THE STUDY OF MICROSTRIP ANTENNA ARRAYS AND RELATED PROBLEMS

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Technical Work

During this period efforts have been expended for the following tasks:

(1) Dual Frequency Microstrip Antennas

The work on rectangular microstrip antennas for dual frequency operation is nearly completed. Briefly, the principle of our approach is based on the excitation of a patch for two or more different modes which correspond to different frequencies. However, for a given geometry, the modal frequencies have a fixed relationship; therefore, the usefulness of such a design is greatly limited. In this study we have contrived three different methods to control the frequency ratio over a wide range. First, as found previously, if shorting pins are inserted at certain locations in the patch, the low frequency can be raised substantially. Second, if slots are cut in the patch, the high frequency can be lowered considerably. By using both techniques, the two frequency ratio can be varied approximately from 3 to 1.3. After that, the addition of more pins or slots becomes ineffective. To reduce the ratio further, we have found a third technique by using a microstrip line wrapped around the patch and connected in parallel with the patch. In so doing, the ratio can be reduced to about 1.07.

For the shorting pins in the patch, a multiple-port theory was developed and found to predict the antenna characteristics very well. As for the case of slots, a hybrid multiple-port theory has been formulated and more recently improved. Again the theory was found to yield results in excellent agreement with those measured. For the third method stated above, due to the strong coupling between the microstrip line and the patch, the cavity model theory appears to be less satisfactory. However, in spite of this, we were surprised to find that the simple theory can still predict very accurately the changes in resonant frequencies due to the microstrip line.

A paper which summarizes the first two methods was prepared for publication during this period. It has been accepted and will appear in the IEEE Transactions on Antennas and Propagation in the near future. A copy of this paper is attached with this report. Since the first part of this work was performed under a RADC grant and the second part under this NASA grant, the supports by both agencies are acknowledged.

(2) CP Microstrip Antennas

One of the most interesting applications of microstrip antennas is in producing circularly polarized (CP) waves with a single feed and without a phase shifting network. A theory as well as design formulas have been developed. Previously, our work considered mostly the feed on certain locations in the patch. But the general theory shows that this need not be so. In fact, for given dimensions of a nearly square patch there exist four loci for the feed, two for producing RCP and two for LCP. ' this period, a general investigation has been made for the purpose of determining whether or not some feed locations could give a wider bandwidth than some others. This work is yet to be completed. However, we are submitting a paper to the Antenna Applications Symposium to be held at the Allerton Park, Illinois, in September 19-21, 1984. A copy of the abstract of this paper is also attached.

(3) Improved Theory for Microstrip Antennas with Multiple Lumped Linear Loads

Previously, to alleviate some difficult problems (of somewhat minor importance) in our cavity model theory, we introduced two factors: the effective feed width and the effective loss tangent. Although they are found to be very useful, the theoretical rigor is often questioned. A rigorous analysis has been made which not only justifies their role, but also provides a much improved theory for multiple port microstrip antennas. This is very significant when the loading of a second port is used for frequency tuning as discussed in (1). A paper is being proposed for publication in Electromagnetics.

(4) Array Design

The work for developing an algorithm for an array module has been started during this period. As of this date, no major results have been obtained. This work will be continued in the next period.

Plans for the Next Period

The work on CP microstrip antennas as described in (2) will be continued. The study on array modules as described in (4) will be more actively pursued.

Travel

Professor Y. T. Lo visited NASA Lewis Research Center on May 3-4, 1984, and presented a progress report and also a seminar on wideband large arrays.

Personnel

During this period the following people have been involved in the work reported above: Y. T. Lo, W. F. Richards, B. Engst, B. F. Wang, and M. L. Oberhart.

MICROSTRIP ANTENNAS FOR DUAL-FREQUENCY OPERATION

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ABSTRACT--Single element microstrip antenna for dual frequency operation have been investigated. By placing shorting pins at appropriate locations in the patch, the ratio of two band frequencies can be varied from 3 to 1.8. In many applications a smaller ratio is desired, and this can be achieved by introducing slots in the patch. In so doing, the ratio can be reduced to less than 1.3. For this type of antenna, a hybrid multiport theory is developed and theoretical results are found to be in excellent agreement with the measured.

I. INTRODUCTION

One of the outstanding features of a thin microstrip antenna is its compactness in structure. Unfortunately it is notoriously narrow-banded unless some degree of compactness can be sacrificed by using a thick substrate. In many applications, it is not operation in a continuous wide-band, but, operation in two or more discrete bands that is required. In this case, a thin patch capable of operating in multiple bands is highly desirable, particularly for large array application where considerable saving in space, weight, material and cost can be achieved. For that goal, a few attempts have been made [1,2], by using two or more patch antennas stacked on top of each other, or placed side by side, or using a complex matching network which takes as much space and weight, if not more, as the element itself. Obviously in all those designs, the advantage of compact structure is sacrificed.

This work is supported in part by RADC/EEAA, Hanscom AFB, MA and NASA Lewis Research Center, Cleveland, OH.

From the cavity-model theory, a single patch antenna can easily be made to resonate at many frequencies associated with various modes. But for most applications, all bands are required to have the same polarization, radiation pattern and input impedance characteristics. It is also desirable to have a single input port and an arbitrary separation of the frequency bands. All of these impose severe constraints on the use of the modes. In this paper we shall describe some methods which can practically achieve all these goals.

An annular patch can have predominantly broadside radiation when excited for the (1,1), (1,2) and even the (1,3) mode. Unlike a circular disc, the frequencies for those modes can be adjusted by choosing the inner and outer radius dimensions. All the aforestated properties can be obtained except that the variation of the two frequency band ratio is somewhat limited [3].

By making use of the difference in the field distributions for various modes, it is possible to practically tune the operating frequencies associated with those modes independent of each other. One method is to place a series of shorting pins at the nodal lines of, for example, the (0,3) modal electrical field of a rectangular patch [3]. These pins will have practically no effect on the (0,3) modal field structure but can have a strong effect on the (0,1) field and thus raise the (0,1) modal frequency. Therefore the low band frequency can be tuned independently. However, the ratio of the two operating frequencies, F_H/F_L, can be varied only from 3 to 2 approximately. On the other hand, if slots are cut in the patch where the magnetic field of the (0,3) mode is maximum, they can have a strong effect on the (0,3) modal field but little on the (0,1) modal field. Thus the operating frequency for the (0,3) mode can be lowered. By using both slots and pins, the two operating bands can be varied over a wide range. In this paper an analytic theory is developed for this type of antenna and then verified by experiment.

II. A MICROSTRIP ANTENNA EXCITED BY A MAGNETIC CURRENT $\underset{\sim}{K}$

First consider a microstrip antenna excited by a magnetic current $\underset{\sim}{K}$ in the slot centered at (x',y') as shown in Figure 1. Following the cavity model theory [4], the antenna can be considered as a cavity bounded by magnetic walls along its perimeter and electric walls at z = 0 and t. Since the substrate thickness t is typically a few hundredths of a wavelength, we can assume that the field excited by the magnetic current

$$\underset{\sim}{K} = \hat{x}[U(x - x' + d_{eff}/2) - U(x - x' - d_{eff}/2)]\,\delta(y - y')\,\delta(z - t)$$

in the slot is approximately the same as that excited by

$$\underset{\sim}{K} = \hat{x}[U(x - x' + d_{eff}/2) - U(x - x' - d_{eff}/2)]\,\delta(y - y')\,[U(z) - U(z - t)]/t$$

where d_{eff} is the effective width of the magnetic current strip of one V/M, and U(•) is the unit step function. The field in the cavity due to $\underset{\sim}{K}$ can then be found by modal-matching as given below:

In region I ($y' \leq y \leq b$)

$$E_{z1} = \frac{d_{eff}}{at}\sum_{m=0}^{\infty}\frac{\sin(\beta_m y')\cos(m\pi x'/a)}{\sin(\beta_m b)}\, j_o\left(\frac{m\pi d_{eff}}{2a}\right)\cos(m\pi x/a)\,\cos[\beta_m(b - y)] \quad ,$$

$$H_{x1} = \frac{jd_{eff}}{at\omega\mu_o}\sum_{m=0}^{\infty}\frac{\beta_m\sin(\beta_m y')\cos(m\pi x'/a)}{\sin(\beta_m b)}\, j_o\left(\frac{m\pi d_{eff}}{2a}\right)\cos(m\pi x/a)\,\sin[\beta_m(b - y)] \ ,$$

$$H_{y1} = \frac{j\pi d_{eff}}{a^2 t\omega\mu_o}\sum_{m=0}^{\infty}\frac{m\sin(\beta_m y')\cos(m\pi x'/a)}{\sin(\beta_m b)}\, j_o\left(\frac{m\pi d_{eff}}{2a}\right)\sin(m\pi x/a)\,\cos[\beta_m(b - y)] \ . \tag{1}$$

In region II ($0 \leq y \leq y'$)

$$E_{z2} = \frac{-d_{eff}}{at}\sum_{m=0}^{\infty}\frac{\sin[\beta_m(b - y')]\cos(m\pi x'/a)}{\sin(\beta_m b)}\, j_o\left(\frac{m\pi d_{eff}}{2a}\right)\cos(m\pi x/a)\,\cos(\beta_m y) \quad ,$$

$$H_{x2} = \frac{jd_{eff}}{at\omega\mu_o} \sum_{m=0}^{\infty} \frac{m\sin[\beta_m(b - y')]\cos(m\pi x'/a)}{\sin(\beta_m b)} j_o\left(\frac{m\pi d_{eff}}{2a}\right)\cos(m\pi x/a)\ \sin(\beta_m y) \quad ,$$

$$H_{y2} = \frac{-j\pi d_{eff}}{a^2 t\omega\mu_o} \sum_{m=0}^{\infty} \frac{m\sin[\beta_m(b - y')]\cos(m\pi x'/a)}{\sin(\beta_m b)} j_o\left(\frac{m\pi d_{eff}}{2a}\right) \sin(m\pi x/a)\ \cos(\beta_m y), \quad (2)$$

where $\beta_m^2 = k^2 - (m\pi/a)^2$, $k^2 = k_o^2\, \varepsilon_r(1 - j\delta_{eff})$, k_o = free space wave number, ε_r = relative dielectric constant of the substrate, δ_{eff} = effective loss tangent [5], μ_o = permeability of free space, $j_o(x) = \sin(x)/x$, and d_{eff} = "effective width" of the magnetic current strip of one V/M. The concept of effective feed width and its implication are discussed in [5]. Examination of Equations (1) and (2) indicates that the resonance occurs when $Re(\beta_m b) \approx n\pi$, $n = 1,2,\ldots$, or $Re(k) \approx [(m\pi/a)^2 + (n\pi/b)^2]^{1/2}$ since $\delta_{eff} << 1$. We shall denote the value β_m for the particular value of n as β_{mn}, and its associated field is called the mnth mode. Clearly in the neighborhood of this resonance the field will be dominated by the term associated with β_{mn}, the value of which depends on the feed location (x',y'). Following the cavity model theory, once the field distribution is found, the Huygen source, $\underset{\sim}{K}(x,y) = \hat{n} \times \hat{z}\, E(x,y)$, along the perimeter can be determined. From $\underset{\sim}{K}$, the far field can then be computed as given below:

$$E_\theta = jk_o(F_x \sin\phi + F_y \cos\phi) \quad ,$$

$$E_\phi = -jk_o(F_x \cos\phi + F_y \sin\phi) \cos\theta \quad , \qquad (3)$$

where

$$F_x = \frac{d_{eff}\, e^{-jk_o r}}{2\pi r} \sum_{m=0}^{\infty} A_m\left\{\sin(\beta_m y')e^{jk_o b\sin\theta\sin\phi} + \sin[\beta_m(b - y')]\right\}$$

$$\cdot \frac{jk_o a \sin\theta\cos\phi}{(m\pi)^2 - (k_o a \sin\theta\cos\phi)^2} \quad , \qquad (4)$$

$$F_y = \frac{bd_{eff}\,e^{-jk_or}}{2\pi ra}\sum_{m=0}^{\infty} A_m\{\sin(\beta_m b)e^{jk_oy'\sin\theta\sin\phi} + \sin(\beta_m y')e^{jk_ob\sin\theta\sin\phi}$$

$$+ \sin[\beta_m(b - y')]\}\ \frac{jk_ob\ \sin\theta\ \sin\phi}{(\beta_m b)^2 - (k_ob\ \sin\theta\ \sin\phi)^2}\ , \tag{5}$$

$$A_m = \frac{\cos(m\pi x'/a)\ j_o(m\pi d_{eff}/2a)}{\sin(\beta_m b)}\ \left[(-1)^m\ e^{jk_oa\sin\theta\cos\phi} - 1\right]\ . \tag{6}$$

Also, from the field in the cavity, the ohmic and dielectric losses as well as the stored energy can be computed and finally the effective loss tangent can be determined.

III. MULTI-PORT ANALYSIS

Let us consider a rectangular microstrip antenna with two ports: port 1 at (x_1,y_1) is fed with an electric current J_1, and port 2 at (x_2,y_2) is fed with a magnetic current K_2 as shown in Fig. 1. The following hybrid matrix [6] can then be used to describe the relationship between the voltage and current at these ports:

$$\begin{bmatrix} V_1 \\ I_2 \end{bmatrix} = \begin{bmatrix} h_{11} & h_{12} \\ h_{21} & h_{22} \end{bmatrix} \begin{bmatrix} I_1 \\ V_2 \end{bmatrix} \tag{7}$$

where $I_1 = d_{1eff}J_1$, d_{1eff} = effective width of source J_1, $V_2 = tK_2$ and the h parameters are given below:

$$h_{11} = -jt\omega\mu_o \sum_{m=0}^{\infty} \frac{\cos^2(m\pi x_1/a)\cos(\beta_m y_1)\cos[\beta_m(b - y_1)]}{a\beta_m \sin(\beta_m b)}\ j_o^2(m\pi d_{1eff}/2a) \tag{8}$$

$$h_{12} = -\frac{d_{2eff}}{a}\sum_{m=0}^{\infty} \frac{\sin[\beta_m(b - y_2)]\cos(\beta_m y_1)}{\sin(\beta_m b)}\cos(m\pi x_1/a)\ \cos\ (m\pi x_2/a)$$

$$\cdot j_o(m\pi d_{1eff}/2a)\ j_o(m\pi d_{2eff}/2a) \tag{9}$$

$$h_{21} = -h_{12} \tag{10}$$

$$h_{22} = \frac{jd_{2eff}^2}{ta\omega\mu_o} \sum_{m=0}^{\infty} \frac{\beta_m \sin(\beta_m y_2)\sin[\beta_m(b - y_2)]}{\sin(\beta_m b)} \cos^2(m\pi x_2/a) j_o^2\left(\frac{m\pi d_{2eff}}{2a}\right) \tag{11}$$

From Equations (8) - (11) all the z-parameters can thus be determined by the relationship between h and z parameters. Then, the input impedance at port 1, Z_{in}, can be computed:

$$Z_{in} = Z_{11} - Z_{12}^2/(Z_{22} + Z_L) \tag{12}$$

where Z_L is the load impedance across the slot terminals at (x_2, y_2). The far field electric vector potential, $\underset{\sim}{F}$, for the two sources can be obtained by superposition as given below:

$$\underset{\sim}{F} = \underset{\sim}{F}_1 + P\underset{\sim}{F}_2 \tag{13}$$

where

$$\underset{\sim}{F}_1 = \frac{jk_o ntbe^{-jk_o r}}{2\pi r} \sum_{m=0}^{\infty} \frac{\varepsilon_{om}\cos(m\pi x_1/a) j_o(m\pi d_{1eff}/2a)}{\beta_m b\sin(\beta_m b)} \left[(-1)^m e^{jk_o a\sin\theta\cos\phi} - 1\right]$$

$$\cdot \left\{\hat{x}\left[\cos(\beta_m y_1)e^{jk_o b\sin\theta\sin\phi} - \cos[\beta_m(b - y_1)]\right] \frac{jk_o a \sin\theta \cos\phi}{(m\pi)^2 - (k_o a \sin\theta \cos\phi)^2}\right.$$

$$- \hat{y}\frac{b}{a}\left[\beta_m b\sin(\beta_m b)e^{jk_o y_1 \sin\theta\sin\phi} + jk_o b\sin\theta\sin\phi\left[\cos(\beta_m y_1)e^{jk_o b\sin\theta\sin\phi}\right.\right.$$

$$\left.\left.\left. - \cos[\beta_m(b-y_1)]\right]\right] \cdot \left[(\beta_m b)^2 - (k_o b \sin\theta \sin\phi)^2\right]^{-1}\right\}, \tag{14}$$

$$\underset{\sim}{E}_2 = \frac{d_{2eff}e^{-jk_or}}{2\pi r} \sum_{m=0}^{\infty} \frac{\cos(m\pi x_2/a)\ j_o(m\pi d_{2eff}/2a)}{\sin(\beta_m b)} \left[(-1)^m e^{jk_o a\sin\theta\cos\phi} - 1\right]$$

$$\cdot\left\{\hat{x}\left[\sin(\beta_m y_2)e^{jk_o b\sin\theta\sin\phi} + \sin[\beta_m(b - y_2)]\right] \frac{jk_o a \sin\theta\cos\phi}{(m\pi)^2 - (k_o a \sin\theta\cos\phi)^2}\right.$$

$$+ \hat{y}\frac{b}{a}\left[\sin(\beta_m b)e^{jk_o y_2\sin\theta\sin\phi} + \sin(\beta_m y_2)e^{jk_o b\sin\theta\sin\phi} + \sin[\beta_m(b - y_2)]\right] \tag{15}$$

$$\left.\frac{j\ k_o b\sin\theta\sin\phi}{(\beta_m b)^2 - (k_o b\sin\theta\sin\phi)^2}\right\} .$$

$$P = j\omega\mu_o \frac{\cos(m\pi x_1/a)\ j_o(m\pi d_{1eff}/2a)}{\beta_m a\sin(\beta_m b)} \cos(\beta_m y_1)\ \cos(m\pi x_2/a)\ \cos[\beta_m(b - y_2)] \tag{16}$$

From these and equation (3), the far field is readily computed. The analysis can be generalized for N slots in a straightforward manner.

A similar theory has been developed for a microstrip antenna with shorting pins [3]. For N pins at N ports, the impedance parameters Z_{ii} and Z_{ij} are given by

$$Z_{ii} = -jk_o t\eta_o \sum_{m=0}^{\infty} \frac{\varepsilon_{om}}{a} \cos^2(m\pi x_i/a) j_o^2\left(\frac{m\pi d_{ieff}}{2a}\right) \frac{\cos(\beta_m y_i)\ \cos[\beta_m(b - y_i)]}{\beta_m \sin(\beta_m b)} \tag{17}$$

$$Z_{ij} = -jk_o t\eta_o \sum_{m=0}^{\infty} \frac{\varepsilon_{om}}{a} \cos(m\pi x_i/a)\cos(m\pi x_j/a)\ j_o^2(m\pi d_{ieff}/2a)$$

$$\cdot\frac{\cos[\beta_m(b - y_j)]\ \cos(\beta_m y_i)}{\beta_m \sin(\beta_m b)} \tag{18}$$

where η_o = 377 ohms, ε_{om} = 1 for m = 0, and 2 otherwise, (x_i,y_i) and (x_j,y_j) are the coordinates of the source J and shorting pin, respectively. For a general case, when the N ports consist of both slots and pins as shown in Fig. 4, the currents and voltages at the N ports can also be written as follow

$$\sum_j I_j Z_{ij} = V_i \quad , \quad i,j = 1,\ldots,N \quad . \tag{19}$$

since the solutions to E and H everywhere in the patch for any J and K have been obtained, one can therefore compute the input impedance Z_{in} at any port, using the same method as discussed above.

IV. THEORETICAL AND EXPERIMENTAL RESULTS

Our approach to the dual-frequency microstrip antenna, as stated earlier, is based on the theoretical argument that shorting pins and slots if placed at appropriate locations in the patch can raise the (0,1) and lower the (0,3) operating frequencies, respectively. In general, with pins and slots, the modal field is no longer pure. The existance of a substantial amount of higher order modes will modify the antenna overall resonant frequency which, as defined in [3], occurs when the reflection coefficient $|\Gamma|$ reaches a minimum.

Several antennas have been constructed and tested to determine the validity of the theory. All of them were made of double copper-clad laminate Rexolite 2200, 1/16" thick. The relative permittivity $\varepsilon_r \approx 2.62$, the loss tangent $\delta \approx 0.001$, and the copper cladding conductivity ≈ 270 KMho. These values were used for theoretical computations.

One of the rectangular microstrip antennas, having the dimensions $a = 19.4$ cm and $b = 14.6$ cm, is fed with a miniature cable at $x_1 = 9.7$ cm and $y_1 = 0$ as shown in Figure 1. A slot of length $\ell = 3.0$ cm and width $w = 0.15$ cm is cut at $x_2 = 9.7$ cm and $y_2 = 7.3$ cm on the patch. The feed location was chosen for a good match to the 50 Ω line for both F_H and F_L bands. The calculated and measured input impedance loci for both bands are shown in Figure 2a and 2b, where for comparison the corresponding loci without slot are also shown by the dashed curves. The calculated and measured radiation patterns are shown in Figure 2c. Similar results for slot length $\ell = 4.5$ cm are shown in Figure 3a and 3b. It is seen that the agreement between theoretical and measured results is excellent for both bands and that the slot has only a minor effect on the low band impedance locus, but a significant effect on the high band impedance locus as expected.

To further reduce the ratio of the operating frequencies of the high and low band, F_H/F_L, in addition to the slots, shorting pins can be inserted along the nodal lines of the (0,3) mode electric field as illustrated in Figure 4. Due to limited space here, only a few typical measured impedance loci and radiation patterns for both bands are shown in Figures 3, 5 and 6. From Figures 3, 5 and 6, it is seen that while the "resonant" frequencies are changed for both bands with pins and slots, in general, the radiation patterns for both bands remain primarily the same. It may also be noted that the input impedance can vary widely with the feed position and one is therefore free to choose the feed position for a desired impedance without undue concern about its effect on the pattern. The measured gains of these microstrip antennas as compared with those of a $\lambda/2$-tuned dipoles, 0.2λ over a ground plane, are -0.5 to -1db for the low band and -1.5 ~ 2db for the high band.

Table 1 summarizes the values of F_H/F_L for six cases. From these results, it is seen that in geneal the slots can lower F_H and shorting pins raise F_L, resulting in a variation of F_H/F_L from 3.02 to 1.31. In fact, this ratio can be reduced even further by adding more pins and slots. However the effectiveness of adding more pins and slots will eventually dminish. Instead, we find that the ratio F_H/F_L can be reduced to about 1.07 by using a C-shaped slot (or a wrapped around microstrip line). This will be reported elsewhere.

TABLE 1

THE OPERATING FREQUENCIES FOR BOTH F_L AND F_H

CASE	F_L(MHz)	F_H(MHz)	F_H/F_L
A. One slot ℓ_1 = 1.0 cm at (9.7,7.3)	628	1900	3.02
B. One slot ℓ_1 = 3.0 cm at (9.7,7.3)	596	1700	2.85
C. Three slots ℓ_1 = 7.0 cm ℓ_2 = ℓ_3 = 3.0 cm at (9.7,2.4), (9.7,7.3) and (9.7,12.2)	555	1420	2.55
D. Three slots ℓ_1 = ℓ_2 = ℓ_3 = 7.0 cm at the same location as in case C.	553	1310	2.36
E. Same as case D but with four pins as shown in Figure 4.	698	1087	1.56
F. Same as case E with six additional pins at (3.7,2.4), (9.7,2.4), (15.7,2.4), (3.7,12.2). (9.7,12.2) and (15.7,12.2)	890	1181	1.31

V. CONCLUSION

This investigation shows that a single rectangular microstrip antenna element can be designed to perform for dual frequency bands corresponding approximately to the (0,1) and (0,3) modes. The frequencies of both bands can be tuned over a wide range, with their ratio from 3 to less than 1.3, by adding shorting pins and slots in the patch. A method for analyzing these antennas has been developed and treats the antenna as a multi-port cavity. The validity of this theory is verified by comparing the computed impedance loci and radiation patterns with the measured for a few simple cases.

As a design guide, in general, the effect of slot on the high-band frequency is stronger if it is placed where the high-order modal magnetic field is stronger, and the effect of short pin on the low-band frequency is stronger if it is placed where the low-order modal electric field is stronger.

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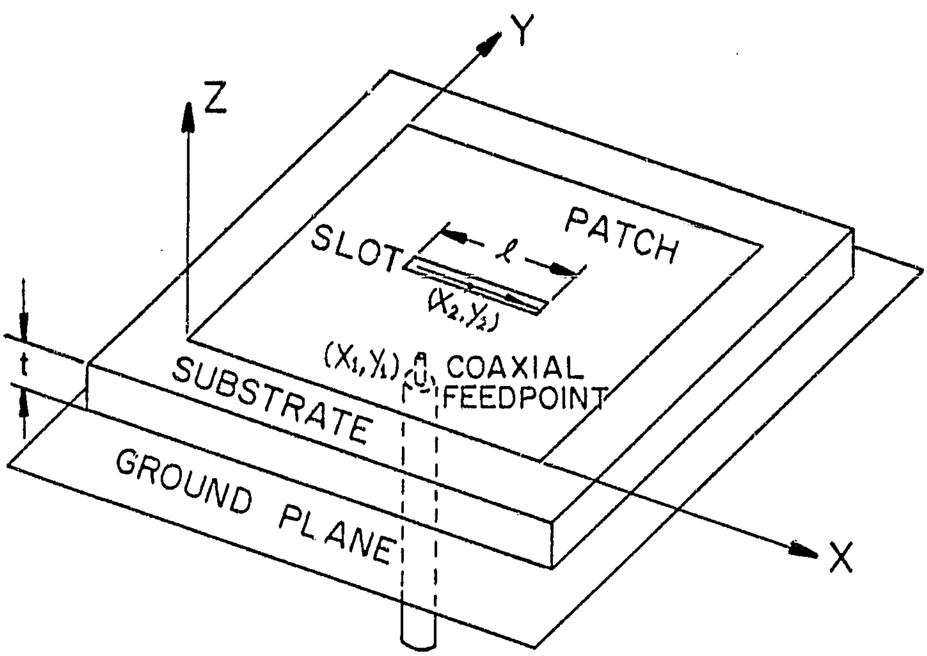
LIST OF FIGURES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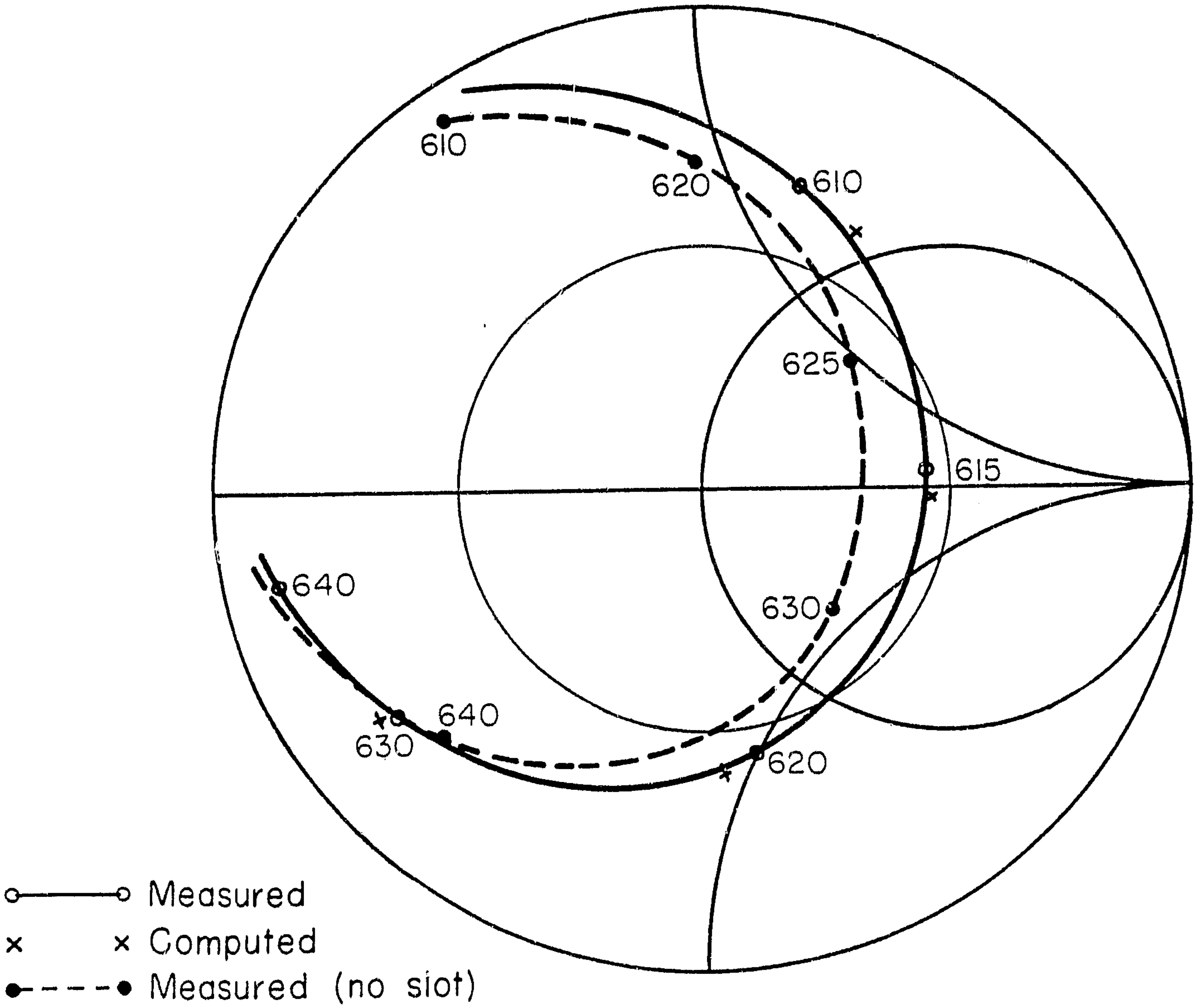
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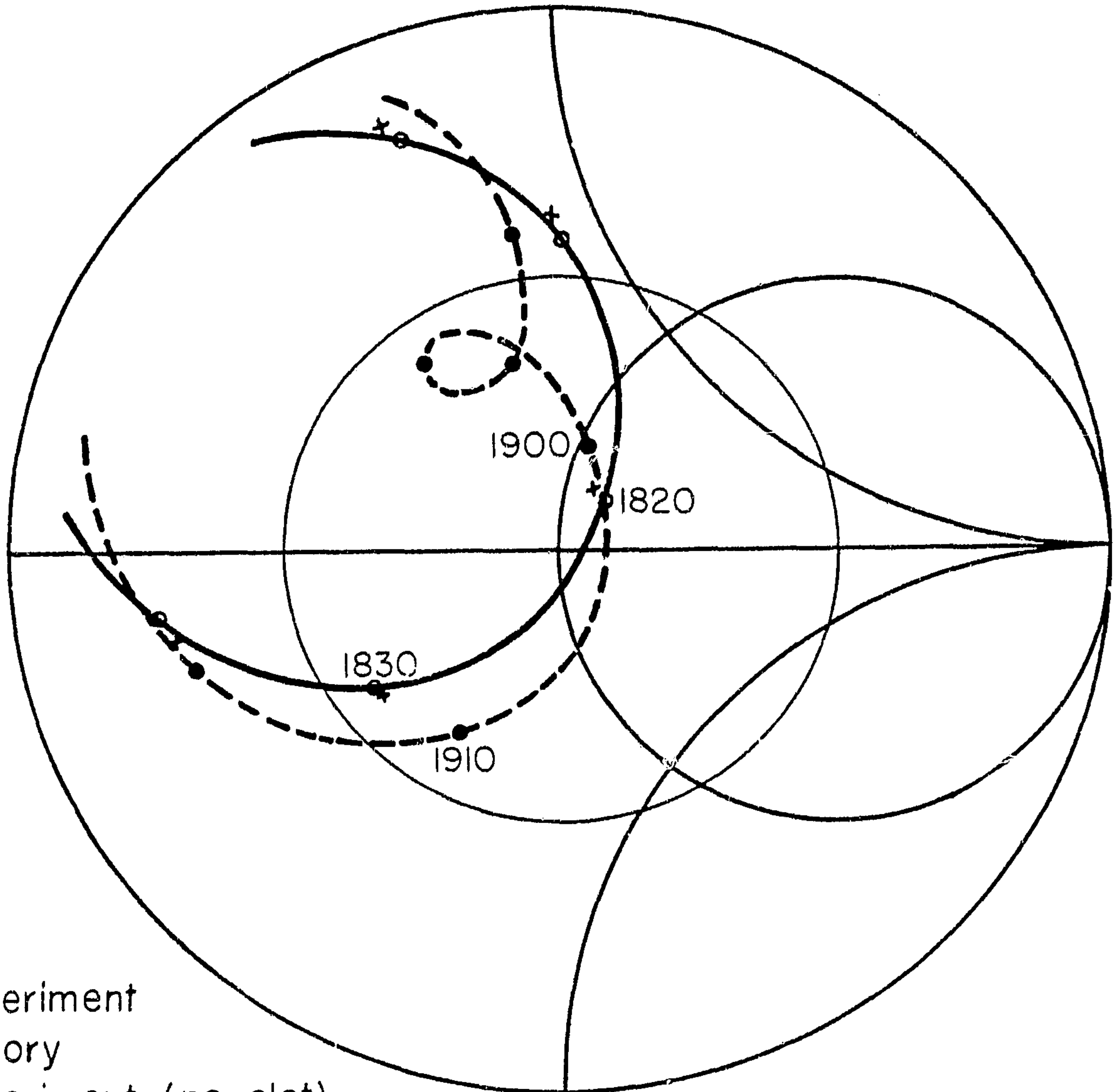
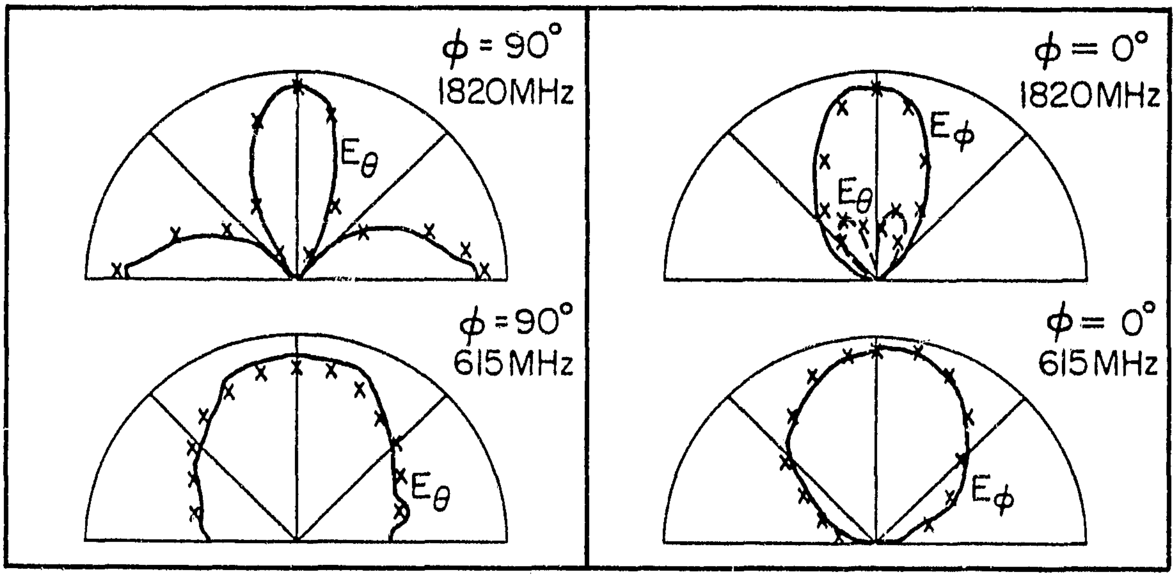
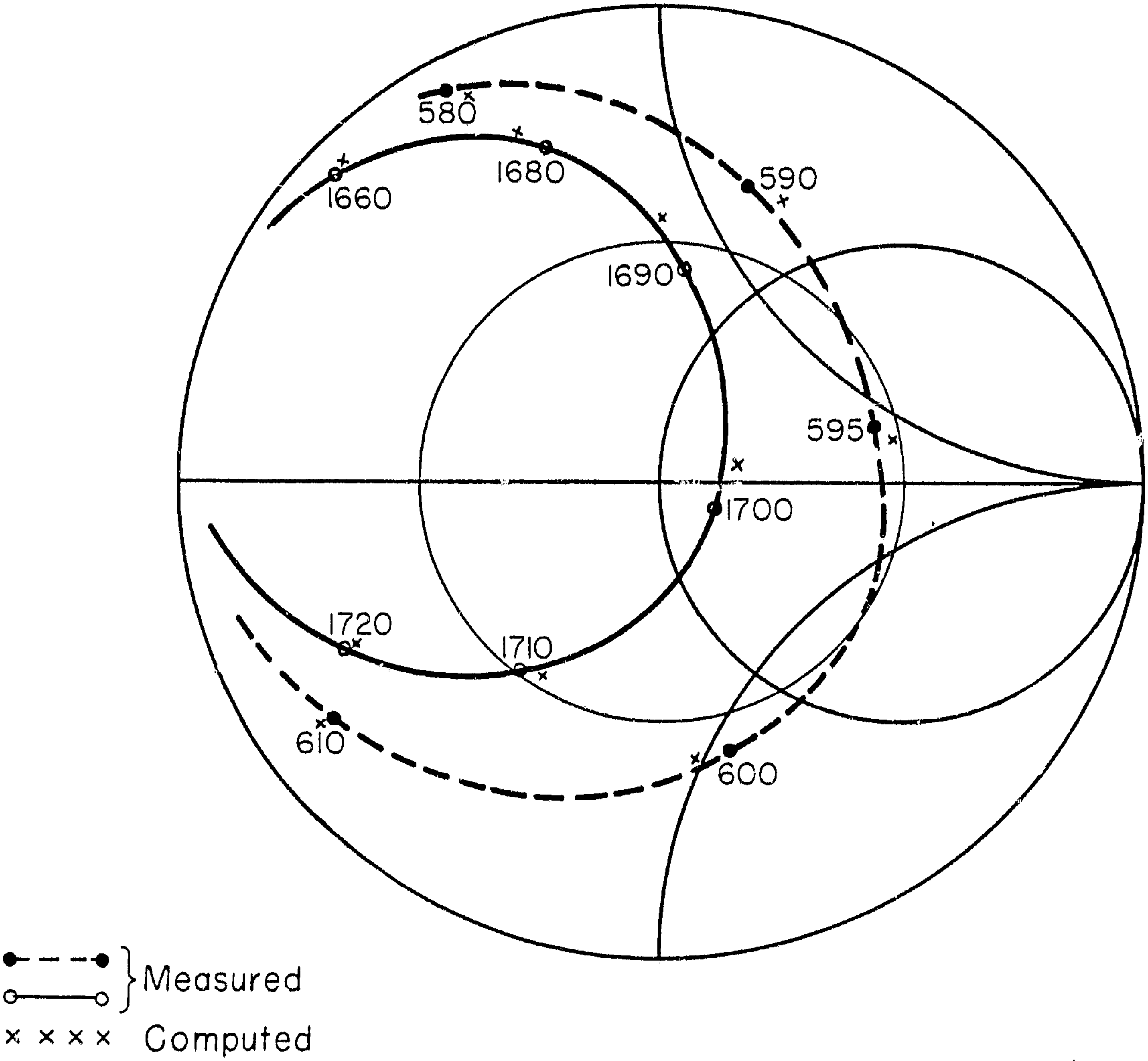
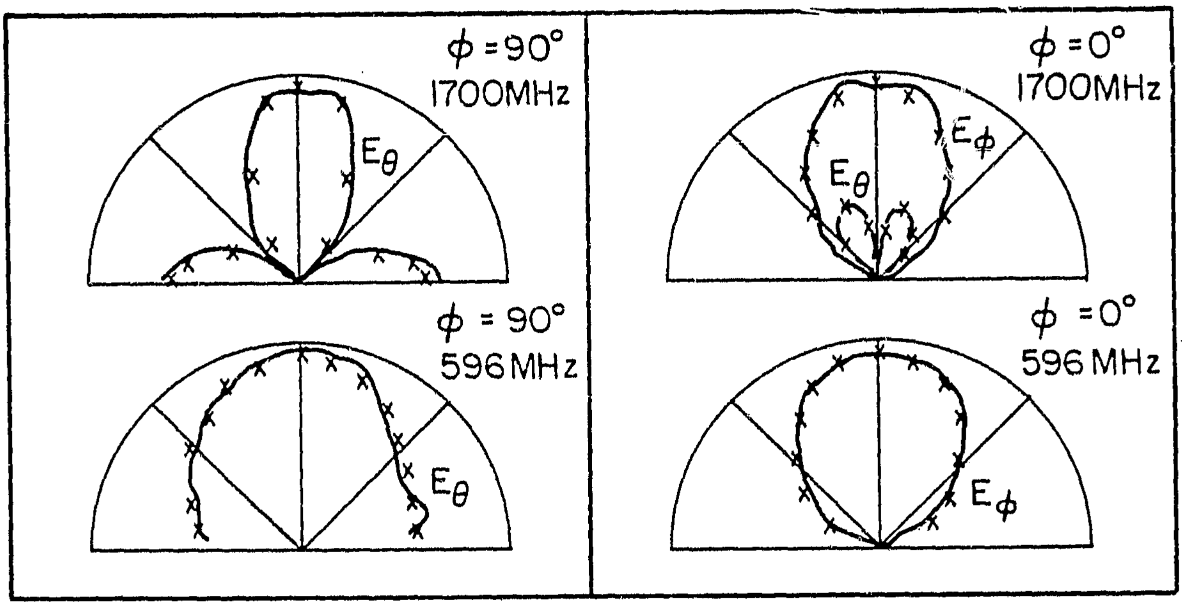


o———o Experiment
× × Theory
●- - - -● Experiment (no slot)
Increment 10 MHz
(Increasing frequency is clockwise)



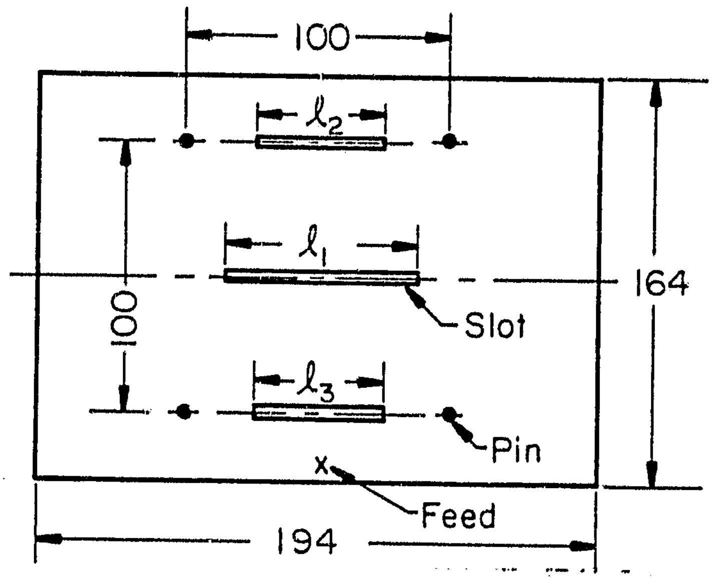
MEASURED xxxx COMPUTED

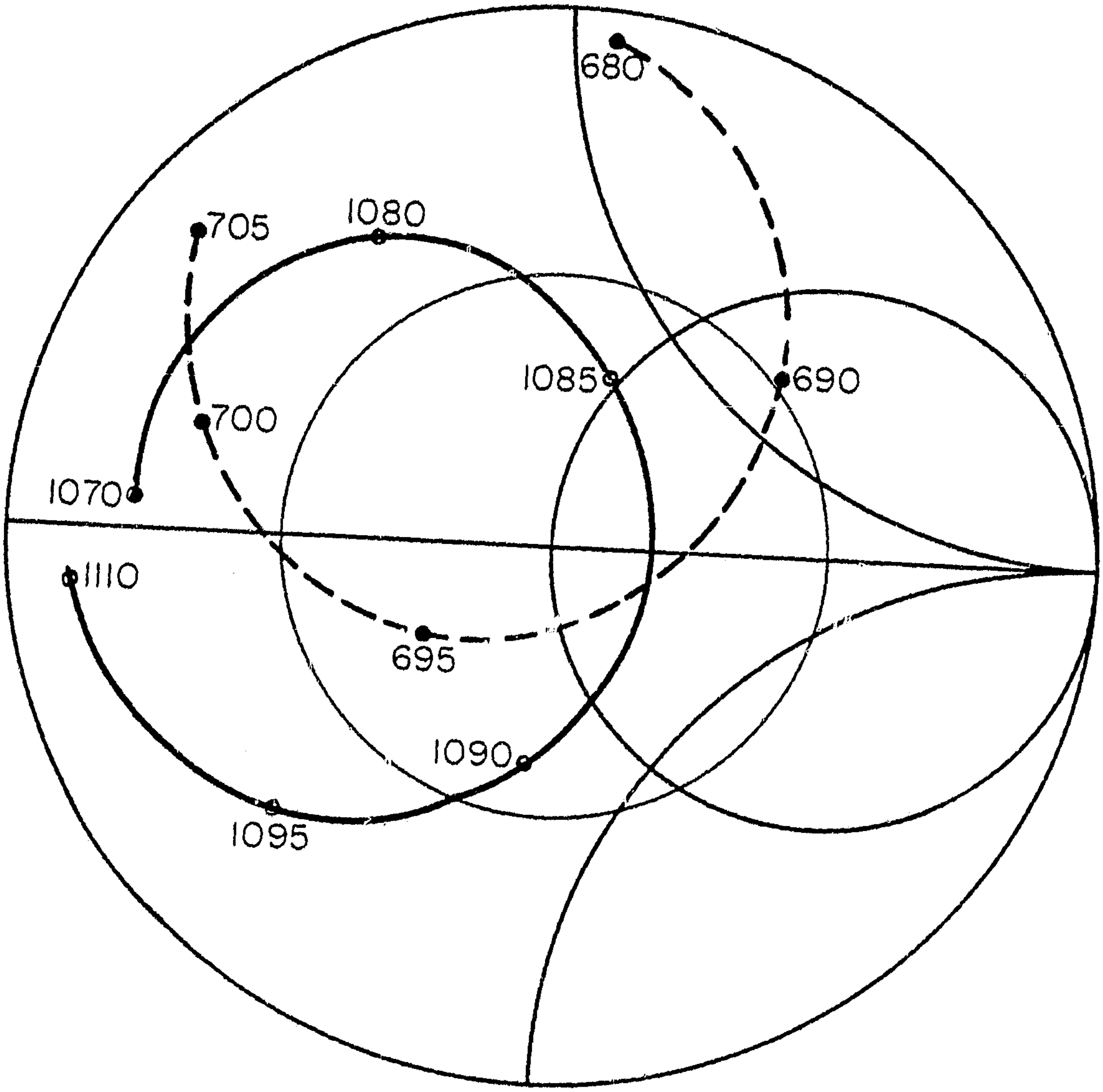

580
590
595
600
610
1660
1680
1690
1700
1710
1720
Measured
Computed

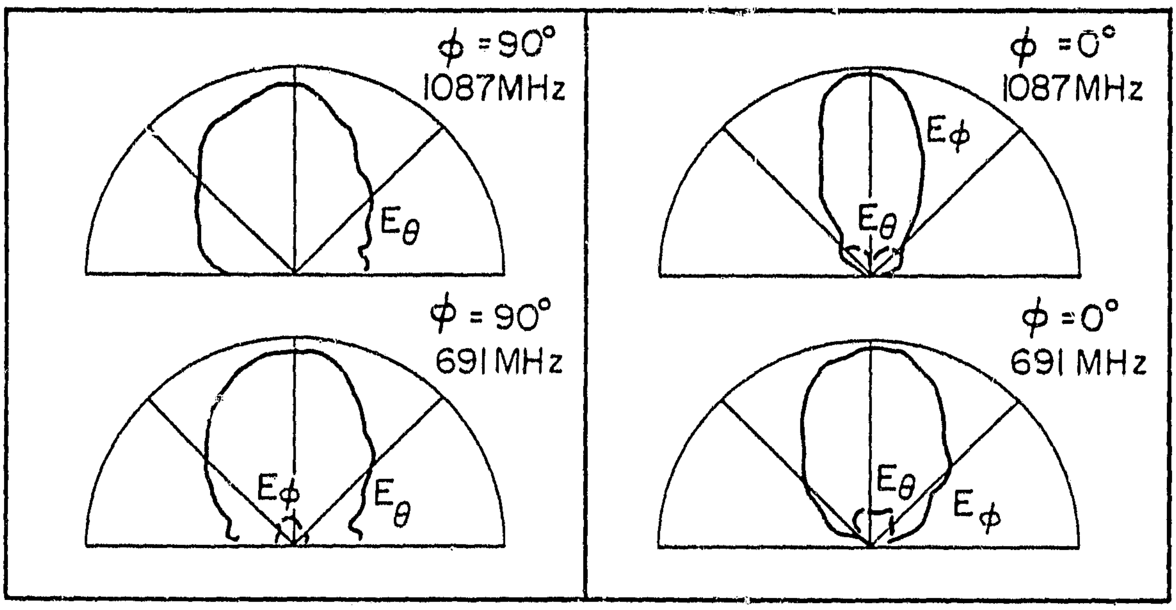


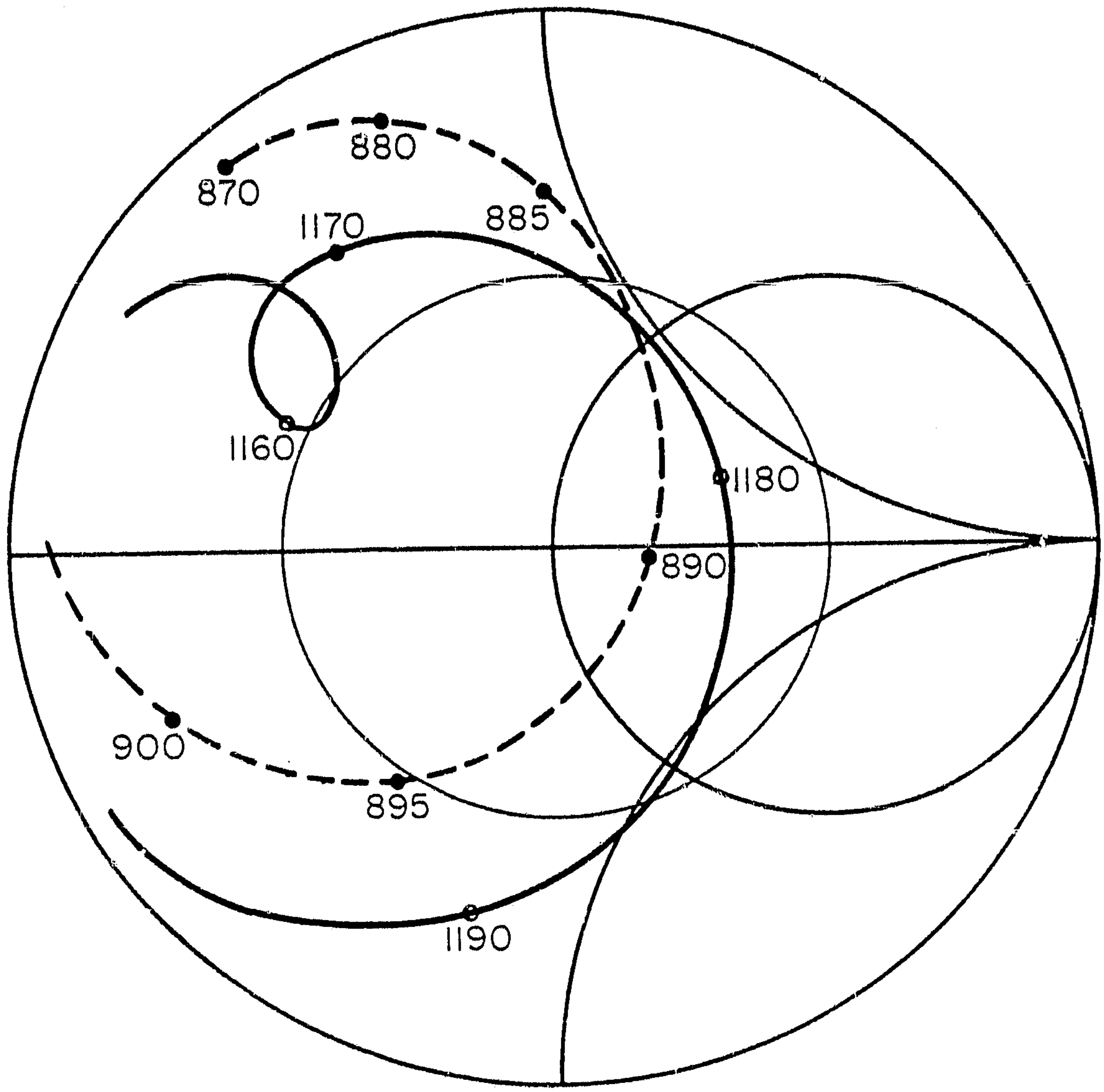
--- MEASURED   x x x x COMPUTED

ORIGINAL PAGE IS
OF POOR QUALITY

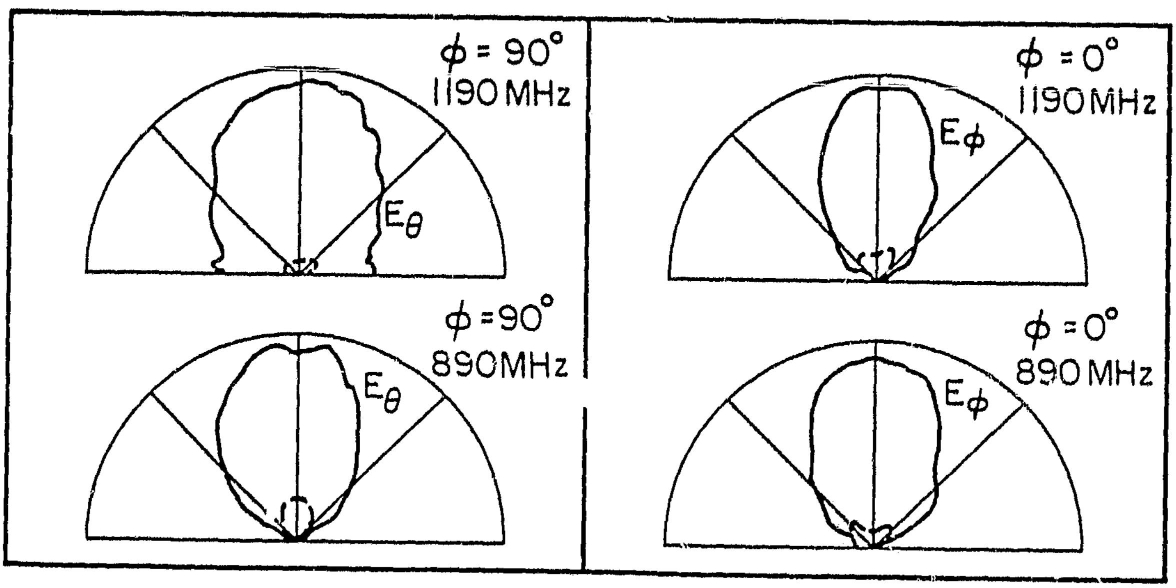



680
705
1080
1085
690
700
1070
1110
695
1090
1095


φ = 90°
1087MHz
Eθ
φ = 90°
691 MHz
Eφ
Eθ
φ = 0°
1087MHz
Eφ
Eθ
φ = 0°
691 MHz
Eθ
Eφ



870
880
885
1170
1160
1180
890
900
895
1190


φ = 90°
1190 MHz
Eθ
φ = 0°
1190 MHz
Eφ
φ = 90°
890 MHz
Eθ
φ = 0°
890 MHz
Eφ