FREQUENCY SELECTIVE LIMITERS FOR HIGH DYNAMIC RANGE MICROWAVE RECEIVERS

J. D. Adam* and S. N. Stitzer**

*Westinghouse Science and Technology Center
Pittsburgh, PA 15235-5098
**Westinghouse Electronic Systems Group
Baltimore, MD 21203

ABSTRACT

A new broadband frequency selective limiter which provides over 14 dB of limiting across more than an octave bandwidth is described. The limiter is fabricated with epitaxially grown YIG films in a stripline configuration and has a threshold power level of below 0 dBm. Multiple FSLs have been cascaded with amplifiers to allow compression of microwave signals with a power range of 60 dB into a range of less than 5 dB.

BACKGROUND

Broadband EW or ESM receiver systems need to intercept signals over wide frequency bands and over a wide dynamic range with a high probability of intercept and with accurate measurement capability. This must be done within a complex signal environment that includes low, medium and high duty cycle and CW signals and other intentional and unintentional interference. The receiver must accurately sense signals in the presence of other signals which may be much stronger. A number of receiver sensor approaches are inherently not compatible with these requirements. Channelized systems, such as the Bragg cell receiver, have the potential to meet these requirements if certain components are developed, and if the system is properly engineered.

All signal processing systems have a finite dynamic range, limited at the low end by signal sensitivity and noise figure and at the high end by errors due to compression, internal generation of spurious responses (intermodulation) and ultimately by permanent damage. In a wideband receiver, strong signals far removed in frequency from a weak desired signal can interfere with measurement of the weaker one. A PIN diode limiter is often used to avoid overload or damage to receivers. When a strong signal causes the limiter to turn on, all signals in the receiver are attenuated by the same amount as the strong signal.

A frequency selective limiter avoids the problem of losing weak signals in the presence of a strong signal by attenuating only those signals in a narrow frequency band around the incident strong signal. Figure 1 illustrates the frequency response of an FSL in the presence of simultaneous strong (f_3) and weak (f_1 and f_2) signals. The most common FSL devices use the nonlinear excitation of magnetic spin waves in an appropriately magnetized ferrite such as yttrium-iron-garnet (YIG). Microwave signals coupled to a sample of YIG produce an essentially linear, low-loss response in the ferrite at low signal levels. Above a critical RF magnetic field strength, the nonlinearity in the precession of the magnetic dipoles in the ferrite becomes strong enough to overcome the natural losses, and energy starts to transfer exponentially to very short wavelength subharmonic spin waves.[1] These "half-frequency" spinwaves can

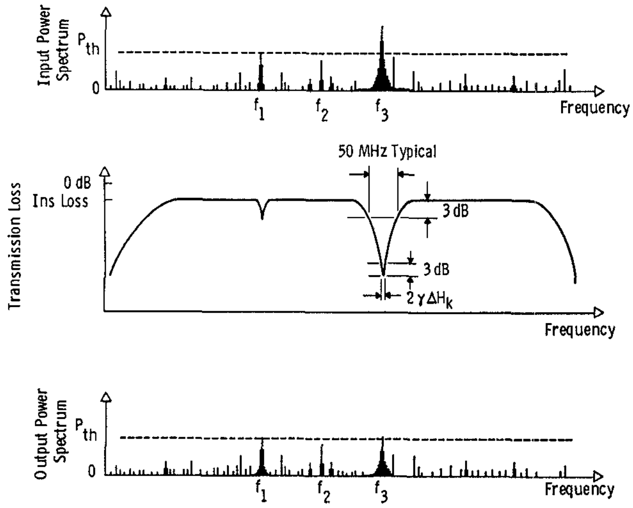

Figure 1 Frequency response of an FSL, and input and output power spectra, for multiple weak and strong signals, showing the frequency selectivity inherent in YIG limiters.

transfer energy readily to the crystal lattice, whereby the excess RF power is dissipated as heat in the ferrite. This nonlinear coupling takes place in a bandwidth on the order of the spinwave linewidth ΔH_k. For high quality single crystal YIG, ΔH_k is about 0.25 oersted, which represents a frequency bandwidth of about 1 MHz. This is the 3 dB bandwidth of the deep absorption notch, Figure 1. Weak signals that are more than a few tens of megahertz away from a strong limited signal will undergo less than 1 dB increase in attenuation above the low level value.

We have previously described a broadband FSL using YIG films coupled to a microstrip transmission line.[2] The YIG films were epitaxially grown on gadolinium gallium garnet (GGG) substrates. A 0.001" (25 micron) wide microstrip was used to develop the necessary RF magnetic field strength at milliwatt power levels. Since microwave power is dissipated continuously along the length of the microstrip, a total of 7 inches of microstrip line was meandered over a 1" x 2" substrate to achieve the desired amount of limiting. A 50 ohm line impedance was used to permit operation over more than two octaves without matching networks. Earlier low threshold FSLs used resonators to enhance the r.f. field within the YIG at the expense of operating bandwidth.[3] However, the meander line FSL suffers from inefficient coupling of the RF magnetic field around the transmission line to the ferrite, uneven limiting along the length of the line, and large size due to the biasing magnets.

Stripline FSL

The most efficient use of the RF magnetic field energy in an FSL takes place when all the RF energy is confined to the ferrite. A ferrite-filled stripline, Figure 2, fits this description. The important dimensions turn out to be similar to those in the meander line device; a 25 micron wide strip generates an RF field strong enough to produce limiting at a power level around one milliwatt. The thickness of the YIG is chosen to set the impedance to 50 ohms, and to provide a sufficient volume of ferrite for a useful dynamic range. The YIG is therefore used in the form of single crystal slabs approximately 100 micron thick.

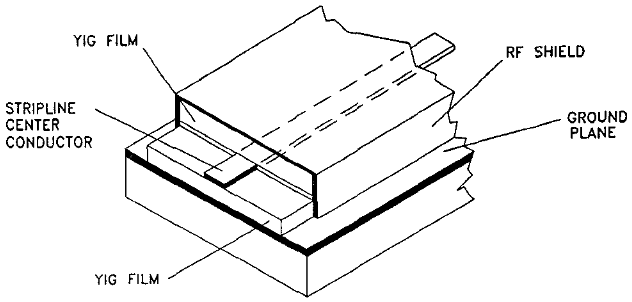

Figure 2 Stripline FSL using epitaxial YIG films.

The variation of above threshold insertion loss normalized to the small signal loss in a 38 mm long stripline FSL over the 2.5 to 5.3 GHz range is shown in Figure 3. Here the small signal loss is approximately 3 dB the threshold power level is just below 0 dBm and more than 14 dB of limiting is achieved for a +19 dBm input power.

As can be seen in Figure 3, limiting decreases at higher power levels. Microwave power is dissipated continuously along the length of the stripline, so in principle more limiting can be obtained by increasing the length. However, once the travelling power falls below threshold, no further limiting takes place in that strip. If more limiting is required, the output of an FSL can be amplified and fed to a second FSL strip. If the amplifier gain exceeds the below-threshold insertion loss, signals that were below threshold in the first strip can be brought into limiting in the second strip. We have built limiter assemblies with five and more strips in this type of cascaded arrangement. The result is that signals within a range of 50 to 60 dB at the input are compressed into a range of just 4 to 6 dB at the output.

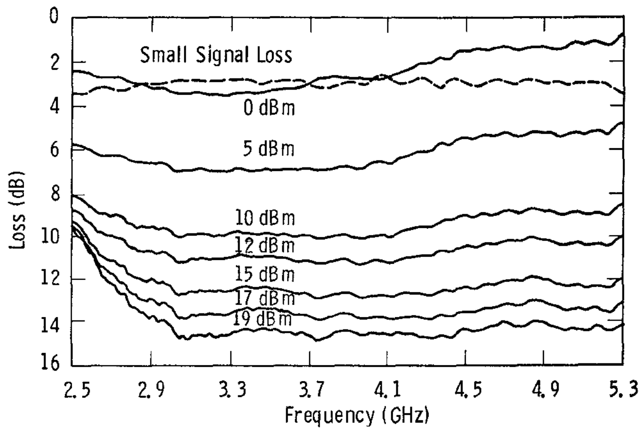

Figure 3 Insertion loss of a stripline FSL over the 2.5 GHz to 5.3 GHz range, for above threshold input signals in the range 0 dBm to +19 dBm, normalized to the below threshold loss (broken line).

CONCLUSIONS

We have developed a compact, broadband stripline frequency selective limiter. Individual limiter strips limit microwave signals in the 1 to 100 milliwatt range across more than an octave bandwidth. Multiple strips have been cascaded with amplifiers to allow compression of signals with a power range of 60 dB to a range of less than 5 dB. The FSL should find many applications in broadband signal processing systems.

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QUASI-OPTICAL REFLECTION CIRCULATOR

B. Lax,* J.A. Weiss,† N.W. Harris, and G.F. Dionne
Lincoln Laboratory, Massachusetts Institute of Technology
Lexington, MA 02173

Abstract — A quasi-optical Faraday rotation ferrite circulator for microwave or millimeter radiation is analyzed by a matrix formalism. Both reflection and transmission configurations at oblique incidence are examined. Numerical results are presented at 35° and 45° incidence across a band centered at 35 GHz. Calculations are also compared with transmission experiments over a 10% bandwidth. Notwithstanding the complexities of oblique incidence, the reflection-type device promises favorable bandwidth, low loss, and isolation comparable to those of the transmission version now in system deployment, and offers the potential for much higher heat dissipation capability.

1. INTRODUCTION

At higher microwave and millimeter frequencies the smaller dimensions of waveguide ferrite circulators limit their power handling capability. To improve the performance of these nonreciprocal components at Ka band, Dionne et al. [1] developed a quasi-optical Faraday rotation transmission isolator or circulator. Employing a free-space optical beam configuration equivalent to that of the classical Faraday-rotation four-port waveguide circulator, the device incorporates an axially magnetized ferrite disc designed to rotate the polarization of the transmitted beam by 45°. Circulator action is achieved through the nonreciprocal rotation within the ferrite element that is positioned between two slanted mutually crossed polarizers. Recently, a millimeter-wave low-loss beam waveguide system with the quasi-optical transmission circulator as an integral component has been demonstrated [2]. In this paper we examine the reflection version of this device, and compare it with its transmission counterpart.

There are significant differences between the transmission and reflection versions of the device. The former is used at normal incidence, whereas the latter requires oblique incidence in order to separate the input and the output beams, which introduces interesting complexities in the analysis. Important practical advantages of the reflection device are that the metal reflector plate structure behind the ferrite disc shown in Fig. 1 may serve both as a heat sink and as a permanent magnet. In addition to providing for more efficient heat removal, the reflection version needs only half the ferrite thickness of the transmission case.

* Member of the Department of Physics, Massachusetts Institute of Technology.

† Also affiliated with the Department of Physics, Worcester Polytechnic Institute.

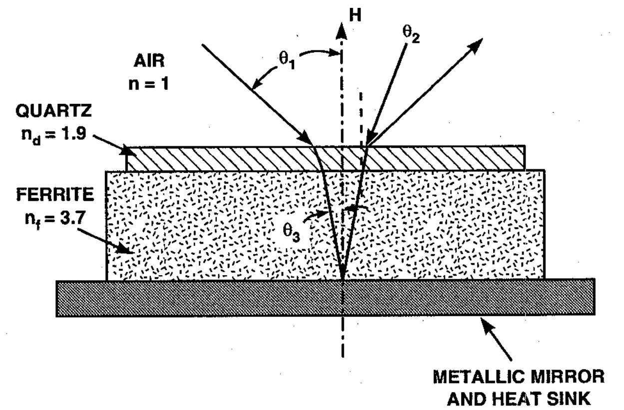

Fig. 1. Configuration for a quasi-optical reflection type microwave circulator. H is the magnetic field required to magnetize the ferrite.

2. BASIC THEORY

Propagation of microwaves at oblique incidence relative to a magnetic field in a ferrite is discussed in the standard texts [3]. The expression can be simplified in our case because the angle of propagation θ_3 in the ferrite medium is relatively small even for large angles of incidence θ_1. For $\theta_1 = 45°$ in air, $\theta_3 = 10°$ in the ferrite. The general expression for the Faraday rotation is given by the following:

$$\Theta_F = n\omega\chi_{xy}T/2c \quad , \qquad (1)$$

where the thickness of the ferrite $T = l\cos\theta_3$, l is the slant length of the beam, n the index of refraction, c the velocity of light in air, ω the angular frequency of the microwave radiation. The susceptibility $\chi_{xy} \approx \omega_M/\omega$ and $\omega_M = \gamma 4\pi M$, the angular frequency corresponding to the magnetization $4\pi M$; γ is the gyromagnetic ratio. In the above approximation, $\chi_{xy} << 1$. Equation (1) then simplifies to:

$$\Theta_F = n\omega_M T/2c \qquad (2)$$

Equation (2) still holds when we take into account the difference in path length due to the birefringence in the tensor medium which splits the two slightly elliptical counter-rotating beams in the ferrite. Thus if the ferrite is ideally matched, the rotation is independent of frequency and the angle of incidence and Θ_F depends only on the magnetization and the thickness of the ferrite.

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For the circulator to operate properly in the transmission mode, $\Theta_F = 45°$; for the reflection mode, $\Theta_F = 22.5°$ per traversal. Since Θ_F is independent of frequency, the circulator is inherently broadband. The ultimate limitation is determined by the requirement of using quarter-wave plates for matching the ferrite to free space.

3. MATRIX TREATMENT

There are two considerations which complicate the analysis of the layered structure shown in Fig. 1. The first is oblique incidence and the second is the nonreciprocal tensor nature of the ferrite. The obliquity requires that we make a distinction between S (perpendicular) and P (parallel) polarizations relative to the plane of incidence. The theory for this is well established and the 2 × 2 matrix formulation for reflection and transmission coefficients is given explicitly for multilayer dielectrics by Yeh [4]. From this analysis it is evident that, except with normal incidence, S and P polarizations cannot be matched simultaneously. But if the match is made for normal incidence at the center frequency, the small mismatch for both polarizations will be identical at that frequency. When the magnetic field is turned on, the ferrite becomes a birefringent medium which splits the incident beam into two slightly elliptical counter-rotating beams separated by a small angle. The splitting is a consequence of the nonreciprocal tensor nature of the magnetized ferrite medium which in effect couples the S and P waves. The emergent output wave is rotated 45° and is also slightly elliptically polarized. The unwanted quadrature polarization represents a small insertion loss.

In order to analyze the complex multilayer structure composed of the quarter-wave plates and the ferrite, it is necessary to represent each by a 4 × 4 matrix. To accomplish this it was necessary to employ the general theory of Zak et al. [5] in order to apply it to layers of any thickness. The published version was adapted to the special case of ultrathin films. Following their prescription, we generate a product of 4 × 4 matrices for each layer and boundary to form a single grand scattering matrix. This is then used to evaluate a set of reflection, transmission, Faraday and Kerr coefficients, as well as the ellipticity for both S and P polarizations of the incident wave. The final grand matrices, represented in symbolic form as scattering matrices, can be written as:

$$S_T = A_a^{-1} \bullet A_d D_d A_d^{-1} \bullet A_f D_f A_f^{-1} \bullet A_d D_d A_d^{-1} \bullet A_a \quad ,$$

$$S_R = A_a^{-1} \bullet A_d D_d A_d^{-1} \bullet A_f D_f A_f^{-1} \bullet A_m \quad , \qquad (3)$$

where S_T is the scattering matrix for the transmission device, a product of eleven 4 × 4 matrices, and S_R is that for the reflection device, a product of eight such matrices. A_a is the surface matrix in air which relates the incident and reflected electric field amplitudes E_S and E_P of the S and P polarized waves to E_x, E_y, H_x, H_y, the tangential electric and magnetic field components at the interface. Similarly A_d and A_d^{-1} are the surface matrices on either side of the dielectric slabs at opposite interfaces and D_d is the transfer matrix which transforms the amplitudes of the S and P waves from one boundary to the other.

The A_f and A_f^{-1} are the ferrite surface matrices which contain off-diagonal terms proportional to the magnetic susceptibility, and D_f is the transfer matrix which, in the product

$$A_f D_f A_f^{-1} = M_f \quad , \qquad (4)$$

incorporates the birefringence and Faraday rotation in the magnetic matrix M_f to take into account all of the consequences of the nonreciprocal tensor medium. The analytical expression for these matrices was derived explicitly [5]. The matrix S_R differs from S_T in having fewer factors and containing a matrix A_m for the metal mirror shown in Fig. 1 (which also serves as a heat sink). Its refractive index is complex and very large; when the reflection and Kerr coefficients are evaluated, it cancels out. It can be eliminated, except for evaluation of the slight loss due to metallic conductivity, and only the first seven factors have to be considered.

4. NUMERICAL RESULTS

Calculations have been carried out for the design and performance of optimized transmission and reflection devices. Parameters chosen for the band centered at 35 GHz are: for the quartz quarter-wave plates $n_d = 1.9$ and for the ferrite $n_f = 3.7$; the magnetization $4\pi M = 1000$ Oe, which gives the ferrite thickness for the transmission device $T_f = 7.2$ mm (3.6 mm for the reflection case). The dielectric quarter-wave plate thickness $T_d = 1.2$ mm. Using these parameters in Eq. (3), we calculated the individual components of the scattering matrix.

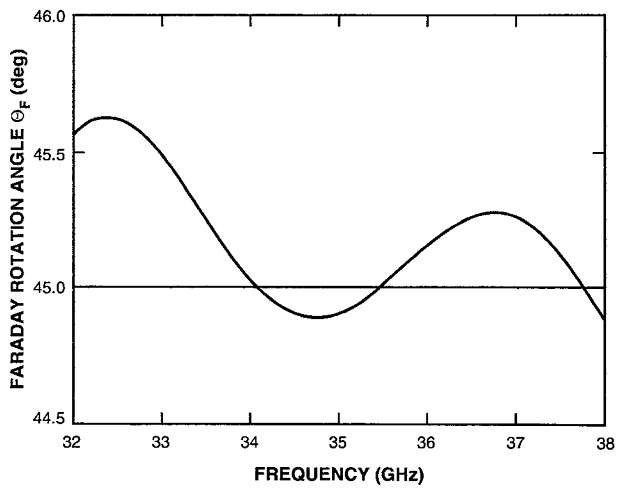

Fig. 2. Calculated Faraday rotation of a transmission circulator at 45° incidence as a function of frequency.

In Fig. 2 the Faraday rotation for the transmission device is plotted from 32 to 38 GHz. The nominal value of the rotation parameter was 45° at center frequency. However, due to the unavoidable mismatch required to accommodate both S and P polarizations equally, there is a small Fabry-Perot interference effect which alters this condition. In practice we would adjust the thickness of the ferrite so that the 45° rotation would be the mean value over the bandwidth. Thus in Fig. 2 we show the rotation which has a deviation of 0.7° over the 6-GHz bandwidth. If the dimension is optimized, the deviation becomes ±0.35°, which is negligible (±1° deviation corresponds to a loss of 0.0013 dB and a cross-polarized component 35.2 dB down).

Figure 3 shows the squared magnitudes of the transmission coefficients T_{ss} and T_{ps}, for the incident (S) and coupled (P) polarizations respectively, as functions of frequency. As expected, the power is nearly split equally between the S and P components at 35 GHz. At the band edges the combined power is reduced by ~0.3 dB, which is the contribution to the insertion loss due to the mismatch of the quarter-wave plates. The ellipticity was also calculated for the transmission circulator and found to be very small (below –30 dB) across the band.

Calculations were also made for the reflection circulator. The Faraday rotation shown in Fig. 4 exhibits a Fabry-Perot effect which causes a variation of ±3.5° between 32 and 38 GHz for 35°and 45° incidence. This is larger than the corresponding values for the transmission device. However it is still small and represents a negligible contribution to the insertion loss, greatest at the band edges. The reflection coefficient is another quantity of interest. Figure 5 illustrates the squared magnitudes of the reflection coefficients R_{ss} and R_{ps}, showing the characteristic Fabry-Perot effect.

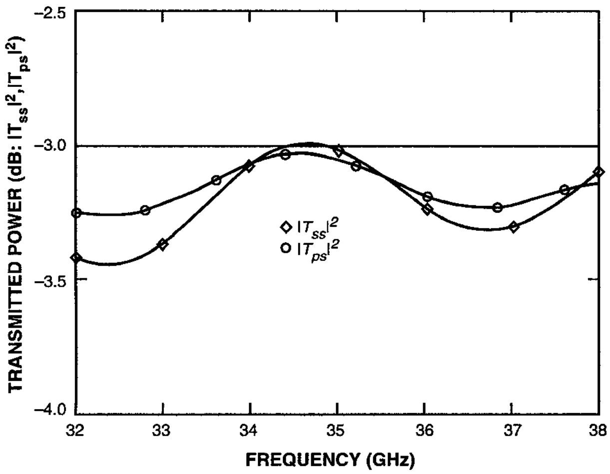

Fig. 3. Calculated T_{ss} and T_{ps} of a transmission circulator at 45° incidence as a function of frequency.

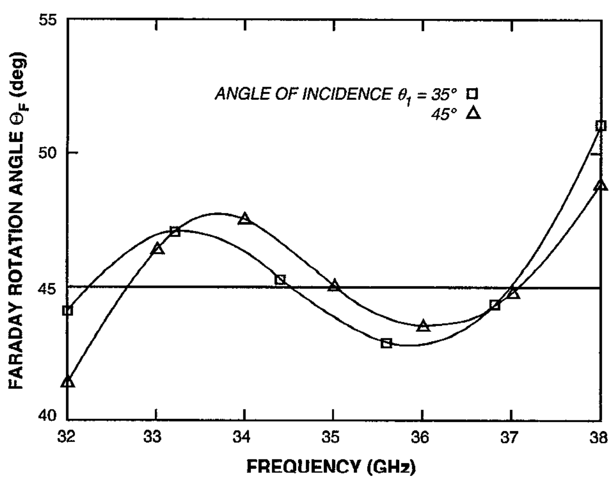

Fig. 4. Calculated Faraday rotation of a reflection circulator at 35° and 45° incidence as a function of frequency.

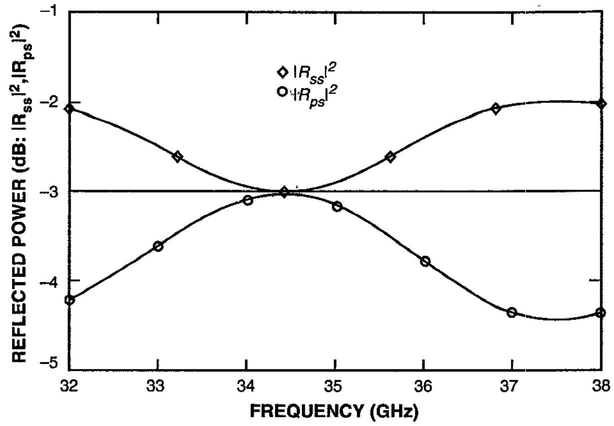

Fig. 5. Calculated R_{ss} and R_{ps} of a reflection circulator at 45° incidence as a function of frequency.

Although the power is not equally split between the S and P components, the total adds to unity (except for minor losses in the ferrite and mirror). Performance limits on the reflection circulator are determined by the rotation and also by the ellipticity, illustrated in Fig. 6. This quantity has been calculated for several values of incidence angle and the plot shows examples at 45° and 35° incidence. It shows, as expected, that the angle of incidence is reduced as the ellipticity is decreased. The corresponding figures of minimum isolation are 16 and 22 dB, respectively. Reducing the angle of incidence, if the system configuration permits, would therefore improve the performance. Means of enhancing the performance, including this option and also by improving the matching structure, are under consideration.

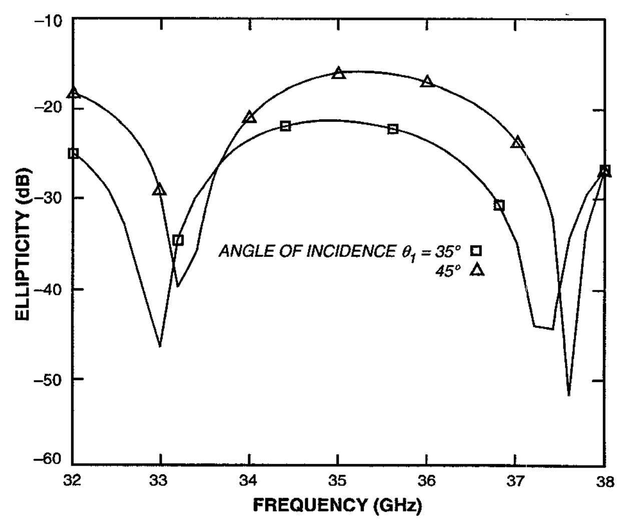

Fig. 6. Calculated ellipticity of a reflection circulator at 35° and 45° incidence as a function of frequency.

5. EXPERIMENTAL RESULTS

Experiments have been performed on a transmission device at oblique incidence to compare with theory. There are internal reflections within the multilayer structure which exaggerate the Fabry-Perot effect. The Faraday rotation over a broad frequency range centered at ~35 GHz is shown in Fig. 7. The parameters for this device are as follows: the magnetization of the ferrite $4\pi M$ = 690 Oe, the dielectric constant of the quartz is $\varepsilon_d = 3.77$, that of the ferrite ε_f = 14.7, the thickness of the ferrite is T_f = 10 mm and that of the quartz T_d = 1.2 mm. The Faraday rotation deviates from the mean value by ±2° over a 10-GHz bandwidth, slightly greater than that expected if the optimum parameters were chosen. In addition, Fig. 8 shows the plot of the ellipticity, again exhibiting higher values at the band edges, but still negligibly small. Although the parameters are not ideal, these results are in good agreement with theory, when the above parameters are used to evaluate the components of the scattering matrix.

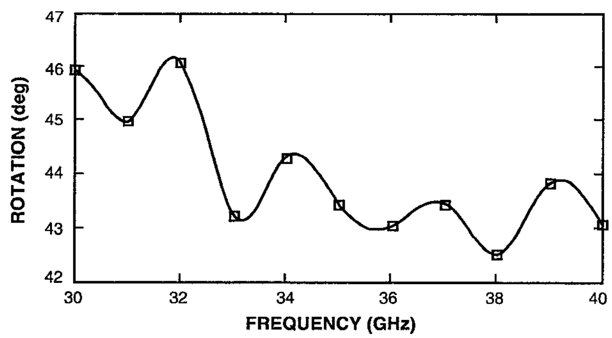

Fig. 7. Experimental values of the Faraday rotation of a transmission circulator at 45° incidence as a function of frequency.

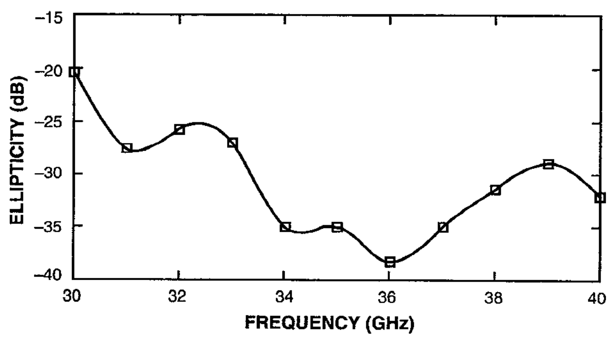

Fig. 8. Experimental values of the ellipticity of a transmission circulator at 45° incidence as a function of frequency.

6. CONCLUSIONS

This paper has outlined the procedure for treating plane-wave propagation with oblique incidence through a complex multilayer structure consisting of scalar dielectric and tensor magnetic media using a matrix formalism adapted from that developed by Zak et al. [5]. The process permits the evaluation of such pertinent quantities as the transmission, reflection, Faraday and Kerr coefficients and the ellipticity. We have applied this technique to analyze both a transmission and a reflection circulator to deduce the insertion losses and the isolation. The reflection circulator, which permits surface cooling and is capable of higher average power, is nearly comparable in microwave performance to that of the transmission device at normal incidence. We have shown that the ferrite layer is inherently broadband and that the limitations on bandwidth are imposed by the quarter-wave plates required for matching. The analysis treated a single layer antireflection medium at oblique incidence. However, it is known from the optical analog that a two-layer structure would improve the performance further. We expect to examine this possibility for our millimeter-wave device.

For the moment we have also neglected the dissipative losses in the dielectric and ferrite media but these can be incorporated by using appropriate complex values for the dielectric constants and the magnetic susceptibilities in the components of the scattering matrix. From previous experience with the transmission circulator we are confident that, though these may add slightly to the insertion loss, they will not affect the isolation or the Faraday rotation significantly. The latter were the main objectives for design considerations of this paper. The final conclusion is that the reflection circulator will satisfy the requirements for a broadband device, but at much higher power than its transmission counterpart.

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