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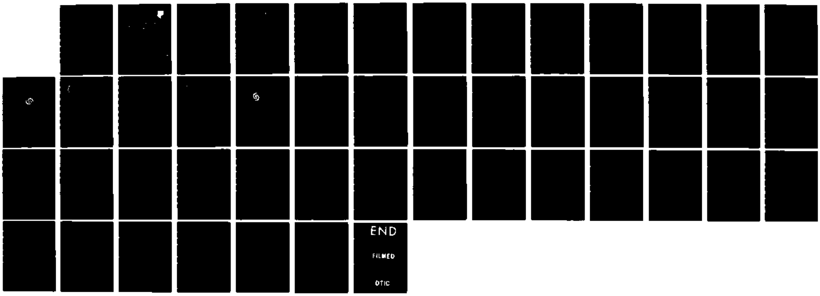

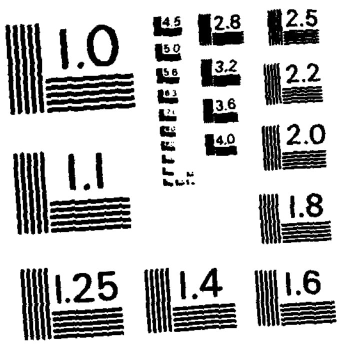

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RADC-TR-85-146
Interim Report
August 1985

A STUDY OF THE SPIRAPHASE AND ANISOTROPIC SUBSTRATES IN MICROSTRIP ANTENNAS

Georgia Institute of Technology

J. J. H. Wang

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UNCLASSIFIED
SECURITY CLASSIFICATION OF THIS PAGE

AD-A159 862

REPORT DOCUMENTATION PAGE

1a. REPORT SECURITY CLASSIFICATION UNCLASSIFIED	1b. RESTRICTIVE MARKINGS N/A
2a. SECURITY CLASSIFICATION AUTHORITY N/A 2b. DECLASSIFICATION/DOWNGRADING SCHEDULE N/A	3. DISTRIBUTION/AVAILABILITY OF REPORT Approved for public release; distribution unlimited
4. PERFORMING ORGANIZATION REPORT NUMBER(S) N/A	5. MONITORING ORGANIZATION REPORT NUMBER(S) RADC-TR-85-146
6a. NAME OF PERFORMING ORGANIZATION Georgia Institute of Technology 6b. OFFICE SYMBOL (If applicable)	7a. NAME OF MONITORING ORGANIZATION Rome Air Development Center (EEA)
6c. ADDRESS (City, State and ZIP Code) Atlanta GA 30332	7b. ADDRESS (City, State and ZIP Code) Hanscom AFB MA 01731
8a. NAME OF FUNDING/SPONSORING ORGANIZATION Rome Air Development Center 8b. OFFICE SYMBOL (If applicable) EEA	9. PROCUREMENT INSTRUMENT IDENTIFICATION NUMBER F19628-84-K-0010
8c. ADDRESS (City, State and ZIP Code) Hanscom AFB MA 01731	10. SOURCE OF FUNDING NOS. PROGRAM ELEMENT NO.: 61102F PROJECT NO.: 2305 TASK NO.: J3 WORK UNIT NO.: 42

11. TITLE (Include Security Classification)
A STUDY OF THE SPIRAPHASE AND ANISOTROPIC SUBSTRATES IN MICROSTRIP ANTENNAS

12. PERSONAL AUTHOR(S)
J. J. H. Wang

13a. TYPE OF REPORT	13b. TIME COVERED	14. DATE OF REPORT (Yr., Mo., Day)	15. PAGE COUNT
Interim	FROM Dec 83 TO Dec 84	August 1985	50

16. SUPPLEMENTARY NOTATION
N/A

17. COSATI CODES FIELD	GROUP	SUB. GR.	18. SUBJECT TERMS (Continue on reverse if necessary and identify by block number)
17	02		anistropic substrate; phased arrays
17	03		spiraphase antenna
			microstrip antenna

19. ABSTRACT (Continue on reverse if necessary and identify by block number)

Two subjects are discussed in this interim report; the spiraphase antenna and the use of anisotropic substrates in microstrip antennas to decouple the radiation and scattering modes. It was observed that the spiraphase, a multi-arm spiral structure can be used either as a reflector or directly as a transmit and/or receive antenna. The bandwidth of the reflector case is much more limited; however, improvements are possible by understanding the mechanism that limits the bandwidth. The broadband nature of the spiraphase used directly as a transmit and/or receive antenna was established. The development of spiraphase into a microstrip antenna is discussed.

The use of anisotropic substrates, or other substrates with "exotic" materials, was found to be useful in decoupling the radiation and scattering modes in the microstrip antenna. This decoupling process is made possible because the use of anisotropic material nullified the foundation on which the reciprocity, or the equivalence (over)

20. DISTRIBUTION/AVAILABILITY OF ABSTRACT UNCLASSIFIED/UNLIMITED ☐ SAME AS RPT ☒ DTIC USERS ☐	21. ABSTRACT SECURITY CLASSIFICATION UNCLASSIFIED
22a. NAME OF RESPONSIBLE INDIVIDUAL Charles J. Drane	22b. TELEPHONE NUMBER (Include Area Code) (617) 861-2051 22c. OFFICE SYMBOL RADC (EEA)

DD FORM 1473, 83 APR EDITION OF 1 JAN 73 IS OBSOLETE.

UNCLASSIFIED
SECURITY CLASSIFICATION OF THIS PAGE

of receiving and transmitting patterns of antenna rests.

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FOREWORD

This Interim Report describes some of the research under SFRC (Short Form Research Contract) No. F19628-84-K-0010, Georgia Tech Project A-3727. A bulk of the research in this project is in the development of a numerical computational algorithm during this interim period, and it will be the subject of the Final Report. The research was performed in the Electromagnetic Effectiveness Division of the Electronics and Computer Systems Laboratory of Georgia Tech Research Institute at Georgia Institute of Technology. Dr. J. J. H. Wang served as the Project Director.

SECTION I

INTRODUCTION

Two subjects, the diode-switched spiraphase antenna and the microstrip antenna on an unconventional (e.g., anisotropic) substrate, constitute the topics of research to which this project is dedicated. This report presents some of the achievements in both subjects for this interim period.

The original spiraphase array concept was based on the assumption that, as a reflector, the spiraphase is similar to a conventional spiral antenna in the following two aspects: that they both have spatial phase (radiation or scattering) patterns and that they are both broadband in their electric performances, especially for the equiangular and logarithmic types [1-6]. However, as has been pointed out [7], this rather intuitive assumption is invalid; thus the experimental broadbanding effort, dedicated and persisting for a period of eight years, has been unsuccessful.

In this study, a truly broadband spiraphase concept was examined. A spiral element was employed; however, it was used directly as an antenna, not as a reflector. This concept had been, however, explored by Kaiser[8] and Sengupta and Smith[9] experimentally. The present effort was devoted to extension of the concept to a microstrip array, in which the basic spiral structure is mounted on a substrate backed by a conducting plane.

Although the spiraphase reflectarray was found to be inherently limited in its frequency bandwidth, some effort was devoted to investigate in greater detail the mechanism which was responsible for this limitation. Much understanding has been gained, and as a result methods for significantly broadening the bandwidth are feasible. The resulting antennas are no longer the conventional spirals but are fundamentally different, even though the spiral structures are retained.

Anisotropy of substrates used in microstrip antennas and circuits has become a subject of increasing importance [10-19]. Most of the work in the literature addresses the related but quite different problem of a dipole in proximity to an anisotropic substrate. Furthermore, little numerical work or experimental results have been reported. In this research, the bulk of the effort is dedicated to the development of a computer algorithm for a Hertzian dipole in a uniaxial anisotropic substrate backed by a conducting plane which will be discussed in its entirety in the Final Report. In this report, the observation that the radiation and scattering modes can be decoupled in an anisotropic medium will be discussed.

SECTION II

SPIRAPHASE AS A BROADBAND ARRAY ELEMENT

It is well known that the spiral antenna has, in its radiation pattern, the factor $\exp(\pm jm\phi)$, where m is the mode index and ϕ is a spherical coordinate with the z-axis parallel to the axis (broadside direction) of the spiral. Normally, the case of m=1 is the only mode of practical interest in spiraphase application. The factor $\exp(\pm jm\phi)$ constitutes a phase pattern which can be utilized for the switching of phase needed in electronic beam-scanning arrays. The first application of this principle was by Kaiser [8], who demonstrated both experimentally and theoretically that beam scan can be accomplished by mechanically rotating individual elements of an eight-element spiral array. Later Phelan [1-6] explored the spiraphase reflectarray concept in which an array of spiral elements is used as a reflector with beam switching capability. Phelan's idea that the use of broadband spirals, such as equiangular spirals, would lead to "frequency-independent" arrays was found to be a misguided intuition. This was pointed out by Wang [7] who observed the inherent limitations in broadbanding the spiraphase reflectarray.

In this research, the possibility of broadbanding for the spiraphase reflectarray was closely examined to learn the precise mechanism by which its bandwidth is limited. With this knowledge, it is possible to propose in this report methods by which the bandwidth of the spiraphase reflectarray can be broadened.

A. Broadband Spiraphase Antenna

The spiraphase array work of Kaiser [8] was conducted just before the frequency-independent spirals became well known. Although broadband potential was implicated in the paper, the schemes and designs conceived by him were primarily based on narrow-band concepts. For example, the side-to-side or back-to-back spacing between spirals were key parameters in some of the designs, which would then be sensitive to the frequency of operation.

Although the phase term in a spiral is associated with a circularly polarized radiation field, a linearly-polarized field with the desired spatial phase term can be generated in several ways as was demonstrated by Kaiser. However, his doublet array element, which attains linear polarization from two adjacent spirals with opposite senses of polarization, can scan in one dimension only. Obviously, this limitation is due to the fact that the spiral doublet has two phase centers, one at each center of the two spirals. The distance between the doublet spirals leads to misalignment of the phase patterns and thus high cross polarization components. Furthermore, if we consider doublet spirals as elements of a planar array, the doublet spirals will be so large in the orthogonal dimension that grating lobes will likely arise.

At the sacrifice of 3 dB, a spiraphase array can achieve linear polarization by placing a screen of parallel resistive strips in front of the array. The polarization of the emerging wave will be perpendicular to the strips. Since most planar and conical spirals are loaded with absorbing materials for broadband matching, the 3 dB loss inherent in the strip filter configuration can often be tolerated.

The spiraphase structure appears to be inherently suitable as a microstrip antenna because of its planar geometry. However, Wood [20] reported that when a 1.59 mm-thick polyguide was used for a spiral antenna, the radiation pattern tends to exhibit large ellipticity or axial ratio. As a result, Wood concluded that the achievement of very wideband operation analogous to the conventional spiral was not feasible, and his further work was concentrated on devices of maximum circumferences of about 1.5 λ, where λ is the free-space wavelength.

It was noticed that the radiation pattern measured by Wood was grossly asymmetrical, which could be primarily due to the asymmetric end-fed configuration employed. The close proximity of the ground plane is not the primary reason for the pattern distortion, but rather provides conditions under which the effect of the asymmetric feed is magnified. The patent of Malagisi [21] uses circular conducting patches as elements in a microstrip array. Each element has a shorting pin at the center of the circular patch. Around the rim of the circular patch, connections are made to the conducting back of the substrate through diodes which are biased at either an "on" or an "off" state. It is governed by the same basic principle as that of the spiraphase antenna. The primary advantage of the spiraphase microstrip array in comparison with the Malagisi array is the large bandwidth that is feasible in the spiraphase concept.

The design of the spiraphase microstrip array begins with the spiral element on a thin substrate on a conducting plane. As shown in Figure 1, a basic equiangular spiral was backed by a dielectric substrate on a ground plane. Broadband impedance measurements were conducted over the frequency range of 4 to 18 GHz. Figure 2 shows the relative reflected power in dB when there is no ground plane, or when d is infinite. The data indicate the level of impedance matching accomplished by the balun depicted in Figure 1. Figures 3 and 4 show the broadband matching for the case of d = ¼ inch and 1/16 inch, respectively. As can be seen, the degradation in impedance matching is about 1 or 2 dB on the average for the ¼ inch substrate backing. When the substrate is reduced to 1/16 inch, the same thickness as in Wood's experiments, the impedance matching is deteriorated by 5 dB on the average in comparison with the case without a ground plane. Thus, the spiral, and the spiraphase, is promising as a microstrip antenna from the viewpoint of impedance matching. As for pattern quality, it does not appear to have any severe inherent limitations but further study is needed before the actual pattern quality can be established.

B. Spiraphase Reflectarray

In a previous paper by this author [7], it was noted that the spiraphase antenna was inherently narrow banded and worked for circular polarization only. However, in this research it was observed that both linear polarization and broader bandwidth were feasible in new types of spiraphase antennas. One example is the utilization of anisotropic substrates which open up a new dimension in the design parameters. Another example is based on a change of the spiral configuration based on an understanding of the mechanism which limits the bandwidth of the spiraphase. This latter approach is discussed here.

Figure 5 shows a single spiraphase element illuminated by a circularly-polarized plane wave $\underline{E}^i$ defined as

$$\underline{E}^i = E_x^i \hat{x} + j\, E_y^i \hat{y} \qquad (1)$$

SECTION IV
CONCLUSIONS AND RECOMMENDATIONS

The spiraphase antenna as a broadband phased array element has been examined and found to be promising. As a direct transmit or receiving antenna, the broadband nature of the spiraphase appears well established. When used in a reflectarray, the inherent bandwidth limitations are found to be remediable by modifying the spiraphase element in a fundamental way so that its antenna mode predominates.

The use of anisotropic matter in antenna design opens a new dimension in antenna design. Its impact on antenna technology is expected to be significant, as evidenced by the discovery that the radiation and scattering modes of antenna can be decoupled. A next generation of antennas based on this approach will likely be developed to meet increasingly more sophisticated systems requirements.

$$E_z = \frac{-k_o^2 J_o}{4\omega\varepsilon_o} H_o^{(2)}(k_o\rho) \tag{3-12}$$

Now, if this line antenna is considered also as a receiving antenna, we can show that the scattered as well as the received power will depend on ε_t. Let the line antenna be illuminated by a plane wave coming from a direction at an angle θ with respect to the z-axis, we see that the electric field of the plane wave cannot, in general, be parallel to that of the z axis. Thus, ε_t will affect the receiving and scattering property of this line antenna.

This case demonstrated that the radiation and scattering modes can be decoupled from each other. In fact, in this ideal case they are totally isolated so that ε_t can be manipulated to alter the scattering and receiving property of the antenna without any effect on its transmitting property.

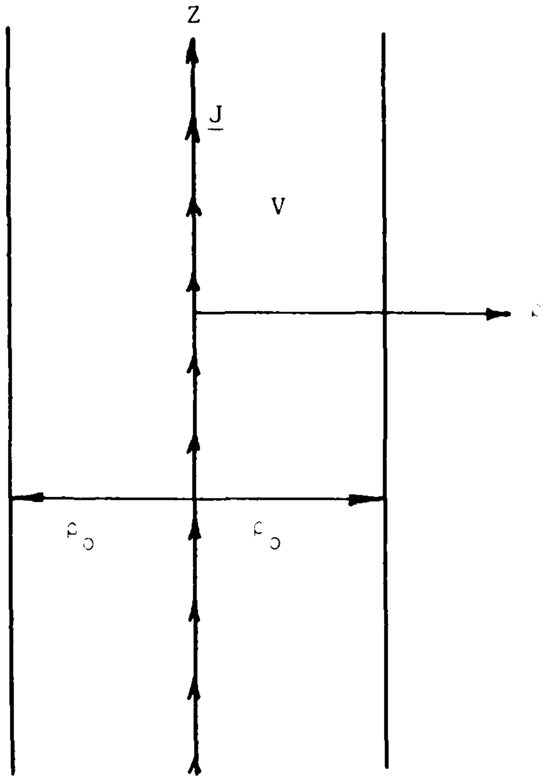

Figure 6. An infinitely long line antenna bounded by a cylindrical volume of anisotropic material.

Figure 6 shows an infinitely long electric line current source along the z-axis which can be expressed as follows

$$\underline{J} = \hat{z}\, J_o\, \delta(\rho) \quad , \tag{3-5}$$

where cylindrical coordinates are employed.

The uniform line current is surrounded by an infinitely long cylindrical volume V of radius ρ_o . The permittivity tensor is,

$$\varepsilon = \begin{bmatrix} \varepsilon_t & o & o \\ o & \varepsilon_t & o \\ o & o & \varepsilon_z \end{bmatrix} \quad , \tag{3-6}$$

where $\varepsilon_t = \varepsilon_z = \varepsilon_o$ for $\rho > \rho_o$ but are complex constants for $\rho \leq \rho_o$.

Let us start with Maxwell's equations,

$$\nabla \times \underline{E} = j\omega\mu_o \underline{H} \text{ , and} \tag{3-7}$$

$$\nabla \times \underline{H} = -j\omega \underline{\underline{\varepsilon}} \cdot \underline{E} + \underline{J} \quad . \tag{3-8}$$

The symmetry of the problem dictates that the fields be independent of z and ϕ coordinates. As a result, Equations (3-7) and (3-8) can be reduced to,

$$E_\rho = 0 \text{ ,} \tag{3-9}$$

$$E_\phi = 0 \text{, and} \tag{3-10}$$

$$\frac{\partial^2 E_z}{\partial \rho^2} + \frac{1}{\rho}\frac{\partial E_z}{\partial \rho} + \omega^2 \mu_o \varepsilon_z E_z = -j\omega\mu_o J_z\, \delta(\rho) . \tag{3-11}$$

Thus, the electric field is independent of ε_t. If we choose $\varepsilon_z = \varepsilon_o$, the radiated field will be the same as that in free-space and be expressed as

TE fields

$$E_x(x,y,z) = E_x^o\,(\sqrt{K_t}\,x, \sqrt{K_t}\,y, \sqrt{K_t}\,z),$$

$$E_y(x,y,z) = E_y^o\,(\sqrt{K_t}\,x, \sqrt{K_t}\,y, \sqrt{K_t}\,z),$$

$$E_z(x,y,z) = E_z^o\,(x,y,z) = 0,$$

$$H_x(x,y,z) = \sqrt{K_t}\,H_x^o\,(\sqrt{K_t}\,x, \sqrt{K_t}\,y, \sqrt{K_t}\,z),$$

$$H_y(x,y,z) = \sqrt{K_t}\,H_y^o\,(\sqrt{K_t}\,x, \sqrt{K_t}\,y, \sqrt{K_t}\,z),$$

$$H_z(x,y,z) = \sqrt{K_t}\,H_z^o\,(\sqrt{K_t}\,x, \sqrt{K_t}\,y, \sqrt{K_t}\,z),$$

$$J_x(x,y,z) = K_t\,J^o{}_x\,(\sqrt{K_t}\,x, \sqrt{K_t}\,y, \sqrt{K_t}\,z),$$

$$J_y(x,y,z) = K_t\,J^o{}_y\,(\sqrt{K_t}\,x, \sqrt{K_t}\,y, \sqrt{K_t}\,z), \text{ and}$$

$$J_z(x,y,z) = K_t\,J^o{}_z\,(\sqrt{K_t}\,x, \sqrt{K_t}\,y, \sqrt{K_t}\,z). \qquad (3\text{-}4)$$

Although Equations (3-3) and (3-4) are originally derived for a medium of infinite extent, they can be expanded into cases in which a finite, or even small, volume of anisotropic matter is used as long as the TE and TM decomposition is valid. This possibility of extension into realistic antenna problems is based on the satisfaction of Maxwell's equations in the region occupied by the source and its immediate region, no matter what size it may be. Another view is that this can be considered as the solution for the primary excitation in dealing with boundary value problems such as for stratified media. An application of these relations is presented in the following subsection.

B. Decoupling of the Radiation and Scattering Modes of Antennas

The preceding subsection provided the background for the present discussion which shows that the radiation and scattering modes of an antenna can sometimes be decoupled from each other by the use of anisotropic matter. In particular, this appears to be feasible for the class of microstrip antennas, which is partially discussed in the Appendix. In this subsection, it will be shown that the decoupling between the radiation and scattering modes is feasible in principle.

The modified reciprocity reveals a very interesting and significant phenomenon. On the one hand, it reveals that the validity of the familiar and fundamental reciprocity is in question when an anisotropic matter is employed in the antenna. On the the other hand, there is a certain orderliness in the way anisotropy affects the characteristics of an antenna. Thus, new types of antennas can be developed by using anisotropic matter. In fact, anisotropy represents a new discussion in antenna design and is expected to have profound impact on antennas and the related technology.

An interesting example is the case of an electromagnetic wave in a medium characterized by the dielectric constant tensor

$$K = \begin{bmatrix} K_t & o & o \\ o & K_t & o \\ o & o & K_z \end{bmatrix} , \qquad (3\text{-}2)$$

in which the Cartesian coordinates are used. Clemmow [30] found that there is a simple relationship between this problem and a related problem which has a vacuum medium and a properly modified current source. We denote the fields $\underline{E}$, $\underline{H}$ and current source $\underline{J}$ of this related problem with superscripts o. Decomposing the problem into the TE and TM fields, he found the following simple relations between these two related problems.

<u>TM fields</u>

$$\begin{aligned}
E_x(x,y,z) &= \sqrt{K_z}\, E_x^o(\sqrt{K_z}\, x, \sqrt{K_z}\, y, \sqrt{K_t}\, z), \\
E_y(x,y,z) &= \sqrt{K_z}\, E_y^o(\sqrt{K_z}\, x, \sqrt{K_z}\, y, \sqrt{K_t}\, z), \\
E_z(x,y,z) &= \sqrt{K_t}\, E_z^o(\sqrt{K_z}\, x, \sqrt{K_z}\, y, \sqrt{K_t}\, z), \\
H_x(x,y,z) &= \sqrt{K_t K_z}\, H_x^o(\sqrt{K_z}\, x, \sqrt{K_z}\, y, \sqrt{K_t}\, z), \\
H_y(x,y,z) &= \sqrt{K_t K_z}\, H_y^o(\sqrt{K_z}\, x, \sqrt{K_z}\, y, \sqrt{K_t}\, z), \\
H_z(x,y,z) &= H_z^o(x,y,z) = 0, \\
J_x(x,y,z) &= K_t \sqrt{K_z}\, J_x^o(\sqrt{K_z}\, x, \sqrt{K_z}\, y, \sqrt{K_t}\, z), \\
J_y(x,y,z) &= K_t \sqrt{K_z}\, J_y^o(\sqrt{K_z}\, x, \sqrt{K_z}\, y, \sqrt{K_t}\, z), \text{ and} \\
J_z(x,y,z) &= K_z \sqrt{K_t}\, J_z^o(\sqrt{K_z}\, x, \sqrt{K_z}\, y, \sqrt{K_t}\, z) .
\end{aligned} \qquad (3\text{-}3)$$

SECTION III

RADIATION AND SCATTERING CHARACTERISTICS OF ANTENNAS USING ANISOTROPIC SUBSTRATES

Antennas have so far been made of linear, isotropic materials, even though some ferrites have been used for electrically small antennas. To meet the ever-increasing requirements demanded in antenna design, one wonders whether the use of anisotropic materials might help. In this research we have found that the use of anisotropic, or even more "exotic", material opens a new dimension in antenna design which could significantly impact antenna technology. In this section, we will emphasize the basic non-reciprocal nature of antennas involving anisotropic matter and proceed to discuss the feasibility of the decoupling of the radiation and scattering modes of antennas followed by a specific case for a microstrip antenna.

A. Modified Reciprocity for Antennas Involving Anisotropic Matter

It is a well known phenomenon, and a basis of much theoretical and experimental work, that the receiving pattern of any antenna constructed of linear isotropic matter is identical to its transmitting pattern [25-26]. However, the use of anisotropic matter can lead to antennas that do not have the reciprocal property often taken for granted by antenna engineers. At this point, it is worthwhile to point out that pattern reciprocity is perhaps the only practical observable regarding reciprocity, at least as far as the practicing engineer is concerned.

A more accurate discussion in this subject can be based on the modified reciprocity theorem [27-29]. From a practical point of view, the following modified reciprocity theorem of Harrington and Villeneuve [27] is most useful.

$$\langle a,b \rangle = \langle b,\tilde{a} \rangle \tag{3-1}$$

In Equation (3-1), $\langle\ \rangle$ denotes reaction, and "$\sim$" denotes that the permittivity and permeability tensors are transposed. In words, Equation (3-1) states that "the reaction of one set of sources on another is equal to the reaction of the latter set on the former when the permittivity tensor $[\epsilon]$ and permeability tensor $[\mu]$ are transposed.

specifically Γ_m, will then become the overwhelming factor in switching $\underline{E}^s$ and thus the array beam. In this case, the incident power will couple into the element termination with impedance Z_ℓ and be reflected back to radiate into space. In the present concept, Z_ℓ will be implemented as distributed loads, such as diode switches. In short, the spiraphase element will thus be modified such that its broadband antenna characteristics will emerge even though it is in a reflecting, or scattering, mode.

For a specific switching state, the scattered field can be expressed, according to Green [22], as

$$\underline{E}^{s}(\theta,\phi;Z_{\ell}) = \underline{E}^{s}(\theta,\phi;Z_a) - I(Z_a^{*})\,\underline{E}^{r}(\theta,\phi)\,\Gamma_m, \tag{2}$$

where

$$\Gamma_m = \frac{Z_{\ell} - Z_a^{*}}{Z_{\ell} + Z_a} \tag{3}$$

$$I(Z_a^{*}) = -\frac{\underline{h}^{r}(\theta,\phi)\cdot\underline{E}^{i}}{2R_a} \tag{4}$$

$$\underline{E}^{r}(\theta,\phi) = -j\,\frac{\eta}{2\lambda}\,\underline{h}^{t}(\theta,\phi)\,\frac{e^{-j\beta r}}{r} \tag{5}$$

Z_a = antenna impedance (the impedance of the antenna to the transmitter when the antenna is used as a transmitting antenna),

Z_{ℓ} = load impedance. (6)

In Equations (4) and (5), $\eta = (\mu_0/\varepsilon_0)^{\frac{1}{2}}$. $\underline{h}^{r}$ and $\underline{h}^{t}$ are the receiving and transmitting antenna heights as defined by Equations (4) and (5), respectively. In the case of back-scattering, $\underline{h}^{r}(\theta,\phi) = \underline{h}^{t}(\theta,\phi)$.

Green's expression of Equation (2) evolved from the earlier works of Harrison and King [23] and Stevenson [24] . Note that the load impedance (of the transmitter or receiver) affects only the second term of the right-hand side of Equation (2). The original spiraphase concept of Phelan is primarily based on the change in $\underline{E}^{s}$, the first term on the right-hand side of Equation (2), to rotate the phase pattern of the reflected wave. The narrow-band nature of this device was demonstrated in Reference 7.

It is possible, however, to modify the present spiraphase reflector to minimize the first term $\underline{E}^{s}(\theta,\phi;Z_a)$ and rely for switching on the second term through the variation of Z_{ℓ} . This calls for the modification of the spiraphase in a fundamental way so that the first term will be a small quantity. The second term on the right-hand side of Equation (2), or

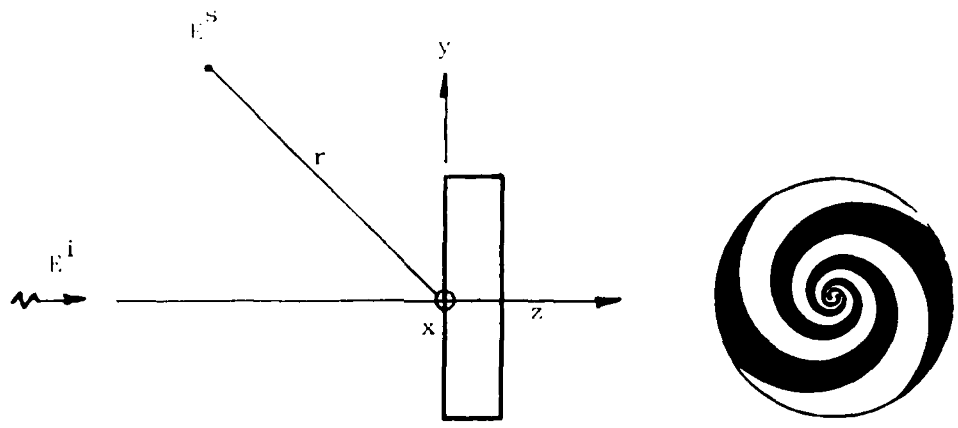

Figure 5. A spiraphase element illuminated by a circularly-polarized plane wave.

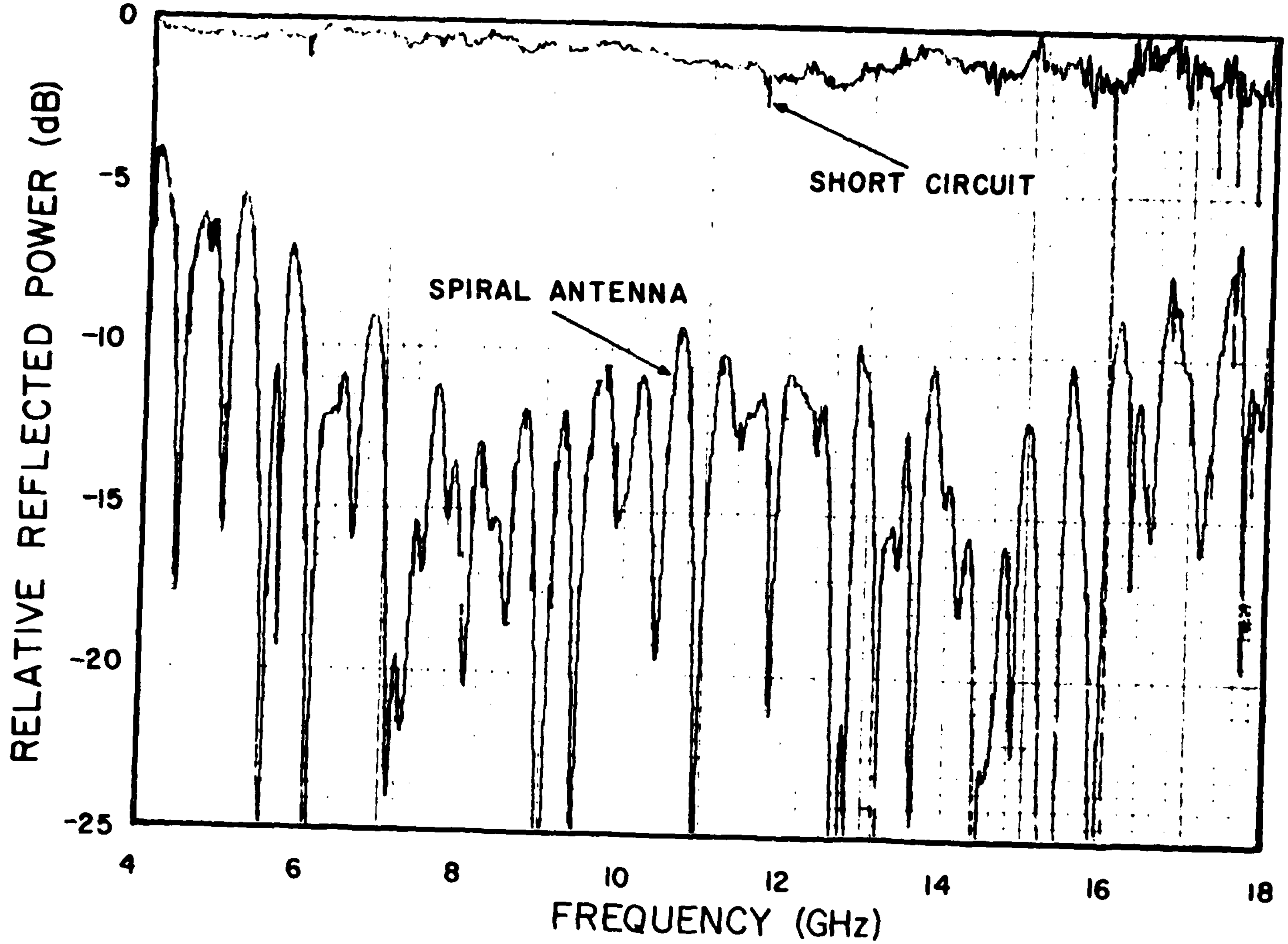

Figure 4. Reflection (impedance) measurement for the spiral of Figure 1 with a ground plane backing at d = 1/16 inches.

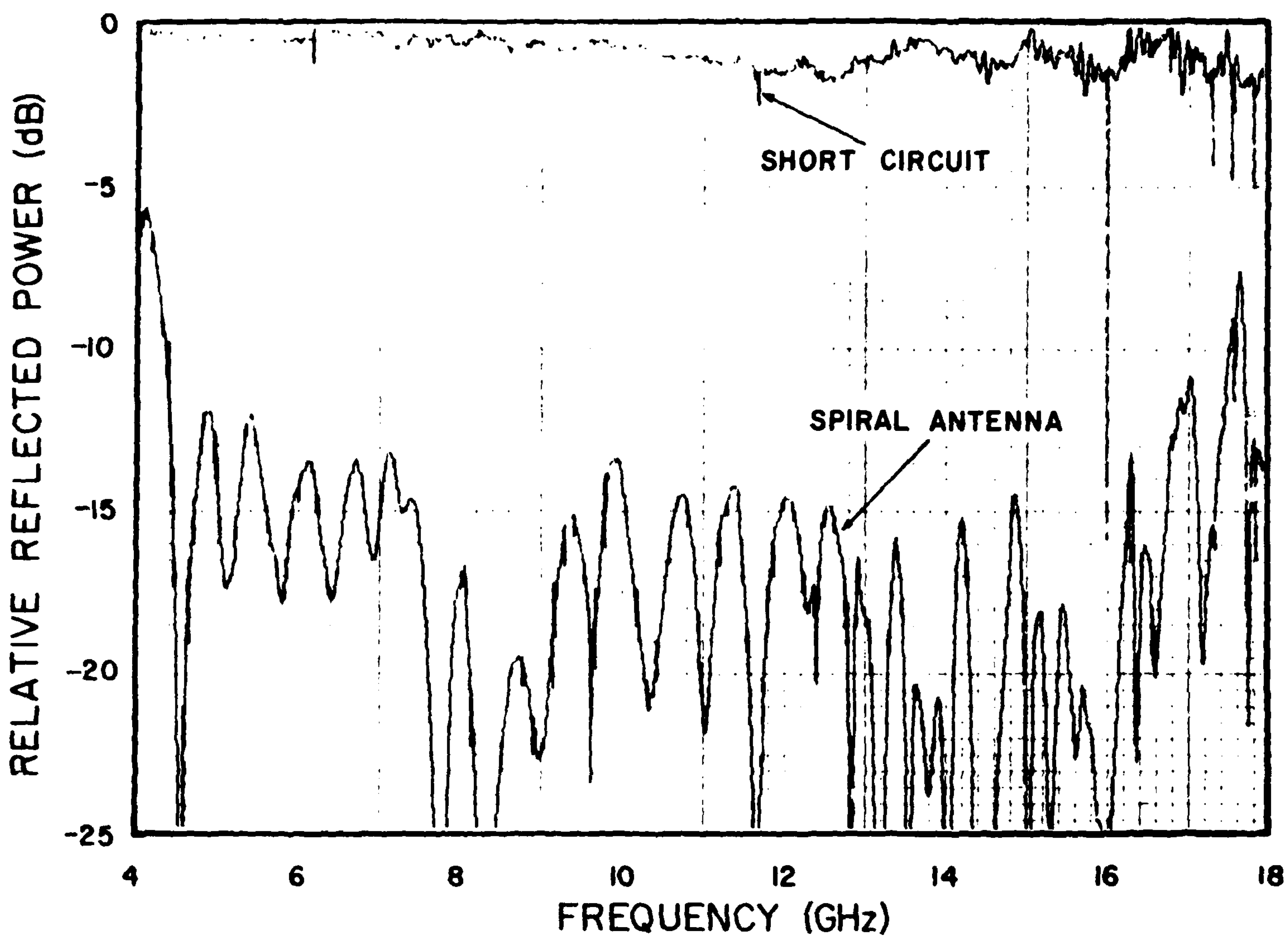

Figure 3. Reflection (impedance) measurement for the spiral of Figure 1 with a ground plane backing at $d = 1/4$ inch.

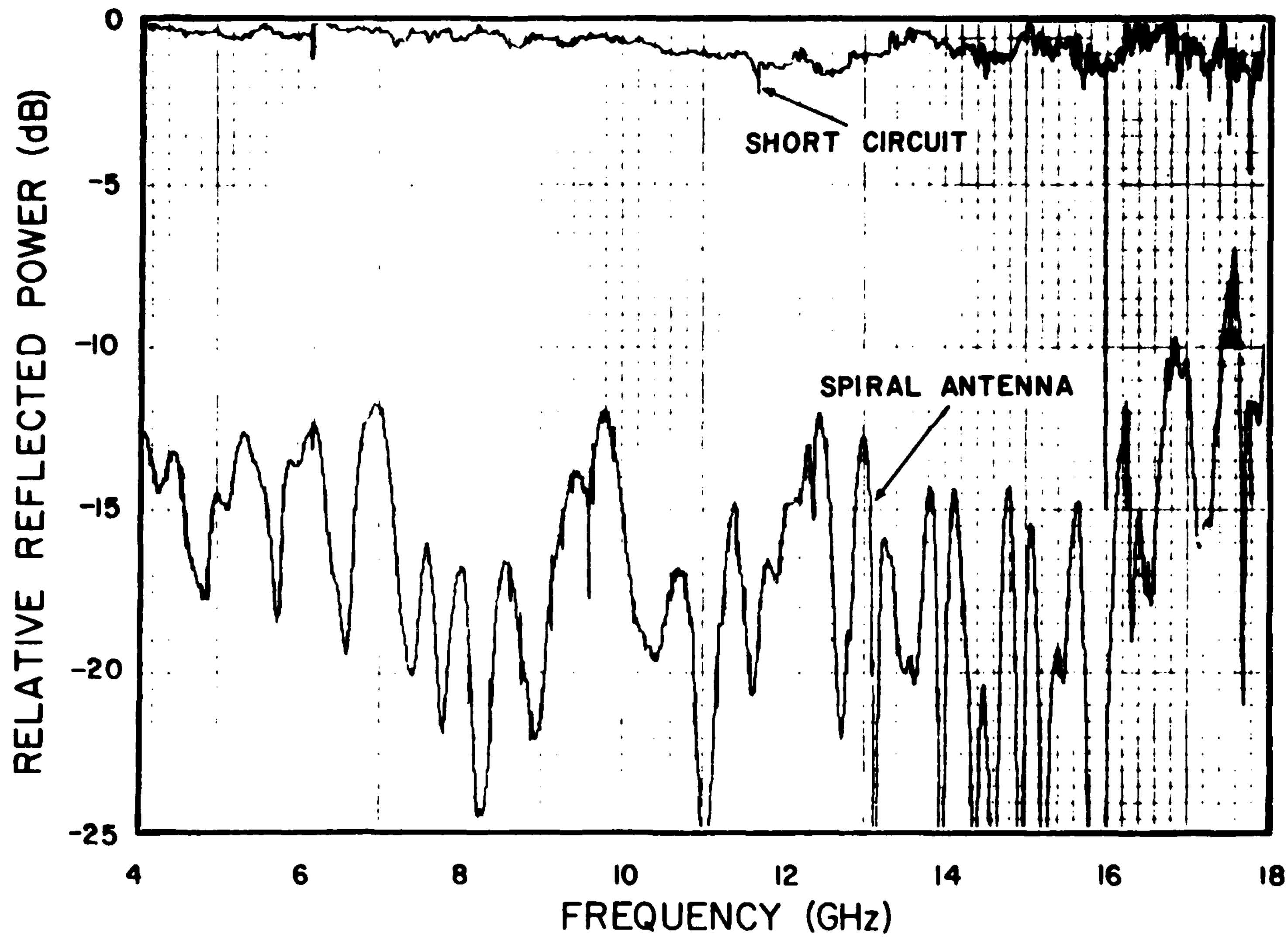

Figure 2. Reflection (impedance) measurement for the spiral of Figure 1 without a ground plane.

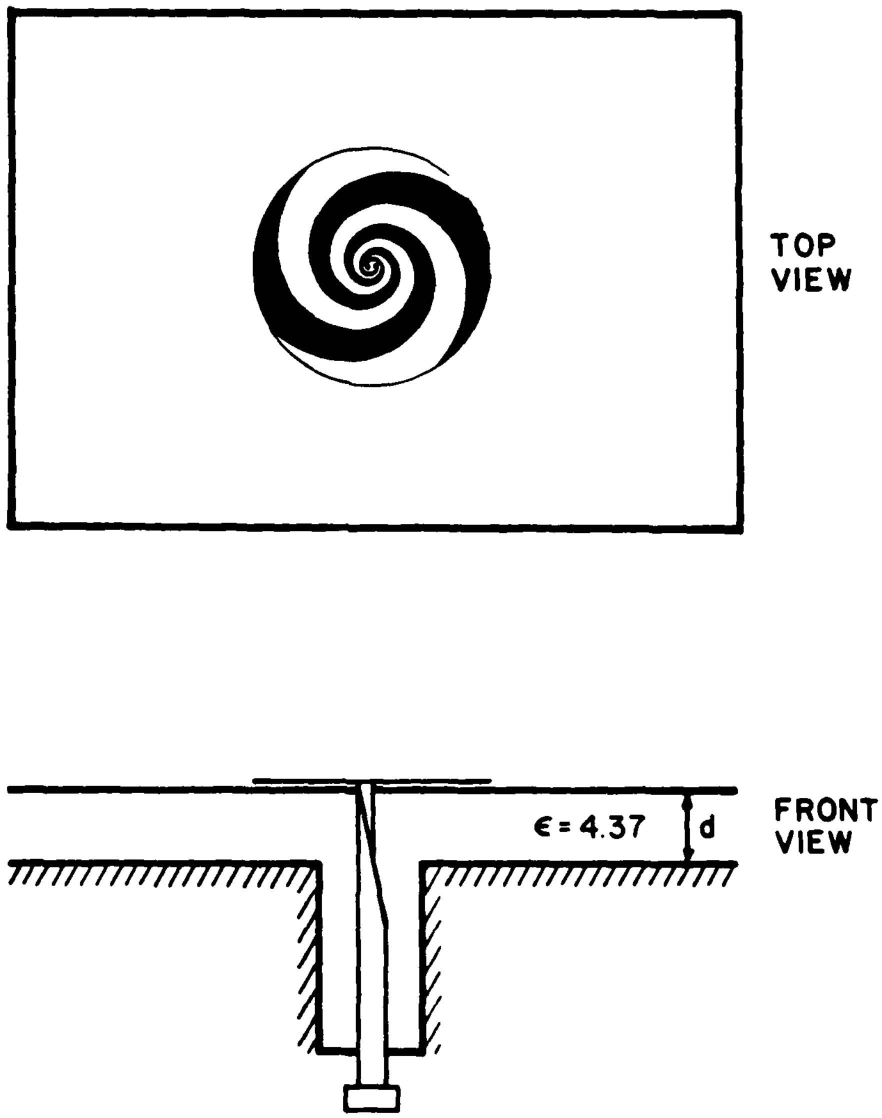

Figure 1. A probe-fed equal-angular spiral microstrip antenna.

SECTION V

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APPENDIX I

A PAPER ENTITLED "RADIATION AND SCATTERING PERFORMANCE OF MICROSTRIP ANTENNAS USING EXOTIC SUBSTRATES," PRESENTED IN THE 1984 ANTENNA APPLICATIONS SYMPOSIUM, UNIVERSITY OF ILLINOIS, SEPTEMBER 19-21, 1984

Radiation and Scattering Performance of Microstrip Antennas Using Exotic Substrates*

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ABSTRACT

Both the scattering and radiation properties are considered in microstrip antenna designs for airborne applications. The use of exotic substrates, instead of simple dielectric substrates, opens up a new dimension, making it much easier to meet both the radiation and scattering requirements. Two types of substrates, the ferromagnetic and the anisotropic dielectric, are being studied because of their commercial availability. It has been observed that the radiation mode and the scattering mode of the microstrip antenna are often distinctly different. Thus, permittivity and permeability tensor of the substrate can be selected to decouple the radiation and scattering modes to achieve simultaneously radiation and scattering performance goals. Analysis techniques used in the design study are also discussed.

1. Introduction

Recently, the use of "exotic" substrates in microstrip

*This research was supported by the Electromagnetic Sciences Division of Rome Air Development Center, Hanscom AFB, USAF Systems Command under Contract F19628-84-K-0010.

antennas has gained recognition as a new way to improve their performances [1,2]. In this paper, we will concentrate in substrates that can act on both the radiation (antenna) and scattering modes of the antenna in different manners so that both the radiation and the scattering performances of the antenna can be simultaneously improved. That this is possible is due to the fact that the radiation and scattering modes of the microstrip antenna are substantially decoupled from each other.

It is conceivable that many possibilities exist in the use of exotic substrates in microstrip antennas. We will, however, focus on the use of uniaxial anisotropic substrates which are suitable for the improvement of radiation and scattering characteristics.

2. Microstrip Antennas Using Uniaxial Anisotropic Substrates

Figure 1 depicts a microstrip antenna covered with a radome. Both the radome and the substrate are, in general, bianisotropic or anisotropic, but is assumed here to be uniaxially anisotropic with permittivity

$$\underline{\underline{\varepsilon}} = \begin{bmatrix} \varepsilon_t & 0 & 0 \\ & \varepsilon_t & \\ 0 & 0 & \varepsilon_z \end{bmatrix} \quad (1)$$

and $\mu = \mu_0$.

Since practical microstrip structures are invariably thin, the electric field in the substrate and the radome has only a z component, that is,

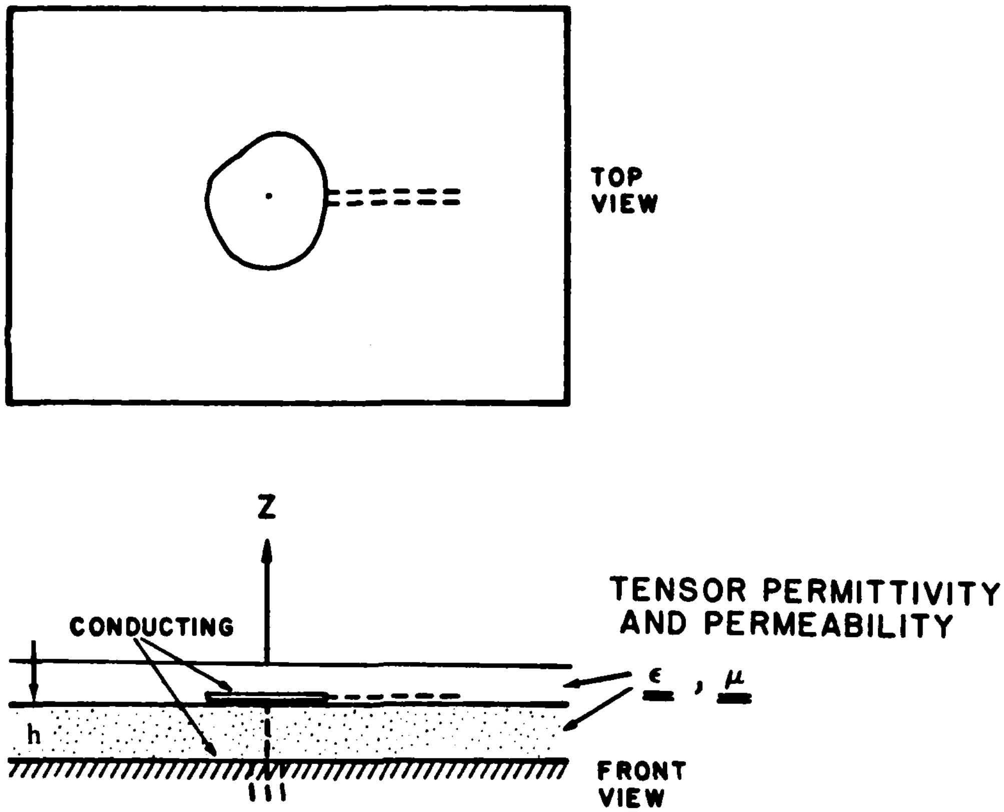

Figure 1. A microstrip antenna with exotic substrate and radome.

$$\underline{E} = \hat{z}\, E \text{ in substrate and radome} \tag{2}$$

The radiation (antenna) mode of the microstrip antenna can be supported by choosing ϵ_z to be of low loss. The scattering mode of the microstrip antenna can be examined by assuming an incident field

$$\underline{E}^i = E^i \,[p\, \hat{t} + (1-p^2)^{1/2}\, \hat{z}\ ] \tag{3}$$

where $\quad 1 \geq p \geq \cos\theta_i$,

and $\hat{t}$ is a unit vector perpendicular to $\hat{z}$. θ_i is the incident angle between $\hat{z}$ and the direction of incident wave propagation. P is a polarization factor, being 1 for TE wave and $\cos\theta_i$ for TM wave. When θ_i is small, $\underline{E}^i$ is primarily $\hat{t}$ polarized. Thus, ϵ_t can be chosen to yield the favored scattering properties without affecting much the radiation property of the microstrip antenna.

3. Analysis of the Radiation Mode

Analysis of a rectangular patch microstrip antenna on a uniaxial anisotropic substrate can be carried out by a combination of the techniques for the case of an isotropic substrate [3,4] and the numerical technique of Kobayashi [5]. Analytical results support the basis of the present design approach; namely the anisotropy of the substrate does not substantially affect the antenna characteristics. Of course, the greater the extent of anisotropy the larger the antenna properties deviate from the case of isotropic substrates.

Although not yet established experimentally, the effects of the anisotropy can probably be handled by using the concept of effective permittivity ε_e [6]. By doing so, we can use the familiar formula developed for isotropic substrates. For example, the resonant frequency can then be related to the patch parameters in the following equations

$$f_r = \frac{c}{2(L + 2\ \Delta\ell)\sqrt{\varepsilon_e}} \tag{4}$$

where

$$\Delta\ell = 0.412\ \frac{(\varepsilon_e + 0.3)(W/h + 0.264)}{(\varepsilon_e - 0.258)(W/h + 0.8)} \tag{5}$$

c = velocity of light in free space,

and L, W and h are dimensions in meters as shown in Figures 1 and 2.

The only unknown in Equation (4) is ε_e, which can be solved numerically by computing the line capacitance C and noting that

$$\varepsilon_e = \frac{C}{C_o} \tag{6}$$

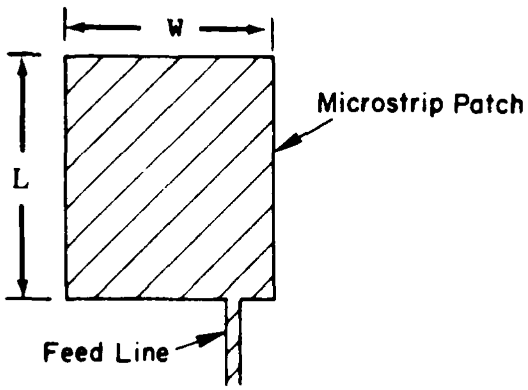

Figure 2. Dimensions of the rectangular patch of a microstrip antenna.

where C_o is the well-known capacitance of the strip patch in the absence of the dielectric substrate; that is, when $\varepsilon = \varepsilon_o$.

The effect of substrate anisotropy can be evaluated by examining the capacitance per unit length for the case of a simple microstrip line. Table 1 shows a comparison of line capacitance per unit length, C/ε_o, between isotropic and anisotropic substrates with the same ε_z. The differences are small, no more than 5%. In fact, a considerable amount of these differences are due to computational errors inherent in each case (the anisotropic cases by Kobayashi [5] , and isotropic cases by Wheeler's method [7]).

In order to reveal the extent of the effects due to substrate anisotropy, the strip line as shown in Figure 3 is examined by the method developed by Shibata, et al [8]. Some of the results are shown in Figures 4-7, in which C denotes the line capacitance per unit for the anisotropic substrate and Cl is the capacitance when $\varepsilon_2 = \varepsilon_1$, the case of isotropic substrate. The ratio C/Cl is plotted versus ε_2 for three normalized line widths (W/B) of 0.01, 0.1 and 1.0. As can be seen, the effect of substrate anisotropy is rather smooth and orderly and does not appear to present serious design difficulties.

It may be desirable to examine the situation when ε_2 is complex. However, there appears to be no reported study for microstrip problems of this nature, especially for the case of a large loss tangent.

Table 1. Comparison of C/ε_o between isotropic and anisotropic substrates

W/h \ $\underline{\underline{\varepsilon}}$	anisotropic $9.4\,\hat{Z} + 11.6\,\hat{t}$	isotropic 9.4	anisotropic $11.6\,\hat{Z} + 9.4\,\hat{t}$	isotropic 11.6
0.1	8.784	8.33	9.260	9.706
1.0	19.689	19.60	22.016	23.24
10.0	105.506	110.31	127.628	130.34

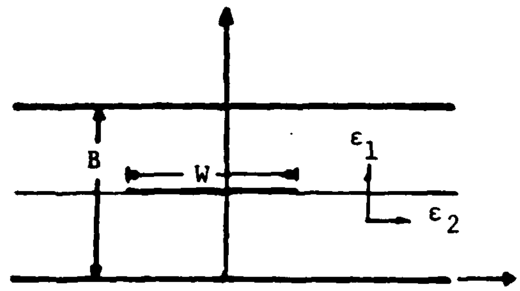

Figure 3. A strip line with anisotropic dielectric.

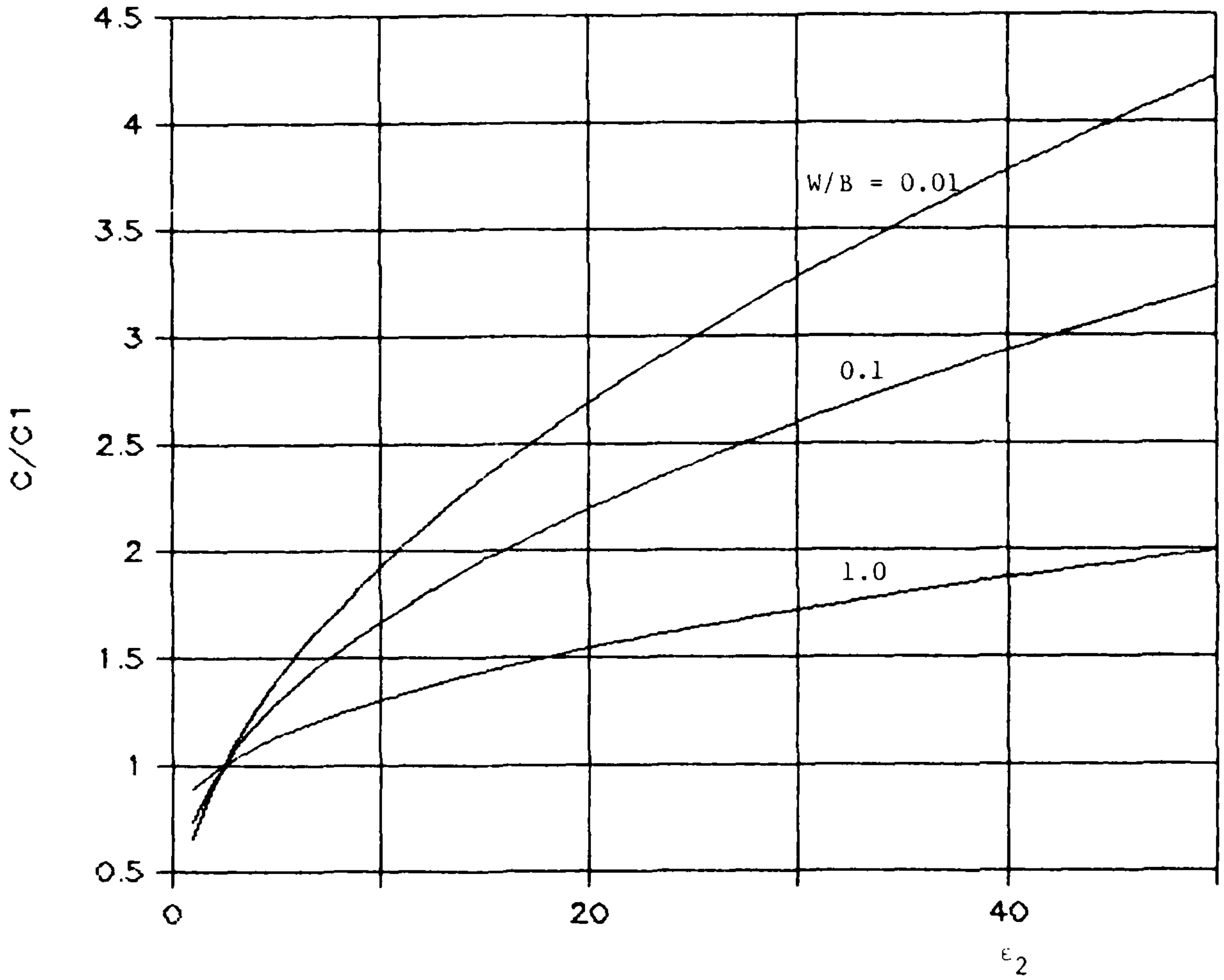

Figure 4. The normalized line capacitance, C/C1, versus ε_2 for the strip line of Figure 3, in which $\varepsilon_1 = 2.5$.

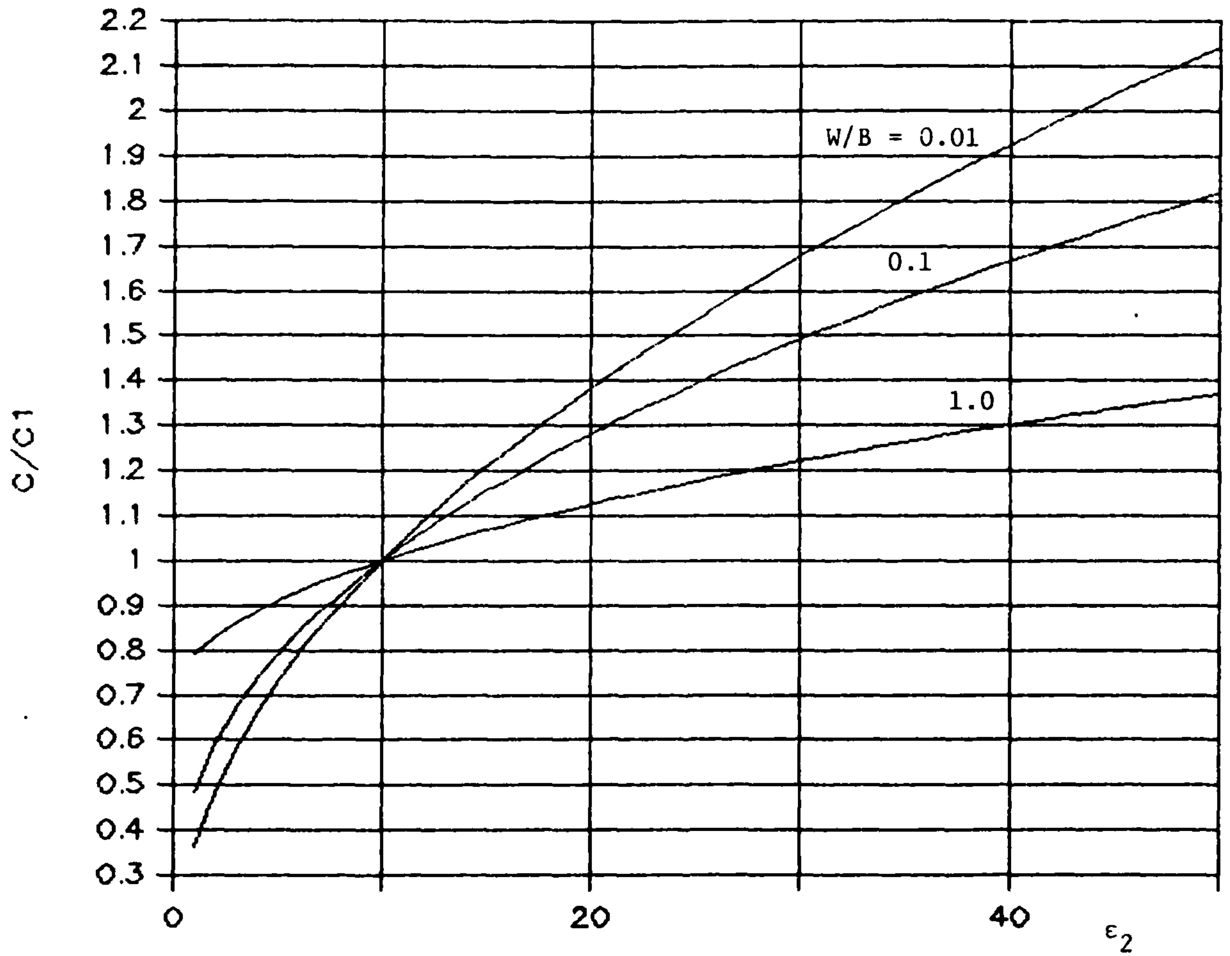

Figure 5. The normalized line capacitance, C/Cl, versus ε_2 for the strip line of Figure 3, in which $\varepsilon_1 = 10$.

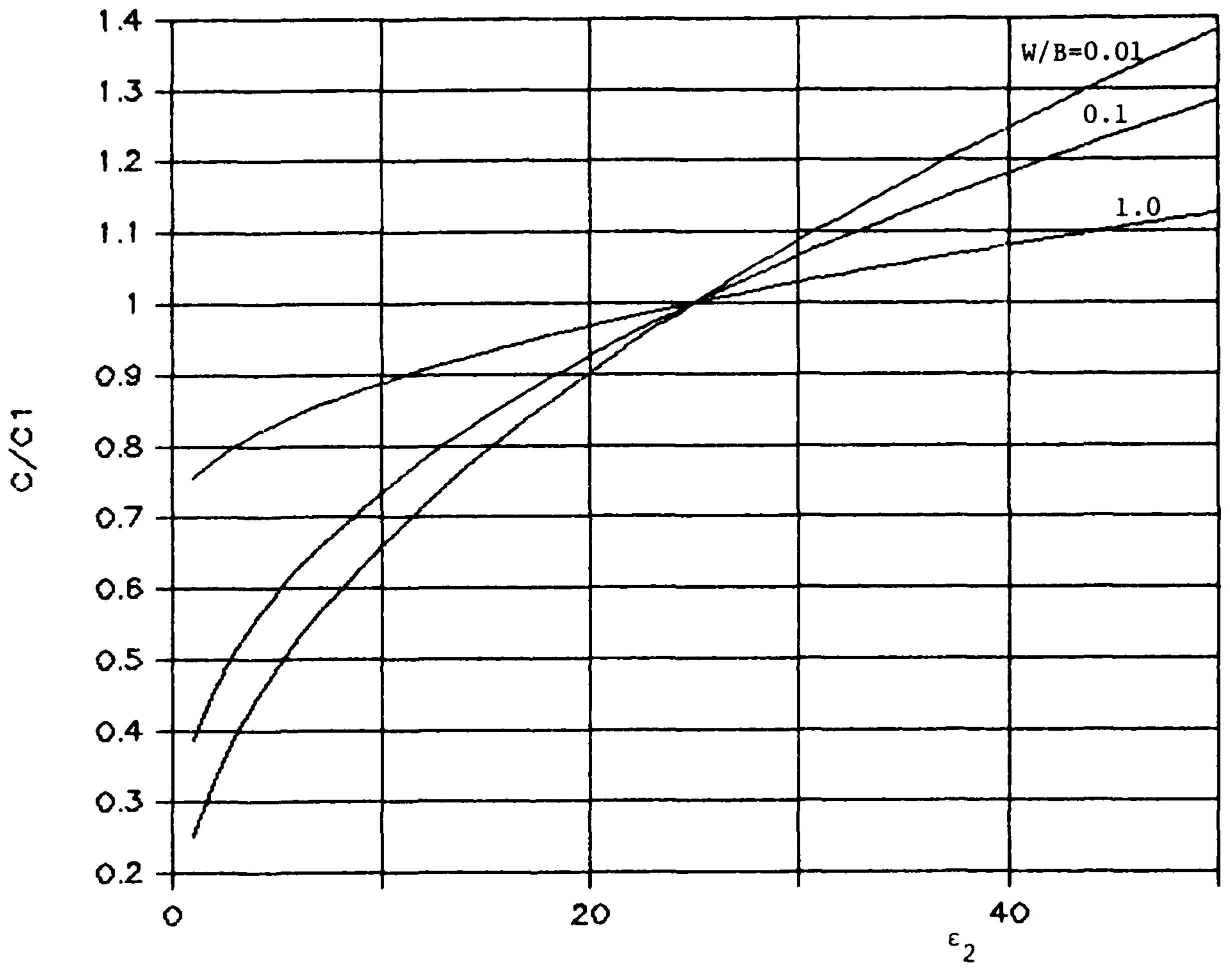

Figure 6. The normalized line capacitance, C/C1, versus ε_2 for the strip line of Figure 3, in which $\varepsilon_1 = 25$.

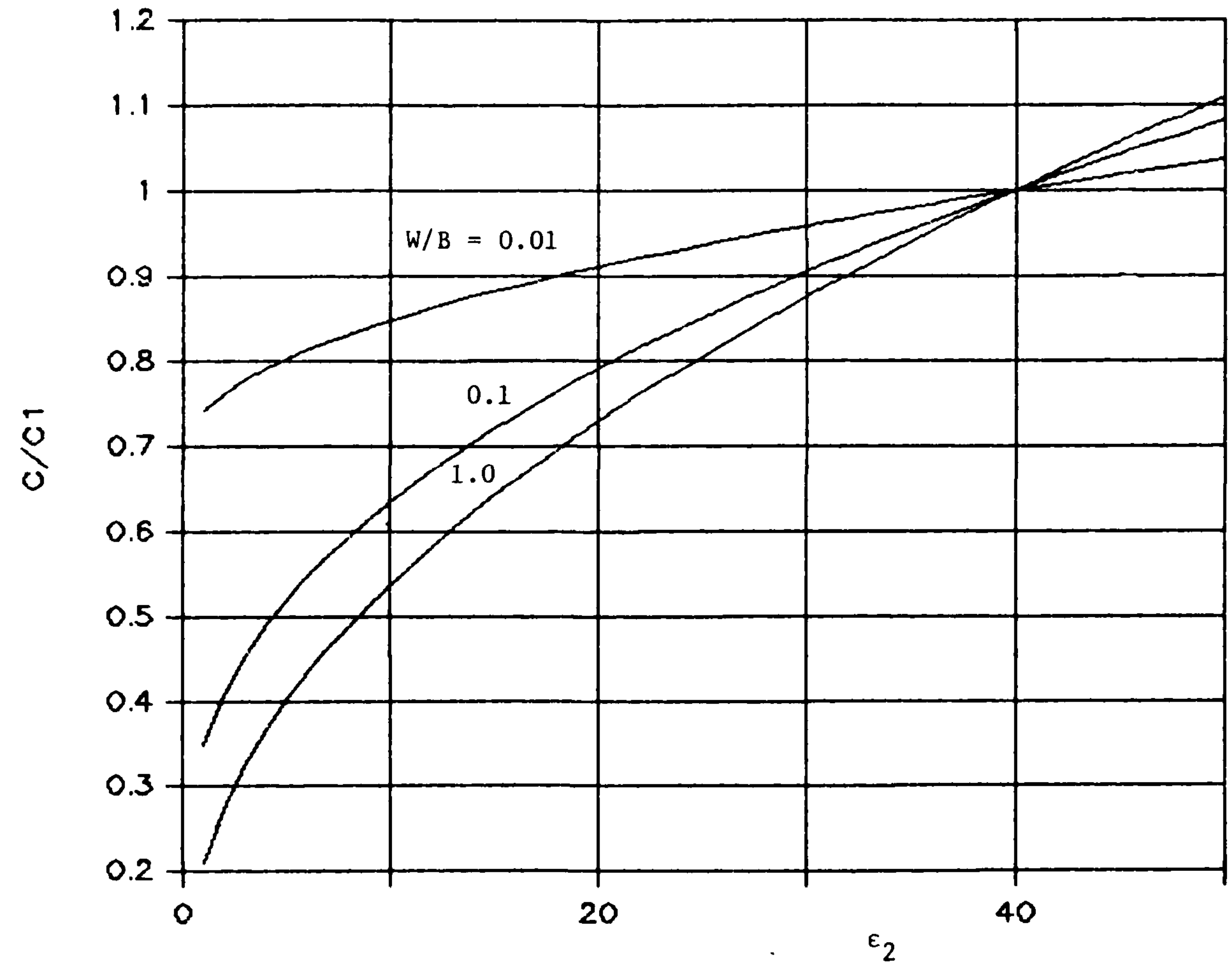

Figure 7. The normalized line capacitance, C/C1, for the strip line of Figure 3, in which $\varepsilon_1 = 40$.

4. Analysis of the Scattering Mode

The scattering mode, in which the microstrip antenna is illuminated by an incident plane wave, can be computed with first-order accuracy by neglecting the presence of the microstrip patch and feed line. The problem is then reduced to the case of reflection from an anisotropic dielectric sheet on a conducting plane, for which direct solution for the half-space problem is available, and the edge effect can often be neglected.

5. Conclusions and Recommendations

We have demonstrated the potential usefulness of microstrip antennas with exotic substrates in applications to advanced aircraft. These exotic substrates are becoming more and more realistic as manufacturing technology progresses. Some experimental work is needed to guide the development of design and analysis techniques.

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