A Short Rugged Ferrite Half-Wave Plate for a Single-Sideband Modulator*

One type of microwave phase shifter consists of two quarter-wave plates between which is placed a half-wave plate whose principal axis is rotatable.[1] The phase shift introduced is directly proportional to the angular displacement of the principal axis of the half-wave plate. Continuous rotation of the principal axis causes continuous advancement or retardation of the phase of the signal traversing the phase shifter. This causes a frequency shift of the signal and the device can thus be used as a single-sideband modulator.[2]

A certain amount of work has been carried out using a tube of ferrite in reduced guide. The ferrite used was an experimental ferrite Type MM3 supplied by Marconi's Wireless Telegraph Company, Ltd. The dimensions of the tube are 0.7 inch O.D. ×0.5 inch I.D.×2 inches long. The ferrite tube was loaded with distrene tubes of 0.5 inch O.D. and various I.D. dimensions. The final half-wave plate arrangement is shown in Fig. 1.

Fig. 2 shows a typical phase shift characteristic obtained. The kinks associated with the ordinary wave phase characteristic are a function of the ferrite tube dielectric loading and the input matching arrangement. The input circuit consisted of a single

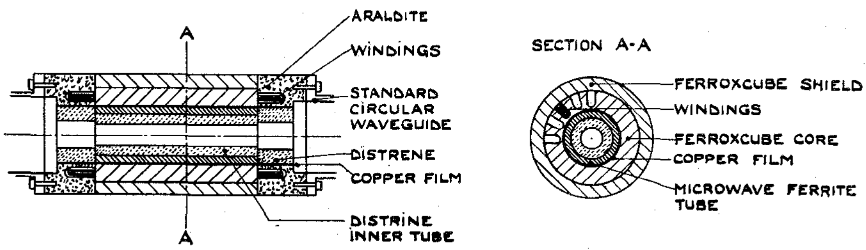

Fig. 1—Cross section of ferrite half-wave plate.

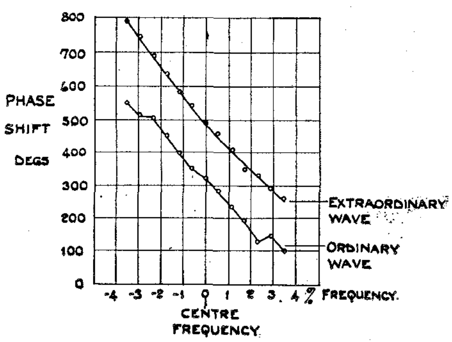

Fig. 2—Phase shift vs frequency characteristic or ordinary and extraordinary wave propagation.

A ferrite half-wave plate in reduced guide has been described by Karayianis and Cacheris.[3] The direction of the applied transverse magnetic field forms the principal axis of the half-wave plate. The advantage of working in reduced guide is that since the guide is more dispersive, a higher differential phase shift per applied field can be obtained. The major disadvantage is the increased difficulties associated with matching the ferrite loaded reduced guide to the normal waveguide run. Matching was attempted by Cacheris and Karayianis by using two 2-inch dielectric tapers. These are difficult to manufacture and are lengthy for some applications.

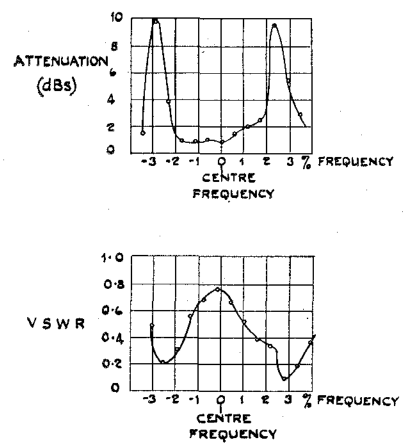

Fig. 3—Attenuation and VSWR vs frequency characteristic of the ferrite half wave plate.

"quarter wavelength" transformer, which was a distrene tube of 0.70 inch O.D. and the I.D. was determined experimentally for any particular frequency range. A typical value was 0.312 inch I.D. The minimum length of the transformer is determined by the winding overlay (Fig. 1). A typical performance curve for this transformer is shown in Fig. 3. It is evident that the insertion loss of the half-wave plate is extremely mismatch sensitive. Where the mismatch is small the insertion loss is approximately 1 db. An impedance plot of this arrangement would not yield a simple theoretical broadband match.

The over-all length of the half-wave plate and matching assembly is less than 3 inches. It could be useful in a single-sideband modulator where a rugged and compact design is desirable, and where narrow band operation is satisfactory.

The author wishes to thank the manager, The English Electric Company, Ltd., Luton, for permission to publish this note and Dr. Benzie of Marconi's Wireless Telegraph Company, Ltd., for supplying the ferrite tubes.

H. I. Glass
Control Dept.
The English Electric Co., Ltd.
Luton, Bedfordshire, Eng.

* Received by the PGMTT, October 9, 1958.

[1] A. G. Fox, "An adjustable wave-guide phase changer," Proc. IRE, vol. 35, pp. 1489–1498; December, 1947.

[2] J. Cacheris, "Microwave single-sideband modulator using ferrites," Proc. IRE, vol. 42, pp. 1242–1247; August, 1954.

[3] N. Karayianis and J. Cacheris, "Birefringence of ferrites in circular waveguide," Proc. IRE, vol. 44, pp. 1414–1421; October, 1956.

A Technique for Minimizing Hysteresis in a 35-DB Ferrite Variable Attenuator*

A requirement arose for a low-power microwave transmitter, the output power of which could be controlled over 40 db with a reset accuracy of 0.5 db for single frequency operation.

The experimental arrangement used is shown in Fig. 1.

The electronically variable short circuit is shown in Fig. 2. This has been described by Scharfman.[1]

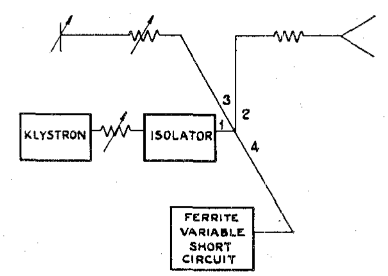

Fig. 1—Variable attenuator.

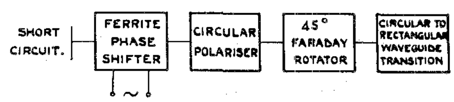

Fig. 2—Electronically variable short circuit.

For one sense of circular polarization the slope of the phase shift vs field curve becomes zero as the ferrite saturates, but for the other sense this is not so marked. This is shown in Fig. 3. Fig. 4 shows the attenuation between arms 1 and 2 vs relative phase difference between arms 3 and 4 of the magic T. It can be seen that the slope of the attenuation vs relative phase shift characteristic curve is extremely steep at the maximum attenuation point. Consider a negatively circularly polarized wave fed into a ferrite loaded section which is subjected to a field sufficiently large to saturate the ferrite. This corresponds to the point P in Fig. 3. The attenuator and short circuit in arm 3 can now be adjusted to give maximum attenuation between arms 1 and 2. This

* Received by the PGMTT, October 9, 1958.

[1] H. Scharfman, "Three new ferrite phase shifters," Proc. IRE, vol. 44, pp. 1456–1459; October, 1956.

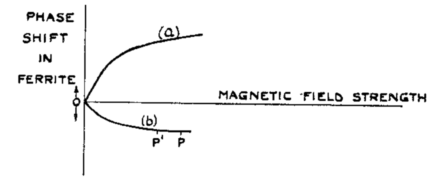

Fig. 3—Phase shift vs applied field. (a) Positively polarized wave. (b) Negatively polarized wave.

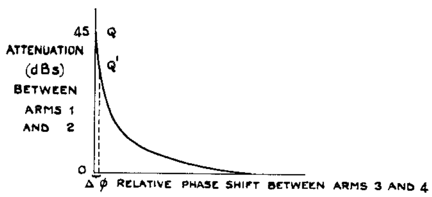

Fig. 4—Attenuation vs relative phase shift.

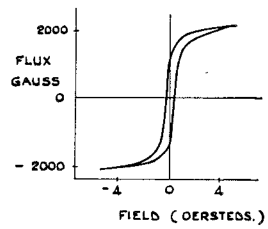

Fig. 5—Hysteresis loop of typical microwave ferrite.

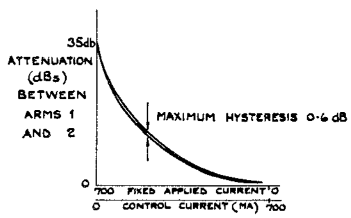

Fig. 6—Attenuator characteristic.

corresponds to the point Q in Fig. 4. If the magnetic field on the ferrite is now changed by a relatively large amount (to point P' in Fig. 3), the actual change in phase in arm 4 is quite small. This results in a small relative phase shift between arms 3 and 4 and this in turn causes the attenuation between arms 1 and 2 to change to the amount corresponding to Q' (Fig. 4). Thus it is obvious that on plotting a curve of attenuation between arms 1 and 2 vs magnetic field applied to the ferrite, a characteristic is obtained whose slope near the maximum attenuation point Q is considerably less steep than that of the curve of Fig. 4. This is shown in Fig. 6.

Since hysteresis is very small near saturation its effect near the steep part of the characteristic of Fig. 6 is very small. Below saturation the hysteresis of the ferrite is more marked (Fig. 5), but since the slope of the characteristic of Fig. 6 is much smaller when the applied field decreases, the effect of this increase in hysteresis is minimized. The final curve for the attenuator is shown in Fig. 6 where it can be seen that the maximum hysteresis measured corresponds to 0.6 db.

The author wishes to thank the manager, the English Electric Company, Ltd., Luton, for permission to publish this note.

H. I. Glass
Control Dept.
The English Electric Co., Ltd.
Luton, Bedfordshire, Eng.

Quarter-Wave Compensation of Resonant Discontinuities*

Introduction

In designing RF transmission line components it is often necessary to place a short-circuited quarter-wavelength stub in parallel with the line or an open-circuited quarter-wavelength stub in series with the line. The stub can be broadbanded by merely changing the characteristic impedance of the line on either side of the stub for a distance of one quarter wavelength.

Broad-Band Stub

It is not generally recognized how broadband a simple stub with quarter-wave transformers can be made. Previous investigators[1] have merely adjusted the transformer impedance for perfect match at two frequencies which depart somewhat from the resonant frequency of the stub without regard for the reflection in the pass band. The following analysis tries to correlate the bandwidth with the allowable reflection in the pass band.

A coaxial broad-band stub is shown in Fig. 1. On each side of the stub the center conductor is enlarged for a length of $\lambda_0/4$ at the center frequency. In these quarter-wave transformers the characteristic impedance is Z_1. The stub is $\lambda_0/4$ long and its characteristic impedance is Z_2. The characteristic impedance of the line is taken as 1 ohm in what follows so that Z_1 and Z_2 are multiples of the characteristic impedance.

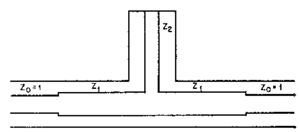

Fig. 1—Coaxial broad-band stub.

The $ABCD$ matrix of the stub plus transformers is

$$\begin{bmatrix} \cos\theta & jZ_1\sin\theta \\ j(1/Z_1)\sin\theta & \cos\theta \end{bmatrix}\begin{bmatrix} 1 & 0 \\ -j(1/Z_2)\operatorname{ctn}\theta & 1 \end{bmatrix}\cdot\begin{bmatrix} \cos\theta & jZ_1\sin\theta \\ j(1/Z_1)\sin\theta & \cos\theta \end{bmatrix}$$

where θ is electrical length of each quarter-wave transformer and the stub. If we let $\theta=\pi/2+\phi$, the over-all matrix becomes

$$\begin{bmatrix} -\sin\phi & jZ_1\cos\phi \\ j(1/Z_1)\cos\phi & -\sin\phi \end{bmatrix}\begin{bmatrix} 1 & 0 \\ j(1/Z_2)\tan\phi & 1 \end{bmatrix}\cdot\begin{bmatrix} -\sin\phi & jZ_1\cos\phi \\ j(1/Z_1)\cos\phi & -\sin\phi \end{bmatrix}$$

which when multiplied gives

$$\begin{Bmatrix} \sin^2\phi-\cos^2\phi+(Z_1/Z_2)\sin^2\phi & -jZ_1\sin\phi\cos\phi(2+Z_1/Z_2) \\ -j(1/Z_1)\sin\phi\cos\phi(2-Z_1/Z_2\tan^2\phi) & \sin^2\phi-\cos^2\phi+(Z_1/Z_2)\sin^2\phi \end{Bmatrix}.$$

The insertion loss is given by[2]

$$\begin{aligned} L &= 10\log_{10}\{1+1/4[(A-D)^2-(B-C)^2]\} \\ &= 10\log_{10}\{1+1/4[(2/Z_1-2Z_1-Z_1^2/Z_2+1/Z_2)\sin\phi\cos\phi-1/Z_2\tan\phi]^2\} \\ &= 10\log_{10}(1+m^2/4) \end{aligned} \quad (1)$$

where

$$mZ_2 = R\sin\phi\cos\phi - \tan\phi \quad (2)$$

$$R = 2Z_2/Z_1 - 2Z_1Z_2 - Z_1^2 + 1. \quad (3)$$

A graph of the magnitude of $|m|Z_2$ is shown in Fig. 2.

$R=1$ gives the maximally flat case with a zero derivative at the origin.

For R greater than 1, a triple peaked response is obtained.

Using some simple trigonometric substitutions it can be shown that $\phi_2=2\phi_1$; also,

$$m_1Z_2 = \tan\phi_2\left(\frac{1-\cos\phi_2}{1+\cos\phi_2}\right) \quad (4)$$

and

$$R = \frac{2}{\cos^2\phi_2+\cos\phi_2} \quad (5)$$

where ϕ_1 is the value of ϕ for worst reflections in the pass band, ϕ_2 is the band edge, and m_1 is the worst value of m in the pass band.

The quantity m_1 is related to the worst voltage standing wave ratio S by

$$m_1 = \frac{S-1}{\sqrt{S}} \quad (6)$$

and the bandwidth is given by

$$BW = 2\phi_2/90. \quad (7)$$

A graph of m_1Z_2 as a function of bandwidth is shown in Fig. 3.

As an example, suppose it is desired to design a stub support for a coaxial line to have a standing wave ratio of no greater than 1.05 over as wide a frequency band as possible. Because of voltage breakdown considerations it is decided that the largest value Z_2 may have is one. Then from (6), $m_1=0.0488$, and from Fig. 3 the bandwidth is 70.4 per cent or a frequency ratio of 2.09:1. R is determined from (5), and Z_1 from (3).

The required value of Z_1 for various values of Z_2 is plotted as a function of bandwidth in Fig. 4. This graph shows that the diameter of the quarter-wave transformers is rather critical. The desired Z_1 is only slightly smaller than the zero bandwidth case.

* Received by the PGMTT, October 27, 1958. The research in this document was supported jointly by the Army, Navy, and Air Force under contract with Mass. Inst. Tech.

[1] G. L. Ragan, "Microwave Transmission Circuits," M.I.T. Rad. Lab. Ser., McGraw-Hill Book Co., Inc., New York, N. Y., vol. 9, pp. 173–176; 1948.

[2] R. M. Fano and A. W. Lawson, "Microwave Transmission Circuits," M.I.T. Rad. Lab. Ser., McGraw-Hill Book Co., Inc., New York, N. Y., vol. 9, ch. 9 and 10; 1948.