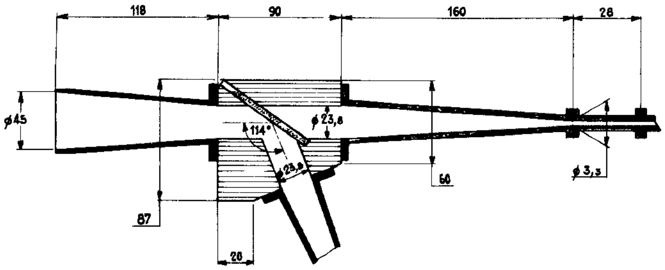

Fig. 3. Cross section of the plate diplexer. (Dimensions are in millimeters.)

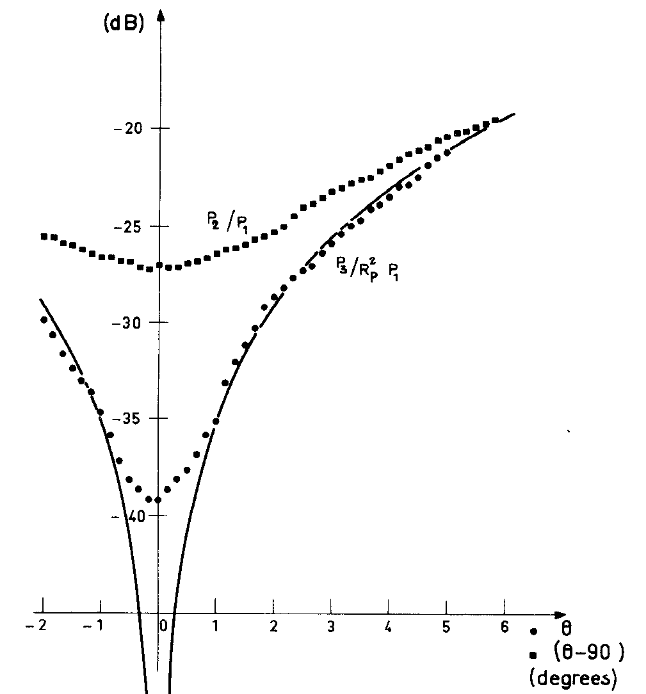

Fig. 4. Normalized power at the reflection output $P_3/R_p^2P_1$ versus θ, and normalized power at the transmission output P_2/P_1 versus $(\pi/2-\theta)$, for the plate diplexer at 70 GHz.

mission lines [3] shows that for 70 GHz, the metal grid with the dielectric substrate gives a reflected power $(P_3/P_1)_0=-12$ dB when $\theta=90°$. But the rectangular waveguide, placed in the right position at the reflection arm, cuts off this spurious power, and experimentally we obtained $(P_3/P_1)_0=-40$ dB.

In future versions of the grid diplexer, more appropriate metal grids must eliminate this small but existing asymmetry (which was not observed at 890 GHz).

We shall now describe the plate diplexer, shown in Fig. 3. It is similar to the grid diplexer with the difference that the grid is replaced by a dielectric polyethylene plate and that the angle of incidence is not 45° but 57°, which is very near the Brewster's angle for polyethylene.

The Brewster's-angle effect is essentially asymmetric because the incident field parallel to the plane of incidence is completely transmitted, but the incident field perpendicular to this plane is only partially reflected. The reflected power is $P_3=R_p^2P_1$ where R_p is the reflection coefficient at the Brewster's angle, and it is maximum when the relation between the thickness of the plate l and the wavelength λ is

$$\frac{l}{\lambda}=\frac{2K+1}{4n\cos i_B}, \qquad K=0,1,2,\cdots$$

where n is the refractive index and i_B is the Brewster's angle. In the best conditions R_p^2 is of the order of 3 dB.

The plate diplexer is then much more asymmetric than the grid diplexer, and we see in Fig. 4 the difference between the zero of P_2/P_1 near $\theta=90°$ and the zero of $P_3/R_p^2P_1$ near $\theta=90°$, measured at 70 GHz. The value $(P_2/P_1)_0=-27$ dB shown in Fig. 4 is justified by the polarizing effect of the rectangular waveguide at the transmission arm, because with a circular-waveguide output it would be $(1-R_p^2)$, say slightly greater than -3 dB.

The value $(P_3/R_p^2P_1)_0=-37$ dB indicates an effective directivity greater than that of the fin-line diplexer. This means that for angles of incidence $\theta=0°$ the plate diplexer has a better accuracy for little $\delta\theta$ variation. The performance of this diplexer is inferior to that of the grid diplexer, but on the other hand it is extremely simple and of negligible cost. Measurements with the HCN laser beam gave $(P_3/R_p^2P_1)_0=-27$ dB.

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Modified H Guide for Millimeter and Submillimeter Wavelengths

A. DOSWELL AND D. J. HARRIS

***Abstract*—A theoretical and experimental investigation is being made of a modified form of H guide as a possible guided wave structure for millimeter and submillimeter wavelengths. The effects of channels in the conducting planes to support a dielectric film has been studied by scaled-up models at 3-cm wavelength. Low losses, even at the shorter wavelengths, are predicted. The channels may be used to filter unwanted higher order modes, and the use of high-permittivity dielectric is suggested to further reduce the guide attenuation.**

I. Introduction

Conventional guiding structures for electromagnetic waves at microwave frequencies present problems of high attenuation and small physical size when scaled for shorter millimeter and submillimeter wavelengths. No guiding structure has yet been devised which is ideal for such wavelengths, but H guide, first described by Tischer [1], has potential application at these wavelengths.

H guide utilizes a dielectric sheet which is perpendicular to two conducting planes, forming a letter H in cross section as in Fig. 1(a), to propagate a surface wave. It can be shown that one mode, the $_0PM_{11}$, which has an electric field distribution mainly parallel to the conducting planes and a magnetic field entirely parallel to the dielectric surface, exhibits a low-loss wall attenuation characteristic as for the TE_{01} mode in circular guide. It is thus an obvious candidate for high-frequency low-attenuation transmission.

II. Attenuation

Total, i.e., wall and dielectric, attenuation of H guide calculated for submillimeter wavelengths using standard techniques is shown in Fig. 2. Curve (a) has been calculated for a guide width of 1 mm, (b) for a guide width of 1.5 mm, and (c) for a guide width of 2 mm. In all cases, a dielectric loss angle of 10^{-3} and an effective metal conductivity of 10^7 S/m has been assumed. The metal conductivity value is of the order indicated by the work of Bled *et al.* [2] and Vershinina and Meriakri [3]. For comparison, the experimental values of attenuation for rectangular guide obtained from [2] and [3] are also shown in Fig. 2.

III. Mode Purity

By choosing a suitable dielectric thickness and by use of an appropriate launching device, it should be possible to excite only the $_0PM_{1n}$ modes where $n=1, 3, 5$. As the cutoff plane separation of

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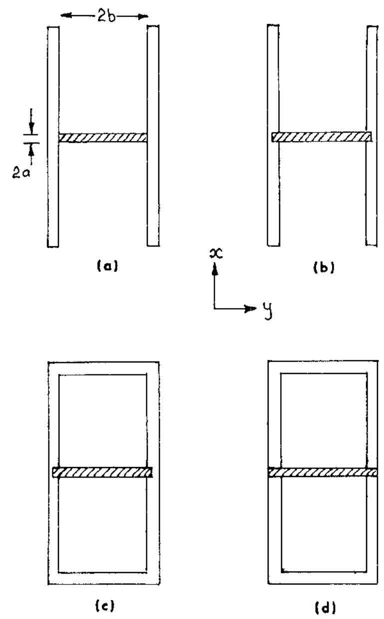

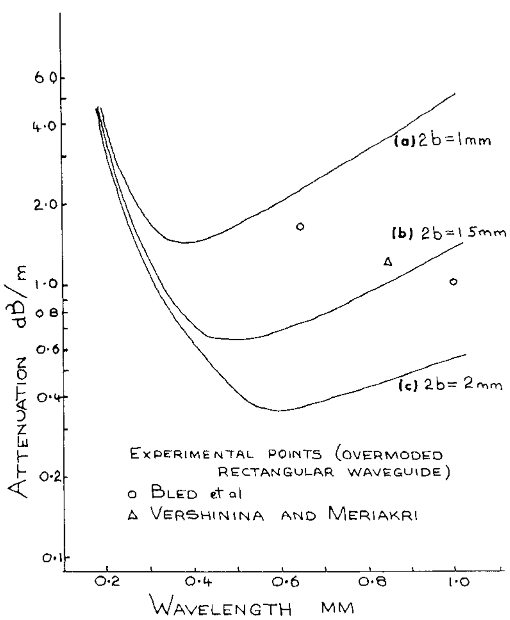

Fig. 2. Total theoretical H-guide attenuation for $\epsilon_r = 2.1$, $\phi_d = 10^{-3}$, $\tau = 10^7$ S/m, and $2a = 15\ \mu$m.

the $_0PM_{1n}$ modes is given approximately by $n\lambda_0/2$, it can be seen that with a 2.0-mm-wide guide operating at 0.5 mm, only the $_0PM_{11}$ and $_0PM_{13}$ modes would be excited. Reduction of the guide width so that only the $_0PM_{11}$ mode propagates would not be feasible because of the high wall attenuation of the $_0PM_{11}$ mode for such small plane separations. Reduction of the $_0PM_{11}$ mode attenuation by increasing the plane separation would introduce further higher order modes.

IV. Guide Design for Submillimeter Wavelengths

H-guide construction for wavelengths between 100 μm and 1 mm requires the use of a thin dielectric film, and the design would have to be modified to include some means for locating and holding the dielectric between the conducting planes. Schematic suggestions for guide construction are shown in Fig. 1(b), (c), and (d).

The design of (c) and (d) makes the guide look like a partially dielectric loaded hollow rectangular guide, and the mathematical treatment is similar for the two types of guide [4]. A design similar to (c), the microguide, has been suggested by Karbowiak [5], but the field mode in the microguide is quite different.

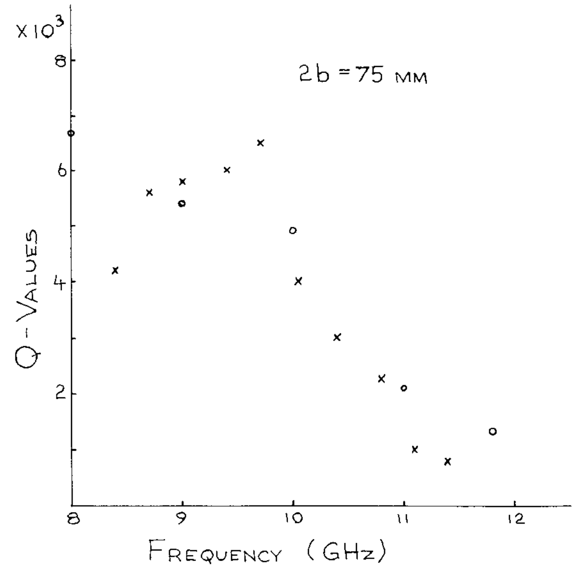

Fig. 3. Experimental and theoretical Q values. X are experimental points and O are values from simplified theory.

The inclusion of the wall parallel to the dielectric alters the H-guide surface-wave exponential damping term outside the dielectric into a hyperbolic function. Physically, the extra wall will have little effect if most of the energy is located close to the dielectric, and experimental work on scaled models of the guide at X band confirms this.

Treating the channels cut into the walls of the guide, Fig. 1(b) and (c), as lengths of hollow dielectric-loaded rectangular guide supporting a TE_{10} mode excited by the z-direction magnetic field of the $_0PM_{11}$ mode, the increased attenuation of the $_0PM_{11}$ mode can be calculated. These calculations indicate that the channel attenuation will generally be small compared to the intrinsic guide attenuation.

The wall power loss of the $_0PM_{13}$ to the $_0PM_{11}$ mode, calculated from the appropriate magnetic field component and the channel impedance, shows that this ratio can be high, suggesting that the $_0PM_{13}$ mode could be filtered out. Experimental work at X band on scaled models of the proposed submillimeter structures, using a resonant length of H guide, has verified the basis of the calculation of the channel attenuation and also the mode filtering property.

Fig. 3 shows experimentally measured Q values of an X-band H-guide resonator which had channels cut in its plates compared to the theoretical Q values of the resonator as calculated from the intrinsic guide and channel attenuation. It can be seen that the simple theory used for the calculations produces reasonably good agreement and justifies the basis of the theoretical approach. The mode filtering property is demonstrated in Fig. 4 where pen recordings of resonance spectra of H-guide resonators with and without the channels present indicate in (a) resonances due only to the $_0PM_{11}$ mode, and in (b) resonances due to both the $_0PM_{11}$ and $_0PM_{13}$ modes. The resonances have been identified with particular modes by comparison with the theoretically calculated resonance spectra, by cutting off the $_0PM_{13}$ mode using narrower plane separations and by use of a probe to investigate the transverse field distributions in the guide.

V. Dielectric and Radiation Loss

Investigation of the dependence of dielectric loss upon relative permittivity and dielectric thickness has shown that increasing the relative permittivity of the dielectric can result in lower dielectric losses. Fig. 5 shows calculated dielectric loss for various thicknesses and relative permittivities of dielectric at constant dielectric loss angle. The thinner the dielectric, the more noticeable the variation of loss with permittivity. The optimum permittivity for a particular thickness can be calculated. Making the dielectric thin ensures that a major part of the dielectric loss is determined by the x-direction component of the electric field. The choice of a suitable relative per-

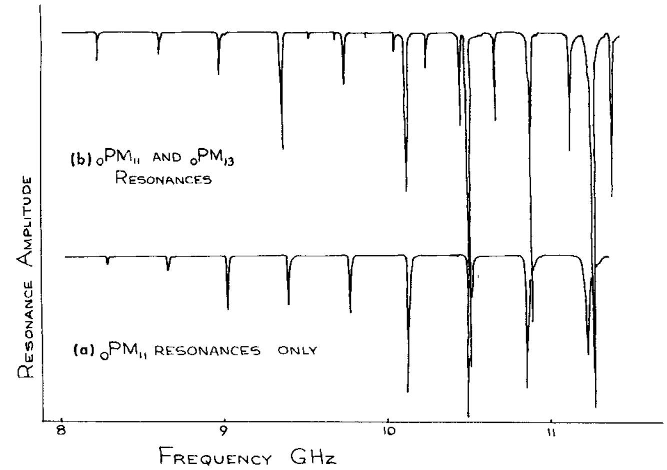

Fig. 4. Resonance spectra in (a) channelled and (b) unchannelled guides.

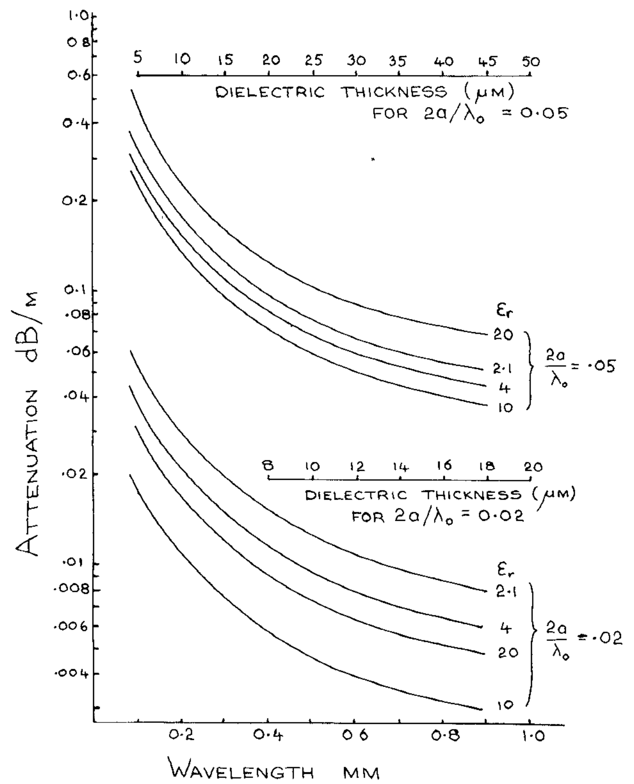

Fig. 5. Variation of dielectric attenuation with thickness and permittivity.

mittivity results in a large part of the electromagnetic energy traveling outside the dielectric and, therefore, not contributing to the dielectric loss.

Radiation loss from the open H-guide structure has been shown to be significant in certain circumstances [6]. It can, however, be reduced to a low value by increasing the height of the guide walls.

VI. Conclusions

Modified H-guide, in which a thin dielectric film is supported across the guide by channels in the conducting planes, appears potentially to offer advantages as a guiding structure for short millimeter and submillimeter wavelengths. The channel may be effective in suppressing higher order modes and allowing wider plane separation and, therefore, lower loss. Possible forms of construction are shown in Fig. 1(c) and (d). The use of a higher permittivity thin film of dielectric may reduce the loss still further. H-guide structures with channels have been investigated experimentally at 3-cm wavelength and the performance outlined verified. Guides are now being constructed for 0.3-mm wavelength operation for investigation using our HCN laser.

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A Simplified Circuit Model for Microstrip

HERBERT J. CARLIN

The advantage of a network model for a physical structure is that the model, if correctly established, implicitly contains the physical constraints of the actual system, and these constraints need not subsequently be called into play for every new case. A recent example is the use of coupled lines [1] to model longitudinally uniform but transversely inhomogeneous waveguides. The network model for a cylindrical waveguide loaded concentrically with a dielectric rod comprised a TE and a TM transmission line coupled together, and the properties of this model demonstrated that, surprisingly, the smooth lossless waveguide structure could support complex eigenvalues as well as backward waves. The general network idea stems from Schelkunoff [2] who established that uniform metallic-bound lossless guide structures can be represented by an infinite number of coupled TE and TM transmission lines. The practical approximating network model is obtained by appropriately truncating the infinite Shelkunoff representation [1].

In this short paper we show how a pair of coupled lines can give an extremely simple model for microstrip dispersion. We take a TEM transmission line and a TE line and form a distributed circuit with these two lines coupled together. The uncoupled lines propagate the ordinary TEM and TE modes. The coupled circuit automatically represents a pair of modes which are no longer TEM or TE, but instead are the two lowest order hybrid modes that exist on the stripline. In effect, circuit theory does the work in producing the required modes.

The pair of coupled lines modeling the microstrip is shown in Fig. 1. The circuit model for the physical structure is based on the fact that TEM- and TE-type modes excite each other by virtue of the presence of the dielectric substrate. It is also assumed in the model that the uncoupled TEM and TE modes propagate at the same velocity at very high frequencies, i.e., there is a common value of $\bar{\epsilon}$ for both of the lines.

The series-impedance and shunt-admittance matrices *per unit length* for the pair of coupled lines in Fig. 1 are

$$Z = p\mu_0 \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix} \qquad Y = p\epsilon_0 \begin{bmatrix} \bar{\epsilon} & c_{12} \\ c_{12} & \bar{\epsilon} \end{bmatrix} + \frac{1}{\mu_0 p} \begin{bmatrix} 0 & 0 \\ 0 & \mathcal{K}^2 \end{bmatrix}. \tag{1}$$

Here $p = \sigma + j\omega$ is complex frequency, and ϵ_0, μ_0 are the constitutive constants of free space. There are therefore only three constants used for the simple circuit model: $\bar{\epsilon}$, the *effective* static dielectric constant; $\mathcal{K}$, the cutoff wavenumber for the uncoupled TE mode; and the coupling capacitance $c_{12} = k\bar{\epsilon}$, where $0 \leq k \leq 1$ is the capacitive coefficient of coupling. The effective dc dielectric constant is given by the static relation

$$\bar{\epsilon} = \left(\frac{z_0}{\bar{z}_0}\right)^2 \tag{2}$$

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