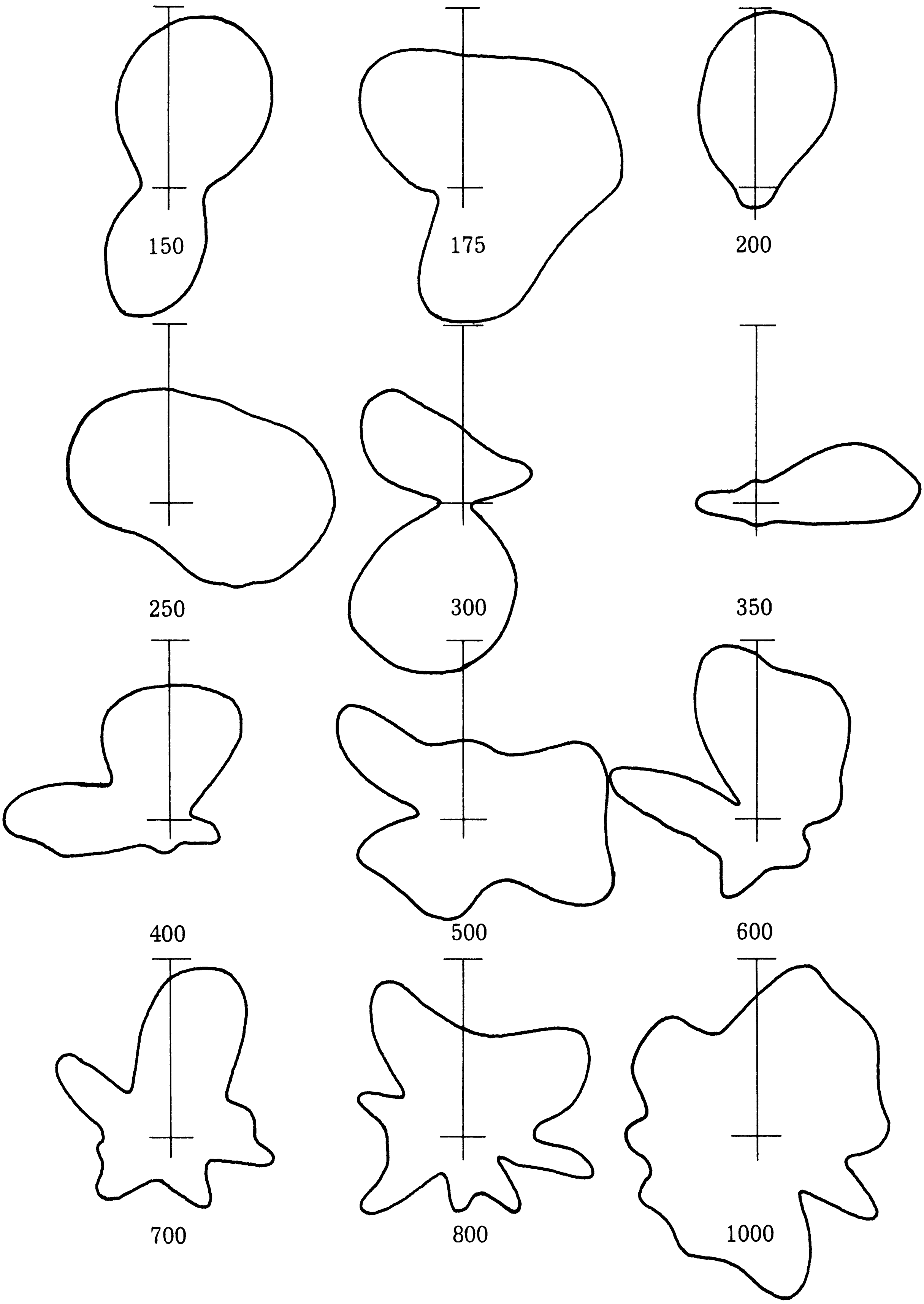


FIG. 2-4: FAR FIELD POWER PATTERNS OF ANTENNA 228, A BIFILAR 14^{o} HELIX WITH A HELICAL WINDING.

III

PHYSICALLY SMALL FERRITE ARRAYS

3.1 General

An interdigital array has been loaded with ferrite and tested. In this report, only ferrite loading of the interdigital array is covered. The complete analysis and results of experimental testing of the interdigital array will be presented in the next quarterly report.

The interdigital array is a very interesting antenna array for several reasons:

1) A compact, flush mounting is possible.
2) The construction and feed are simple.
3) The antenna is wideband with relatively high gain.
4) The elements may be loaded.

3.2 Experimental Results

The ferrite powder EAF-2 was placed between the array surface and the ground plane for all models (A-1, A-2, A-3, B-1, and B-2) of the interdigital array; see Figs. 3-2 through 3-6. The specification of the models is repeated below for convenience, where 2N+1 is the number of elements and 2a, b, d, and ℓ are as defined in Fig. 3-1.

TABLE I

Parameter	Antenna Type				
	A-1	A-2	A-3	B-1	B-2
2N+1	13	13	13	21	21
(2a)*	0.15	0.15	0.15	0.15	0.15
b	1.2	1.8	2.4	1.2	1.8
d	1.0	1.0	1.0	1.0	1.0
ℓ	8.0	8.0	8.0	8.0	8.0

* equivalent, all dimensions are in cm.

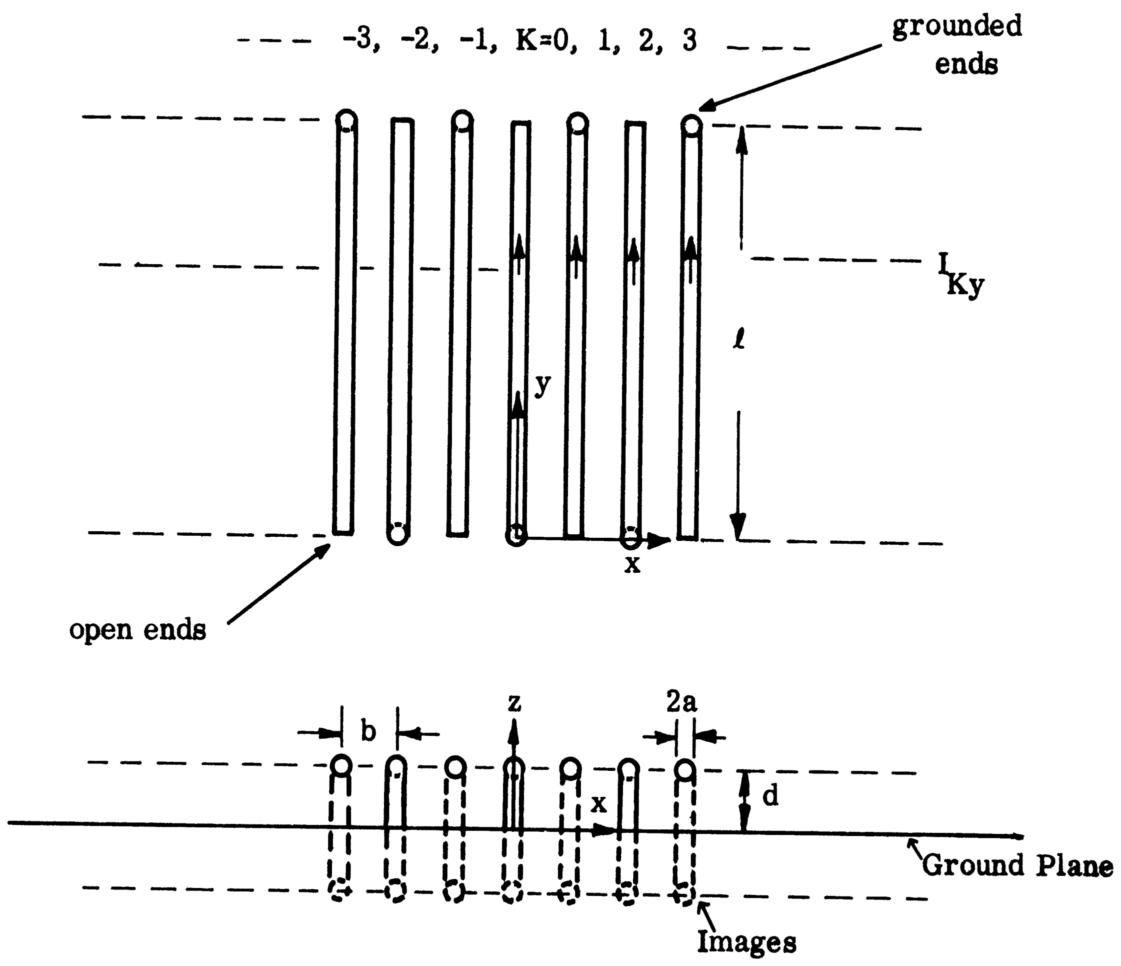


FIG. 3-1: THE GEOMETRY OF THE CONDUCTING ARRAY ELEMENTS.

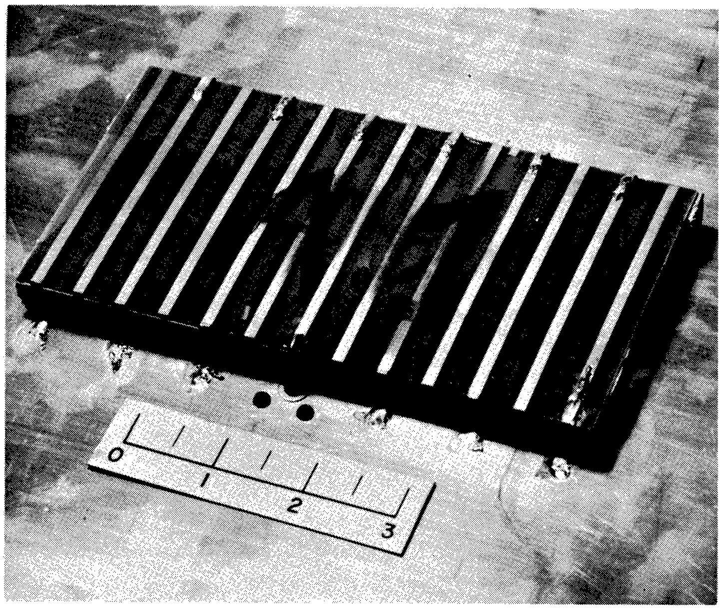


FIG. 3-2: THE INTERDIGITAL ARRAY ANTENNA A-1.
$\ell = 8\,\mathrm{cm}$, $d/\ell = 0.125$, $b/\ell = 0.15$, $a/\ell = 0.01$, $N = 6$.

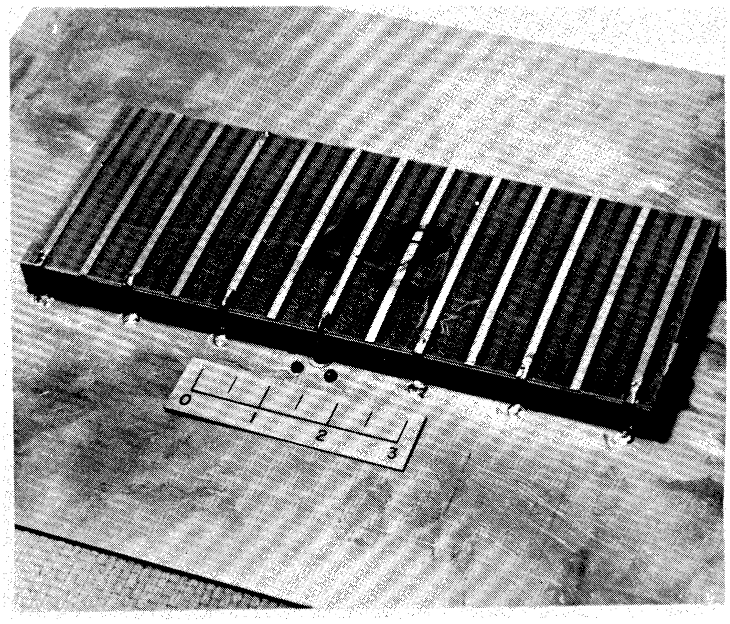


FIG. 3-3: THE INTERDIGITAL ARRAY ANTENNA A-2.
$\ell = 8\,\mathrm{cm}$, $d/\ell = 0.125$, $b/\ell = 0.225$, $a/\ell = 0.01$, $N = 6$.

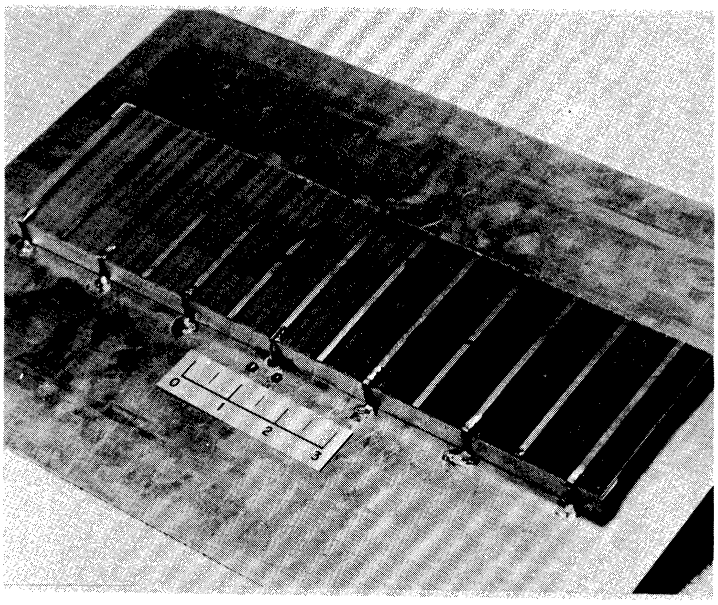


FIG. 3-4: THE INTERDIGITAL ARRAY ANTENNA A-3.
$\ell = 8\,\text{cm}$, $d/\ell = 0.125$, $b/\ell = 0.30$, $a/\ell = 0.01$, $N = 6$.

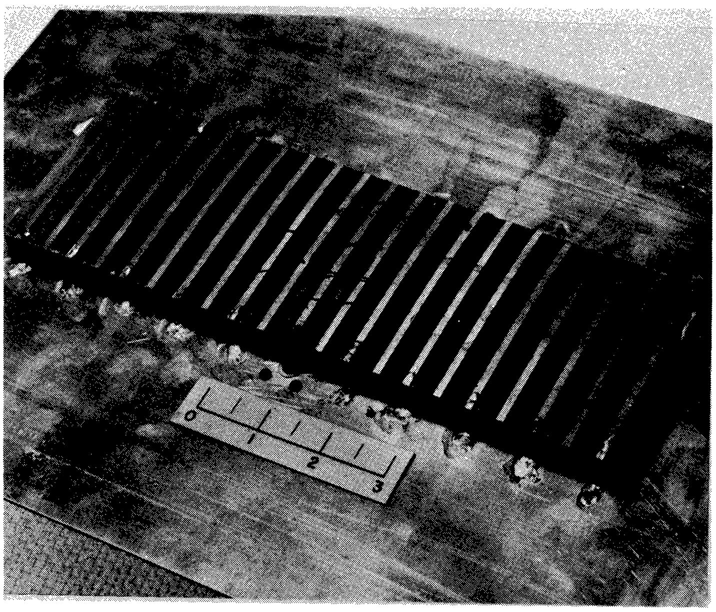


FIG. 3-5: THE INTERDIGITAL ARRAY ANTENNA B-1.
$\ell = 8\,\text{cm}$, $d/\ell = 0.125$, $b/\ell = 0.15$, $a/\ell = 0.01$, $N = 0$.

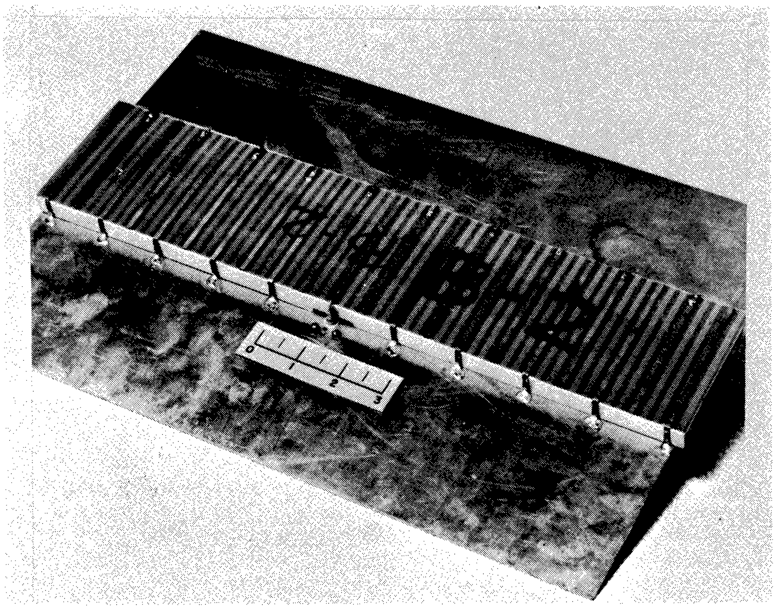


FIG. 3-6: THE INTERDIGITAL ARRAY ANTENNA B-2.
$\ell = 8\,\mathrm{cm}$, $d/\ell = 0.125$, $b/\ell = 0.225$, $a/\ell = 0.01$, $N = 10$.

The far field patterns were taken and are shown in Figs. 3-7 through 3-11 along with the unloaded patterns at corresponding frequencies. Figure 3-7 shows the far field patterns of the antenna A-1. It is seen that the unloaded antenna has a fairly broad bandwidth above 350 MHz. This seems to indicate a shift in center frequency as well as a narrower bandwidth when the antenna is loaded with a ferrite material.

Similar effects are seen for antenna A-2 (Fig. 3-8). Antenna A-3 seems to have a lower center frequency and a lower upper frequency bound (450 MHz) than the other models (Fig. 3-9). Antenna B-1 (Fig. 3-10) is good up to 600 MHz, a little higher than antenna A-1, while B-2 (Fig. 3-11) is good up to only 550 MHz, a little lower than antenna A-2.

Although a conclusion cannot be drawn from the above observations, it can be said in general that the effect of loading an interdigital array with ferrite material is to lower the center frequency and to decrease the bandwidth of the structure. A shift of the lower frequency bound is seen to be from 350 MHz down to 250 MHz; this is quite significant if the decrease in bandwidth is not considered to be a serious disadvantage.

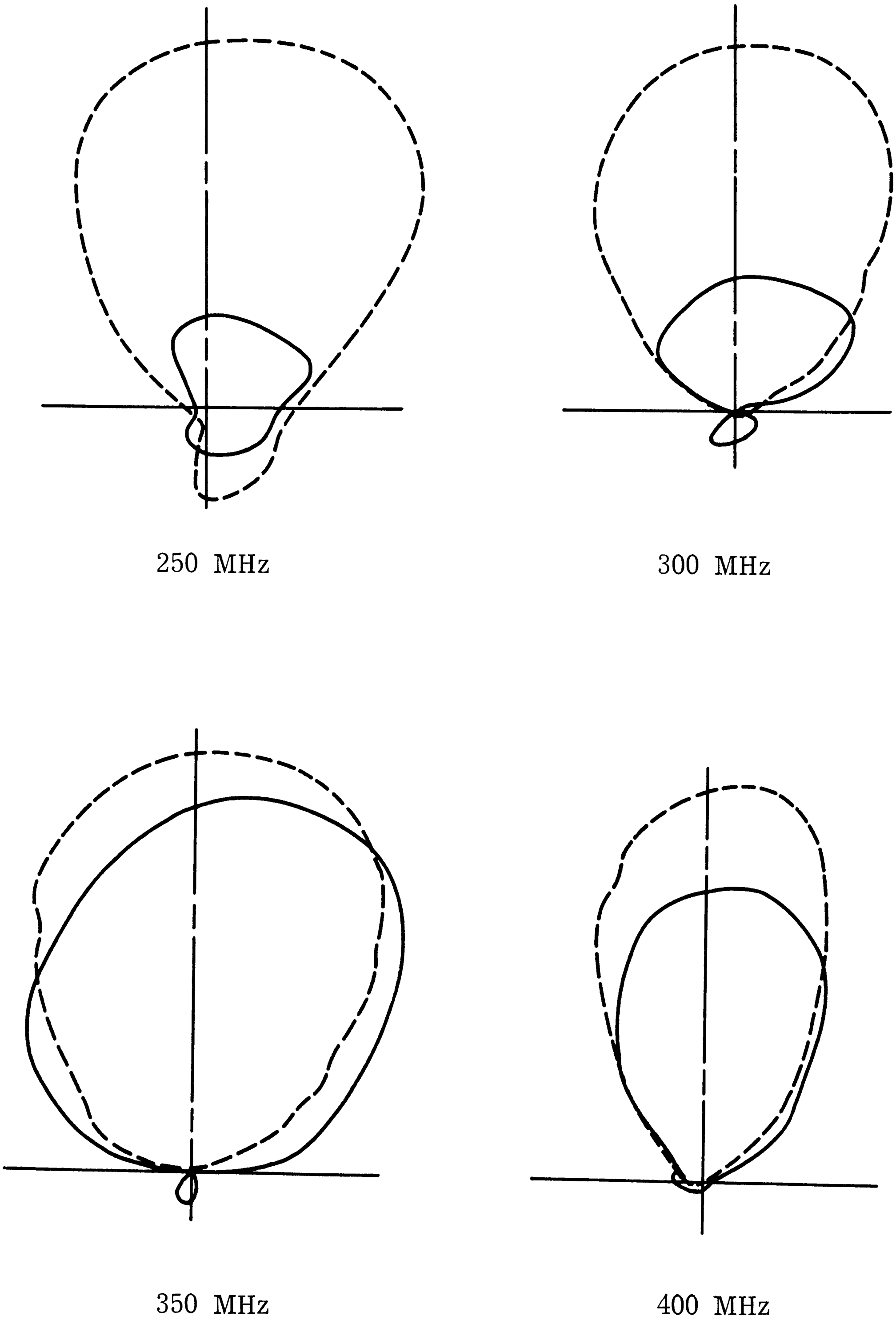


FIG. 3-7-a: THE FAR FIELD PATTERNS OF THE INTERDIGITAL ARRAY ANTENNA A-1 LOADED AND UNLOADED WITH EAF-2 FERRITES. (—) Unloaded (---) Loaded.

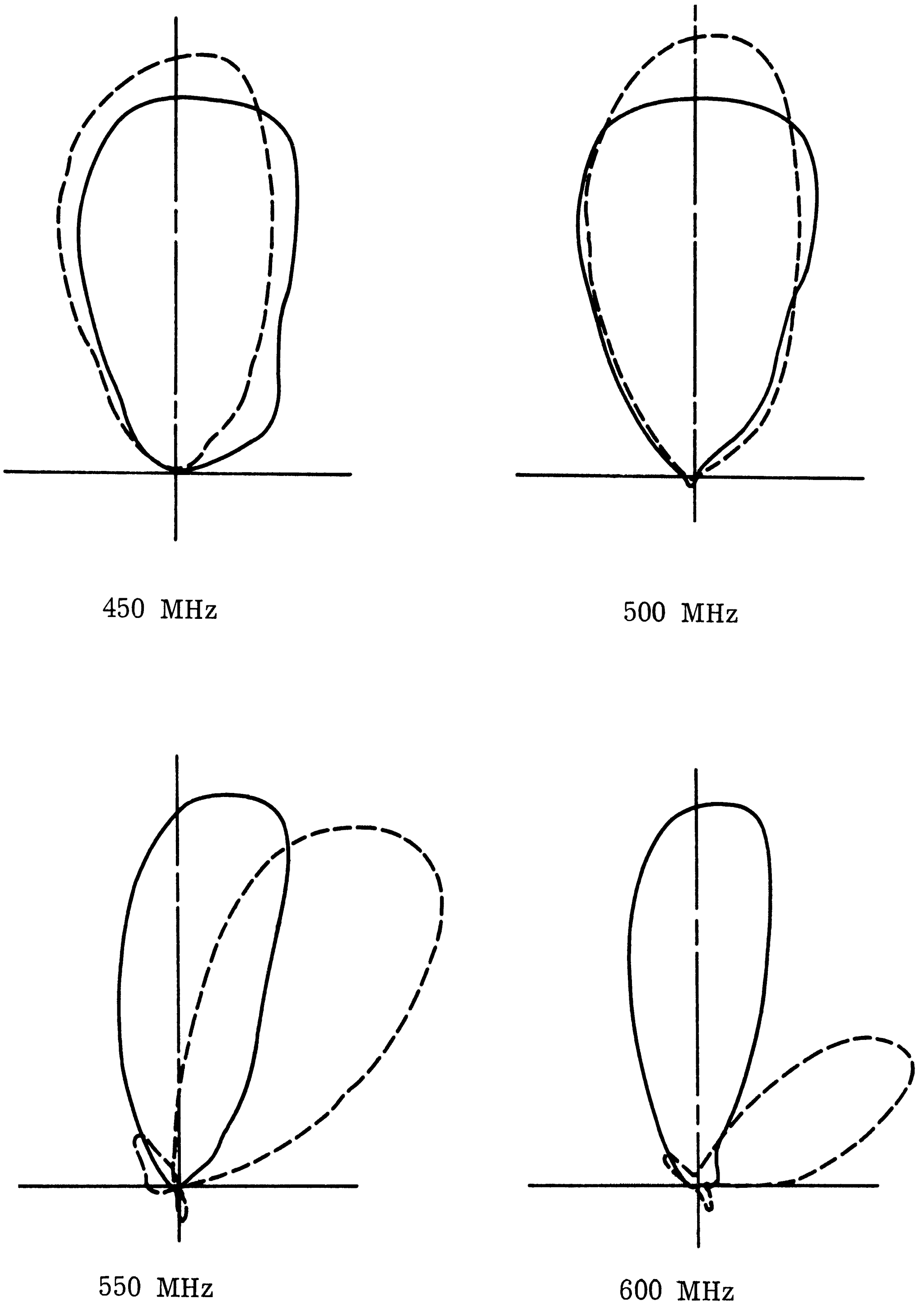


FIG. 3-7-b: THE FAR FIELD PATTERNS OF THE INTERDIGITAL ARRAY ANTENNA A-1 LOADED AND UNLOADED WITH EAF-2 FERRITE. (—) Unloaded, (---) Loaded.

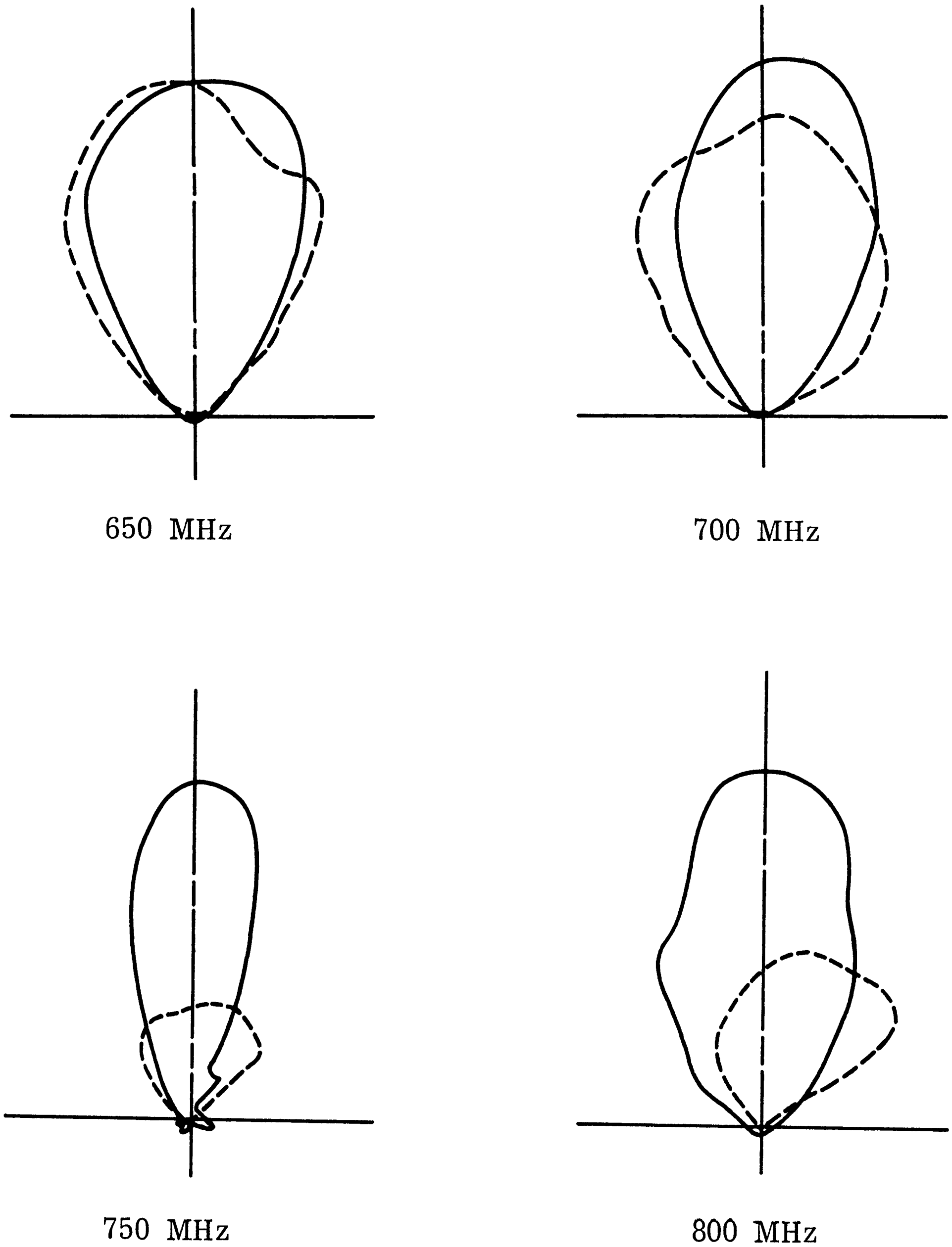


FIG. 3-7-c: THE FAR FIELD PATTERNS OF THE INTERDIGITAL ARRAY ANTENNA A-1 LOADED AND UNLOADED WITH EAF-2 FERRITE. (—) Unloaded, (---) Loaded.

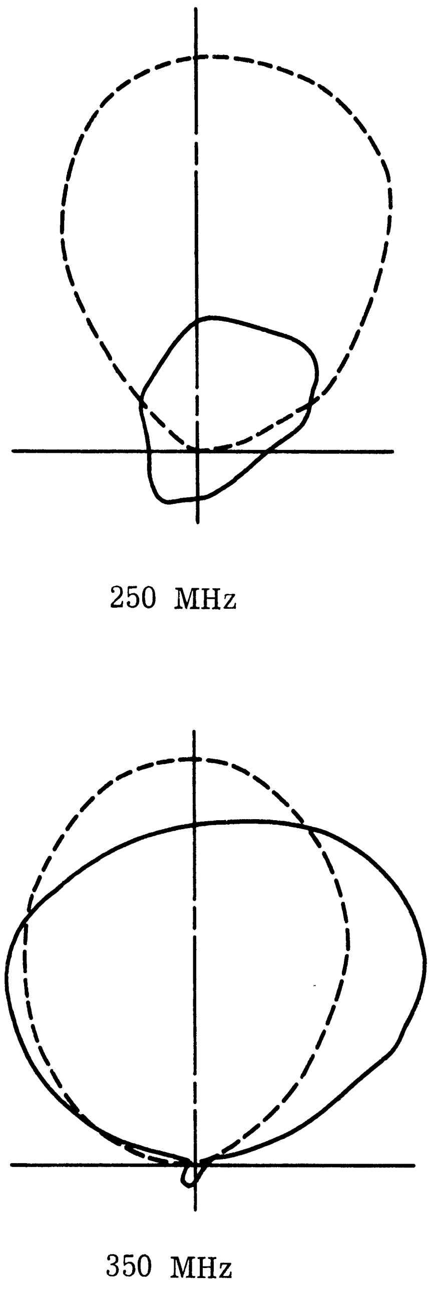


FIG. 3-8-a: THE FAR FIELD PATTERNS OF THE INTERDIGITAL ARRAY ANTENNA A-2 LOADED AND UNLOADED WITH EAF-2 FERRITE. (—) Unloaded, (---) Loaded.

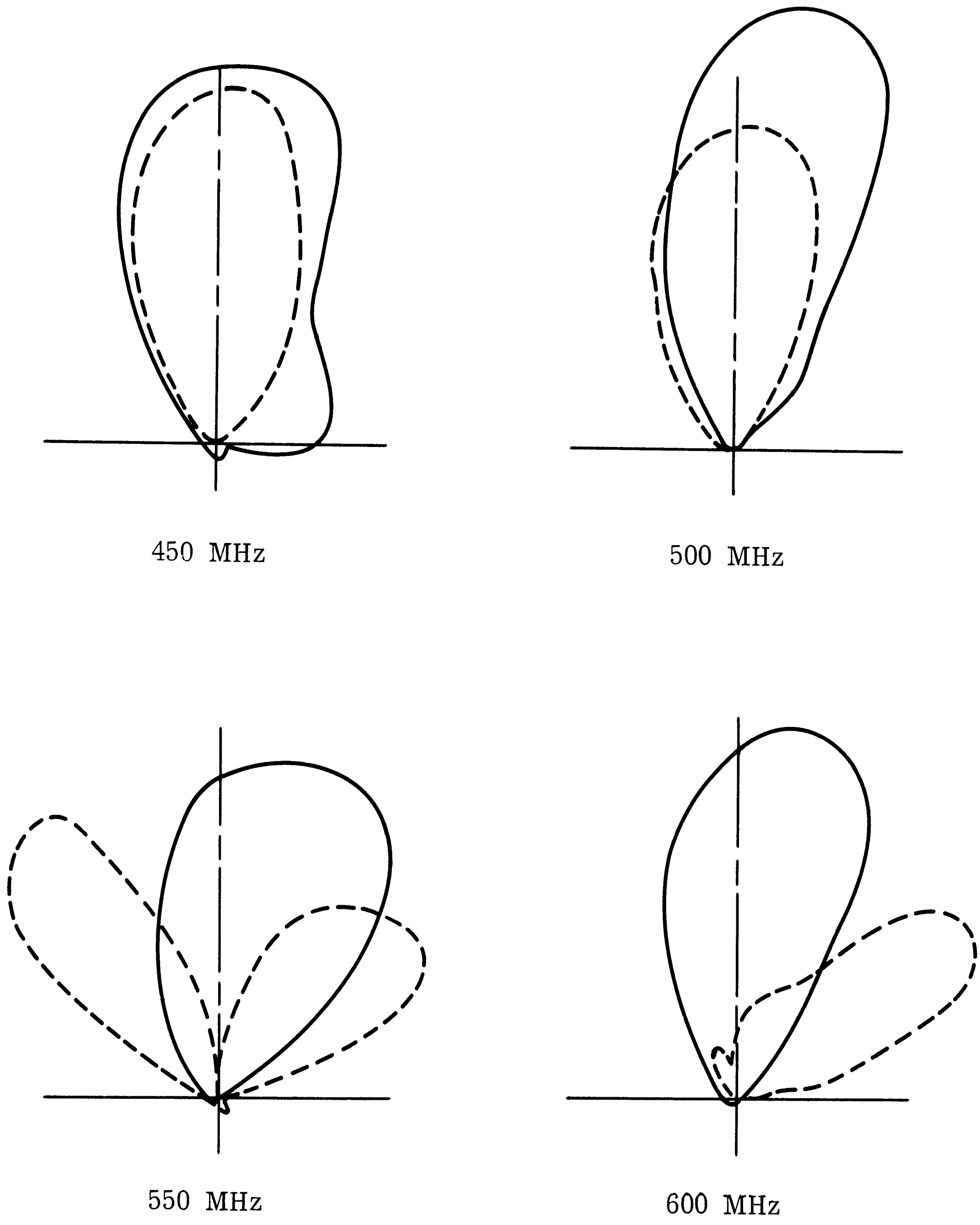
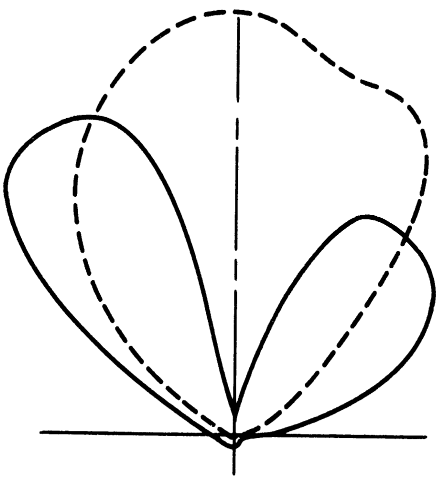
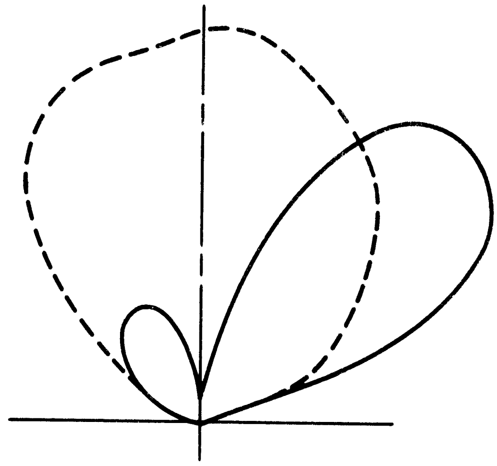


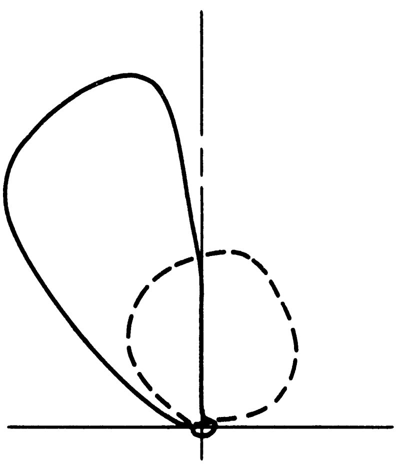
FIG. 3-8-b: THE FAR FIELD PATTERNS OF THE INTERDIGITAL ARRAY ANTENNA A-2 LOADED AND UNLOADED WITH EAF-2 FERRITE. (—) Unloaded, (---) Loaded.



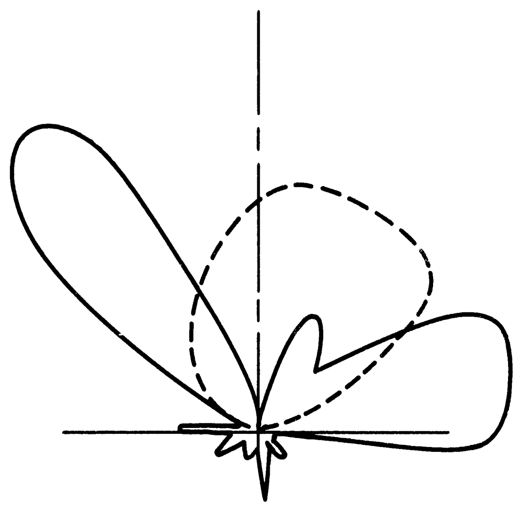
650 MHz



700 MHz

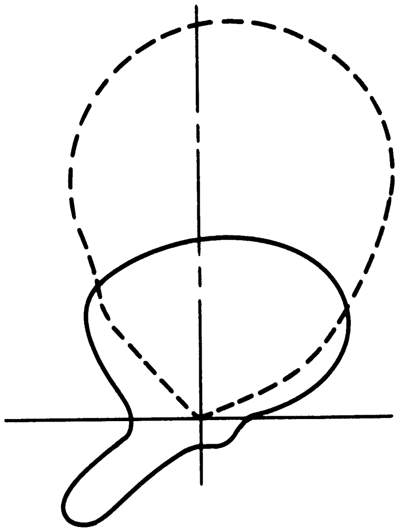


750 MHz

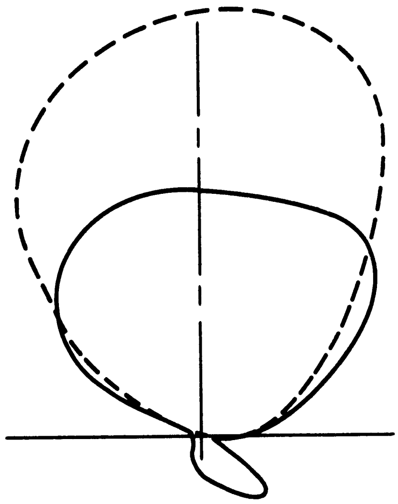


800 MHz

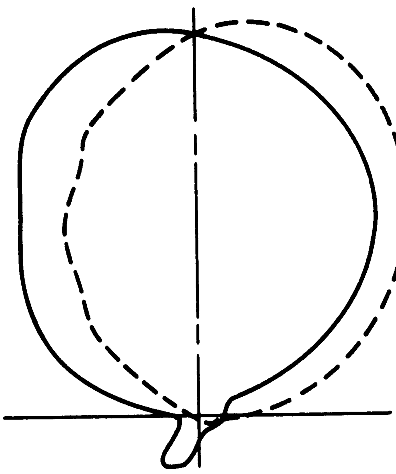
FIG. 3-8-c: THE FAR FIELD PATTERNS OF THE INTERDIGITAL ARRAY ANTENNA A-2 LOADED AND UNLOADED WITH EAF-2 FERRITE. (——) Unloaded, (---) Loaded.



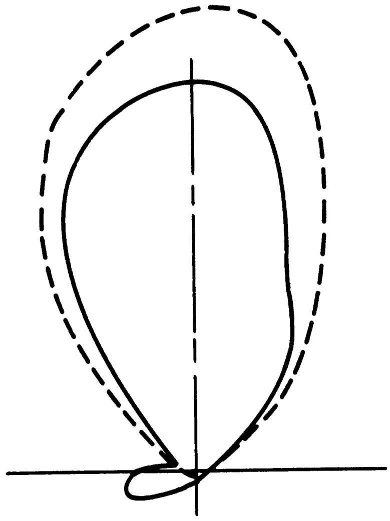
250 MHz



300 MHz



350 MHz



400 MHz

FIG. 3-9-a: THE FAR FIELD PATTERNS OF ANTENNA A-3 LOADED AND UNLOADED WITH EAF-2 FERRITE. (——) Unloaded, (---) Loaded.

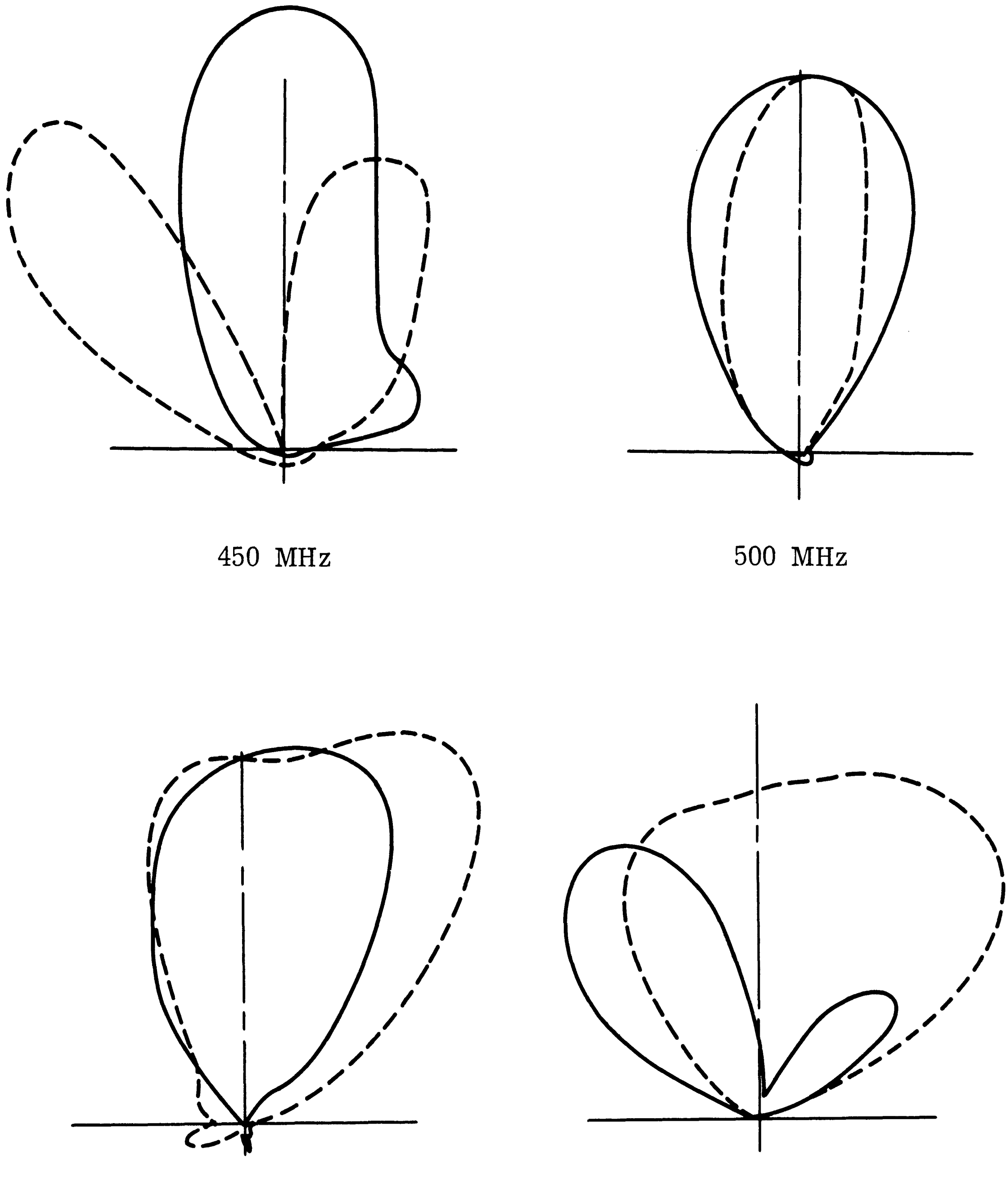


FIG. 3-9-b: THE FAR FIELD PATTERNS OF ANTENNA A-3 LOADED AND UNLOADED WITH EAF-2 FERRITE. (—) Unloaded, (---) Loaded.

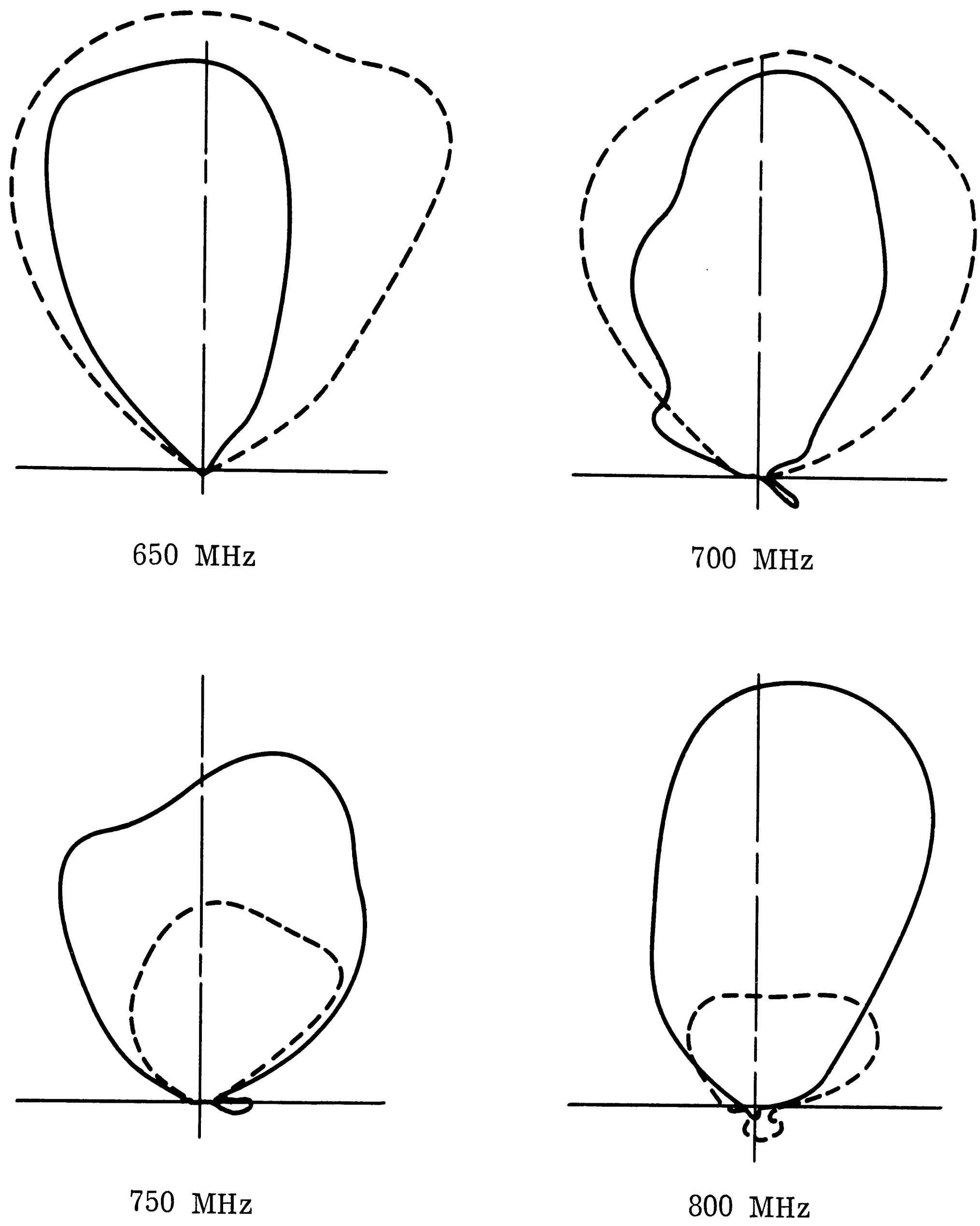
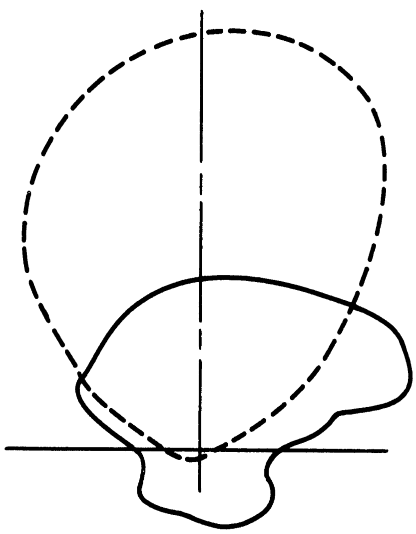
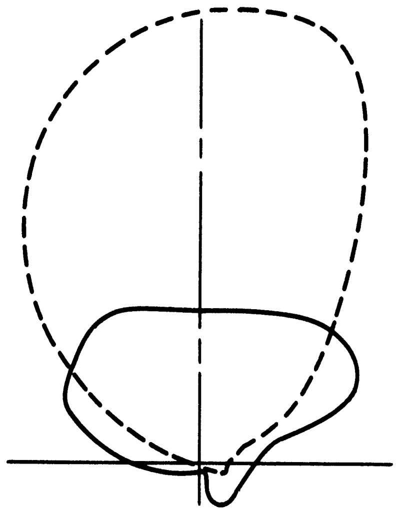


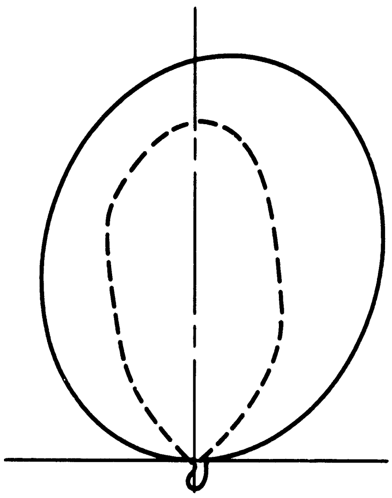
FIG. 3-9-c: THE FAR FIELD PATTERNS OF ANTENNA A-3 LOADED AND UNLOADED WITH EAF-2 FERRITE. (——) Unloaded, (---) Loaded.



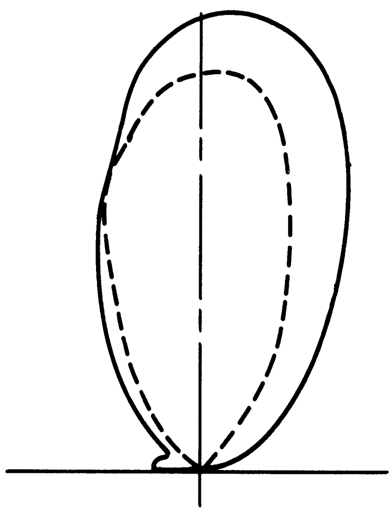
250 MHz



300 MHz



350 MHz



400 MHz

FIG. 3-10-a: THE FAR FIELD PATTERNS OF ANTENNA B-1 LOADED AND UNLOADED WITH EAF-2 FERRITE. (—) Unloaded, (---) Loaded.

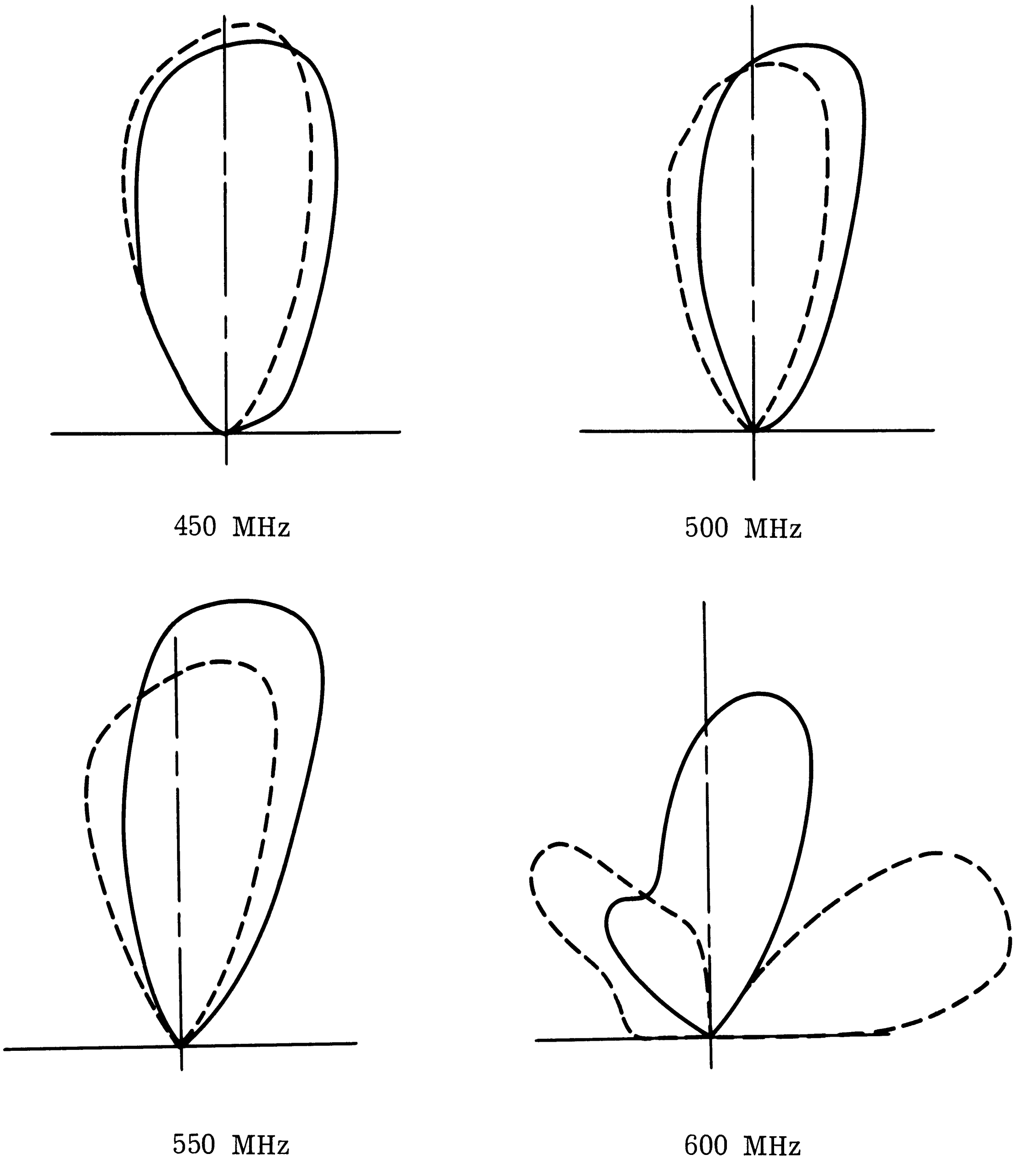


FIG. 3-10-b: THE FAR FIELD PATTERNS OF ANTENNA B-1 LOADED AND UNLOADED WITH EAF-2 FERRITE. (—) Unloaded, (---) Loaded.

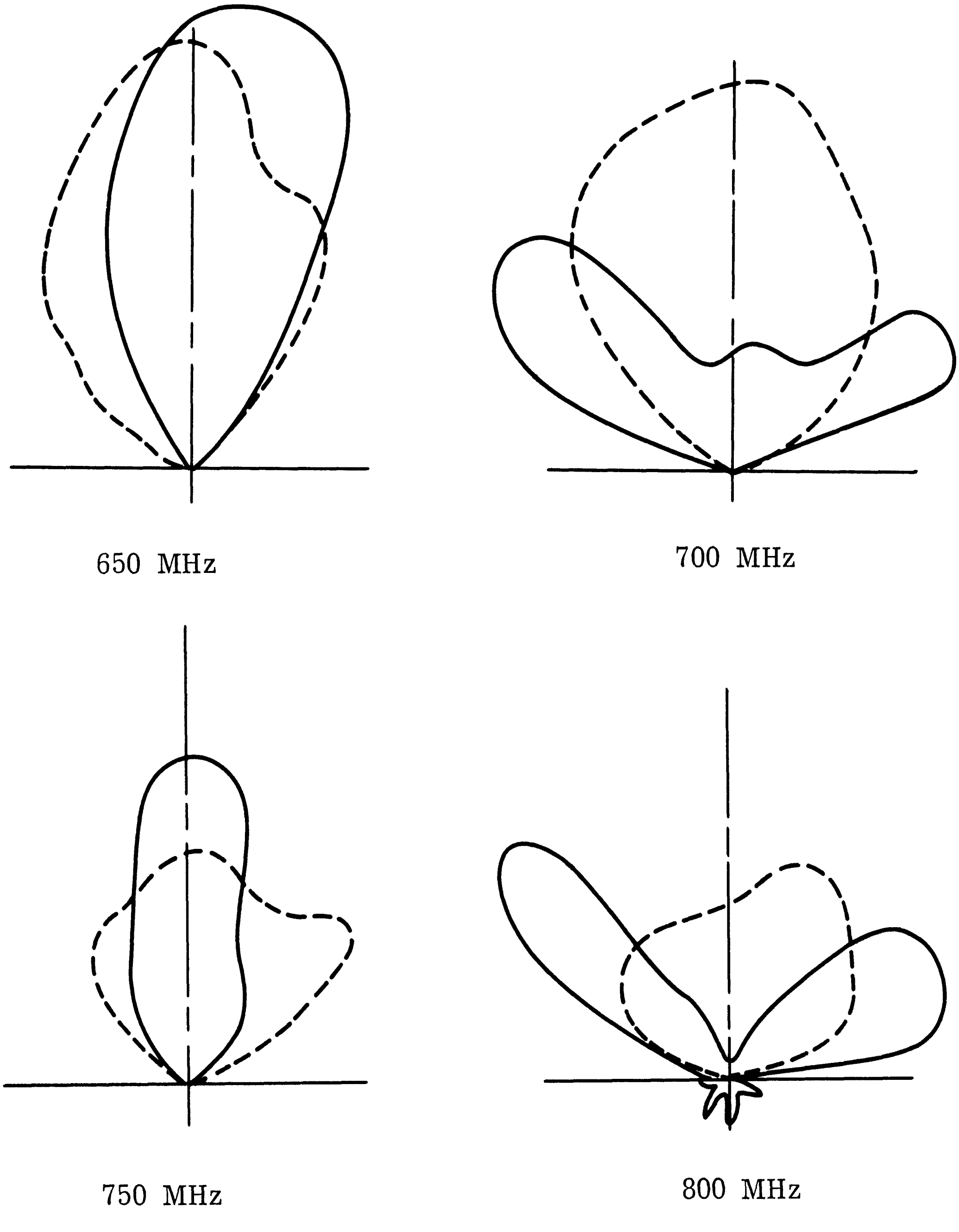


FIG. 3-10-c: THE FAR FIELD PATTERNS OF ANTENNA B-1 LOADED AND UNLOADED WITH EAF-2 FERRITE. (——) Unloaded, (---) Loaded.

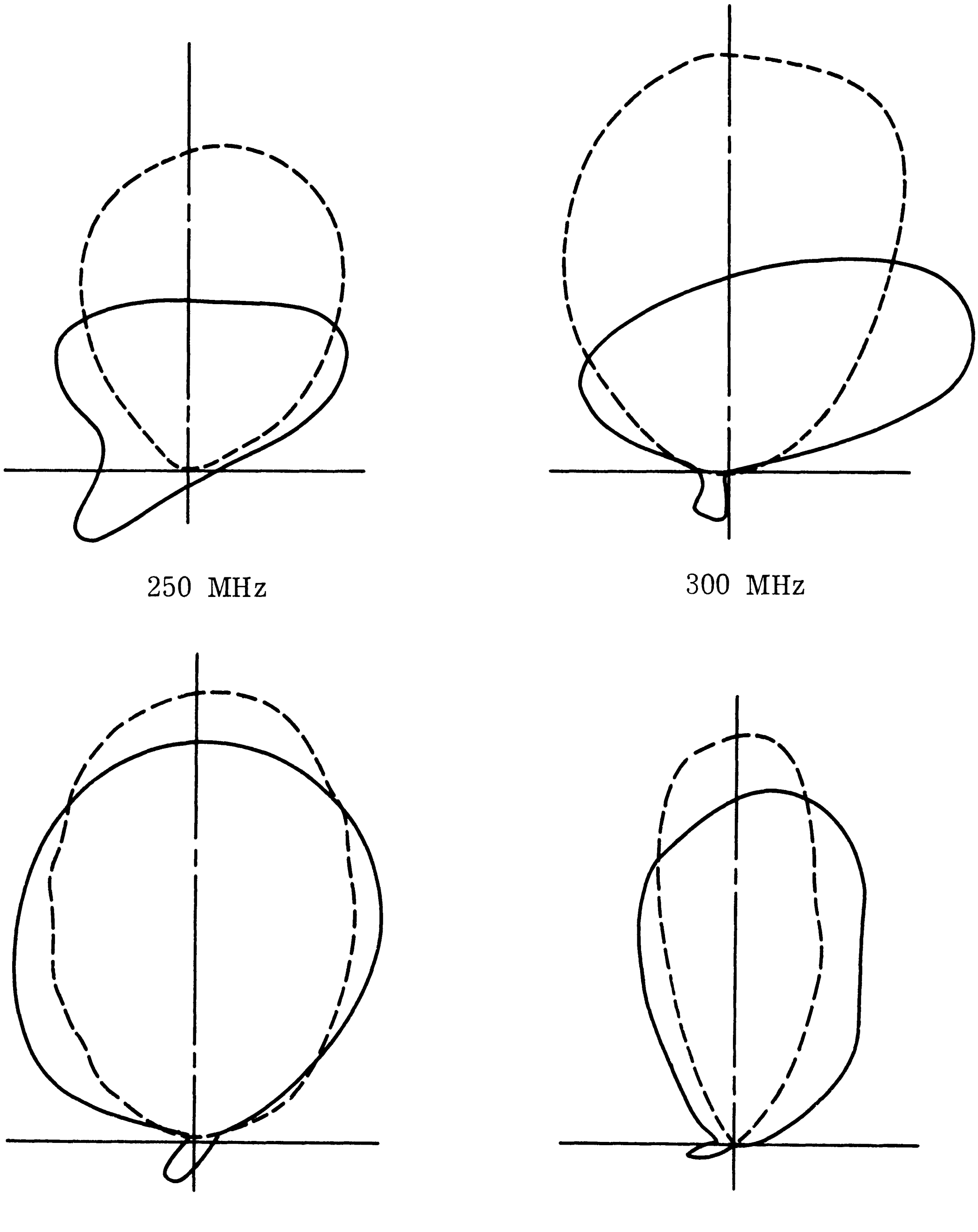


FIG. 3-11-a: THE FAR FIELD PATTERNS OF ANTENNA B-2 LOADED AND UNLOADED WITH EAF-2 FERRITE. (——) Unloaded, (---) Loaded.

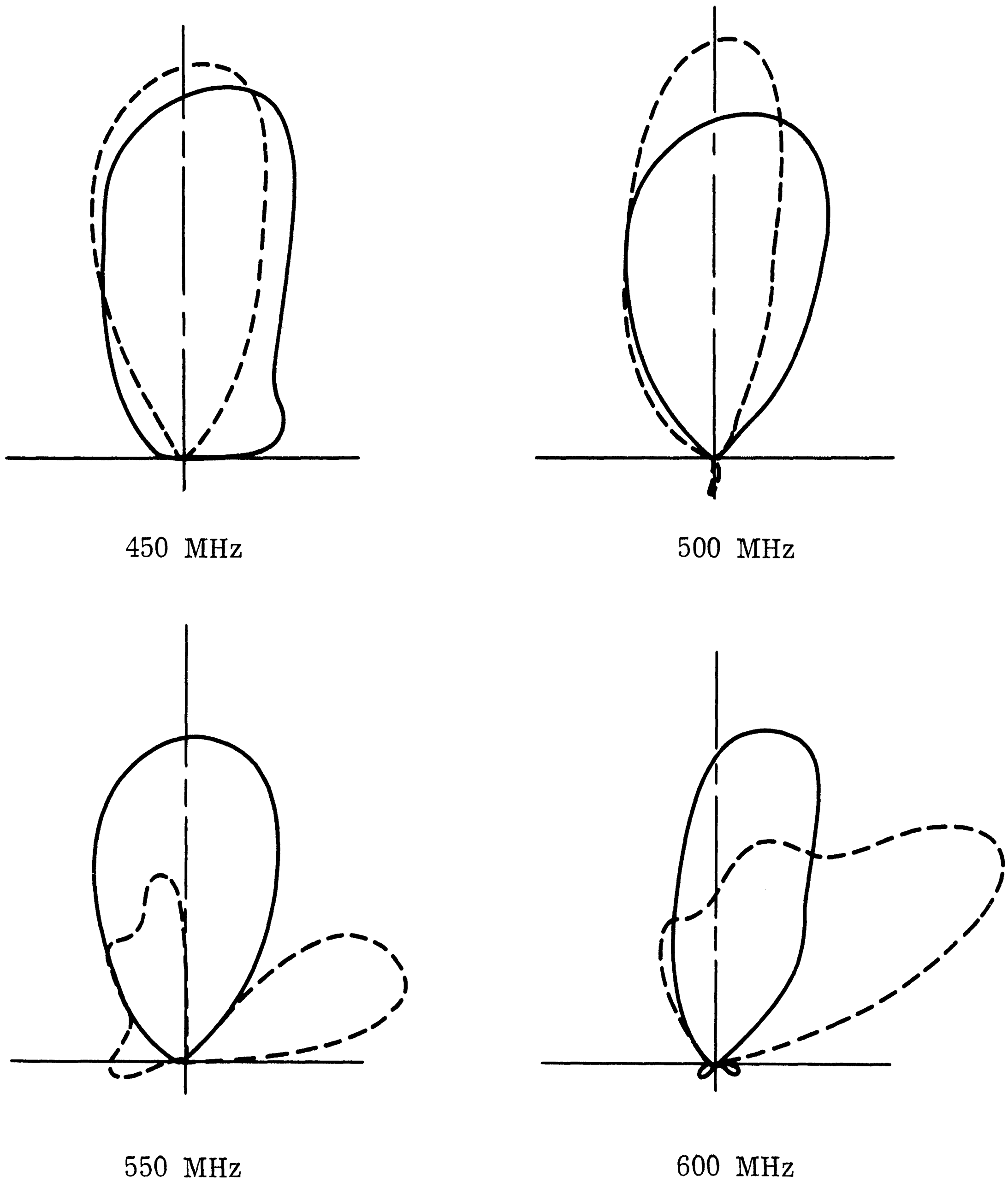


FIG. 3-11-b: THE FAR FIELD PATTERNS OF ANTENNA B-2 LOADED AND UNLOADED WITH EAF-2 FERRITE. (——) Unloaded, (---) Loaded.

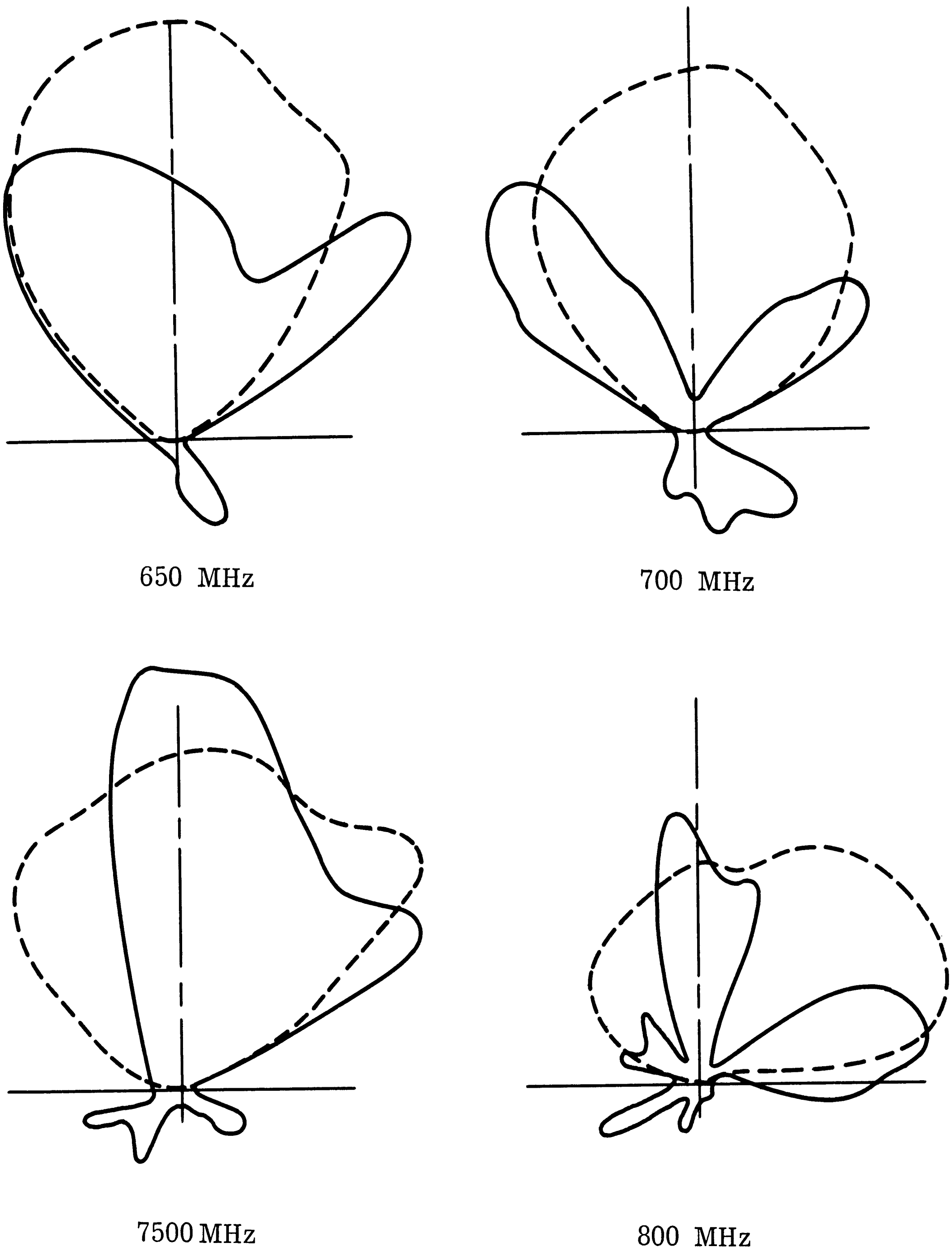


FIG. 3-11-c: THE FAR FIELD PATTERNS OF ANTENNA B-2 LOADED AND UNLOADED WITH EAF-2 FERRITE. (—) Unloaded, (---) Loaded.

IV

FERRITE ROD ANTENNAS

4.1 Analysis

In a previous report by this laboratory (Lyon et al, 1966), a general outline was described which is employed to analyze the characteristics for different magnetic and dielectric properties of the medium used for construction of a rod radiator. A rod infinite in length was assumed and the boundary value problem was solved for the case of the hybrid mode, HE_{11}. A determinantal equation was derived of the form:

$$\left(\frac{2\pi a}{\lambda}\right)^2 \frac{1}{(k_0a)^2}\left\{\frac{H_1'^{(2)}(k_0a)}{H_1^{(2)}(k_0a)}\right\}^2 + \left(\frac{2\pi a}{\lambda}\right)^2 \frac{\mu_r\epsilon_r}{(k_1a)^2}\left\{\frac{J_1'(k_1a)}{J_1(k_0a)}\right\}^2$$

$$-\left(\frac{2\pi a}{\lambda}\right)^2\left(\frac{\mu_r+\epsilon_r}{(k_0a)(k_1a)}\right)\left\{\frac{J_1'(k_1a)H_1'^{(2)}(k_0a)}{J_1(k_1a)H_1^{(2)}(k_0a)}\right\}$$

$$= -\left\{\frac{1}{(k_1a)^2} - \frac{1}{(k_0a)^2}\right\}^2\left\{\frac{(k_0a)^2\mu_r\epsilon_r - (k_1a)^2}{\mu_r\epsilon_r - 1}\right\} \tag{4.1}$$

with γ defined as

$$\left.\begin{aligned} \gamma^2 &= \left(\frac{2\pi}{\lambda}\right)^2 - k_0^2 \qquad k_0 \text{ is eigenvalue for outside region} \\ \gamma^2 &= \left(\frac{2\pi}{\lambda}\right)^2 \mu_r\epsilon_r - k_1^2 \qquad k_1 \text{ is eigenvalue for inside region} \end{aligned}\right\}. \tag{4.2}$$

The above equations (4.2) yield:

$$(k_1a)^2 - (k_0a)^2 = \left(\frac{2a}{\lambda}\right)^2 \pi^2 \left[\mu_r\epsilon_r - 1\right] \quad . \tag{4.3}$$

Then an expression relating the axial phase velocity, v, of the HE_{11} mode to the speed of light, c, the dimensions of the rod and wavelength, λ, is given by:

$$c/v = \frac{1}{\left(\frac{2a}{\lambda}\right)\pi} \sqrt{\frac{(k_1a)^2 - \mu_r\epsilon_r(k_0a)^2}{\mu_r\epsilon_r - 1}} \quad . \tag{4.4}$$

It is this latter expression giving c/v as a function of $2a/\lambda$ which serves an engineering purpose in the design of a rod for a particular free space wavelength λ, when k_1, k_0, μ_r, ϵ_r, are given. With Eq. (4.4) in mind, Eqs. (4.1) and (4.2) were solved simultaneously in the computer for k_0a and k_1a a given μ_r, ϵ_r and $2a/\lambda$. Once k_0a and k_1a were known, plots of c/v vs. $2a/\lambda$ were drawn.

4.2 Design Interpretations

It is to be observed that this analysis assumes a rod of infinite length. The HE_{11} mode is studied because of the no-cutoff properties which it exhibits and the assumption of yielding end-fire radiation when a finite length is considered. It should be emphasized that when a finite rod is used as an antenna, no simple HE_{11} mode is present; moreover none of the other simple mode descriptions of the rod which apply to an infinite waveguide are present. Thus, in the case of the rod antenna it would be safer to speak of the HE_{11} mode component. For the finite rod it is assumed that the HE_{11} component is the most important part of the field configuration.

From the analysis it is seen that the diameter of the rod is directly related to the axial phase velocity and that, in fact, a diameter is desired such that $c/v \simeq 1$ for a particular frequency. This criterion is necessary for propagation of the mode in the axial direction with most of the energy propagating exterior to the rod, yet directed by it. In the case where c/v is substantially greater than one the mode propagates inside the rod, thus contributing to excess losses (if the material is lossy) and reflections due to discontinuities at the ends.

In Figs. 4-1 and 4-2, graphs of c/v vs $2a/\lambda$ are presented for different values of μ_r and ϵ_r. Any of these graphs would describe the behavior of any ferrite material with proper μ_r and ϵ_r values. From previous knowledge of dielectric rods the bandwidth of the antenna is limited by the feeding system and ϵ_r. In fact, the higher ϵ_r, the smaller the bandwidth of operation with $c/v \simeq 1$ for a range of values of $2a/\lambda$. This seems to be the case for the ferrite rod also.

4.3 Experimentation

An experiment is being employed where the feeding system is a helix antenna. The ferrite material used is EAF-2 ferrite powder which limits the frequency of operation of the antenna to a center frequency of 400 MHz. Other feeding arrangements will also be employed.

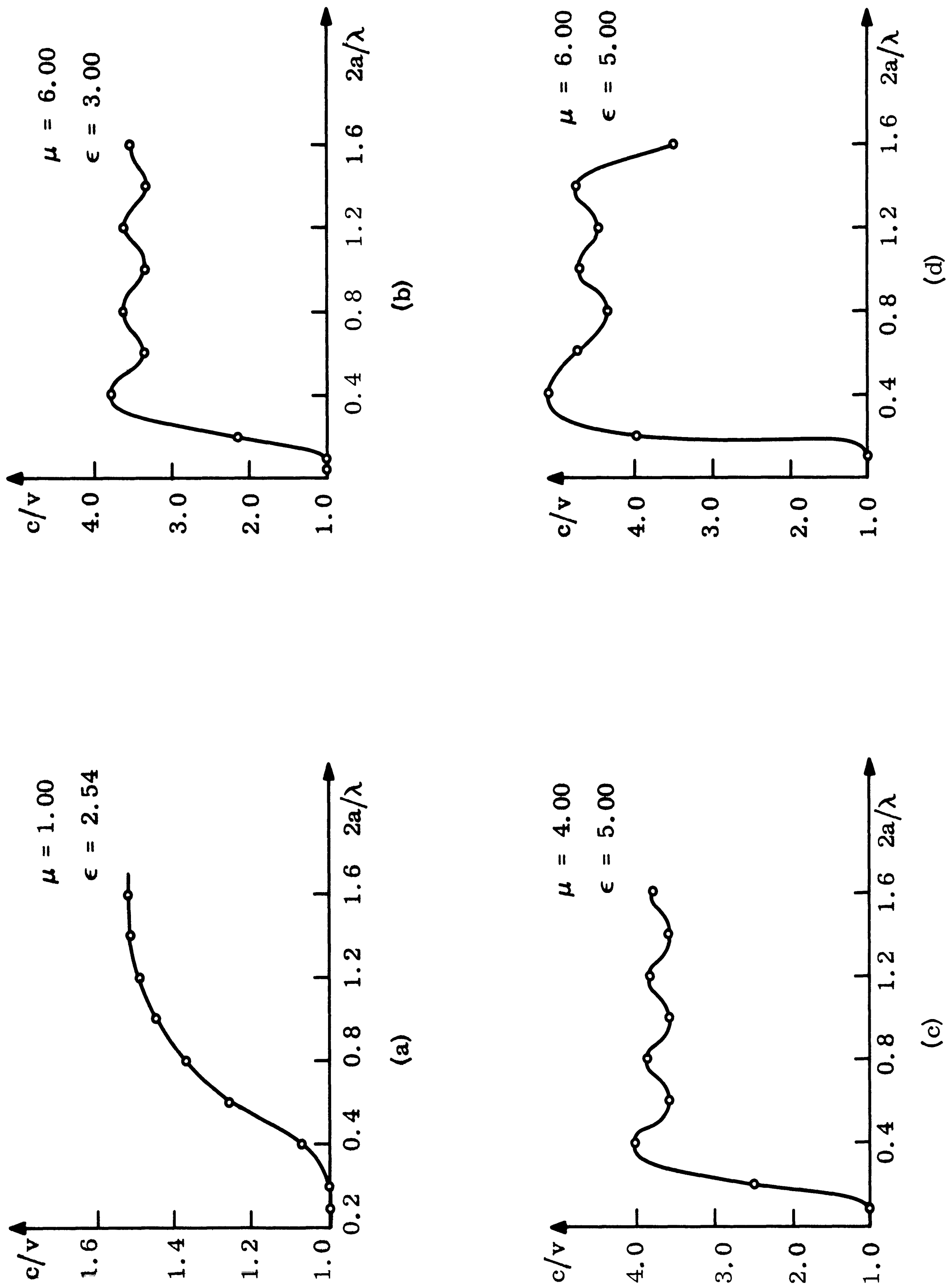


FIG. 4-1: VELOCITY CURVES FOR SELECTED FERRITE RODS.

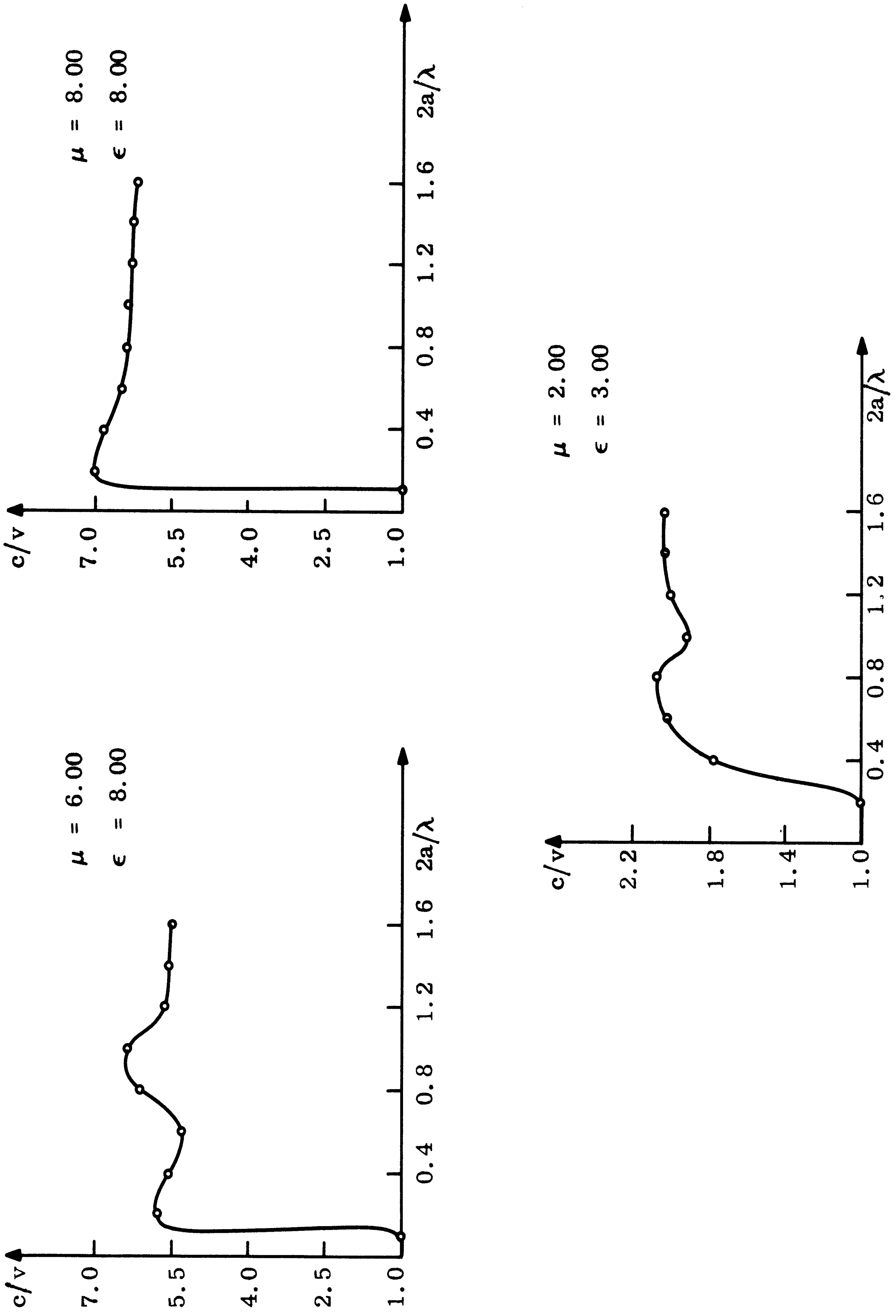


FIG. 4-2: VELOCITY CURVES FOR SELECTED FERRITE RODS.

V

LOW FREQUENCY FERRITE ANTENNAS

The objective of this task is to investigate the design feasibility of new types of ferrite antennas that are useable at frequencies as low as 30 MHz. For convenience, the types of ferrite antennas have been classified according to the structure of their radiating elements as follows: 1) linear, 2) loop, and 3) frequency independent. The work reported below has been directed toward ferrite loading of linear radiating elements.

An effort has been made to identify realistic applications of ferrite loading to linear radiating elements. Accordingly, the investigation has focused upon applications which improve the performance of antennas that are relatively small. The range of sizes considered are $0.1\lambda < 2h < 0.5\lambda$, where λ = free space wavelength, and h = element half-length. The low end was chosen so as to avoid severe supergain limitations in element performance, while the high end was considered to be a practical size limit for a loaded 30 MHz element. Moreover, the detailed discussion is limited to center fed (or ground plane imaged) elements which support standing wave current distributions.

5.1 General Discussion on Linear Elements

Parameters which are considered most important in describing element performance are: (a) radiation pattern, (b) radiation efficiency, (c) impedance bandwidth, and (d) element weight. The applicability of ferrite loading is assessed by its influence upon these parameters. As the relative importance of these parameters vary according to the intended antenna application, so does the relative applicability of ferrite loading. The remainder of this section discusses the relationship between ferrite loading and the parameters, and points out the manner in which the parameters may be enhanced by the use of ferrite.

5.1.1 Radiation Pattern

The E-field radiation pattern of a linear element having a standing wave current distribution for which $2h < .5\lambda$ is always a single lobe. The lobe has its maximum in the plane normal to the dipole axis, and possesses E-plane symmetry. E-plane beamwidth depends (1) upon element length, and (2) upon phase velocity (Stephenson and Mayes, 1966). The radiation pattern for small element is proportional to $|\sin\theta|$. The pattern becomes somewhat more directive as both h (element half-length) is increased, and v_p (phase velocity along the element) is decreased. The directivity increase is rather insignificant for practical values of h and v_p. Consequently, the addition of ferrite loading to linear elements, with its inherent effect upon v_p, will not appreciably effect the radiation pattern. This result allows the use of ferrite enhancing other element parameters without fear of causing pattern deterioration.

5.1.2 Radiation Efficiency

While the radiation pattern is fairly insensitive to h, v_p, and current distribution, the radiation efficiency is quite dependent upon these factors. Radiation efficiency is determined from the ratio of radiation resistance to ohmic resistance, the efficiency being greatest when the ratio is large. Since a certain amount of ohmic resistance is unavoidable, it is important to maintain high radiation resistance. Attainment of this objective becomes increasingly difficult for very short radiation elements. For an unloaded short linear element, radiation resistance is proportional to h^2, whereas the conductive ohmic resistance is proportional to h.

It is well known that the maximum radiation resistance for a given element length $2h < .5\lambda$, assuming the value of current at the feed is greater than or equal to the value anywhere else on the antenna, occurs when the standing wave current distribution is uniform; this corresponds to $I(z) = I_o$ (constant amplitude), over the entire element length. However, to a fair approximation, the current distribution for an unloaded thin element is: $I(z) = I_m \sin k(h - |z|)$. For a short element, this distribution is essentially triangular. A conventional technique for optimizing the short element current distribution is to end load the element. This tends to create the desired uniform current distribution, and hence increase the radiation resistance for a given length, h. Another effective technique sometimes used is to uniformly load the element, so as to decrease the phase velocity of the current. This results in a current distribution approximated by: $I(z) = I_m \sin \beta(h - |z|)$, where $\beta = \omega/v_p > k = \omega/c$. This distribution has a larger current moment, and hence a larger radiation resistance, than that of an unloaded element of the same length.

In employing these techniques, the ohmic resistance is also increased somewhat. Hence, a trade-off is usually reached in obtaining the maximum ratio of radiation resistance to ohmic resistance. This same trade-off exists when lumped and distributed ferrite loadings are incorporated into the techniques described above. Depending upon the quantity and placement of ferrite loading, the radiation efficiency parameter can vary appreciably.

5.1.3 Impedance Bandwidth.

The input impedance of a linear element can be represented for small frequency deviations about any frequency, except for anti-resonance, by a series resistance and reactance. Moreover, , the derivative with respect to frequency of this reactance will always be positive, unless some impedance compensation network is acting upon the element terminals. When a linear element is operated at a frequency away from self-resonance, a pure reactance is sometimes placed in series with the terminals of the element to make the total impedance series resonant. Since the frequency derivative of any passive purely reactive single element impedance is positive, the frequency sensitivity of the resulting series resonant circuit will be increased. For this situation it may be said that any element loading which has the effect of adding series reactance will increase the frequency sensitivity, and hence decrease the elements impedance bandwidth.

An alternative approach which tends to increase the impedance bandwidth is the insertion of a parallel resonant circuit across the element terminals. In this case, the susceptances of the element and the parallel resonant circuit tend to cancel over some frequency band. An optimum configuration using a parallel resonant circuit has been treated by Shnitkin and Levy (1962). This approach works well at VHF-UHF frequencies where stubby dipoles having low Q may conveniently be constructed. For applications at lower frequencies, its usefulness decreases since elements are generally of higher Q.

In summary, a single linear element of fixed length can have its impedance bandwidth increased by external compensating techniques. The simplest of these compensating techniques is a parallel resonant circuit across the terminals of the element. At the expense of added complexity in the external compensating circuit, one can obtain only successively smaller incremental improvements in impedance bandwidth.

Multiple linear elements which are tightly coupled may be interconnected to serve a dual purpose. That is, in addition to serving as a linear radiator, the elements can function as transmission line impedance compensators. The sections of transmission line essentially function as the elements in the compensating cirucit discussed previously. An example of such a configuration is the conventional $\lambda/2$ folded dipole. The details of the compensation will be described in Section 5.2.2.

As mentioned under the discussion of radiation efficiency, the application of ferrite loading can alter the element current distribution, and hence the element input impedance. In particular, ferrite may be used to advantage in raising the radiation resistance. Experience has indicated that any increase in radiation resistance of a single linear element due to ferrite loading can be accompanied by a corresponding increase in reactance at some frequencies. Consequently there may be no net increase in impedance bandwidth. A net decrease in bandwidth may sometimes occur (Taylor, 1955). Similar dependencies may apply for multiple linear elements, in consideration of overall frequency sensitivity. However, one may trade improved impedance bandwidth over some frequency interval for accentuated frequency sensitivity elsewhere. In view of the built-in impedance compensation ability of multiple linear elements, it is theoretically possible to improve impedance bandwidth over some finite frequency interval. It is expected that this impedance compensating ability may be enhanced by an appropriate application of ferrite loading. Potentially promising configurations will be discussed in Section 5.2.2.

5.1.4 Element Weight.

The application of ferrite is but one technique for optimizing the three electrical parameters previously considered. For instance, dielectric material loading can often be used to advantage. However, ferrite may be preferred over dielectric loading in order to obtain a wider frequency band. Moreover, phase velocities can be varied by utilizing a small diameter helix (coiled conductor) which serves as an approximation to a linear element. The merits of ferrite loading must be assessed in relation to its effectiveness compared to these other methods of loading. To this end, element weight is considered a constraint upon which a comparison of methods may be made.

5.2 Advantages of Multiple Linear Elements

It was pointed out in Section 5.1.3 that multiple linear elements which are tightly coupled can be interconnected to serve a dual purpose. In addition to performing as linear radiators, the elements also serve as sections of transmission line. This latter function can be utilized to provide both impedance matching and impedance broadbanding inherent within the antenna structure. Moreover, the applications of material loading can be used to accentuate this matching and broadbanding capability. Sections 5.2.1 and 5.2.2 discuss the specific manner in which these capabilities may be enhanced.

The essential features of multiple linear elements are manifested in the transition from one single element to two coupled elements. For this reason the study has focused upon the two element case, although the same principles apply to three or more elements. Appendix A outlines the analysis of two parallel linear elements, and furnishes the appropriate equations for the case of tight coupling. Tight coupling is essential in order to apply transmission line theory to our advantage. As the appendix explains, the antenna excitation can be decomposed into a symmetric (radiating) mode, and asymmetric (transmission line) mode. Since radiation from the asymmetric current is negligible (due to the close element spacing), this mode may be altered to our advantage in controlling the impedance variations without experiencing detrimental effects upon the radiation pattern. Various geometrical arrangements exist for altering the asymmetric mode without effecting the symmetric (radiation) mode. Additional freedom is obtained by applying material loading techniques, since the material has dissimilar electrical effects upon the two modes. In particular the current phase velocity differs for the two modes. This is a result of the interaction between the fixed material and dissimilar field distributions of the two modes.

5.2.1 Internal Impedance Matching

To illustrate the use of multiple linear elements for impedance matching, consider first a structure corresponding to the familiar $\lambda/2$ folded dipole. This is a special case of the formulation in the Appendix A corresponding to: $Z_L = Z_T = 0$, $\beta_s = \beta_a = k$, $h = \ell = \lambda/4$. Neglecting losses, the pertinent relations from equations (A.3), (A.6), and A.8) are:

$$Z_{in} = \frac{2Z_s Z_a}{Z_s + Z_a}$$

$$Z_s = 2\,Z_{11}\,(a_e)$$

$$Z_a = j\,Z_o \tan kh = j\,\infty .$$

The above yield the familiar result of $Z_{in} = 4\,Z_{11}(a_e)$, where $Z_{11}(a_e) \simeq 73\Omega$. The factor of 4 in Z_{in} (which for N linear elements is equal to N^2) is also a consequence of the two elements having equal radii ($a_1 = a_2 = a$), and may be varied by making $a_1/a_2 \neq 1$. Of more importance is the fact that this factor is unchanged if $h \neq \ell \neq \lambda/4$, provided only that $|Z_a| = \infty$. This can be accomplished in either of two ways: 1) by adding material loading so that $\pi/2 = kh = \beta_a \ell$; or 2) by adding a parallel capacitor across the antenna terminals which will resonate with $Z_a = 2jZ_o \tan k\ell$. The factor 2 arises since there are two transmission line segments of length $k\ell$ connected in series across the antenna terminals. Since the factor of 4 can be maintained independent of h, one can foreshorten the antenna to any extent and still have Z_{in} equal 4 times the impedance of a single foreshortened element. This technique is an efficient way of raising the inherently low impedance of foreshortened elements.

As a second example consider the slightly more general case where $h = \lambda/4$, while ℓ is allowed to vary ($0 \leq \ell \leq h$). The input impedance Z_{in} is seen to be twice the parallel combination of $Z_s \simeq 146\,\Omega$ and $Z_a = jZ_o \tan k\ell$. Since $0 \leq |Z_a| \leq \infty$ as $0 \leq \ell \leq h$, $|Z_{in}|$ may be adjusted continuously from 0 to $296\,\Omega$. Moreover, $R_e\left\{Z_{in}\right\}$ may be set at any desired level from 0 to $296\,\Omega$ by proper adjustment of ℓ, and $\mathrm{Im}\left\{Z_{in}\right\}$ made zero by a slight readjustment of h, or by adding a lumped reactance at the antenna terminals. This versatility allows a wide range of input impedances with a minimum of external matching. A number of curves illustrating the effect have been calculated by Chang, and appear in a reference (King, 1956, p. 345). A popular version of this fundamental principal is the so-called "T"-match. Similar results are obtained when material loading is incorporated.

While it was indicated in the first example the material loading was a practical way of maintaining $|Z_a| = \infty$ (via adjusting $\beta_a \ell$ to equal $\pi/2$), the unique advantage of material loading is its ability to enhance frequency compensation, as described in the next section.

5.2.2 Internal Impedance Broadbanding

The impedance broadbanding capability of multiple linear elements can best be illustrated by again considering first the familiar $\lambda/2$ folded dipole, described in Section 5.2.1, and then its generalization. Two distinct mechanisms tend to increase the impedance bandwidth of this structure. First, the impedance associated with the symmetric excitation mode, $Z_s = 2Z_{11}(a_e)$, corresponds to twice that of a single linear element having an effective radius $a_e = \sqrt{ab}$. Since this effective radius is substantially larger than the radius of an individual element, the reactance variation with respect to frequency about the resonant point is reduced from what it would be for either element operating alone. The second mechanism is that of susceptance cancellation between the parallel impedances of the symmetric mode and the asymmetric (transmission line) mode. Take, for example, the situation when the structure is operated slightly below the normal $\lambda/2$ resonant frequency. Z_s is capacitive since the elements are electrically shorter than $\lambda/2$ resonance, while Z_a is inductive since the elements function as a transmission line of electrical length $< \pi/2$ terminated in a short circuit. Slightly above the resonance the reverse situation occurs.

While the susceptances of the two modes oppose each other, the cancellation is not complete. This is because the susceptance of Z_a which is given by $\left(\frac{1}{jZ_o} \cot \frac{\pi f}{2f_o}\right)$ is much smaller near resonance than that of Z_s. However, for appreciable frequency excursions from resonance (perhaps 25 percent) the susceptances will cancel, thus forming two additional resonant frequencies. The impedance characteristics about these resonant frequencies are described in references (Beam et al, 1952 and King, 1956). It can be observed that the impedance variation with frequency change is high at the lower resonant point, whereas the impedance variation is more attractive at the higher resonant frequency. In particular, at this higher resonant frequency the input resistance is only moderately higher than at normal $\lambda/2$ resonance. Moreover, the reactance change with frequency is broad and negative, which offers the possibility of easy compensation. The primary disadvantage of operating at this resonance point seems to be the necessary increase in element length.

The foregoing discussion has been on the ordinary $\lambda/2$ folded dipole for which $\beta_s = \beta_a = k$. It is anticipated that the application of material loading will improve the utility of the folded dipole. For instance, applying coatings of ferrite and dielectric material to the elements results in $\beta_a = \omega/v_{pa} > k$. The appropriate formula for describing the phase velocity reduction was presented in Section 2.2 in connection with the helix discussion. While the material coatings will also increase β_s, the quantitative effects of the increase will be dissimilar, since the near zone field distributions resulting from the symmetric and asymmetric current modes are unlike. Hence, this affords an independent control of the resonant frequencies for Z_s and Z_a. In addition the coatings affect the frequency sensitivity of Z_a, and also the value of Z_o. The interplay between these various parameters is being investigated, with the expectation that the increase versatility can be put to advantage in improving the impedance bandwidth. Several proposed applications are now awaiting experimental verification. These will be described in detail along with measured results in a future report.

VI

FERRITE AND OTHER LOADING MATERIALS

One of the difficulties in using ferrite materials in antennas is that most ferrites become exceedingly lossy in the UHF range. Some powdered ferrites, such as the EAF-2, can be used up to 600 MHz (Lyon et al, 1966). However, the relative permeability is low. This is a common characteristic of all ferrites: to get low loss at higher UHF, the relative permeability must be small (Lyon et al, 1965).

As was reported earlier, (Lyon et al, 1966), Emerson and Cuming Eccosorb CR, a microwave casting resin absorber, turns out to be a fairly good magnetic material in the UHF range. To determine if other microwave absorbing materials would be good magnetic materials in the UHF range, a survey was conducted of all producers listed under "Absorbers, Microwave" on page 11 of the EEM for 1966-67 (Electronic Engineers Master, 1966). Several companies replied to the questionnaire and some of these organizations do produce absorbers that are good magnetic materials in the UHF range.

6.1 Useful Materials

Emerson and Cuming appears to produce the largest selection of materials that can be used in the UHF range. Their Eccosorb MF series includes six materials that have a "Q" of about 20 and a relative permeability ranging from 1 to 5. The relative dielectric constant ranges from about 2 to 27. These materials have a specific gravity of about 4.0 and are quite strong physically, although brittle.

Stock sizes of these materials are 1, 2, and 3 in. diameter rods that are 12 in. long, and 12 by 12 in. sheets that are 1/2, 1 and 2 in. thick. The Eccosorb CR mentioned earlier is a casting resin identical to one member of this family, except it can be cast into any desired shape.

6.2 Possible Useful Materials.

Four companies supplied information about microwave absorbers; there is some indication that these products may be good over part of the UHF band, although the useful upper frequency would be near 300 to 500 MHz. However, the data supplied is not sufficient to determine precisely what the upper limit would be.

Custom Components, Inc. produces a type MP2312 Polyiron that has an attenuation of 160 db per inch at 3.3 GHz. More than likely, this material would be too lossy even at the lower UHF band, although there is not enough information to determine this precisely.

Co-Ax Devices Inc. produces a microwave absorber that is recommended for use between 500 and 2000 MHz as an absorber. So possibly, below 500 MHz, this would be a low loss material. Again, however, insufficient information was supplied.

Core-Tronics Inc. produces several microwave absorbers that, unfortunately, appear to be good absorbers throughout most of the UHF band. These materials would probably be useful only in the VHF band. One company, Magnetic Core Corporation, supplied information on torroid core materials that are useful up to about 250 MHz. However, no information was supplied on materials sold expressly as absorbers.

Three companies, Raytheon Company, Sperry Microwave Electronics Company, and Countis Industries, sell microwave ferrite material, but their data does not include relative permeability or magnetic loss tangent. These materials are intended primarily for use in circulators, phase shifters, isolators, and other microwave ferrite components. Possibly, these materials could be of use as low loss materials at UHF. However, insufficient information is available at this time. These materials may be no better than existing ferrites.

One company, Sponge Products Division of B. F. Goodrich Co., has delayed in forwarding information. Hence the acceptability of their materials cannot be determined.

6.3 Non-Useful Materials.

The following companies replied to the survey, but offered no information on suitable materials: Silk City Industrial Ceramics Inc., Waveline Inc., Microtech Division of Ovitron Corp., Custom Materials Inc., and Conductron Corp. The latter company does make UHF ferrites and they were helpful in discussing general characteristics of ferrite materials.

6.4 Summary.

At the present time, only the Eccosorb MF series produced by Emerson and Cuming is useful over a sizable portion of the UHF spectrum. To cover the higher end of the range, however, very low relative permeabilities would have to be accepted to obtain a "Q" greater than 10. In a conversation at Dayton, Ohio last August by one of the authors with Dr. E. F. Buckley, of Emerson and Cuming it was pointed out that the properties of the MF series of materials could be expected to be within 10 percent of the published values. Consequently, the Eccosorb MF series appears to be a promising line of materials.

VII

CONCLUSION

Progress has been made under each of the four assigned task areas. Under Task 1, on the loaded log conical antenna, sufficient study has been made so that a developmental model is now possible. Under Task 2, a new antenna called the interdigital array has been studied. It is believed that this antenna has certain features which will make it useful on aerospace systems where there is a strict limitation on available space. This antenna can be called an array since it has several elements; on the other hand, since the individual elements cannot be subjected to program control, it is more appropriate to think of it as a separate antenna to be mounted in the surface of an aerospace vehicle. This antenna would provide broadside radiation from the vehicle. It also can be operated to produce substantial endfire radiation.

On the basis of the analysis shown in this report, it can be concluded that useful endfire rod antennas, suitable for the 300 to 1000 MHz range, can be produced. The required dimensions are now known and models are under construction.

Under Task 4, the basic studies related to antenna types useful at 30 MHz have been carried out in considerable detail. It is apparent from these studies that there are inherent limitations on the use of linear elements and groups of linear elements as far as bandwidth, radiation pattern, efficiency, and weight and size are concerned. In some instances, it has been found that methods of reducing size have been very detrimental to the frequency bandwidth possible.

VIII

FUTURE EFFORT

In the continuation of the study of UHF-VHF antennas utilizing ferrite loading, it is expected that more emphasis will be put on Task 2. This is in accordance with recent conversations with the project monitor. This concentration of effort will be directed largely toward exploiting the advantages of the relatively small ferrite-loaded slots as elements in an array. Some advantage may be obtained because of the constancy of the driving point impedance as frequency is varied. Another possibility to be considered is the use of an array within an array, since each of the elements will require less space than the usual slot element.

It is anticipated that considerably more experimental data will be obtained for Task 3, on endfire ferrite rod antennas, and for Task 4, on ferrite antennas suitable for operation at 30 MHz. For both of these tasks, emphasis will be on experimental work.

The work under Task 1, involving the physically small log conical spiral antenna, will be continued with further study of slow wave conductors. It is also expected that lumped ferrite inductance loading will continue.

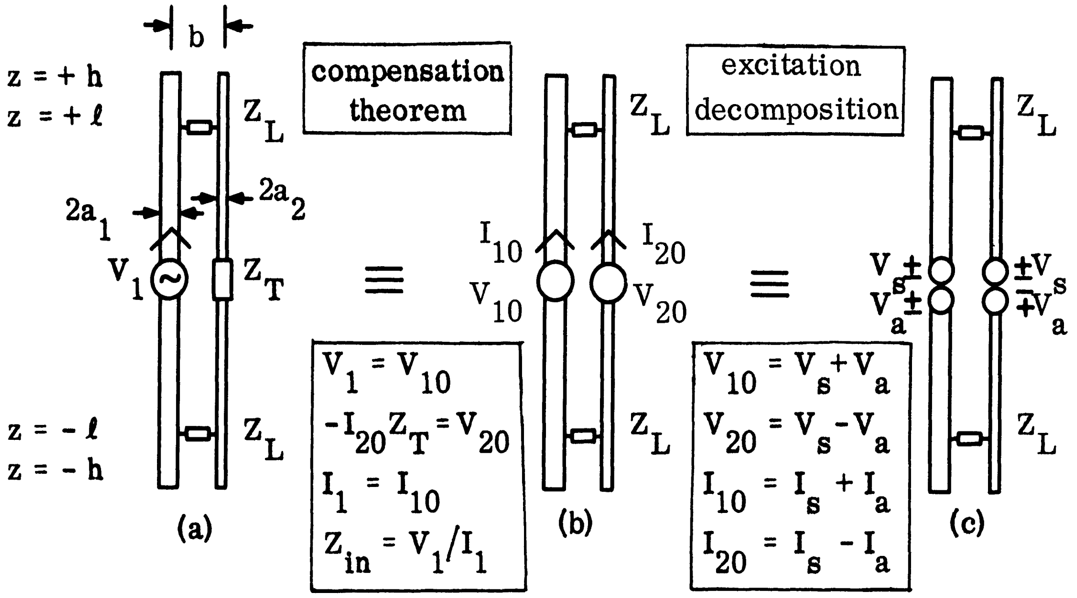
APPENDIX A

ANALYSIS OF TWO PARALLEL LINEAR ELEMENTS

A.1 General Formulation.

Consider two parallel linear elements of half-length h and separation distance b as shown in Fig. A-1 (a). Element 1 is center driven with a voltage source V_1, while element 2 is center loaded with an arbitrary terminating impedance Z_T. In addition, arbitrary impedances Z_L are placed between the two elements a distance ℓ either side of center. This problem may be analyzed by: (1) applying the compensation theorem to replace Z_T by an equivalent generator; and (2) decomposing the excitation into symmetric and asymmetric excitation components. These steps are shown in Fig. A-1 (b) and (c), respectively. By applying the principle of superposition, one may solve for the current due to the symmetric and asymmetric excitation as two separate problems and add the results to obtain the desired solution. Strictly speaking, the current distribution resulting from each mode of excitation should be determined as a boundary value problem. Under certain limiting conditions, results having acceptable accuracy may be obtained by using a simpler analysis. In any event, the distribution of currents is uniquely determined by the geometrical structure of the element, which is independent of the excitation magnitude. The sum of the resulting two current distributions is then the solution to the original problem. The current on each element is given by Eq. (A.1), which appears with Fig. A-1. A symmetric impedance Z_s and an asymmetric impedance Z_a, may be indentified with each mode of excitation (Eq. A.2). These mode impedances along with Z_T determine the input impedance Z_{in} (Eq. A.3). Moreover, the actual magnitude of mode excitation is determined by the impedances and the voltage source V_1 (Eq. A.4).

A mathematical simplification results if the linear elements are thin ($h^2 \gg b^2 \gg a^2$), identical ($a_1 = a_2 = a$), and closely spaced ($b/h \ll \lambda$). Although not strictly necessary for some statements, this situation is assumed in all



$$I_1(z) = I_s(z) + I_a(z) \tag{A.1}$$

$$I_2(z) = I_s(z) - I_a(z)$$

$$Z_s \equiv \frac{V_s}{I_s(o)} \qquad Z_a \equiv \frac{V_a}{I_a(o)} \tag{A.2}$$

$$Z_{in} = \frac{2Z_sZ_a + (Z_s + Z_a)\,Z_T}{Z_s + Z_a + 2Z_T} \tag{A.3}$$

$$V_s = \left[\frac{Z_s\,(Z_a + Z_T)}{2Z_sZ_a + (Z_s + Z_a)\,Z_T}\right] V_1$$

$$V_a = \left[\frac{Z_a\,(Z_s + Z_T)}{2Z_sZ_a + (Z_s + Z_a)\,Z_T}\right] V_1 \tag{A.4}$$

FIG. A-1: ELEMENT EXCITATION DECOMPOSITION, AND FUNDAMENTAL EQUATIONS.

that follows. In such cases the solutions simplify to those of two tightly coupled antennas, for the symmetric excitation, and a substantially non-radiating transmission line for the asymmetric excitation. The effect of Z_L upon the symmetric excitation is negligible since its ends are at equipotential points. The effect of Z_L upon the asymmetric excitation depends upon its value and position on the structure. The solutions to the two parts of the formulated problem are now outlined.

The problem of coupled equal length linear elements not incorporating material loading, but with an arbitrary terminating impedance Z_T, has been thoroughly treated, among others, by Tai (1948). Tai has an intergral equation formulation which yields accurate results on the impedance and the current distribution. Analytic results are also presented in graphical form for selected element lengths and thicknesses, as a function of element spacing. While Tai's formulation may be used where precision is warranted, its immediate usefulness is in establishing the appropriateness of invoking less exact but simpler formulations. In particular, the results indicate that the EMF method, which assumes a sinusoidal current distribution, yields acceptably accurate impedance information for summetric excitation over the range of element lengths of present interest. Furthermore, for close element spacing the exact analysis for asymmetric excitation coincides with the solution derived from transmission line theory. With the foregoing as justification, the viewpoint is taken that suffi cient accuracy is obtained by using the EMF method, and transmission line results for the symmetric and asymmetric excitation portions of the problem, respectively.

A.2 Symmetric Excitation Mode.

Uniform material loading is depicted in the formulation as follows: The first order effect of the material is to alter the phase velocity of the symmetric currents flowing on the structure. While certain second order aspects of this alteration may in some instances be difficult to evaluate in an analytical fashion, they are amendable to experimental determination, as is the justification of assumptions involved. Accordingly, the symmetric excitation current distribution for each element is approximated by:

$$I_s(z) = \frac{V_s}{Z_s} \sin \beta_s (h - |z|), \qquad 0 \leq |z| \leq h \tag{A.5}$$

where:

$\beta_s = \omega/v_{ps} > k$, symmetric excitation wave number.

The symmetric excitation impedance is given by:

$$Z_s = 2Z_{11}(a_e) \tag{A.6}$$

where:

$Z_{11}(a_e)$ = input impedance of a single linear element of length 2h with effective radius $a_e = \sqrt{ab}$.

Numerical values of Z_{11} are obtainable from the EMF method. Curves are presented by Moore and Beam (1963) for the case $\beta_s = 2k$. Since impedance information for general β_s is not widely available, one may need to rely on measurements for specific information. Nevertheless, the general behavior of Z_{11} for $\beta_s \neq k$ is evident from the reference.

A.3 Asymmetric Excitation Mode.

The solution for asymmetric excitation is obtainable directly from transmission line theory. A general formulation is used which takes into account ohmic losses, since in certain instances relatively high transmission line currents may exist. The results are presented below, whre many of the symbols have been previously defined. The asymmetric excitation current distribution is given by:

$$I_a(z) = \begin{cases} I_o \left[e^{\gamma(\ell - |z|)} - \Gamma_o \, e^{-\gamma(\ell - |z|)} \right], & 0 \leq |z| \, \ell \\ I_o \dfrac{1 + \Gamma_o}{\cosh \gamma(h - \ell)} \quad \sinh \gamma(h - |z|), & \ell < |z| \leq h \end{cases} \tag{A.7}$$

where

$$I_o \equiv \frac{V_a}{Z_o} \frac{1}{\left[e^{\gamma\ell} + \Gamma_o e^{-\gamma\ell} \right]} ,$$

Z_o = characteristic impedance of the transmission line,

$\gamma = \alpha + j\beta_a$ complex propagation constant, and

$$\Gamma_o = \frac{Z(\ell) - Z_o}{Z(\ell) + Z_o} \quad \text{complex reflection coefficient.}$$

The asymmetric impedance is given by

$$Z_a = Z_o \frac{Z(\ell) + Z_o \tanh \gamma\ell}{Z_o + Z(\ell) \tanh \gamma\ell} \tag{A.8}$$

where

$$Z(\ell) = \frac{Z_{oc}(\ell) Z_L}{Z_{oc}(\ell) + Z_L} , \text{ and}$$

$$Z_{oc}(\ell) = Z_o \coth \gamma \, (h - \ell).$$

Knowing Eqs. (A.6) and (A.8) allow the determination of Z_{in}, Eq. (A.3). Moreover, since V_s and V_a are now known by Eq. (A.4), the element currents have been determined through Eq. (A.1), with (A.5) and (A.7).

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Under Task 1. a log conical spiral antenna having a diameter of 21 cm at the base and a length of 65 cm along the axis is described. Various possible loading techniques are discussed which may meet the frequency bandwidth requirements for the antenna. One objective is to obtain a conductor which in itself is a slow wave structure.

Under Task 2, considerable progress has been made in obtaining an antenna which is a series of finger-like elements. This antenna is called an interdigital array. Effects of loading on this interdigital array are discussed. Under this same task, an array with ferrite filled slots capable of being controlled magnetically is also under consideration. During this report period, very little effort has been put on this type of array. However, in continuing Task 2, a major part of the effort will be placed on such an array using ferrite filled slots as elements.

Under Task 3, an analytical effort is described providing the basic criteria for rod type antennas. The results of this analysis can be used for either ferrite rod or dielectric rod antennas. Experimental models of ferrite rod antennas are under constructio

Under Task 4, a survey has been made on various possibilities of reducing size and obtaining required bandwidth performance. Linear elements and combinations of linear elements have been considered in some detail.

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