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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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.

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.

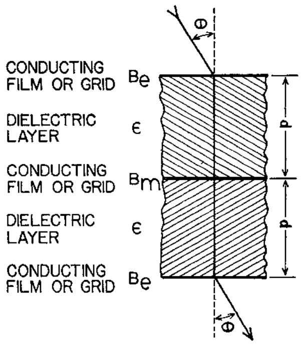

Fig. 1. Symmetrical sandwich.

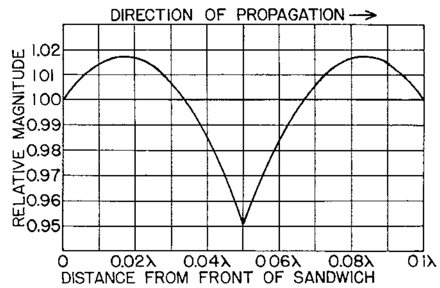

Fig. 2. Relative magnitude of the electric field within the sandwich.

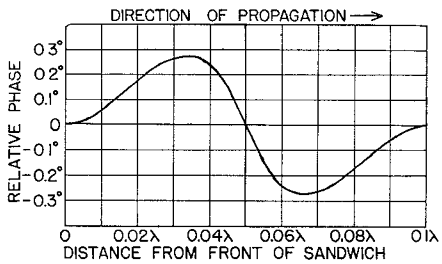

Fig. 3. Relative phase of the electric field within the sandwich.

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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

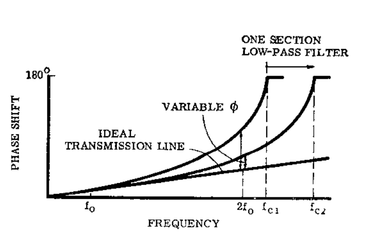

Fig. 1. Phase vs. frequency characteristics of different transmission networks.

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