# Propagation Properties of Ferromagnetic Insular Guide

JIQING XIA, STUDENT MEMBER, IEEE, PETER P. TOULIOS, MEMBER, IEEE, AND CARMINE VITTORIA, SENIOR MEMBER, IEEE

*Abstract* —**In order to realize new ferrite planar devices for applications at millimeter-wavelength frequencies, we have considered in our development of nonreciprocal ferrite devices schemes of incorporating ferromagnetic layers in dielectric insular guide geometries. Our research program involves both the calculation and the measurement of device characteristics. For the calculational part a method of effective permeability is introduced to characterize the ferrite material in which the permeability is a tensor. The propagation properties of the insular guide were calculated by using a single-mode approach. Our calculational results of dispersion, dielectric, and conductivity losses show resonant behavior with the application of a magnetic bias field for a guide configuration in which the ferrite replaces the insular dielectric. Ferrite phase shifters, filters, isolators, and circulators are potential applications of this guide configuration. For the experimental part wave dispersion and attenuation were measured in a purely dielectric insular guide from 26.5 to 40 GHz. In addition to these experiments wave attenuation was measured as a function of magnetic bias fields for the case where a hexagonal ferrite platelet was mounted on the ground plane near the insular guide. General agreement is found between calculations and measured attenuation.**

## I. Introduction

MILLIMETER-WAVE systems have wide applications in radar, communications, radiometry, and instrumentation. Presently dielectric waveguides and microstrip lines are available for millimeter-wave integrated circuit applications, and these structures are reciprocal. For many applications at high frequencies there is a need for nonreciprocal ferrite devices. In order to design new devices with a nonreciprocal property or frequency selectivity, such as isolators, filters, and directional couplers, new guide structures as well as new ferrite materials are needed to meet the device development requirements. In this research project, a hexagonal ferrite material has been chosen as a potential candidate for new magnetic materials operational at high frequencies. In particular the hexagonal ferrite layer is incorporated in the dielectric insular guide structure. We find that this guide structure offers an ideal structure for the incorporation of ferrite layers and slabs, for example.

Dielectric insulated image guide (insular guide), Fig. 1(a), is one of many planar devices developed for millimeter-wave integrated circuits. The main advantage of this particular guide configuration over other image guide structures is that it reduces conductivity propagation losses, since the wave propagates mostly in the dielectric guide "insulated" from the conducting plane (see Fig. 1(a)). In addition to conductivity losses there are dielectric and radiation losses. Radiation losses are usually minimal if there are no sharp bends in the guide structure or if propagation of electromagnetic waves in the fundamental mode is assumed in the guide.

Dielectric losses scale nearly linearly with frequency so that low-loss tangent materials are desirable in the "insulated" dielectric. In considering a practical guide structure we believe that the insular dielectric guide is ideal for reducing conductivity propagation losses but not the dielectric losses. Although the insular guide geometry involves higher fabrication cost over a single dielectric guide, the advantages of this guide far outweigh the disadvantages. For example, this geometry allows for the inclusion of ferrite slabs in the insular region where the microwave magnetic field is a maximum. This means that maximum coupling is possible between the ferrite and the device. Throughout our theoretical development an insulating ferromagnetic layer with low loss tangent is assumed. Thus, when applying a dc magnetic field to this structure, ferromagnetic resonance will occur in the layer. Hence, frequency selectivity is possible in the ferrite device.

The propagation properties of dielectric image guide have been calculated in our previous work [1]. There are now exact and approximate solutions of propagation characteristics in dielectric image guides. Exact solutions require significant computation time and are impractical in engineering design considerations. There are various methods used to approximate wave propagation characteristics in dielectric image guides [2]–[9]. Among these approximate methods the single-mode approach is a relatively simple method with good accuracy in comparison to the so-called exact method [4].

Previous calculations involving ferrite materials include slot devices with magnetic substrates. These devices have

Manuscript received October 17, 1988; revised April 20, 1989. This work was supported in part by the U.S. Office of Naval Research under Contracts N00014-86-K-0534 and N00014-88-k-0185.

J. Xia and C. Vittoria are with the Department of Electrical Engineering, Northeastern University, Boston, MA 02115.

P. P. Toulios is with the Epsilon–Lambda Electronics Corporation, 427 Stevens St., Geneva, IL 60134.

IEEE Log Number 8929898.

0018-9480/89/1000-1547$01.00 

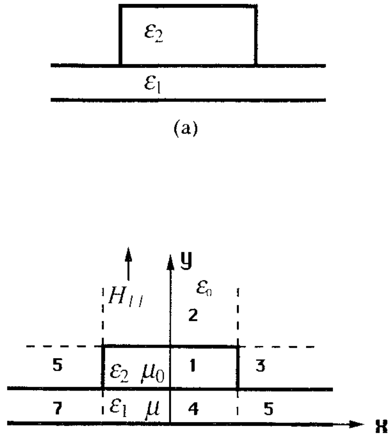


Fig. 1. (a) Geometrical configurations of insular image guide. (b) Ferromagnetic insular guide, where the insulating dielectric layer of the insular guide is replaced by a ferromagnetic layer. The magnetic field is applied normal to the ground plane.

been analyzed by using a full-wave analysis based on a Green's function [10]. To provide a simple and direct solution of wave propagation characteristics in a ferrite insular guide, we introduce an effective permeability method in the analysis. Using a previous calculational approach [1], phase propagation, attenuation, and other propagation characteristics have been calculated for the ferrite insular guide geometry shown in Fig. 1(b). We find that it may be possible to magnetically bias the ferrite layer to ferromagnetic resonance (FMR) and still maintain wave propagation in the $E^y$ fundamental mode. Naturally, propagtion losses are maximum at FMR (6 dB/cm). In addition to magnetic losses we find that dielectric losses scale linearly with frequency at high frequencies. This result is remarkable in view of the fact that there is more than one layer in the insular guide structure. A linear dependence of dielectric losses has been calculated before for single-layer guiding structures only.

We have tested an insular guided structure in which a single-crystal hexagonal ferrite material is placed on the ground plane next to the insular layer. The coupling is maximized when the ferrite platelet is placed on the ground plane in physical contact with the insular guide. Minimum coupling is realized when the ferrite is placed on top of the guide structure since the microwave magnetic field is also minimum there. In summary, we believe that these results will form the theoretical basis for the development of new ferrite insular image guide devices, such as phase shifters, modulators, isolators, narrow-band filters, and nonreciprocal couplers at millimeter-wave frequencies.

In Section II, the effective permeability method is discussed and used as the fundamental theory to characterize the propagation properties of the ferrite insular guide. Another fundamental method in this analysis, the single-mode method, is discussed in Section III. Section IV presents the calculation results of both ferromagnetic insular guide and dielectric insular guide. Experimental results are shown and discussed in Section V.

## II. Effective Permeability of Ferromagnetic Material

A ferrite insular guide (FIG) is a novel device in which a ferromagnetic layer replaces the insulating dielectric of an insular guide. Strictly speaking, a ferrimagnetic material will do just as well. Fig. 1(b) shows the cross section of this guide structure.

In a ferromagnetic material, Maxwell's equations may be written in the time harmonic forms

$$\nabla \times \vec{e} = -j\omega\mu_0\vec{\mu}\cdot\vec{h} \tag{1}$$

$$\nabla \times \vec{h} = j\omega\epsilon_0\epsilon\vec{e} \tag{2}$$

$$\nabla\cdot\vec{e} = 0 \tag{3}$$

$$\nabla\cdot\left(\mu_0\vec{\mu}\cdot\vec{h}\right) = 0 \tag{4}$$

where $\vec{e}$ and $\vec{h}$ are the electric and magnetic field intensities, respectively; $\omega = 2\pi f$; $f$ is the frequency; $\mu$ is the relative permeability tensor of the ferrite; $\epsilon$ is the relative permittivity of the material; and $\mu_0$ and $\epsilon_0$ are the corresponding values for $\mu$ and $\epsilon$ in free space.

As in previous approximate methods of solution there are two fundamental modes of propagation in a rectangular dielectric image guide: $E^y$ and $E^x$ modes. Throughout this paper we will assume an $E^y$ mode of propagation in the structure, which implies

$$h_y = 0. \tag{5}$$

The $\vec{\mu}$ is defined as

$$[\mu] = \begin{bmatrix} \mu_{xx} & 0 & -j\mu_{xz} \\ 0 & 1 & 0 \\ +j\mu_{zx} & 0 & \mu_{zz} \end{bmatrix} \tag{6}$$

where $\mu_{xx}$, $\mu_{xz}$, $\mu_{zx}$, and $\mu_{zz}$ are the matrix elements, which can be obtained from the equation of motion for magnetization using Gilbert's damping term, for example.

For the case of an applied magnetic field normal to the plane of a ferromagnetic layer and along the $y$ axis we write

$$\mu_{xx} = \mu_{zz} = 1 + \frac{4\pi M_s\gamma/f_0 + j4\pi M_s\gamma f\Delta f/2f_0^3}{1-(f/f_0)^2 + jf\Delta f/f_0^2} \tag{7}$$

$$\mu_{xz} = \mu_{zx} = \frac{4\pi M_s\gamma f/f_0^2}{1-(f/f_0)^2 + jf\Delta f/f_0^2} \tag{8}$$

where $f$ is the microwave frequency, $f_0$ is the FMR frequency, $\Delta f$ is the FMR frequency line width, and $\gamma$ is the gyromagnetic ratio of the ferrite. In most oxide ferrites $\gamma = 2.8$ MHZ/Oe. The quantity $4\pi M_s$ is the saturation magnetization [11].

We assume that the ferrite is hexagonal in crystal structure with the $C$ axis perpendicular to the ferrite layer plane. When a dc magnetic field is applied parallel to the

$C$ axis, the FMR condition can be derived by minimizing the free energy of the ferrite material. The following is the FMR condition for single-domain excitation:

$$\frac{f_0}{\gamma} = H_{\|} + H_A - 4\pi M_s, \qquad H_{\|} \geqslant 4\pi M. \tag{9}$$

Here $H_A$ is the anisotropy field, and $H_{\|}$ is the external dc magnetic field parallel to the $C$ axis. Equation (9) is valid for external fields greater than $4\pi M$.

From (9) one can see that the resonant frequency $f_0$ is determined by the external magnetic field $H_{\|}$. This means that the resonant frequency can be tuned by changing the external magnetic field. In the case of electromagnets, the resonant frequency can be changed simply by adjusting the driving current in the electromagnets.

For the propagtion of the $E^y$ mode, based on the permeability tensor we have obtained, the propagation constant can be derived as follows. Combining (1) and (2), we have

$$\nabla(\nabla \cdot \vec{h}) - \nabla^2 \vec{h} = \omega^2 \mu_0 \epsilon_0 \epsilon \vec{\mu} \cdot \vec{h}. \tag{10}$$

Expanding $\vec{\mu}$ in terms of its matrix elements we get

$$\vec{\mu} \cdot \vec{h} = (\mu_{xx} h_x - j\mu_{xz} h_z)\vec{a}_x + (\mu_{xx} h_z + j\mu_{xz} h_x)\vec{a}_z. \tag{11}$$

Substituting (11) into (10), we obtain equations for $h_x$ and $h_z$ only:

$$[\omega^2 \mu_0 \epsilon_0 \epsilon \mu_{xx} + (k_x^2 - k^2)] h_x + [k_x k_z - j\mu_{xz} \omega^2 \mu_0 \epsilon_0 \epsilon] h_z = 0 \tag{12}$$

$$[k_x k_z + j\mu_{xz} \omega^2 \mu_0 \epsilon_0 \epsilon] h_x + [\omega^2 \mu_0 \epsilon_0 \epsilon \mu_{xx} + (k_z^2 - k^2)] h_z = 0. \tag{13}$$

It is noted that there is no component of $\vec{h}$ in the $y$ direction. This is a result of our assumption that propagation occurs in the $E^y$ mode. For a nontrivial solution of $h_x$ and $h_z$, the determinant of the parameter matrix equation of the above equations is set to zero. Approximating $k_y = 0$ ($k_y^2 \ll k^2$), we obtain the following relation:

$$k^2 = \frac{\mu_{xx}^2 - \mu_{xz}^2}{\mu_{xx}} \omega^2 \mu_0 \epsilon_0 \epsilon. \tag{14}$$

Define

$$\mu_{\text{eff}} = [\mu_{xx}^2 - \mu_{xz}^2]/\mu_{xx} \tag{15}$$

as an effective permeability, then

$$k^2 = \omega^2 \mu_0 \epsilon_0 \epsilon \mu_{\text{eff}}. \tag{16}$$

Hence, the propagation constant of the ferrite can be expressed in terms of a simple effective permeability constant. This is the so-called effective permeability method, which will be used throughout our analysis.

As an example, Fig. 2 shows $\mu_{\text{eff}}$ as a function of frequency with resonant frequency $f_0 = 60$ GHz, $4\pi M_s = 4$ kgauss, and $\Delta f/\gamma = 1$ kOe. The values of $4\pi M$ and $\Delta f$ are typical of hexagonal ferrite polycrystalline materials.

It is pointed out that if we had chosen the magnetic field to be applied in the ferrite layer plane, the assumption $h_y = 0$ would not be valid. We would have to consider an admixture of $E^y$ and $E^x$ modes of propagation in our analysis.

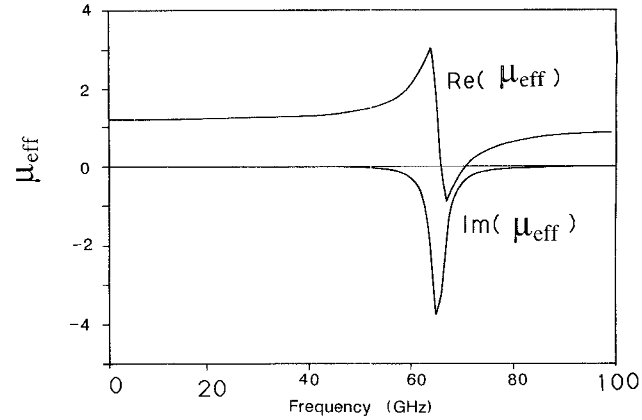


Fig. 2. The effective permeability, $\mu_{\text{eff}}$, is plotted as a function of frequency. For definition of $\mu_{\text{eff}}$, see text.

## III. Wave Propagation in Ferrite Insular Guide

The single-mode method assumes that a single fundamental mode, the $E^y$ mode, can propagate in the dielectric waveguide of Fig. 1. Field distributions are assumed in different regions. These are regions from 1 to 7 in Fig. 1(b). The expressions of the field assumption are included in the Appendix. For the $E_y^{11}$ mode, the $h_x$ component is maximum at the ferrite layer. By matching boundary conditions of electric and magnetic fields at the boundaries of the dielectric guide (which is region 1 in Fig. 1(b)), the characteristic equations relating the propagation constants may be obtained [1], [2], [3], [7].

For $E^y$ modes, the characteristic equation of the transverse propagation constant $k_y$ may be determined upon application of boundary conditions [7], [12], whereby we require tangential electric and magnetic fields at $y = 0$, $h$, and $h + b$ to be continuous across boundaries:

$$\begin{aligned} &\epsilon_1 \epsilon_2 k_{y0} k_y \cosh(k_{y1} h) \cos(k_y b) \\ &\quad + \epsilon_2^2 k_{y0} k_{y1} \sinh(k_{y1} h) \sin(k_y b) \\ &\quad - \epsilon_1 k_y^2 \cosh(k_{y1} h) \sin(k_y b) \\ &\quad + \epsilon_2 k_{y1} k_y \sinh(k_{y1} h) \cos(k_y b) = 0 \end{aligned} \tag{17}$$

where $\epsilon_1$ and $\epsilon_2$ are the relative dielectric constants of the insulating layer and the dielectric guide, respectively.

Our calculations are of sufficient generality so that both gyromagnetic and purely dielectric effects may be calculated by simply setting $\mu_{\text{eff}} \neq 1$ and $\mu_{\text{eff}} = 1$, respectively. For the case in which the insular guide is replaced by a ferrite $\epsilon_1$ may be written as $\epsilon_1 \mu_{\text{eff}}$. Otherwise, $\mu_{\text{eff}} = 1$, as shown in Fig. 1(a). Here $b$ and $h$ are the thicknesses of the guide and the layer next to the conductive plane, respectively (see Fig. 1(b)). In addition, we have

$$k_{y1}^2 = (\epsilon_2 - \epsilon_1) k_0^2 - k_y^2 \tag{18}$$

$$k_{y0}^2 = (\epsilon_2 - 1) k_0^2 - k_y^2. \tag{19}$$

The characteristic equation of the other transverse propagation constant, $k_x$, is again obtained using Marcatili's method of calculation [3]:

$$\tan(k_x a) = \frac{k_{x0}}{k_x} \tag{20}$$

with

$$k_{x0}^2 = (\epsilon_2 - 1)k_0^2 - k_x^2 \tag{21}$$

where $2a$ is the width of the dielectric guide.

Interestingly, there is no explicit dependence of $k_x$ on $\mu_{\text{eff}}$. This is due to the fact that the boundary conditions for obtaining (21) do not involve fields in the ferrite material. Only the fields in regions 1, 3, and 5 in Fig. 1(b) are used (see Fig. 1(b)).

The propagation constant in the propagation direction, $k_z$, may be obtained after solving (17) and (20) for $k_x$ and $k_y$ so that we have

$$k_z^2 = \epsilon_2 k_0^2 - k_x^2 - k_y^2 \tag{22}$$

where $k_0 = 2\pi/\lambda_0$ and $\lambda_0 = c/f$, $c$ being the speed of light in free space. The guide wavelength of the insular guide is defined as

$$\lambda_g = 2\pi/\text{Re}(k_z). \tag{23}$$

This $\lambda_g$ describes the spatial period of the propagating wave inside the guide. The phase velocity of the wave is determined by the guide wavelength. Since the propagation constant in the ferrite medium includes an effective permeability which has a resonance near that FMR frequency, the guide wavelength will also have resonance at the frequency.

By definition, the attenuation due to both dielectrics in the insular guide may be calculable from

$$\alpha_d = -\text{Im}(k_z). \tag{24}$$

We find that for a multiple-dielectric-layer guided structure it is convenient to define attenuation due to complex dielectric constants in each layer by (24).

Since there is only one conductive surface, previous analyses [2] of conductivity propagation losses are appropriate to the case considered here. Ferrite layers in the insular region do not alter the analysis. Hence, we may define the conductivity loss parameter as before as

$$\alpha_c = W_1/2W_t \tag{25}$$

where $W_1$ is the loss per unit length along the propagation direction due to the finite conductivity of the ground plane:

$$W_1 = \frac{R_s}{2}\int_0^1 dz \int_{-\infty}^{+\infty} \left[|H_x|^2 + |H_z|^2\right] dx. \tag{26}$$

Here $R_s$ is the surface impedance of the ground plane [2]:

$$R_s = \sqrt{\omega\mu_0/2\sigma_c}\,. \tag{27}$$

$H_x$ and $H_z$ are the fields at the $y = 0$ surface in regions 4, 6, and 7 (see Fig. 1). $W_t$ is the total propagation power [1], [2]:

$$W_t = \int_0^{+\infty} dy \int_{-\infty}^{+\infty} \left[\frac{1}{2}\vec{E}_y \times \vec{H}_x^*\right] dx. \tag{28}$$

$W_1$ and $W_t$ are calculated by using the field distribution in different regions. The formulas are shown in the Appendix.

The insertion of the ferrite layer affects the fields in regions 4, 6, and 7 in Fig. 1(b). The resonant behavior of the $\mu_{\text{eff}}$ introduces a resonance into the conductivity and dielectric losses.

## IV. Calculational Results

The computational work involved the solutions of the characteristic equations for the transverse propagation constants $k_x$ and $k_y$, equations (17) and (20). These are complex transcendental equations, which can have many roots. Each root of the characteristic equation indicates a propagating mode in the waveguide. The main source of difficulty in this calculation is searching for the proper roots of $k_x$ and $k_y$. To find a solution of a specific mode, e.g., the fundamental $E_y^{11}$ mode, one has to properly choose the initial values of $k_x$ and $k_y$ in the iterative computer procedure that we have used to ensure the right branch of the root. A graphic searching technique is found to be useful in finding the initial values. Plotting the curve of the equation indicates the rough position of each mode on the graph. After $k_x$ and $k_y$ are obtained, the guide wavelength $\lambda_g$, dielectric loss $\alpha_d$, and conductivity loss $\alpha_c$ are calculated using the above definition.

For an insular guide loaded with a ferrite material we have $\mu_{\text{eff}} \neq 1$.

### A. Ferromagnetic Insular Guide

For this case one would expect the ferrite material to couple with the microwave fields in the region where the magnetic microwave field is maximum. This region is located in the insulating layer (see Fig. 1(b)). The propagation properties of a ferromagnetic insular guide have been calculated using the single-mode method. For this calculation we chose $f_0 = 60$ GHz, and $\mu_{\text{eff}}$ is obtained from Fig. 2. $\lambda_0/\lambda_g$ is plotted in Fig. 3. This curve is referred to as the dispersion of the waveguide or the velocity ratio by certain authors [2]. The quantity $\lambda_0/\lambda_g$ changes by about 10 percent near $f_0$. This means that at FMR, the guide wavelength is smaller and then larger than usual and $\lambda_0/\lambda_g$ is clearly tunable by changing the external magnetic field. This result implies that the phase of the signal may be modulated using ferrite materials.

Figs. 4 and 5 show the corresponding dielectric loss and conductivity loss of the FIG structure. The attenuation reaches a maximum at frequencies near FMR, since the guide structure is strongly coupled to the ferrite. The attenuation near FMR is approximately 20 dB higher than

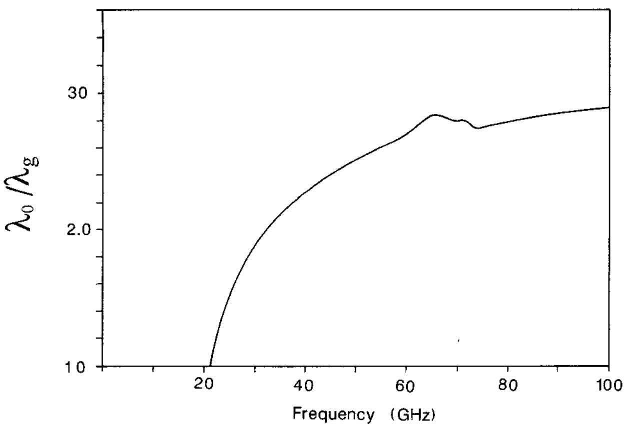


Fig. 3. Guide wavelength, $\lambda_g$, is plotted as a function of frequency for a ferrite-loaded insular guide.

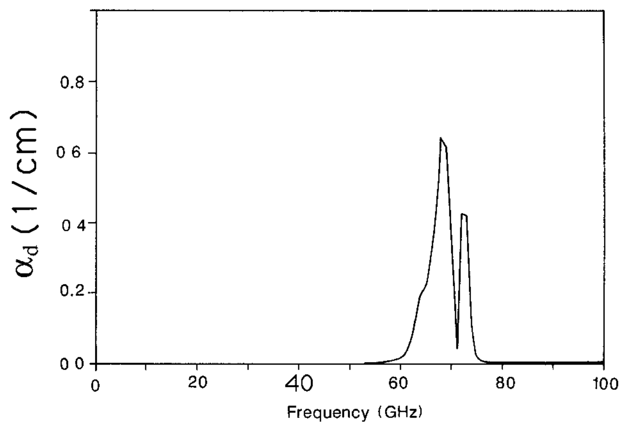


Fig. 4. Dielectric loss constant plotted as a function of frequency assuming the $\mu_{\text{eff}}$ of Fig. 2.

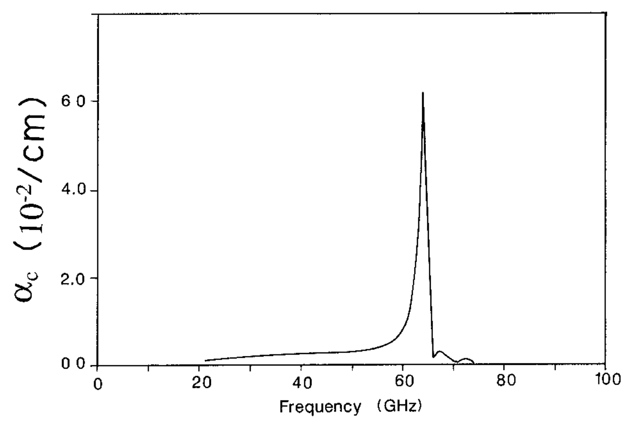


Fig. 5. Conductivity loss constant plotted as a function of frequency assuming the $\mu_{\text{eff}}$ of Fig. 2.

usual. The results show a very sharp attenuation peak which has a possible application for a narrow-band filter. Although there is a strong interaction between the ferrite material and the microwave fields in the guide, the fundamental mode of propagation is still the $E^y$ mode. We suspect that if the external field is applied parallel to the ground plane, one could not maintain $h_y = 0$ as required

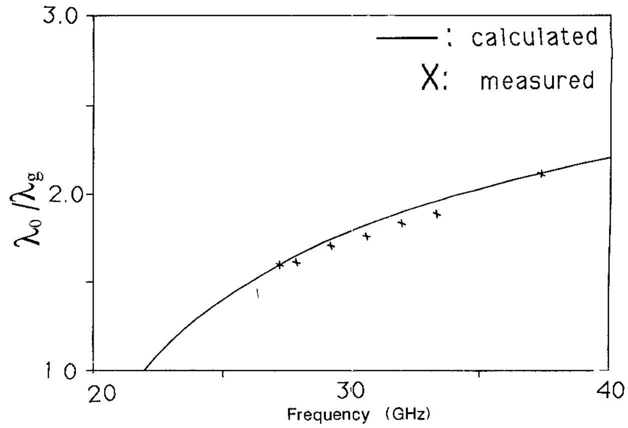


Fig. 6. Measured and calculated dispersions as a function of frequency for a purely dielectric insular guide (see text for assumed parameters).

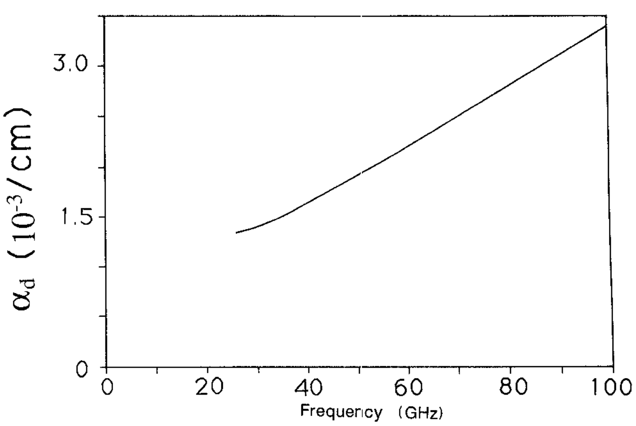


Fig. 7. Dielectric loss constant plotted as a function of frequency for a purely dielectric insular guide.

for this fundamental mode of propagation. This method of excitation enhances the ferrite interaction with the device without affecting the character of the fundamental mode.

### *B. Dielectric Insular Guide ($\mu_{\text{eff}} = 1$)*

The dispersion curve of the insular dielectric guide in Fig. 1(a) is calculated when $\mu_{\text{eff}} = 1$ (shown in Fig. 6). The parameters $\epsilon_1$, $\epsilon_2$, $a$, $b$, and $h$ and the conductivity of the ground plane, $\sigma_c$, are chosen from a real device fabricated by Epsilon-Lambda ($\epsilon-\lambda$) Electronic Corporation (Geneva, IL) where $a = 0.095$ cm, $b = 0.170$ cm, $h = 0.015$ cm, $\epsilon_1 = 2.25(1 - j1\times10^{-4})$, and $\epsilon_2 = 9.5(1 - j1\times10^{-4})$. The length of the dielectric waveguide is 11.2 cm. The insular guide operated from 26.5 to 40 GHz. As shown in Fig. 6, this insular guide has a cutoff frequency at 22 GHz. The cutoff concept here has the following meaning: $k_{y0}$ or $k_{x0}$ becomes mainly imaginary, which means that the power will radiate outside the dielectric guide instead of propagating inside. This implies that $k_z$ becomes mainly imaginary so that the wave will attenuate along the propagation direction.

Fig. 7 shows the dielectric loss versus frequency. It can be seen that the dielectric attenuation increases as the

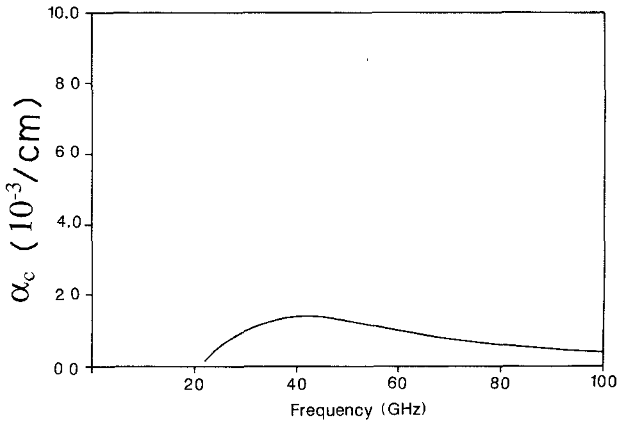


Fig. 8. Conductivity loss constant plotted as a function of frequency for a purely dielectric insular guide.

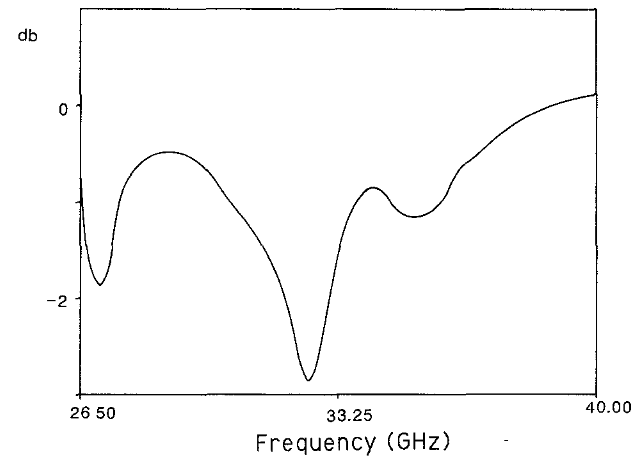


Fig. 9. Output voltage of insular guide with a hexagonal ferrite platelet measured as a function of frequency. The minimum in transmission corresponds to ferrimagnetic resonance of the ferrite.

frequency increases. At high frequencies this relationship is approximately linear. The conductivity loss is shown as a function of frequency in Fig. 8. Because of the large conductivity $\sigma_c$ ($=6.17\times10^5$) of the ground plane, $\alpha_c$ is quite small in this case ($\approx 1\times10^1$/cm). Thus, for most practical insular guides, the dielectric loss is the dominant source of propagation losses. It is clear that all types of loss are additive so that a FIG structure would never be able to reduce conductivity or dielectric losses.

It should be pointed out that the single-mode method used in this calculation is an approximation. For more complicated structures and more accurate results the exact method should be used though it will involve significantly more computational complexities.

## V. Experimental Results

Experiments have been performed on the insular guide from 26.5 to 40 GHz. A simple electric wire probe was used to measure the standing wave ratio of the insular guide. The center conductor of a coaxial line was linearly driven by a micrometer directly over the insular guide. When translating the electric probe along the dielectric guide, we were able to measure reliably the voltage or electric field as a function of the linear displacement of the probe. This technique is very useful in measuring the position of the minima and maxima of the signal voltage for determining the guide wavelength [7].

Results of these measurements are shown in Fig. 6. It shows that the calculation results are in good agreement with the experimental results.

However, we have not been able to develop a technique to separate dielectric loss, conductivity loss, and radiation loss from each other, although total insertion losses were measured by conventional techniques (see Fig. 9).

A single-crystal hexagonal ferrite platelet was mounted on the same insular guide structure tested above. Magnetic parameters of this hexagonal ferrite platelet were measured by a vibrating sample magnetometer (VSM) and FMR techniques. In summary the magnetic parameters of this ferrite were a saturation magnetization $4\pi M_s = 3$ kG, a uniaxial anisotropy field $H_A = 9.2$ kOe, and $g = 1.9$. For high-frequency applications hexagonal ferrites are preferred over spinel ferrites, since they require lesser magnetic biasing fields in comparison to spinel ferrites. Using a hexagonal ferrite platelet with $C$ axis normal to the platelet plane we were able to tune it to FMR with fields less than 3 kOe and dramatically affect wave propagation in an insular guide. In some applications where magnetic losses are to be avoided it may not be necessary to tune the ferrite to FMR. For a spinel ferrite we would need biasing fields on the order of 10–15 kOe to tune it to FMR at 26–30 GHz.

Hexagonal ferrite platelets characterized by us were inserted in the insular guided structure. The platelet was inserted next to the dielectric guide and on the ground plane in order to couple to fringing microwave magnetic fields. The coupling was found to be small if the ferrite was placed on top of the guide. The external field was applied normal to the plane of the platelet in order to maintain the integrity of the fundamental $E^y$ mode of propagation.

Although this geometry is not identical to that of Fig. 1, we believe that this geometry is useful in the design of nonreciprocal directional couplers, filters, circulators, isolators, etc. We applied a dc magnetic field, ranging from 0 to 20 kOe, to the hexagonal ferrite, and absorption resonances in the propagation of power were measured (see Fig. 9). The resonant frequency was tunable by the external magnetic field. This, we believe, supports the notion that ferromagnetic insular guides can serve as a basis for new devices with frequency selectivity.

For $0 < H < 4\pi M_s$, we surmise magnetic domains are nucleated in the hexagonal ferrite. Hence, multidomain resonances may be possible [13] at low fields. Nevertheless, the device is operational at high frequencies and at rela-

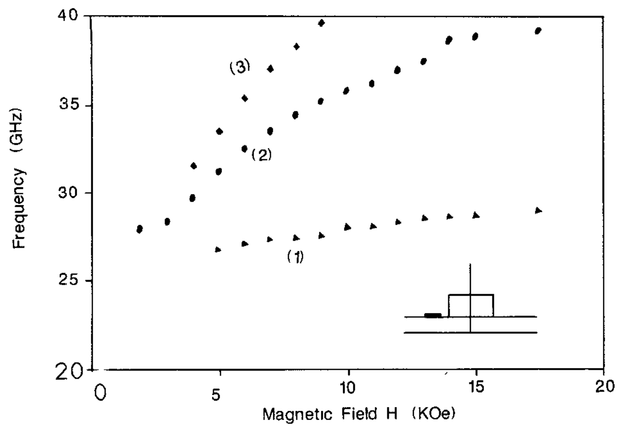


Fig. 10. Ferrimagnetic resonance frequency of the ferrite (measured in Fig. 9) is plotted as a function of bias magnetic field. Curve *3* is the dispersion of the main or uniform precession of ferrimagnetic resonance ($g = 1.9$). Curves *1* and *2* are subsidiary resonance modes and are attributed magnetostatic mode excitations.

tively low fields. As shown in curve *(3)* of Fig. 10, for $H > 4\pi M_s$ the resonance frequency scales linearly with frequency, as one would expect for single-domain excitations in ferrites (see (9)). However, there are other resonances in the guide structures (see Fig. 10). We believe that these subsidiary resonances are magnetostatic mode excitations, since they also scale with frequency and are characterized by integers.

## VI. Conclusions

A new structure involving hexagonal ferrites and insular guide is proposed for device applications at millimeter-wavelength frequencies. An effective permeability method is used to analyze the propagation properties of the ferrite insular guide. Dispersion, dielectric loss, and conductivity loss of both dielectric insular and ferromagnetic insular guides have been calculated by using the single-mode approximation. Experimental measurements were performed from 26.5 to 40 GHz on a dielectric insular guide. A hexagonal ferrite is incorporated in the insular guide for measuring total insertion losses of the device. General agreement was found between measured and calculated values of insertion loss.

The results show that this ferromagnetic insular guide structure has resonant behavior in its propagation properties and that the resonant frequency is tunable by the external magnetic field. For new ferrite device designs this structure is useful, since the $E^y$ mode of propagation is maintained throughout the analysis.

In this paper we have developed a scheme by which ferrite can be incorporated in dielectric image guides without affecting the purity of the fundamental mode of propagation. We believe that this scheme is practical for two reasons: (1) The ferrite can be easily incorporated in the structure or near the guided structure so that fabrication costs will be minimal; (2) hexagonal ferrites require small biasing fields, which can readily be attained along with the permanent magnet materials now available.

## Appendix

### A. Field Distribution Assumptions

In Fig. 1(b) the RF magnetic field components $H_x$ in different regions are assumed in the following form for an $E^y$ mode of propagation:

$$H_x = \begin{cases} M_1 \cos(k_x x)\cos(k_y y + \beta) & \text{region 1} \\ M_2 \cos(k_x x)\exp[-k_{y0}(y - h - b)] & \text{region 2} \\ M_3 \exp[-k_{x0}(x - a)]\cos(k_y y + \beta) & \text{region 3} \\ M_4 \cos(k_x x)\cos(k_{y1} y) & \text{region 4} \\ M_5 \exp[k_{x0}(x + a)]\cos(k_y y + \beta) & \text{region 5} \\ M_6 \exp[-k_{x0}(x - a)]\cos(k_{y1} y) & \text{region 6} \\ M_7 \exp[k_{x0}(x + a)]\cos(k_{y1} y) & \text{region 7} \end{cases}$$

where the $M$'s are constants.

### B. $W_l$ and $W_t$ Expressions

The $W_1$ integral (26) is calculated in regions 4, 6, and 7:

$$W_1 = I_1 + I_2$$

$$I_1 = \frac{aR_s M_4^2}{2}\left[1 + \frac{\sin(2k_x a)}{2k_x a} + \frac{\cos^2(k_x a)}{k_{x0} a}\right]$$

$$I_2 = \frac{aR_s M_4^2}{2}\left\{\left(\frac{k_x}{k_z}\right)^2\left[1 - \frac{\sin(2k_x a)}{2k_x a}\right] + \left(\frac{k_{x0}}{k_z}\right)^2 \frac{\cos^2(k_x a)}{k_{x0} a}\right\}.$$

$W_t$ is calculated in regions 1 to 7. Using symmetry,

$$W_t = P_1 + P_2 + 2P_3 + 2P_6 + P_4$$

$$P_1 = \frac{abM_1^2}{4} z_1 \left[1 + \frac{\sin(2k_x a)}{2k_x a}\right] \cdot \left[1 + \frac{1}{k_y b}\sin(k_y b)\cos(2k_y h + k_y b + 2\beta)\right]$$

where

$$\beta = \tan^{-1}\left(\frac{\epsilon_2 k_{y0}}{k_y}\right) - k_y(h + b)$$

$$P_2 = \frac{z_2 M_2^2 a}{4k_{y0}}\left[1 + \frac{\sin(2k_x a)}{2k_x a}\right]$$

$$P_3 = \frac{z_3 M_3^2}{4k_{x0}}\left[\frac{b}{2} + \frac{1}{2k_y}\sin(k_y b)\cos(2k_y h + k_y b + 2\beta)\right]$$

$$P_4 = \frac{z_4 M_4^2 ah}{4}\left[1 + \frac{\sin(2k_{y1} h)}{2k_{y1} h}\right]\left[1 + \frac{\sin(2k_x a)}{2k_x a}\right]$$

$$P_6 = \frac{z_6 M_6^2 h}{8k_{x0}}\left[1 + \frac{\sin(2k_{y1} h)}{2k_{y1} h}\right].$$

The $z$'s are wave impedances:

$$z_1 = \frac{\epsilon_2 k_0^2 - k_y^2}{\omega \epsilon_0 \epsilon_2 k_z}$$

$$z_2 = \frac{k_0^2 + k_{y0}^2}{\omega \epsilon_0 k_z}$$

$$z_3 = \frac{k_0^2 - k_y^2}{\omega \epsilon_0 k_z}$$

$$z_6 = z_4 = \frac{\epsilon_1 k_0^2 - k_{y1}^2}{\omega \epsilon_1 \epsilon_0 k_z}.$$

$M_1$, $M_2$, $M_3$, and $M_6$ may be expressed in terms of $M_4$. Since $\alpha_c$ involves the ratio of $W_1$ and $W_t$, $M_4$ does not show up in the final expression for $\alpha_c$.


## Acknowledgment

The authors wish to express their appreciation to Dr. S. W. McKnight for his very helpful discussions.


## References


[1] J. Xia, S. W. McKnight, and C. Vittoria, "Propagation losses in dielectric image guides," *IEEE Trans. Microwave Theory Tech.*, vol. 36, pp. 155–158, Jan. 1988.

[2] P. P. Toulios and R. M. Knox, "Rectangular dielectric image lines for millimeter wave integrated circuits," in *Proc. Wescon Conf.* (Los Angeles, CA), 1970. pp. 1–10.

[3] E. A. J. Marcatili, "Dielectric rectangular waveguide and direction coupler for integrated optics," *Bell Syst. Tech. J.*, vol. 48, no. 7, pp. 2079–2102, Sept. 1969.

[4] J. E. Goell, "A circular-harmonic computer analysis of rectangular dielectric waveguides," *Bell Syst. Tech. J.*, vol. 48, no. 7, pp. 2133–2160, Sept. 1969.

[5] R. M. Knox, and P. Toulios, "Integrated circuits for the millimeter through optical frequency range," in *Proc. Symp. Submillimeter Waves* (New York), Apr. 1–2, 1970, pp. 497–516.

[6] W. McLevige, T. Itoh, and R. Mittra, "New waveguide structures for millimeter wave and optical integrated circuits," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-23, pp. 788–794, Oct. 1975.

[7] P. Bhartia and I. J. Bahl, *Millimeter Wave Engineering and Applications*. New York: Wiley, 1984, ch. 6.

[8] T. Itoh, "Inverted strip dielectric waveguide for millimeter-wave integrated circuits," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-24, pp. 831–827, Nov. 1976.

[9] K. Solbach and I. Wolff, "The electromagnetic fields and phase constants of dielectric image lines," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-26, pp. 266–274, Apr. 1978.

[10] E. El-Sharawy and R. W. Jackson, "Coplanar waveguide and slot line on magnetic substrates: Analysis and experiment," *IEEE Trans. Microwave Theory Tech.*, vol. 36, pp. 1071–1079, June 1988.

[11] F. J. Rachford, P. Lubitz, and C. Vittoria, "Microwave resonance and propagation in nonsaturated ferromagnetic media. I. Magnetic resonance in single crystal ferrite platelets," *J. Appl. Phys.*, vol. 53, no. 12, pp. 8940–8951, Dec. 1982.

[12] A. K. Jordan, "Some effects of dielectric loading on ferrite phase shifters in rectangular waveguide," *IRE Trans. Microwave Theory Tech.*, Jan. 1962.

[13] F. J. Rachford, P. Lubitz, and C. Vittoria, "Magnetic multi-domain resonance n single crystal ferrite platelets," *J. Appl. Phys.*, vol. 52, no. 3, pp. 2259–2260, Mar. 1981.


**Jiqing Xia** (S'87) was born in Hunan, China, on June 24, 1964. He received the B.S. degree from Peking University, Beijing, China, in 1984 and the M.S.E.E. degree from Northeastern University, Boston, MA, in 1988. In the summer of 1988, he was with Schlumberger-Doll Research, Ridgefield, CT. Currently, he is working towards the Ph.D. degree in electrical engineering at the Massachusetts Institute of Technology, Cambridge, MA. His research interests concern the applications of both electromagnetics and solid-state material science to microwave integrated circuits.

**Peter P. Toulios** (S'60–M'61) was born in Kastoria, Greece, on January 16, 1934. He received the B.S. and M.S. degrees in electrical engineering from the University of Illinois, Urbana, in 1960 and 1961, respectively. In 1968, he received the Ph.D. degree in electrical engineering from the Illinois Institute of Technology, Chicago.

From 1961 to 1974, he worked as a microwave engineer for IIT Research Institute, Chicago, IL, on microwave component design, including integrated circuits using dielectric waveguides. In 1974, he cofounded Epsilon Lambda Electronics Corporation, Geneva, IL, where he has been working as an engineering adviser on millimeter-wave integrated circuits and subsystems, employing various transmission-line media.

Dr. Toulios has authored or coauthored 14 papers and holds six U.S. patents. He served as a cochairman of the 1972 MTT-S Symposium Technical Program Committee and he is a member of Sigma Xi.

**Carmine Vittoria** (S'62–M'63–SM'83), photograph and biography not available at the time of publication.