ferrite geometry. At low RF power levels the limiter exhibits a maximum insertion loss of 0.9 db and a maximum VSWR of 1.13 over the band.

Conclusion

In this paper several techniques for increasing the insertion loss of ferrite-loaded waveguide structures at high RF power levels have been presented, and the operating characteristics of a ferrite microwave limiter have been described. The most important problem appears to be the distortion of the RF pulse waveform by the ferrite limiter. While the plateau of the limited RF pulse has been reduced to 11 watts and further improvement can reasonably be expected, the leading edge spike, because of its large amplitude and long duration, is a more severe problem. Perhaps as we learn more about the mechanism of the ferrite's nonlinear behavior, engineering techniques can be found which will solve the spike problem by an appropriate waveguide-ferrite configuration or through the use of new ferromagnetic materials.

Acknowledgment

The author wishes to thank the various laboratory personnel who contributed suggestions in a number of the experiments. The assistance of W. Asp in the laboratory measurements is especially appreciated.

Nonreciprocity in Dielectric Loaded TEM Mode Transmission Lines*

D. FLERI† AND G. HANLEY†

Summary—An analysis is presented of partially dielectric loaded strip transmission line from the point of view of ferrite applications. It is shown that the microwave magnetic field is elliptically polarized both at the dielectric surface and within the dielectric. The degree of elliptical polarization is expressed analytically as a function of the dielectric constant, the degree of dielectric loading, and the frequency. For specific values of dielectric constant and loading, a high degree of circularity may be made to exist at the dielectric surface over extremely broad frequency bands. Experimental data are presented which are in accord with the theoretical predictions.

Introduction

A GREAT variety of nonreciprocal propagation characteristics has been achieved at microwave frequencies through the use of ferrites. A necessary requirement for nonreciprocity is that the microwave magnetic field in the region of the ferrite be circularly polarized.[1] This requirement is easily met in rectangular waveguide propagating the dominant mode[2–4] and in circular waveguide propagating the circularly polarized TE_{11} mode.[3–4] In coaxial line and strip transmission line propagating the TEM mode, however, the microwave magnetic field is linearly polarized at all points, and therefore any ferrite effects will be completely reciprocal. It has been reported previously[5–6] that partially filling the cross section of coaxial line with a dielectric serves to distort the mode pattern and create an almost true sense of circular polarization at the air-dielectric interface. This mode distortion technique thereby renders coaxial line suitable for nonreciprocal applications. In a similar manner, polarization conversion may be effected in strip transmission line by appropriate dielectric loading.[7] Analysis of this latter transmission line structure forms the substance of this paper.

The dielectric loaded strip transmission line configuration is shown in Fig. 1. The co-ordinate axes are chosen so that Z represents the direction of propagation, and Y represents what will be referred to subsequently as the transverse direction. It will be shown that the polarization is elliptical both at the dielectric surface and within the dielectric. The degree of elliptical polarization at the dielectric surface is a function of the dielec-

* Manuscript received by the PGMTT, July 7, 1958; revised manuscript received, September 2, 1958. This work was supported by the U. S. Air Force Cambridge Res. Center, Contract No. AF19(604)-2248.

† Sperry Gyroscope Co., Div. of Sperry Rand Corp., Great Neck, N. Y.

[1] C. L. Hogan, "The ferromagnetic Faraday effect at microwave frequencies and its applications: The microwave gyrator," *Bell Sys. Tech. J.*, vol. 31, pp. 1–31; January, 1952.

[2] M. L. Kales, N. H. Chait, and N. G. Sakiotis, "A nonreciprocal microwave component," *J. Appl. Phys.*, vol. 23, pp. 816–817; June, 1953.

[3] J. H. Rowen, "Ferrites in microwave applications," *Bell Sys. Tech. J.*, vol. 32, pp. 1333–1369; November, 1953.

[4] A. G. Fox, S. E. Miller, and M. T. Weiss, "Behavior and applications of ferrites in the microwave region," *Bell Sys. Tech. J.*, vol. 34, pp. 5–103; January, 1955.

[5] B. J. Duncan, L. Swern, K. Tomiyasu, and J. Hannwacker, "Design considerations for broadband ferrite coaxial line isolators," Proc. IRE, vol. 45, pp. 483–490; April, 1957.

[6] H. Seidel, "Ferrite slabs in transverse electric mode waveguide," *J. Appl. Phys.*, vol. 28, pp. 218–226; February, 1957.

[7] R. S. Mangiaracina and B. J. Duncan, "Nonreciprocal ferrite devices in TEM mode transmission line," presented at Natl. PGMTT Symp., New York, N. Y.; May, 1957.

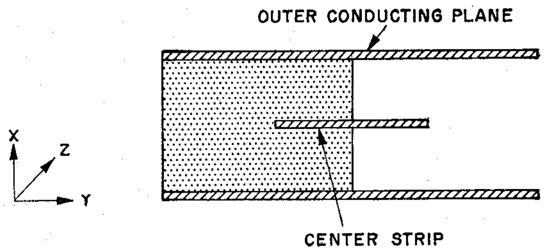


Fig. 1—Cross-sectional view of dielectric loaded strip transmission line structure.

tric constant, the degree of dielectric loading, and the frequency. It is the purpose of this paper to derive analytically the equation which relates these quantities and to discuss, on the basis of the derived result, the choice of parameters which renders the structure suitable for broad-band applications.

Although the analysis is presented in terms of strip transmission line, it is valid within engineering accuracy for dielectric loaded coaxial line as well.

Discussion

The microwave electric and magnetic fields in balanced strip transmission line propagating the TEM mode are purely transverse. However, if the structure is partially loaded with dielectric, a longitudinal component of the microwave magnetic field is required in order to satisfy the boundary conditions. This Z-directed H field interacts with the X component of the microwave E field to generate a Poynting vector in the transverse direction. Since a propagating mode exists in the transverse direction, the transverse resonance formulation may be used to determine the propagation constants of the transmission line system.[8]

For ease of analysis, the dielectric loaded strip transmission line structure is replaced by a simpler equivalent circuit. This circuit and the relationship it bears to the original structure is shown in Fig. 2(a) and 2(b). Symmetry allows the equivalent representation to be further simplified as shown in Fig. 2(c). A discussion of the various steps involved in formulating this equivalent representation is contained in the reference previously cited.[8]

The equivalent width of the transmission line structure, D, and of the dielectric, d, may be determined from the dimensions of the original structure as shown in Fig. 2. The ground plane separation, b, strip width, w, and the quantity, t, are directly measurable. S is the distance of the open circuit plane from the edge of the center strip. In general, S is a function of frequency, but if the operating conditions are such that $b/\lambda g$ never exceeds approximately 0.2, S may be represented by a constant independent of frequency. Under these conditions S is equal to (b/π) ln 2. Thus,

[8] G. Hanley, "The determination of propagation constants for partially filled strip line," unpublished thesis, Polytechnic Inst. of Brooklyn, Brooklyn, N. Y.; June, 1957.

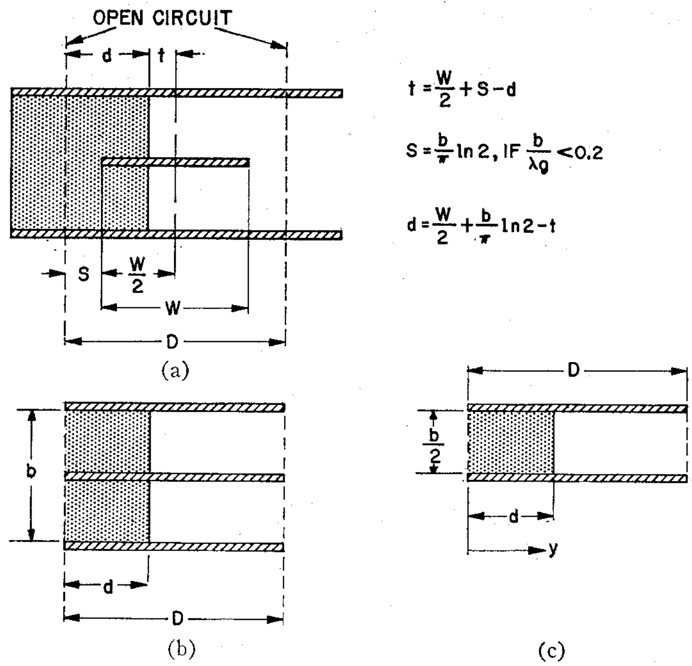


Fig. 2—Equivalent representation of the dielectric loaded strip transmission line structure.

$$D = w + \frac{2b}{\pi} \ln 2 \tag{1}$$

$$d = \frac{w}{2} + \frac{b}{\pi} \ln 2 - t. \tag{2}$$

The ratio, D/d, is referred to as the degree of loading.

By the method of transverse resonance the field equations for the dominant mode may be derived. These are obtained from a knowledge of the resonance voltage and current distribution and the mode functions of the uniform waveguides in the transverse direction.

There are two sets of field equations, one valid in the dielectric region, and the other valid in the air region.[8]

In the dielectric region:

$$H_X = 0 \tag{3a}$$

$$H_Y = I(Z)NV(0) \cos [k_{\epsilon Y} Y] \tag{3b}$$

$$H_Z = j \frac{k_{\epsilon Y}}{k_Z} I(Z)NV(0) \sin [k_{\epsilon Y} Y]. \tag{3c}$$

In the air region:

$$H_X = 0 \tag{4a}$$

$$H_Y = \frac{I(Z)NV(0) \cos [k_{\epsilon Y} d] \cosh [|k_Y| (D - Y)]}{\cosh [|k_Y| (D - d)]} \tag{4b}$$

$$H_Z = j \frac{|k_Y| I(Z)NV(0) \cos [k_{\epsilon Y} d] \sinh [|k_Y| (D - Y)]}{k_Z \cosh [|k_Y| (D - d)]} \tag{4c}$$

where

$I(Z)$ = current distribution in direction of propagation,

N = normalization factor,

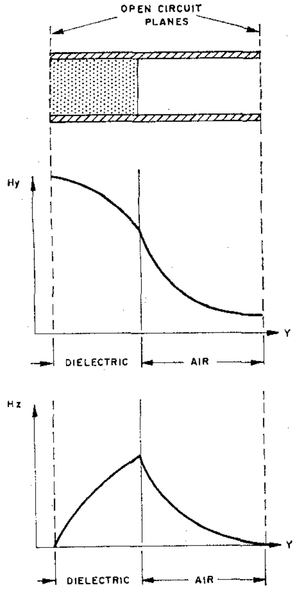


Fig. 3—The microwave magnetic field configuration in the dielectric loaded strip transmission line.

$V(0)$ = amplitude factor,
k_Z = guide wave number $= 2\pi/\lambda_g$,
k_Y = transverse wave number in air,
$= \sqrt{(2\pi/\lambda_0)^2 - (2\pi/\lambda_g)^2}$,
$k\epsilon_Y$ = transverse wave number in dielectric,
$= \sqrt{(2\pi/\lambda_0)^2\epsilon - (2\pi/\lambda_g)^2}$,
D = equivalent width of transmission line structure,
d = equivalent width of dielectric,
Y = as defined in Fig. 2.

In the dielectric region, the magnetic field variation is sinusoidal whereas in the air region the variation exhibits a hyperbolic dependence on transverse position. The field components are shown graphically in Fig. 3. The dominant mode is characterized by a guide wavelength smaller than that of free space.

The degree of elliptical polarization may be represented by the ratio of H_Y to H_Z. This ratio is termed the polarization factor. The polarization factor depends on the dielectric constant, the degree of loading, and the frequency. The functional dependence of the polarization on these parameters is easily found by forming the ratio H_Y to H_Z in both regions of the structure.

In the dielectric region:

$$\alpha_\epsilon = \left|\frac{H_Y}{H_Z}\right| = \frac{k_Z}{k_{\epsilon Y}} \cot\,[k_{\epsilon Y} Y]. \tag{5}$$

In the air region:

$$\alpha_a = \left|\frac{H_Y}{H_Z}\right| = \frac{k_Z}{|k_Y|} \coth\,[\,|k_Y|\,(D - Y)]. \tag{6}$$

α_a is the polarization factor in the air region and α_ϵ is the polarization factor in the dielectric region. Circular polarization corresponds to a factor of unity. It is clear from (6) that the polarization factor may never be equal to unity in the air region since the quantity, $k_Z/|k_Y|$, is always greater than unity (since $\lambda_g < \lambda_0$) and the hyperbolic cotangent is never less than unity. Therefore, circular polarization does not exist in this region.

Since the Y and Z components of the magnetic field are continuous at the dielectric boundary, H_Y must be greater than H_Z in the dielectric as well as in the air region. Hence, circular polarization does not exist at any point in this structure. It is obvious from the field configuration (Fig. 3) that the best sense of circularity exists at the air-dielectric interface.

Eq. (5) provides the functional dependence of the degree of elliptical polarization on the dielectric constant, the degree of loading, and the frequency. The nature of this dependence is more easily illustrated graphically.

Fig. 4 depicts the variation of the polarization factor at the dielectric surface as a function of wavelength (normalized with respect to the parameter, d) for several degrees of loading. At higher frequencies, *i.e.*, for $\lambda/2d$ less than approximately 3, the polarization factor is independent of the degree of loading.

As the wavelength increases, the curves diverge. For a loading factor of one third, *i.e.*, for a D/d ratio of 3, the polarization factor remains less than 1.2 for $\lambda/2d$ between 1 and 8.5. This represents more than an 8 to 1 frequency band.[9] A polarization factor of 1.2 corresponds to a resonance attenuation ratio of about 20 to 1 db if it is assumed that the only loss mechanism is the absorption by the ferrite of negative circularly polarized energy. A D/d ratio of 2 yields a polarization of 1.2 or less over approximately 5 to 1 frequency band. The band becomes increasingly narrow as the degree of loading is increased.

Fig. 5 shows the variation of the polarization factor with normalized wavelength for various values of dielectric constant. The degree of loading for this case corresponds to a D/d ratio of 3. It is seen that the polarization factor increases without limit in the large wavelength region of the curve. This may be expected for at zero frequency, *i.e.*, in the static case, H_Z is equal to zero, and therefore, the ratio of H_Y to H_Z is infinite. The dual condition exists at short wavelengths and the ellipticity tends to infinity in this region as well. Between these two extremes the polarization factor remains close to unity over a wide range of wavelengths. The higher the dielectric constant the lower and broader the characteristic becomes. This trend is offset, however, by the inception of higher order modes. The higher the dielectric constant, the lower is the frequency at which these modes may propagate. In many applications a dielectric constant of 10 represents a good compromise between a well-defined sense of circularity over a wide band and suitable mode purity.

[9] Mode pure operation exists only for $\lambda/2d$ greater than approximately 4.5. When operation corresponds to $\lambda/2d$ less than 4.5 special effort must be made not to excite higher order modes.

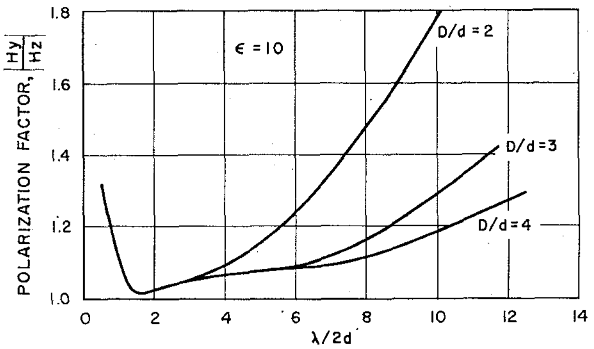


Fig. 4—The polarization factor as a function of normalized wavelength for various degrees of dielectric loading.

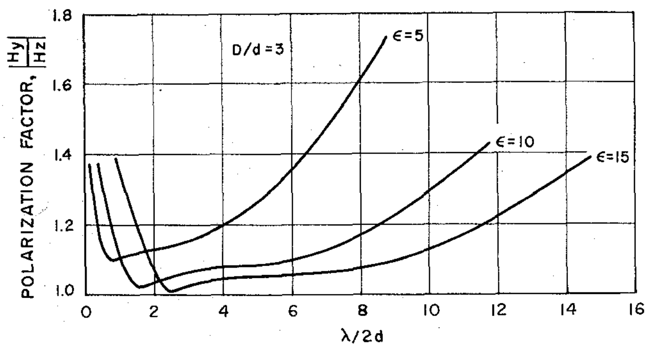


Fig. 5—The polarization factor as a function of normalized wavelength for various values of dielectric constant.

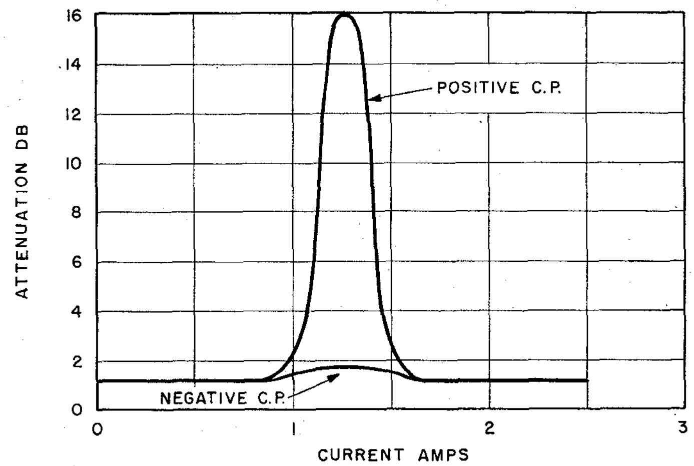
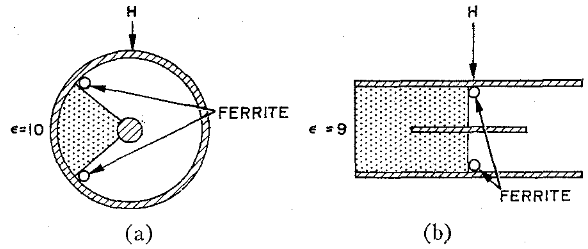


Fig. 6—Typical attenuation characteristic of ferrite dielectric loaded strip transmission line structure.

The foregoing analysis was conducted in an effort to determine the parameters which render dielectric loaded strip transmission line suitable for broad-band nonreciprocal ferrite applications. It was seen that for a dielectric constant of 10 and a D/d ratio of 3, the polarization factor at the air-dielectric interface remained less than 1.2 over an extremely broad frequency band. As was pointed out, a polarization factor of 1.2 corresponded to a resonance attenuation ratio in db of 20 to 1 db.

Experimental data are now presented which tend to confirm the results of this analysis.



V (mc)	Resonance Attenuation Ratio (db)	
	A	B
1000	12:1	9:1
1500	50:1	20:1
2000	40:1	50:1
3000	30:1	—
4000	25:1	50:1
4500	35:1	—

Fig. 7—Representation of ferrite dielectric loaded coaxial and strip transmission line test structure and tabulation of experimental results.

Ferrite was positioned at the air-dielectric interface and forward and reverse wave attenuation was measured as a function of applied magnetic field. Of course, these measurements cannot be used to test the theoretical results quantitatively as ferrite loading causes considerable perturbation of the microwave field configuration.

Fig. 6 is representative of the data taken. The ferrite was a low $4\pi M_s$ aluminate characterized by a narrow linewidth and extremely low loss away from resonance. The loss away from resonance was primarily that of the mode distorting dielectric. If this is subtracted from the forward and reverse loss at resonance, the resulting attenuation ratio is a measure of the degree of elliptical polarization of the microwave magnetic field within the ferrite. Measurements were made at several frequencies extending from 1000 to 4500 mc. Similar data were taken in coaxial line since the analysis of strip transmission line was approximately valid for this type of line as well.

The resonance attenuation ratios are tabulated in Fig. 7 below the corresponding structures.

It is apparent from the data of Fig. 7 that a well-defined sense of circularity existed at the air-dielectric interface in both coaxial and strip transmission lines. In the coaxial structure the D/d ratio was approximately 3 and the dielectric constant was 10. A high degree of circularity existed from 1500 to 4500 mc—a 3 to 1 frequency band. The dielectric loading of the strip transmission line structure corresponded to a D/d ratio of 2. The dielectric constant of the mode distorting medium was 9. The measurements indicated that a well-defined sense of circularity existed from 1500 mc to at least 4000 mc. It was not feasible to make measurements at higher frequencies, but it is fairly safe to predict that this structure is suitable for applications extending over at least a 3 to 1 frequency band.

Conclusions

It has been shown that in partially dielectric loaded strip transmission line the microwave magnetic field was elliptically polarized at the air-dielectric interface. The degree of elliptical polarization was a function of the dielectric constant, the degree of loading, and the frequency. An expression relating these quantities has been derived. For specific values of dielectric constant and loading, the polarization factor at the dielectric surface was shown to be 1.2 or less over very broad frequency bands. Measurements in ferrite loaded strip transmission line and coaxial structures indicated the existence of a high sense of circularity at the dielectric interface over at least a 3 to 1 band. It is clear from the foregoing analysis and measurements that dielectric loaded strip transmission line and coaxial line are very well suited for broad-band nonreciprocal ferrite applications.

Acknowledgment

The authors are indebted to B. J. Duncan for suggesting the analysis presented in this paper and to L. Swern for his advice during the course of this work.

Ferrite Phase Shifter for the UHF Region*

C. M. JOHNSON†

Summary—**An extremely compact, low-loss, ferrite phase shifter has been developed for the 200 to 800-mc region. It consists of a folded stripline structure approximately 6½ inches long and less than 1 inch square in cross section. The device requires a longitudinal magnetic field of sufficient intensity to place the operating region above resonance. For field swings of about 900 oersteds (from 430 to 1250 oersteds at 400 mc), 360° change in phase shift can be obtained with about 1 db of loss. The phase shifter is reciprocal and shows identical low-power and high-power characteristics up to at least 10-kw peak. Some additional data are included on the operation of the phase shifter down to 10 mc and up to 2000 mc.**

Introduction

Several different types of electronically controllable ferrite phase shifters have been successfully developed for the microwave region. In the UHF region, however, ferrite devices capable of 360° phase shift have usually proven too lossy and bulky to be practical.

In an attempt to overcome these difficulties in the UHF region, a compact, folded stripline, ferrite phase shifter has been developed which produces 360° change in phase shift with very low loss over an extremely wide frequency region.

This phase shifter is operated on the high-field side of resonance, requiring a relatively large magnetic field. In compensation, however, operation in this region eliminates the nonlinear effects usually observed at high RF power levels. This characteristic, along with the fact that the phase shifter is reciprocal, permits its use in both transmitting and receiving systems.

* Manuscript received by the PGMTT, June 2, 1958; revised manuscript received, August 20, 1958. This research was supported by the AF Cambridge Res. Center under Contract No. AF 19(604)-2407.

† Electronic Communications, Inc., Timonium, Md.

Description

Construction

Fig. 1 shows the phase shifter construction. The device is about 6½ inches long and consists of 5 layers of stripline. Each layer is loaded with two 0.40-inch×0.05-inch×6-inch strips of ferrite, one on each side of the center conductor. The center conductor is folded as shown to provide continuity between layers. Thus, the total length of ferrite through which the wave must travel is 32.2 inches, or 82 cm.

Fig. 2 shows a photograph of the complete phase shifter. Input and output lines are standard RG-8/U cables. The transition to stripline is made simply by slotting the center conductor of the coax to receive the rectangular center conductor of the stripline and connecting the coax shield to the stripline ground plates.

In the model shown here the layers of stripline are fastened together by bolts spaced along the sides. These bolts also serve as electrical shorts between the ground plates.

The dimensions of the stripline shown in Fig. 1 are such that

$$\sqrt{\frac{\epsilon_1}{\mu_1}} \qquad Z_0 = 115 \text{ ohms},$$

where Z_0 is the characteristic impedance, ϵ_1 the relative dielectric constant, and μ_1 the relative permeability. For $\epsilon_1 = 11.5$ and $\mu_1 \approx 2$, typical of the ferrite material used, Z_0 is approximately 50 ohms. The permeability, of course, is a variable here, and therefore it is not possible to attain an exact 50-ohm impedance over the full range of phase shift.