change. However, the stray coupling of the initiating pulse was reduced by 30 dB. This invariance of the delay modes with changes in coupling orientation indicates that these spin waves are circularly polarized [4].

The sample employed in this experiment was a parallelepiped, with dimensions of $0.440 \times 0.142 \times 0.117$ inch and oriented so that the [111]-direction was parallel to the rod axis. The static magnetic field was applied parallel to the rod axis and the measurements were conducted at a frequency of 2 Gc/s, using pulses of 0.5-μs width and 1-mW peak power.

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References

[1] W. Strauss, "Magnetoelastic waves in Yttrium iron garnet," *J. Appl. Phys.*, vol. 36, p. 118, January 1965.
[2] J. R. Eshbach, "Spin-wave propagation and the magnetoelastic interaction in Yttrium iron garnet," *J. Appl. Phys.*, vol. 34, p. 1298, April 1963.
[3] F. A. Olson, and L. D. Buchmiller, "A two-port microwave delay line," *1964 IRE PTGMTT International Symp. Digest*, p. 80.
[4] W. Strauss, E. Strauss, and G. Frieder, "Evidence for polarization reversal of magnetoelastic waves," *Appl. Phys. Letters*, vol. 6, p. 18, January 1965.

Synthesizing Air with a Radome Sandwich

It is possible to design lossless radome sandwiches which have a transmission coefficient of unity and an insertion phase angle of zero. If the magnitude and phase of the electric field within the sandwich approximates that of the electric field in air, the edge diffraction of the sandwich should be small.

Consider the symmetrical sandwich, shown in Fig. 1, consisting of three lossless conducting films or grids of negligible thickness and two layers of an ideal dielectric. Let ϵ denote the relative dielectric constant of the dielectric layers, B_e denote the susceptance of the outside films or grids, B_m denote the susceptance of the center film or grid, θ denote the angle of incidence in air, and λ denote the wavelength in air. It is assumed that the relative permeability of the dielectric is unity. For convenience in notation, let

$$D = 2\pi d/\lambda,$$
$$K = \sqrt{\epsilon - \sin^2\theta},$$
$$k = \cos\theta.$$

For perpendicular polarization, the transmission coefficient is unity and the insertion phase angle is zero if

$$B_e = K \cot KD - k \cot kD,$$
$$B_m = 2K \cot KD - \frac{2K^2 \cos kD \sin kD}{k \sin^2 KD}.$$

The corresponding values for parallel polarization are

$$B_e = (\epsilon/K) \cot KD - (1/k) \cot kD,$$
$$B_m = 2(\epsilon/K) \cot KD - \frac{2k\epsilon^2 \cos kD \sin kD}{K^2 \sin^2 KD}.$$

These equations can be derived by using the transmission-line analogy for radome sandwiches [1], [2].

For $\epsilon = 4$ and $\theta = 0$, these equations become

$$B_e = -\tan D,$$
$$B_m = 4B_e.$$

A sandwich may be designed for use at a frequency of 9375 Mc/s, using fiberglass laminate with $\epsilon = 4$ and parallel wires. Let $d = 0.05\lambda = 0.062$ inch. Then $B_e = -0.325$, which corresponds to 0.005-inch wires separated approximately 0.45 inch, and $B_m = -1.3$, which corresponds to 0.005-inch wires separated approximately 0.19 inch. Although such wire grids do not satisfy the ideal assumptions, these parameters indicate that an ideal sandwich can be approximated.

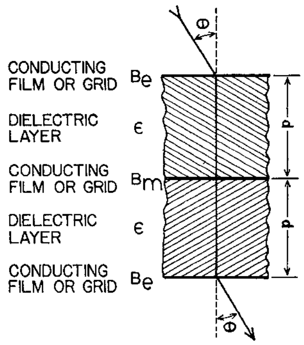

Fig. 1. Symmetrical sandwich.

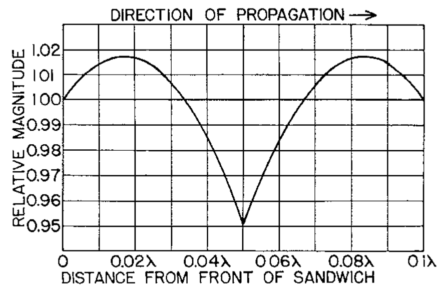

Fig. 2. Relative magnitude of the electric field within the sandwich.

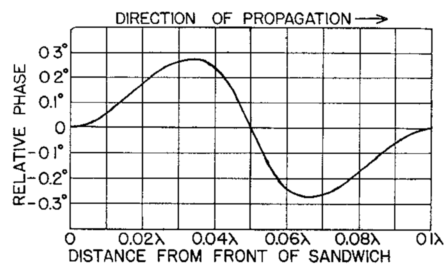

Fig. 3. Relative phase of the electric field within the sandwich.

The relative magnitude and phase of the electric field within an ideal sandwich with $\epsilon = 4$, $\theta = 0$, $d = 0.05\lambda$, $B_e = -0.325$, and $B_m = -1.4$ compared to that in air are shown in Figs. 2 and 3. It should be observed that the electric field within the sandwich does not deviate greatly from what it would be in air.

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References

[1] H. F. Mathis, "Transmission characteristics of sandwiches," *IRE Trans. on Microwave Theory and Techniques*, vol. MTT-3, pp. 57–58, October 1955.
[2] H. F. Mathis, "Transmission line analogy for sandwich propagation," *Electronics*, vol. 33, pp. 100, 102, May 20, 1960.

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A Variable Harmonic Phase Delay Coaxial Network

A simple, adjustable, harmonic phase delay, equalizer network has been developed in coaxial transmission line. A harmonic equalizer network is defined as a circuit whereby the phase shift or phase delay through the device between two harmonically related or widely separated frequencies differs from an ideal dispersionless circuit. This circuit makes use of the phase shift properties of a reactive cutoff type network as the cutoff frequency is varied. A coaxial low-pass filter type network was chosen because of the mechanical simplicity that this type of transmission medium yields, in addition to the broad range of operating bandwidths available. Figure 1 shows the dispersion relationship between an ideal transmission line and a low-pass filter. Whatever the phase relationship of a fundamental signal f_0 with respect to its harmonic $2f_0$ at the input to the filter, the output phase relationship will be changed by $\phi°$. This relative phase shift is realized due to the characteristics of the filter network near cutoff. For example, in a Constant-k design, the total phase shift through the filter network is a function of the number n of LC filter sections employed and the proximity of the frequency f of interest with respect to the cutoff frequency f_c. If the cutoff frequency is varied as shown in Fig. 1 from f_{c1} to f_{c2}, the relative harmonic phase shift ϕ may be computed, depending on the cutoff condition, according to the general relation

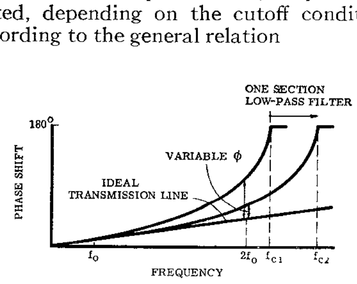

Fig. 1. Phase vs. frequency characteristics of different transmission networks.

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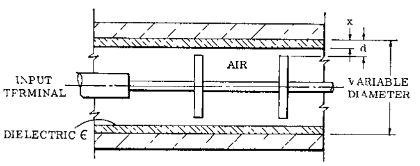
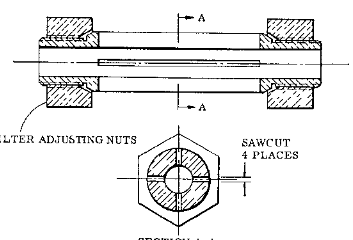


(a)



(b)

Fig. 2. (a) Open condition of variable cutoff low-pass coaxial filter. (b) Mechanical diagram showing the simplicity of the filter adjustment mechanism.

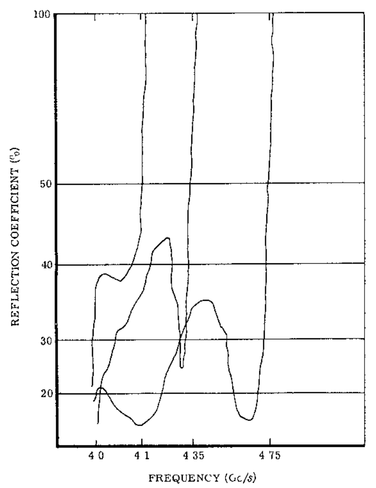


Fig. 3. Reflection coefficient for three conditions of a variable cutoff low-pass filter.

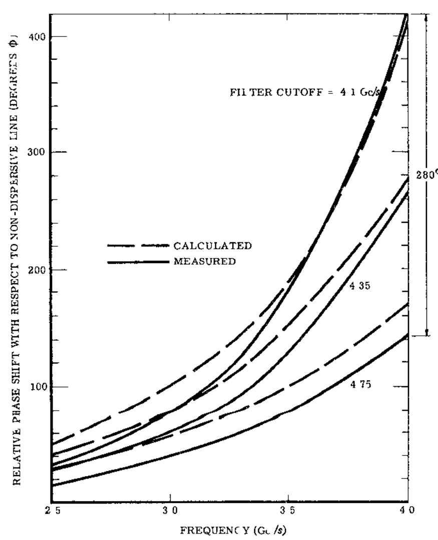


Fig. 4. Correlation of theoretical and measured phase shift data for three conditions of the variable cutoff low-pass filter.

$$\phi = 114.6n\left[\operatorname{Arcsin}\left(\frac{f}{f_c}\right) - \frac{f}{f_c}\right] \text{degrees.} \quad (1)$$

For the case $f=2f_0$ in (1), the change in ϕ at f_0 is assumed negligible because of the frequency separation of f_0 with respect to either f_{c1} or f_{c2}. In almost any application, it must be noted that the frequency of interest must not be selected too close to the filter's cutoff frequency if the reflection coefficient is to be acceptable.

Figure 2(a) exhibits the practical coaxial low-pass network. The smaller center conductor high impedance line sections represent the required series inductance whereas the low impedance ring sections produce the desired amount of shunt capacitance. Upon cutting four equally spaced axial slots in the outer conductor, as shown in Fig. 2(b), this outer diameter was made quite flexible by applying concentric external pressure through the use of adjusting nuts. The dielectric shown was a sheet wrapped inside the outer coax cylinder to partially fill up the gap between the center ring conductors and the outer coaxial shell. Thus, by varying the designated air gap $(d-x)$, the shunt capacitance is made adjustable with a consequent change in the filter cutoff frequency. The ratio of the cutoff frequencies for the two limiting conditions (fully open and fully closed) can be approximately equated as,

$$\frac{f_c(\text{open})}{f_c(\text{closed})} \cong \sqrt{\frac{\epsilon}{1 + \frac{x}{d}(\epsilon - 1)}}. \quad (2)$$

Normally, the lowest cutoff frequency f_c (closed) is initially determined when designing the filter network. Then the upper cutoff frequency f_c (open) may be obtained from (2) which should be good for ratios of f_c (open) to f_c (closed) up to at least 1.25.

An S-band harmonic phase delay network was designed to exhibit a maximum cutoff variation from about 4.0 to 4.8 Gc/s through the use of (2). The dielectric material used in this ten section filter was Aclar, which has a relative dielectric constant of approximately 2.5. Three of the cutoff conditions are shown in Fig. 3 as a function of the reflection coefficient characteristics from 4 to 5 Gc/s. The reflection coefficient was under 15 percent at all frequencies below 3.9 Gc/s for each cutoff condition shown.

Correlation of the theoretical with measured relative phase shift data for the three cutoff conditions is graphically related in Fig. 4. The theoretical phase shift was determined by (1). Considering that small mismatches occurred over the band, 2.5 to 4.0 Gc/s, each correlation is quite good. Thus, a continuously controllable difference in phase shift of about 280°, at 4 Gc/s with respect to 2 Gc/s, was observed in this simple coaxial configuration which proved to be adequate for a particular circuit application. However, it should be observed that a 360° variation could have readily been realized by increasing the number of filter sections. In addition, it should also be noted that the frequencies of concern do not necessarily have to be harmonically related. However, the closer the frequency separation, the smaller will be the relative phase difference in any given filter. If both frequencies of concern were closely spaced near the filter's cutoff frequency, then, the actual phase shift or phase delay would be the relative angle difference $\phi_2-\phi_1$.

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A Semi-Automatic Technique for Tuning a Reflectometer

Introduction

Tuned microwave reflectometers are used in single-frequency high-precision microwave measurement applications. The tuners are adjusted such that a constant detector output is achieved when the reflectometer is terminated by loads of constant reflection coefficient magnitude and variable phase. At present, the phase variation is achieved by sliding the loads inside a terminating section of uniform waveguide.

The tuning procedure is accomplished in two steps: 1) tuner x, as shown in Fig. 1, is adjusted for no variation in the detected output as a low reflection load is moved inside the terminating waveguide, and 2) tuner y is adjusted for no output variation as a short-circuit is moved inside this waveguide [1]–[3].

This correspondence describes a method which substantially shortens the time required for tuning the reflectometer by automatically driving the load inside the uniform waveguide and synchronously displaying the tuning response on an oscilloscope. The application of this method to existing microwave systems employing the reflectometer technique requires the utilization of additional equipment, which is commercially available, except for one item, the carriage.

Description of Equipment Used

Figure 1 shows the simplified diagram of the reflectometer in the left part and the added equipment required for the application of this technique in the right part.

In the tuning process two types of loads are used, a large reflection and a low reflection termination, as in Fig. 2. Teflon spacers are used to support the loads inside the waveguide and to reduce wear of the guide.

The reciprocating motion of the load is accomplished by attaching the load to a driven carriage. The carriage, shown in Fig. 3, was made in the National Bureau of Standards instrument shop.[1] It converts the oscillating rotational motion of a commercially available "sweep drive" output into reciprocating longitudinal motion. Adjustments are provided on the carriage for alignment.

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[1] This carriage is more elaborate than required but was available and satisfactory for demonstrating the technique.