AD-A112 607 CALIFORNIA UNIV LOS ANGELES INTEGRATED ELECTROMAGNET--ETC F/G 9/5
ON THE EFFECT OF SUBSTRATE THICKNESS AND PERMITTIVITY ON PRINTE--ETC(U)
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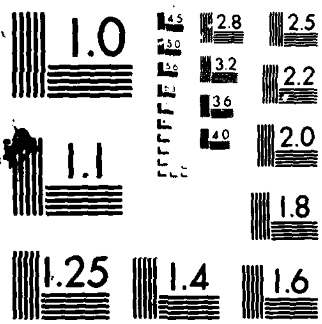

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"ON THE EFFECT OF SUBSTRATE THICKNESS AND PERMITTIVITY ON PRINTED CIRCUIT DIPOLE PROPERTIES"

by

P.B. Katehi and
N.G. Alexopoulos

Sponsored by:

U.S. Army Contract DAAG 29-79-C-0050
U.S. Navy Contract N00014-79-C-0856

SECURITY CLASSIFICATION OF THIS PAGE (When Data Entered)

REPORT DOCUMENTATION PAGE

READ INSTRUCTIONS BEFORE COMPLETING FORM

1. REPORT NUMBER

2. GOVT ACCESSION NO. AD-A112 607

3. RECIPIENT'S CATALOG NUMBER

4. TITLE (and Subtitle)
On the Effect of Substrate Thickness and Permittivity on Printed Circuit Dipole Properties.

5. TYPE OF REPORT & PERIOD COVERED
Technical Laboratory Report

6. PERFORMING ORG. REPORT NUMBER

7. AUTHOR(s)
P. B. Katehi and N. G. Alexopoulos

8. CONTRACT OR GRANT NUMBER(s)
DAAG 29-79-C-0050
N0014-79-C-0856

9. PERFORMING ORGANIZATION NAME AND ADDRESS
Electrical Engineering Department
UCLA
Los Angeles, CA 90024

10. PROGRAM ELEMENT, PROJECT, TASK AREA & WORK UNIT NUMBERS

11. CONTROLLING OFFICE NAME AND ADDRESS
Dr. James Mink
ARO - P 4800-C-81

12. REPORT DATE
December 1981

13. NUMBER OF PAGES
31

14. MONITORING AGENCY NAME & ADDRESS (if different from Controlling Office)
Army Research Office
Research Triangle Park
North Caroline

15. SECURITY CLASS. (of this report)
Unclassified

15a. DECLASSIFICATION/DOWNGRADING SCHEDULE

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19. KEY WORDS (Continue on reverse side if necessary and identify by block number)
Substrate Permittivity
Printed Circuit Dipole

epsilon sub r

20. ABSTRACT (Continue on reverse side if necessary and identify by block number)
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S/N 0102-LF-014-6601

SECURITY CLASSIFICATION OF THIS PAGE (When Data Entered)

Classification 20. (Continued) epsilon sub r

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On the Effect of Substrate Thickness and Permittivity on Printed Circuit Dipole Properties.

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This work was supported by the U.S. Army under Contract DAAG29-79-C-0050 and by the U.S. Navy under Contract N00014-79-C-0856.

Abstract

The effect of substrate thickness and relative permittivity on the radiation properties of printed circuit dipoles (PCD's) is investigated. A trade-off between substrate thickness and resonant input resistance, bandwidth and radiation efficiency is presented for a PTFE glass random fiber substrate. If is found that for a fixed substrate thickness B, the resonant length and directivity decrease with increasing relative dielectric constant ε_r. The $\bar{E}$-plane normalized power pattern is also examined as a function of ε_r and B. It is shown that even for very thin substrates, multiple beam radiation can result for certain values of ε_r by the excitation of surface waves. Multiple beam patterns can also be obtained with increasing B for a given ε_r. In fact, as B increases it is determined that the resonant length, bandwidth and resonant resistance approach the apparent value of a PCD on a dielectric half space.

Introduction.

Printed circuit dipoles (PCDs) have been successfully employed in conformal arrays [1] - [3]. Recently, Elliott and Stern [4] and Stern and Elliott [5] demonstrated a design procedure for the construction of small arrays consisting of electromagnetically coupled microstrip dipoles. Implicit in their approach is the accurate knowledge of mutual coupling and array element optimization. In the aforementioned applications the dipoles were printed on very thin substrates (compared to free space wavelength). In the millimeter and submillimeter frequency range PCDs have been utilized in detector and imaging arrays with several wavelength thick substrates (see Rutledge et al. [6] - [8]). The performance characteristics of PCDs depend in a very crucial way on the substrate thickness B and substrate relative dielectric constant ϵ_r. This is due to the fact that ϵ_r and B control the amount of energy coupled into surface wave modes guided by the substrate [9] - [12]. Consequently, the radiation efficiency, (i.e., the ratio of energy radiated by the antenna into free space to that coupled in the substrate), input impedance, radiation pattern and bandwidth need to be investigated as functions of ϵ_r and B.

In this paper, a trade-off study of ϵ_r and B on the PCD properties is carried out. As a specific example of substrate thickness effect on antenna properties a PTFE glass random fiber substrate material is considered with $\epsilon_r = 2.35$. The results presented herein are based on refined analytical and numerical extentions of the information on printed dipoles contained in [9] - [11].

II. Analytical Formulation.

A precise understanding of the effect of substrate thickness and relative permittivity on the characteristics of printed circuit dipoles requires the development of integral equation solutions for the antenna current distribution. The integral equation for the PCD is given in its general form by (see figure 1)

$$\overline{E}(x,y,z) = \int_0^L \left[k_0^2 \overline{\overline{I}} + \nabla\nabla\right] \cdot \overline{\overline{G}}(\overline{r}/\overline{r}') \cdot \overline{J}(\overline{r}')dr' \qquad (1)$$

where L is the length of the dipole, $\overline{\overline{I}}$ is the unit dyadic $\overline{\overline{I}} = \hat{x}\hat{x} + \hat{y}\hat{y} + \hat{z}\hat{z}$, $\overline{J}(\overline{r}')$ is the unknown current density distribution and $\overline{\overline{G}}(\overline{r}/\overline{r}')$ is the dyadic Green's function [9] - [10] for the problem of interest. For the case of a PCD oriented along the $\hat{x}$-direction and with h = 0 [12] equation (1) simplifies to

$$E_x = \int_0^L \left[k_0^2\, G_x + \frac{\partial^2}{\partial x^2}(G_x - G)\right] J_x dx' \qquad (2)$$

where [10], [12]

$$G_x = -\frac{j\omega\mu_o}{2\pi k_o^2}(1-\epsilon_r)\int_0^\infty J_o(\lambda\rho)\left(\frac{\sinh(uB)}{f_1(\lambda,B)}\right)\left(\frac{\sinh(uB)}{f_2(\lambda,B)}\right)u\lambda d\lambda \qquad (3)$$

$$G = -\frac{j\omega\mu_o}{2\pi k_o^2}(\epsilon_r-1)\int_0^\infty J_o(\lambda\rho)\left(\frac{\sinh(uB)}{f_1(\lambda,B)}\right)\left(\frac{\cosh(uB)}{f_2(\lambda,B)}\right)u_o\lambda d\lambda \qquad (4)$$

$$f_1(\lambda,B) = u_o \sinh(uB) + u\cosh(uB) \qquad (5)$$

$$f_2(\lambda,B) = \epsilon_r u_o \cosh(uB) + u\sinh(uB) \qquad (6)$$

$$\rho = \left[(x-x')^2 + (y-y')^2\right]^{1/2} \qquad (7)$$

and

$$u_o = \left[\lambda^2 - k_o^2\right]^{1/2} \quad , \quad u = \left[\lambda^2 - k^2\right]^{1/2} \qquad (8)$$

The zeros of $f_1(\lambda,B)$, $f_2(\lambda,B)$ in the denominator of equations (4), (5) for $k_o < \lambda < k$ correspond to TM, TE surface wave modes [9] - [12] respectively with a TM mode having zero cut-off. The number of excited surface wave modes depends on the substrate thickness B and relative dielectric constant ε_r. By employing the method of moments [13] with a basis set of overlapping piecewise-simusoidal currents the integral equation as given by equation (2) can be discretized into a matrix form [V] = [Z] [I] where the elements of [Z] can be written as

$$Z_{ij} = \frac{1}{\left[\sin k_o l_x\right]^2} \cdot \left\{ \sum_{\nu=-1,0} \sum_{\nu'=-1,0} k_o^2 \int_{x_{i+\nu}}^{x_{i+\nu+1}} dx \int_{x_{j+\nu'}}^{x_{j+\nu'+1}} dx' \cdot \right.$$

$$\sin\left[k_o \mid l_x \nu + (x_{i+\nu+1} - x) \mid\right] \sin\left[k_o \mid l_x \nu' + (x_{j+\nu+1} - x') \mid\right] \cdot$$

$$G\left(\sqrt{(x-x')^2 + (y-y')^2}\right) + \sum_{\nu=-1,0} \sum_{\nu'=-1,0,1} k_o \left[1 - (2\cos k_o l_x + 1)\delta(\nu')\right]$$

$$\cdot \int_{x_{i+\nu}}^{x_{i+\nu+1}} \sin\left[k_o \mid l_x \nu + (x_{k+\nu+1} - x) \mid\right] \cdot \left[G_x \left(\sqrt{(x-x')^2 + (y-y')^2}\right) - \right.$$

$$\left.\left. - G\left(\sqrt{(x-x_{j+\nu'})^2 + (y-y')^2}\right)\right]\right\} \qquad [9]$$

In equation (9) $l_x = L/N$ where N is the total number of subsections the dipole is divided in. By inversion the current distributions can be obtained for arbitrary parameters ε_r, B and L for center-fed thin (radius = $10^{-4}\lambda_o$) dipoles.

Computation of the matrix elements Z_{ij} involves integration of Sommerfeld type of integrals along the real λ-axis [9], [10] for which efficient routines have been written [12].

III. Effect of Substrate Permittivity Variation

The relative permittivity is varied for a fixed substrate thickness of $B = 0.1016\lambda_0$. For values of ε_r up to $\varepsilon_r = 45$ and for the chosen B there are three surface wave modes that can be excited. The antenna resonant length L_r is shown in figure 2 as a function of ε_r. As more energy is coupled into guided waves inside the substrate the resonant length decreases with a cusp discontinuity exhibited at every value of ε_r where the onset of the next surface wave mode occurs. Figure 2 implies that the radiation efficiency of the PCD decreases with increasing ε_r. The dependence of the input impedance Z_{in} on ε_r is demonstrated in figures 3 - 5, where Z_{in} is plotted vs. antenna length L for $\varepsilon_r = 2, 10, 35$ respectively and for $B = 0.1016\lambda_0$. As ε_r increases the following effects can be observed: (a) The value of the total input resistance R decreases. (b) The reactance becomes increasingly capacitive with a reduced number of resonances.

IV. Substrate Thickness Variation.

A PTFE glass random fiber substrate is selected with $\varepsilon_r = 2.35$ to analyze the effect of increasing B. The variation of PCD resonant length vs. B is demonstrated in figure 6 wherefrom it is concluded that as $B \rightarrow \infty$, $L_r \rightarrow 0.375\lambda_0$ which is anticipated to be the resonant length of a wire dipole of radius

$a = 10^{-4}\lambda_0$ on a dielectric half-space with $\epsilon_r = 2.35$. Figure 7 demonstrates the dependence of PCD resonant resistance on B. It is worth mentioning that for $B = 1.2\lambda_0$ there exist 6 excited modes in the substrate. As $B \rightarrow \infty$, $R_r \rightarrow 50$ ohms. Figure 8 shows the radiation efficiency $\eta = \frac{R_{rr}}{R_{rs}}$ where $R_r = R_{rr} + R_{rs}$, R_{rr} being the radiation resonant resistance and R_{rs} the surface wave resonant resistance respectively. The bandwidth of the PCD is exhibited in figure 9 vs. B. The bandwidth has been defined here as

$$BW = \frac{1}{L_r}\left[\frac{2R_r}{\left.\frac{dX}{d(L/\lambda)}\right|_{L_r}}\right] \qquad [10]$$

These curves provide a trade-off analysis for a PCD as e.g. if we wish to choose a maximum radiation efficiency dipole with $\eta \equiv 4$ at $B = 0.2\lambda_0$, then the resonant length is $L_r = 0.369345\lambda_0$, $R_r = 90$ ohms and BW ≈ 18%. The corresponding $\bar{E}$-plane normalized radiation pattern is shown in figure 10, where it is observed that the half-power beamwidth is approximately 70 degrees. On the other hand if an input $R_r = 50$ ohms is desired then $B \equiv 0.12\lambda_0$, BW = 8% and $\eta \equiv 2.8$.

V. Radiation Pattern

It is of interest to investigate now the dependence of the far-field pattern on substrate properties. The far-zone electromagnetic field can be obtained by a saddle-point method and the electric field components are given by

$$E_\theta = 2k_\theta^2 \Pi_\theta \tag{11}$$

and

$$E_\phi = 2k_o^2 \Pi_\phi \tag{12}$$

where

$$\Pi_\theta = -j100 \frac{e^{-jk_oR}}{R} \cos\phi \cdot \left(\frac{\cos[k_o \ell_x \sin\theta\cos\phi] - \cos(k_o\ell_x)}{k_o \sin(k_o \ell_x [1 - \sin^2\theta\cos^2\phi])} \right)$$

$$\cdot \left[\cos\theta\ \phi(\varepsilon_r, B, \theta) + (\varepsilon_r - 1)\sin\theta\Psi(\varepsilon_r, B, \theta) \right] \sum_{n=1}^{N} I_n e^{jnk_o\ell_x \sin\theta\cos\phi}, \tag{13}$$

$$\Pi_\phi = j100 \frac{e^{-jk_oR}}{R} \sin\phi \cdot \left(\frac{\cos[k_o \ell_x \sin\theta\cos\phi] - \cos(k_o\ell_x)}{k_o \sin(k_o \ell_x [1 - \sin^2\theta\cos^2\phi])} \right) \cdot$$

$$\phi(\varepsilon_r, B, \theta) \sum_{n=1}^{N} I_n e^{jnk_o\ell_x \sin\theta\cos\phi}, \tag{14}$$

$$\phi(\varepsilon_r, B, \theta) = \frac{1}{k_o} \left\{ \frac{1}{1 - j \frac{\sqrt{\varepsilon_r - \sin\theta}}{\cos\theta} \cot(k_o \sqrt{\varepsilon_r - \sin^2\theta})} \right\} \tag{15}$$

and

$$\Psi(\varepsilon_r,B,\theta) = \frac{\tan\theta}{k_0}\left(\frac{1}{1 - j\,\frac{\sqrt{\varepsilon_r - \sin^2\theta}}{\cos\theta}\cot(k_0\sqrt{\varepsilon_r - \sin^2\theta}\;B)}\right)$$

$$\cdot\left(\frac{1}{\varepsilon_r + j\,\frac{\sqrt{\varepsilon_r - \sin^2\theta}}{\cos\theta}\tan(k_0\sqrt{\varepsilon_r - \sin^2\theta}\;B)}\right) \tag{16}$$

or

$$\Psi(\varepsilon_r,B,\theta) = \tan\theta\,\Phi(\varepsilon_r,B,\theta)\;\Lambda(\varepsilon_r,B,\theta) \tag{17}$$

with

$$\Lambda(\varepsilon_r,B,\theta) = \frac{1}{\varepsilon_r + j\,\frac{\sqrt{\varepsilon_r - \sin^2\theta}}{\cos\theta}\tan(k_0\sqrt{\varepsilon_r - \sin^2\theta}\;B)} \tag{18}$$

An investigation of the expressions given for the far-zone electric field indicates that the effect of the substrate properties on the radiation pattern is controlled by the factors $\Phi(\varepsilon_r,B,\theta)$ and $\Lambda(\varepsilon_r,B,\theta)$. Furthermore, it is verified that $\Phi(\varepsilon_r,B,\theta)$ is a result of the substrate guided TM modes, while $\Lambda(\varepsilon_r,B,\theta)$ is due to the TE modes. A thorough analysis of Φ and Λ indicates that Φ determines the position of nulls, principal as well as secondary maxima of the pattern for $\theta < \pi/2$. On the other hand the factor Λ affects the sidelobe maxima only and in general it smooths out the pattern. In summary, the following results can be stated for the dependence of the radiation pattern on ε_r and B.

A. Fixed Substrate Thickness.

For fixed substrate thickness B and varying relative dielectric constant ϵ_r we find:

1. If $1 + (\frac{n}{2})^2 (\frac{\lambda_o}{B})^2 \le \epsilon_r < (\frac{n+1}{2})^2 (\frac{\lambda_o}{B})^2$, integer n (19)

the radiation pattern consists of one lobe only.

2. If $(\frac{n}{2})^2 (\frac{\lambda_o}{B})^2 \le \epsilon_r < 1 + (\frac{n}{2})^2 (\frac{\lambda_o}{B})^2$, integer n (20)

there may exist more than one beam. Specifically if

$$[\![2 \sqrt{\epsilon_r} \frac{B}{\lambda_o}]\!] = [\![2 \sqrt{\epsilon_r - 1} \frac{B}{\lambda_o}]\!] + N \tag{21}$$

and

$$2 \sqrt{\epsilon_r} \frac{B}{\lambda_o} = [\![2 \sqrt{\epsilon_r - 1} \frac{B}{\lambda_o}]\!] + N + a \tag{22}$$

where $[\![\quad]\!]$ indicates integer value of, N is an integer and a an arbitrary real positive number then

i) If N = 0 (a > 0), there exists a single lobe only.

ii) If N > 0, a > 0, there exist 2N + 1 lobes with one of the maxima always at $\theta = 0$.

iii) If N > 0, a = 0, there exist 2N lobes, with a null at $\theta = 0$.

B. Fixed Substrate Permittivity.

In this case the following conclusions can be drawn:

1. If $(\frac{n}{2}) \frac{1}{\sqrt{\epsilon_r - 1}} \le \frac{B}{\lambda_o} < (\frac{n+1}{2}) \frac{1}{\sqrt{\epsilon_r}}$ (23)

with n an integerthere is always one lobe at most.

2. If $(\frac{n}{2})\frac{1}{\sqrt{\epsilon_r}} < \frac{B}{\lambda_o} < \frac{n}{2}\frac{1}{\sqrt{\epsilon_r - 1}}$, (24)

n an integer, there may exist more than one lobes.

Again relations (21) and (22) hold. The pattern nulls as determined by $\Phi(\epsilon_r,B,\theta)$ occur at

$$\theta_n = \pm \sin^{-1}\left[\epsilon_r - (\frac{n\lambda_o}{2B})^2\right]^{1/2} \qquad (25)$$

n an integer.

The dependence of the $\bar{E}$-plane normalized radiation power pattern on ϵ_r and B is now investigated. We consider again the PTFE substrate with $\epsilon_r = 2.35$. For the optimum radiation efficiency of figure 8 the substrate thickness is $B = 0.2\lambda_o$. In this case equation (22) is satisfied for N = 0 and a = 0.613 i.e. the radiation pattern consists of a single lobe as shown in figure 10. In light of equations (21) and (22) it can be verified that for $B = 0.2\lambda_o$, $0.975\lambda_o$ and $1.05\lambda_o$ the $\bar{E}$-plane normalized power pattern will have respectively one, two and three lobes as shown in figures 11 - 13.

If the substrate thickness B is fixed, e.g. at $B = 0.1016\lambda_o$ then the $\bar{E}$-plane normalized power pattern is shown in figures 14 - 16 for $\epsilon_r = 2,10,35$ respectively. With increasing ϵ_r we observe that the PCD directivity is decreased because more energy is radiated close to the $\theta = \pi/2$ direction along the length of the antenna as the number of modes guided in ths substrate increases. It is further observed in figure 17, that when $\epsilon_r = 25$ and for $B = 0.1016\lambda_o$ there exist three lobes and according to equation (25) the nulls are at $\theta = \pm 62.111^o$. This last case is of special interest since for the given ϵ_r and B values there exist three guided modes in the substrate. The zero of $f_1(u,\beta)$ pertaining to the dominant TM mode can be shown to occur near the branch point k_0. This explains the nature of the pattern near $\theta = \pi/2$, i.e. a strong surface wave mode is launched along the ends of the dipole and it propagates as a cylindrical wave on the substrate interface [14].

Conclusions.

It has been demonstrated in this paper that it is possible to make a precise trade-off analysis for the design of optimum printed circuit dipole elements. This has been achieved by determining the dependence of the antenna element properties on substrate thickness and relative dielectric constant. For a substrate chosen with $\epsilon_r = 2.35$, an optimum design for a center fed dipole has been determined to require a substrate thickness of $0.2\lambda_0$. This yields an optimum radiation efficiency $\eta = 4$, i.e. the power radiated in free space is four times that which is coupled in guided modes in the substrate. This design also gives an optimum bandwidth of 18% and an input resonant resistance of 90 ohms for a resonant length of $L_r = 0.369345\lambda_0$.

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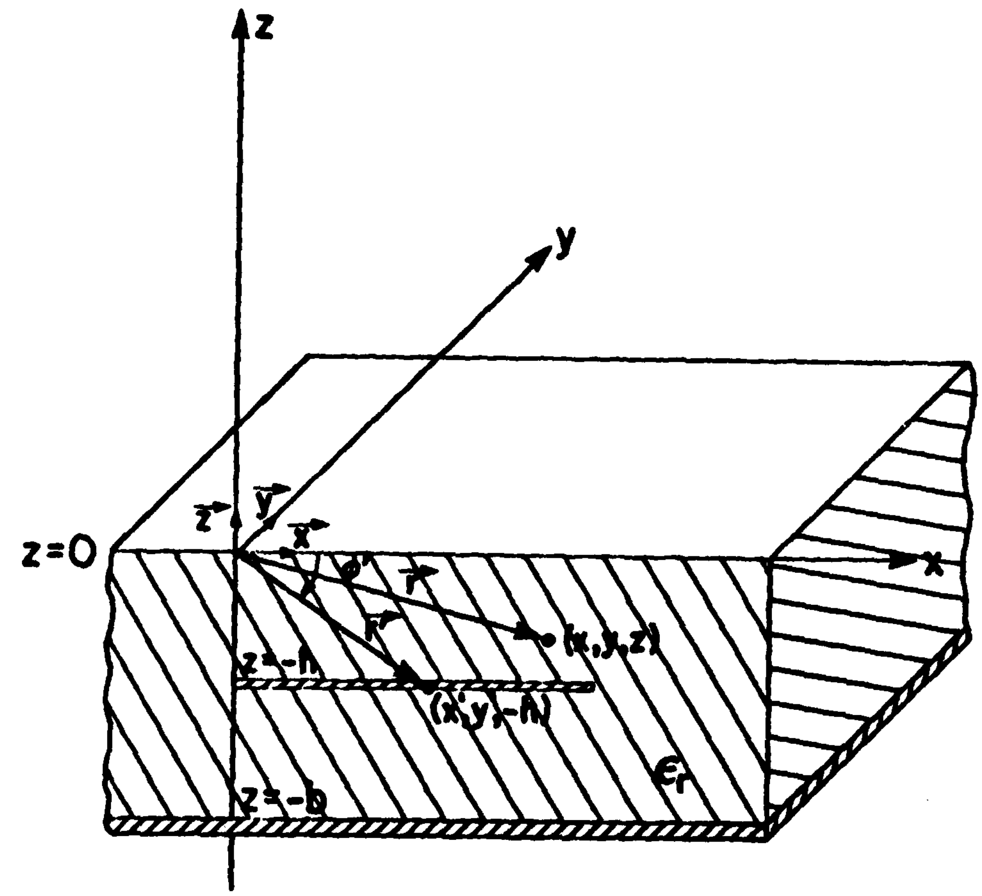
Acknowledgements:

The authors wish to thank Professor H.J. Orchard for helpfull discussions on bandwidth definitions.



$$\vec{r}_x = \frac{\vec{r}'\cos\phi'}{|\vec{r}'\cos\phi'|}$$

Figure 1 : Wire Dipole Embedded in a Grounded Dielectric Slab

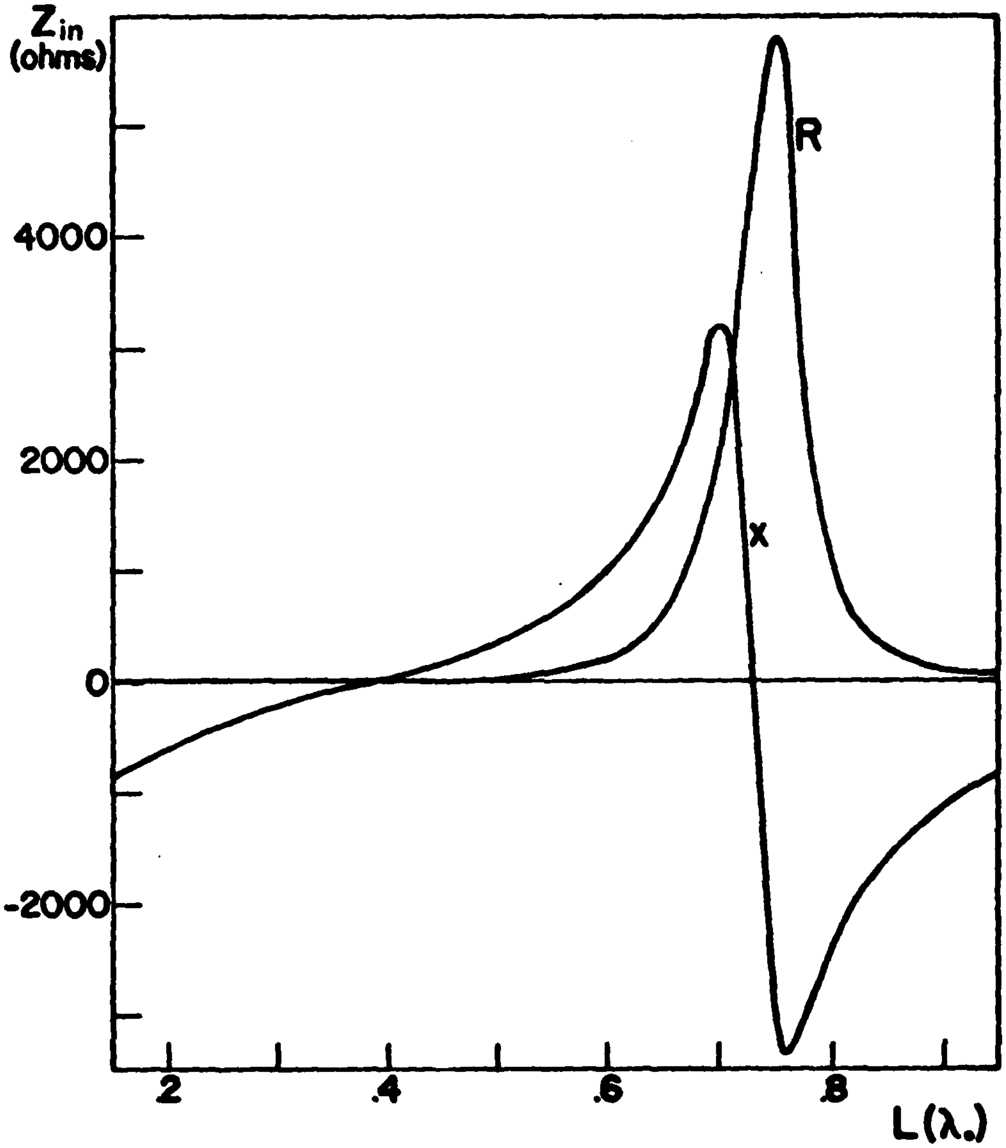


Figure 2 : Input Impedance for a Printed Dipole with $\epsilon_r=2$ and $b=0.1016\lambda_o$ Resonant Length $L_r=0.38\lambda_o$

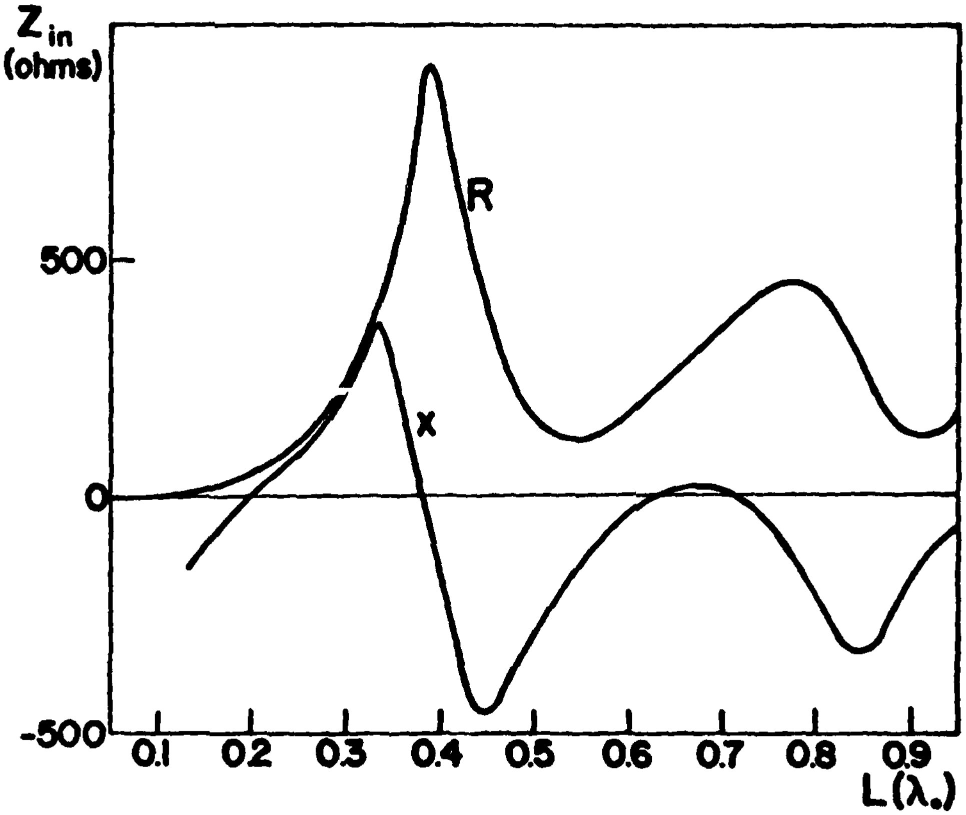


Figure 3 : Input Impedance for a Printed Dipole with ϵ_r=10 and b=0.1016λ_o Resonant Length L_r=0.205λ_o

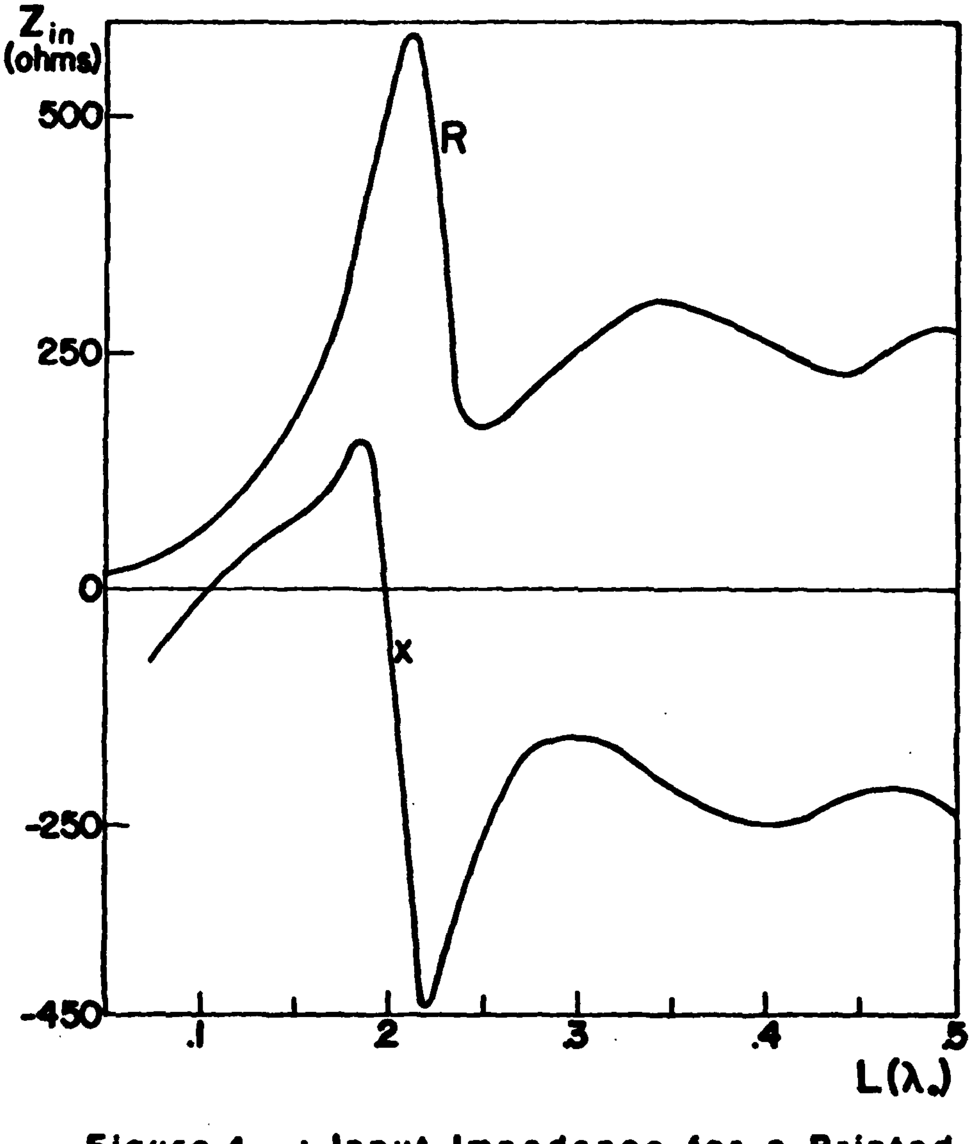


Figure 4 .: Input Impedance for a Printed Dipole with ϵ_r=35 and b=0.1016λ_o Resonant Length L_r=0.1025λ_o

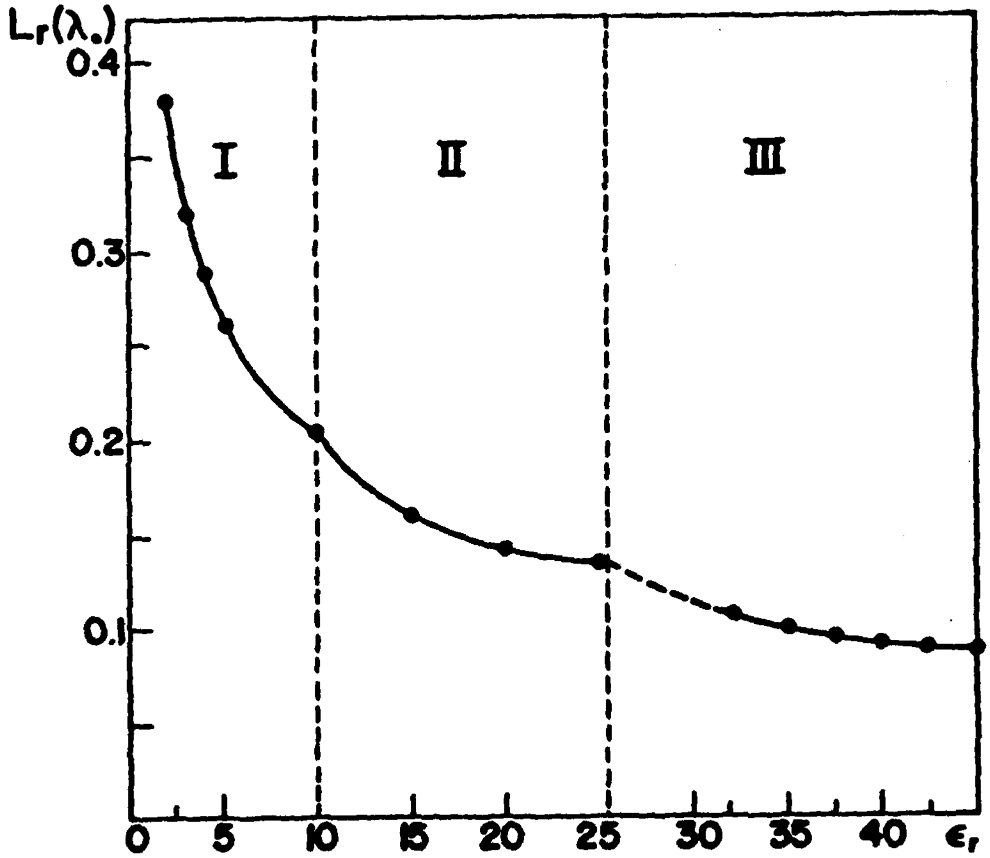


Figure 5 : Resonant Length vs. Relative Dielectric Constant ϵ_r for the Printed Dipole with $b=0.1016\lambda_o$

I: One Surface Wave

II: Two Surface Waves

III: Three Surface Waves

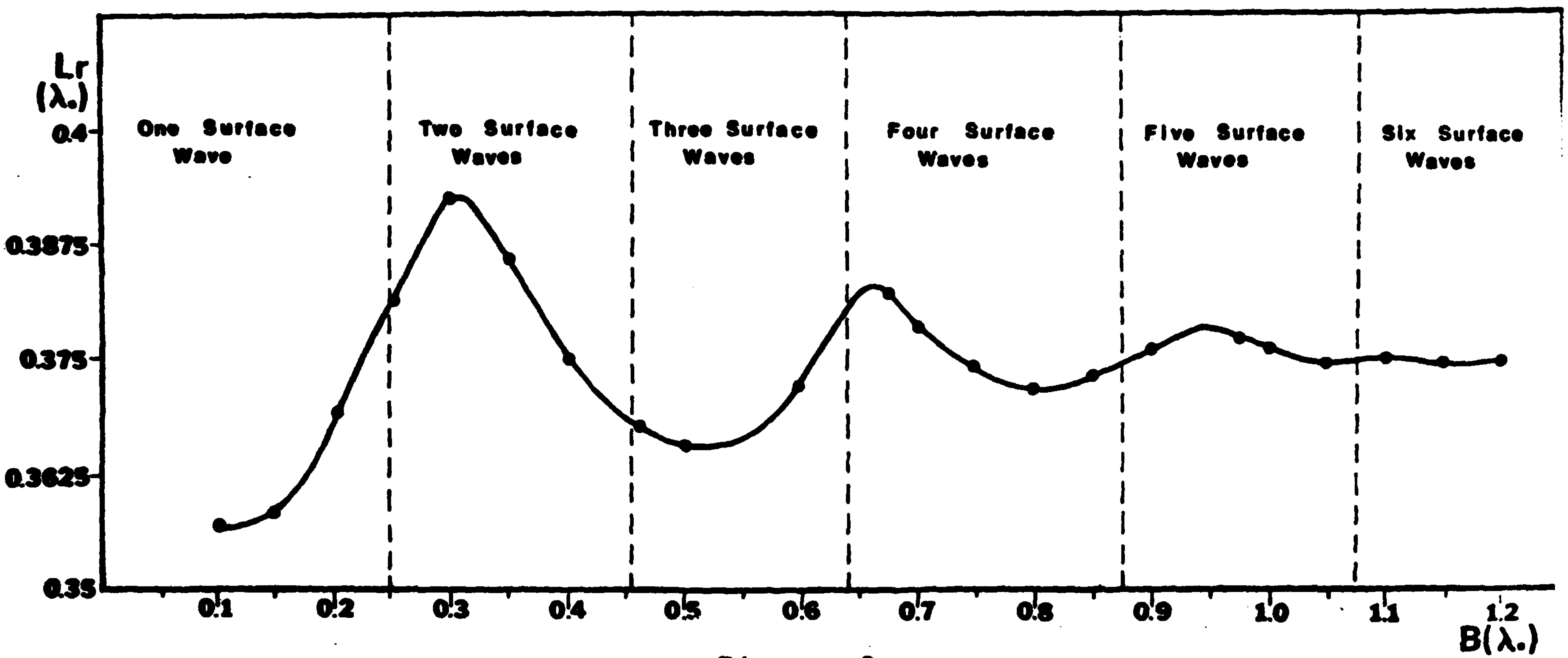


Figure 6

Resonant Length L vs. Substrate Thickness B

($\epsilon_r = 2.35$)

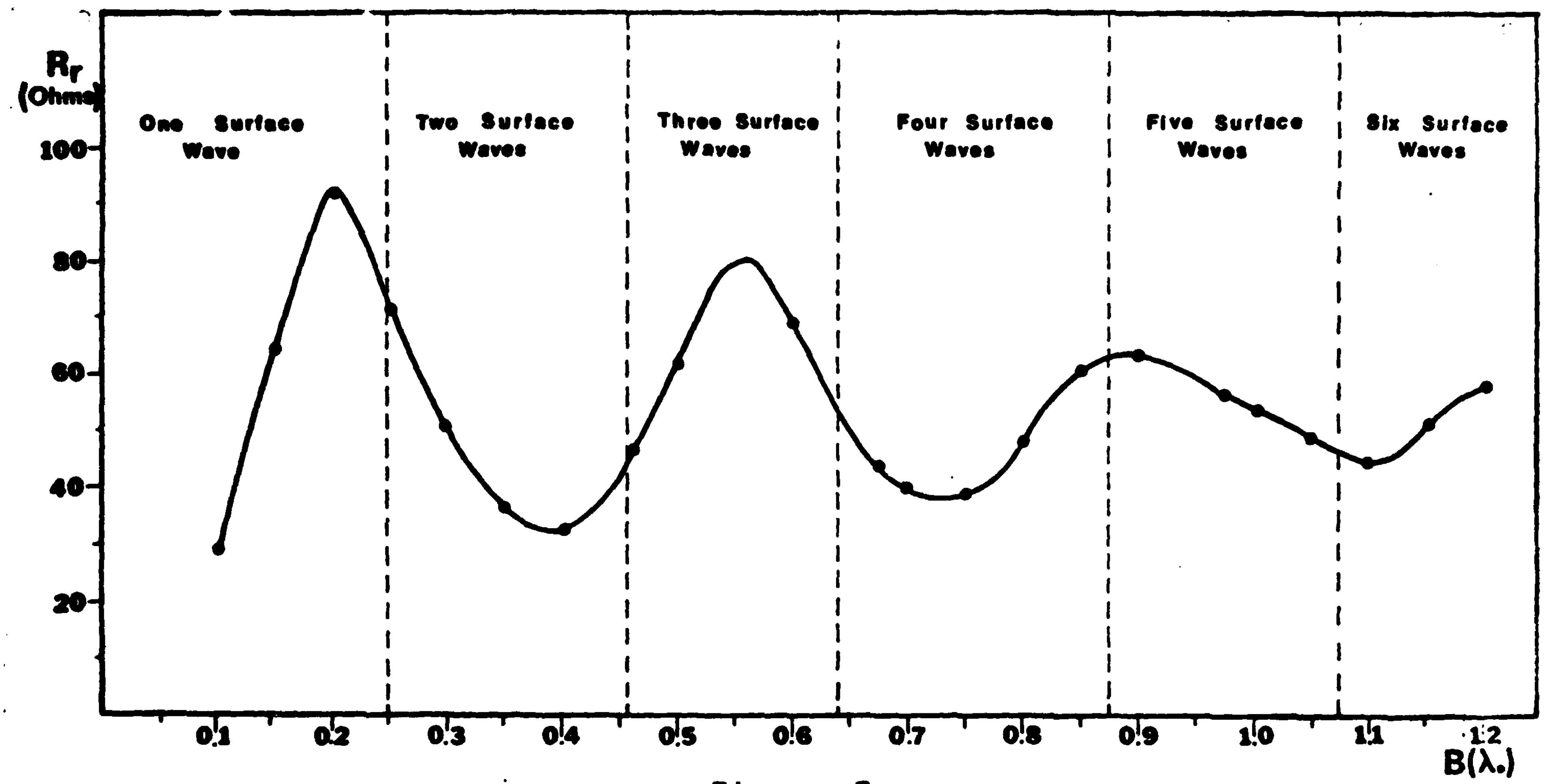


Figure 7

Resonant Resistance vs. Substrate Thickness B

$(\epsilon_r = 2.35)$

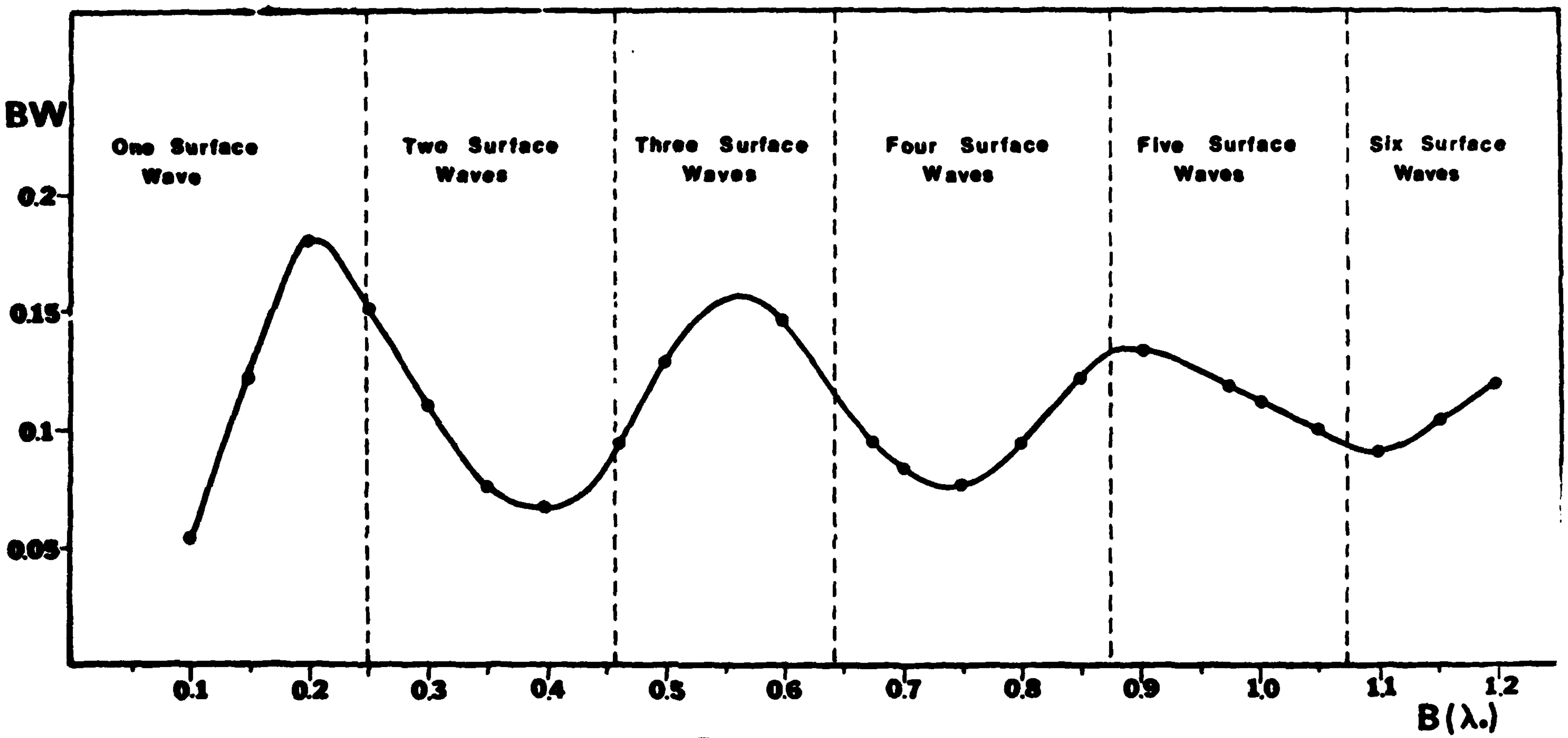


Figure 8

Bandwidth BW vs. Substrate Thickness B

($\epsilon_r = 2.35$)

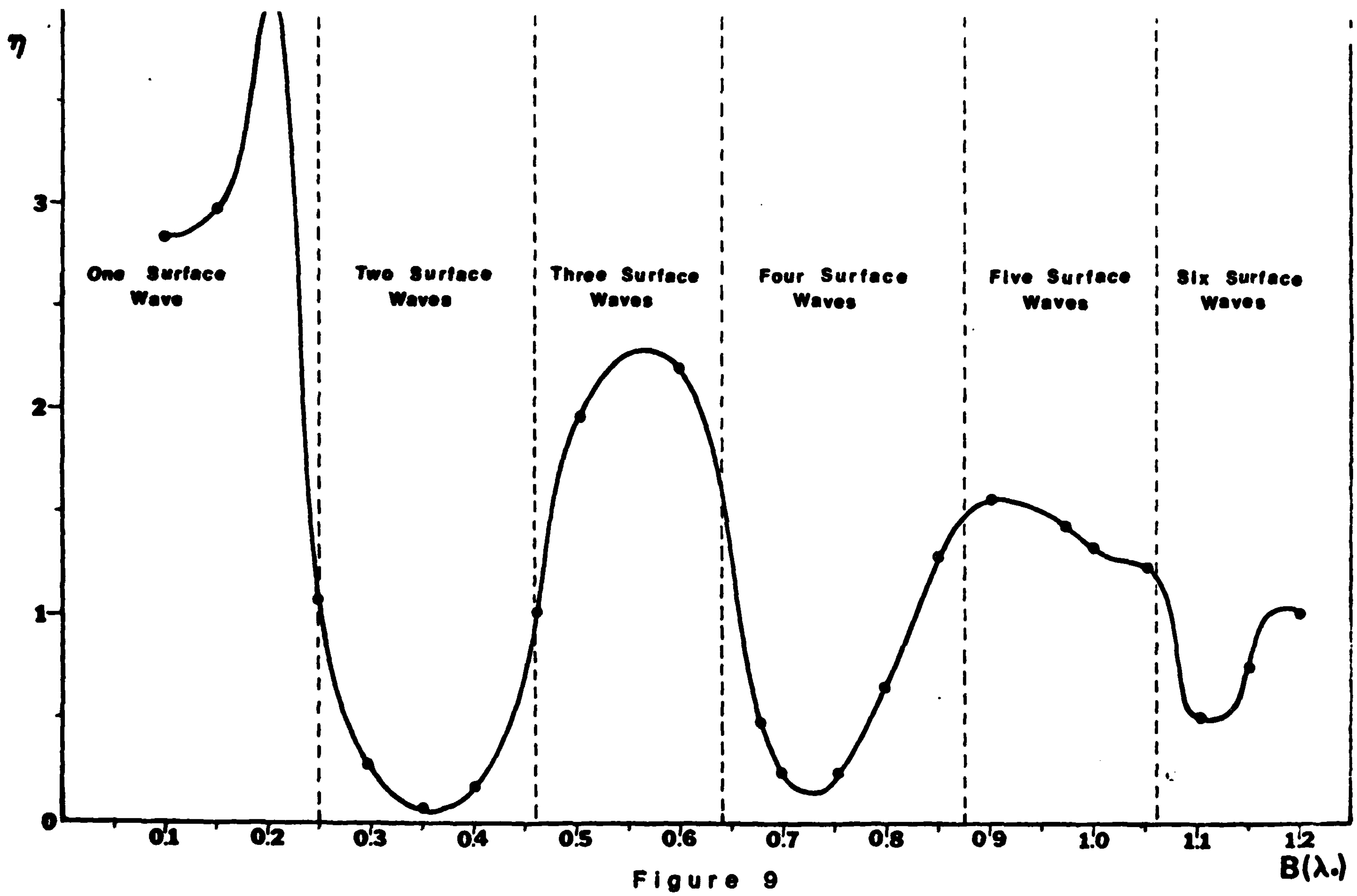


Figure 9

Efficiency η vs. Substrate Thickness B ($\epsilon_r = 2.35$)

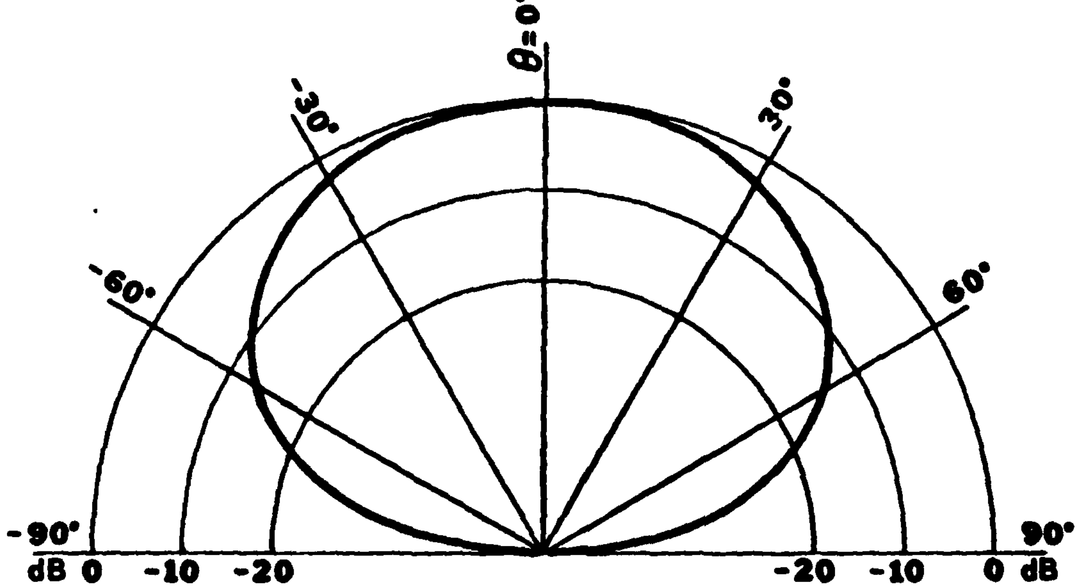


Figure 10

E-Plane Normalized Power Pattern

(ϵ_r=2.35, B=0.20λ_o, L=0.369345λ_o)

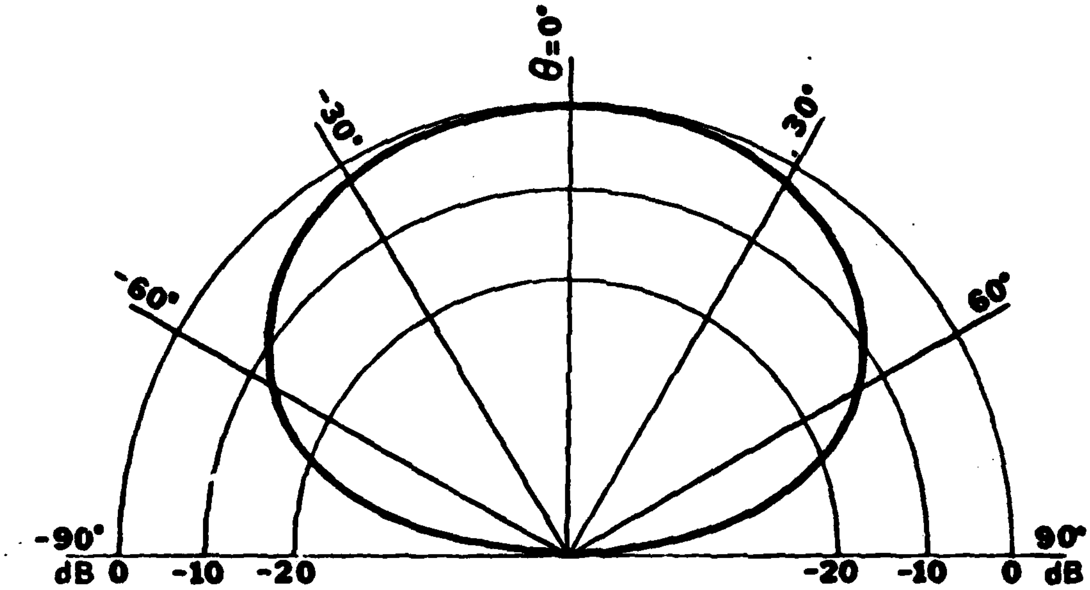


Figure 11

E-Plane Normalized Power Pattern

(ϵ_r=2.35, B=0.2λ_o, L=0.3λ_o)

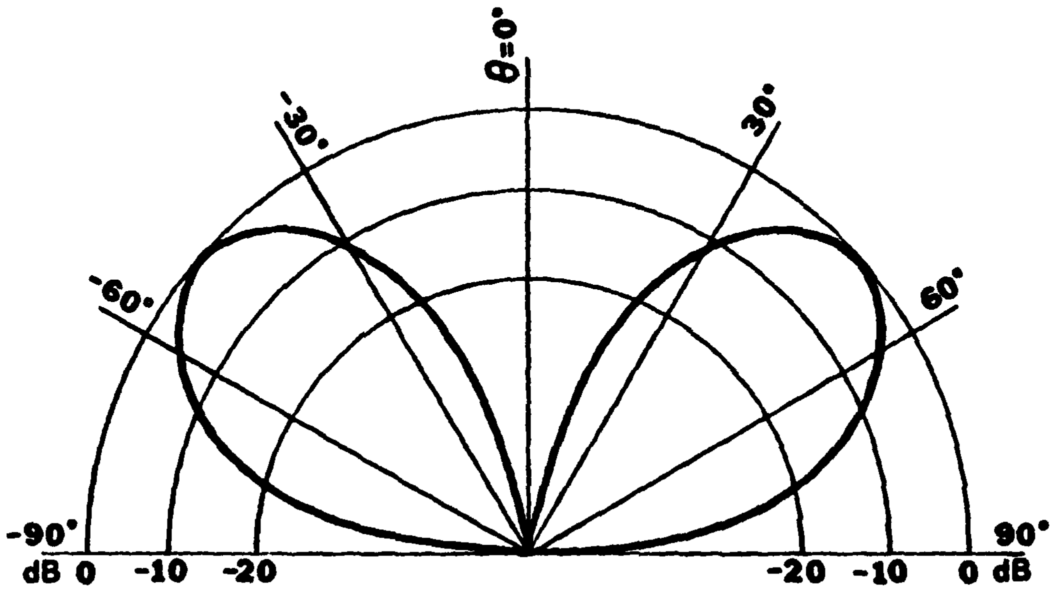


Figure 12

E-Plane Normalized Power Pattern

(ϵ_r=2.35, B=0.975λ_o, L=0.3λ_o)

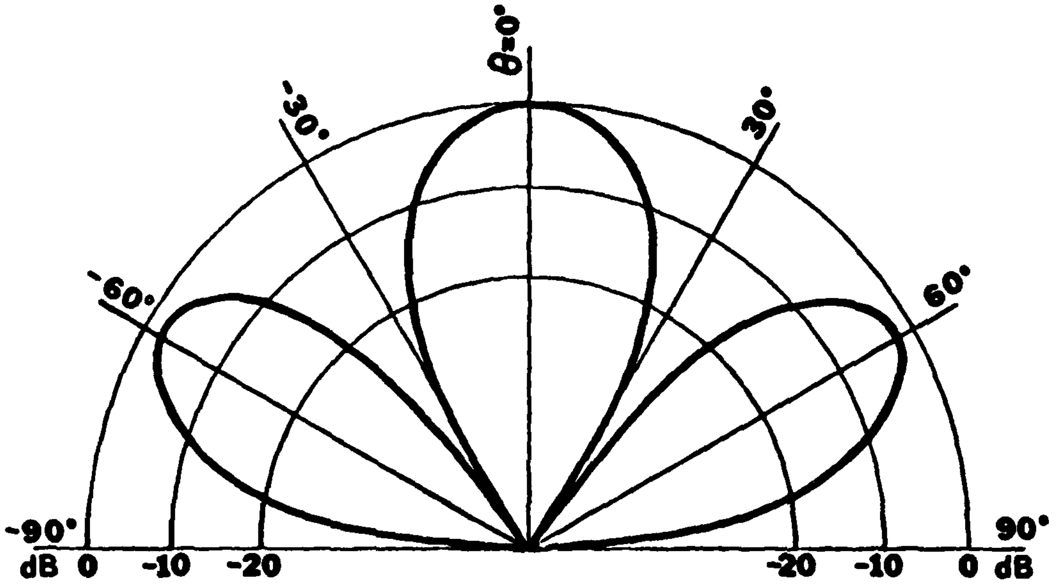


Figure 13

E-Plane Normalized Power Pattern
($\epsilon_r = 2.35, B = 1.05\lambda_o, L = 0.3\lambda_o$)

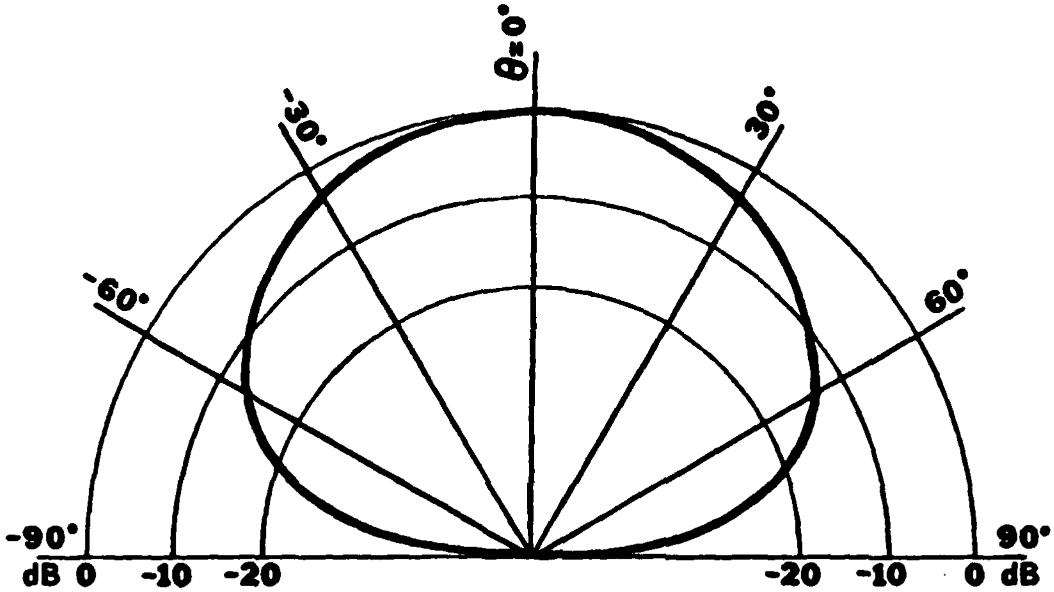


Figure 14

E-Plane Normalized Power Pattern

(ϵ_r=2.35, B=0.1016λ_o, L=0.3λ_o)

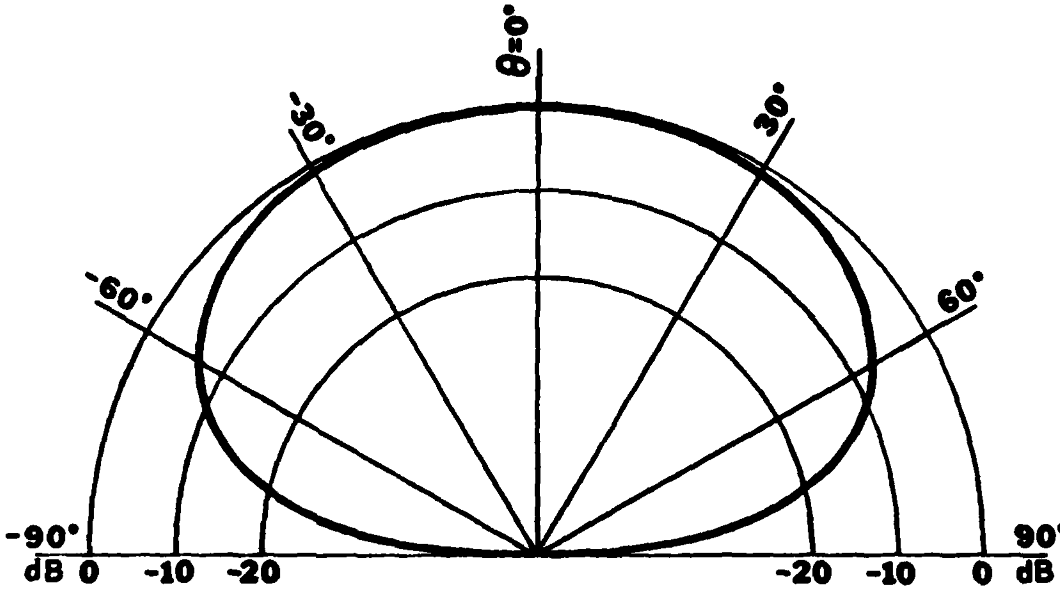


Figure 15

E-Plane Normalized Power Pattern

(ϵ_r=10.0, B=0.1016λ_o, L=0.3λ_o)

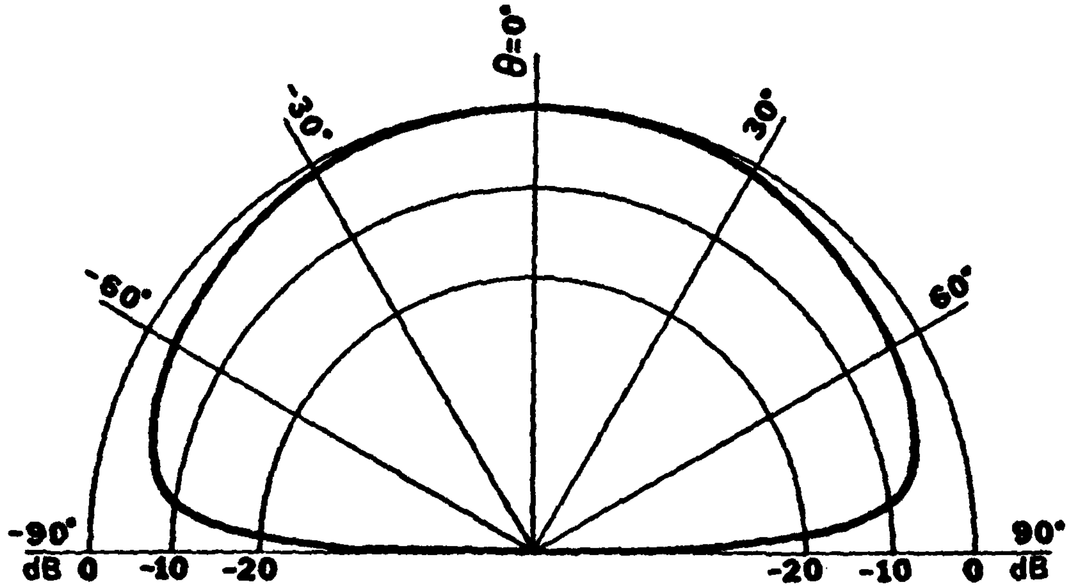


Figure 16

E-Plane Normalized Power Pattern

(ϵ_r=35.0,B=0.1016λ_o ,L=0.3λ_o)

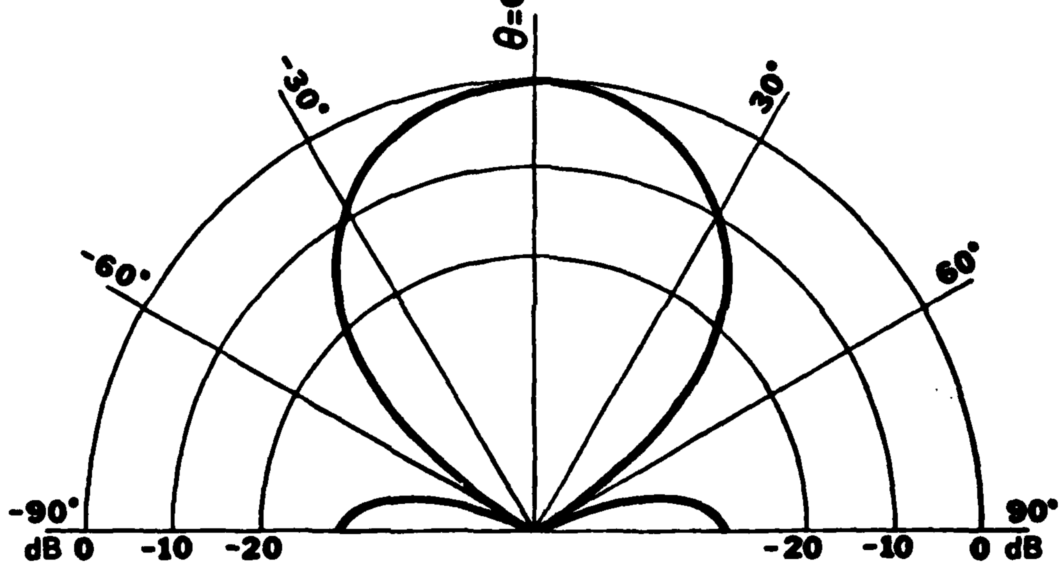


Figure 17

E-Plane Normalized Power Pattern

($\epsilon_r = 25$, $B = 0.1016\lambda_o$, $L = 0.3\lambda_o$)