Wide-Band Resonance Isolator*

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Summary—A parallel-plate transmission line loaded with capacitors or high dielectric constant material along a narrow strip has a circularly polarized RF magnetic field everywhere external to the loading over a very broad band. The magnetic resonance line of a narrow linewidth ferrite was inhomogeneously broadened by a very inhomogeneous magnetic field to provide resonance absorption over a wide frequency range. A prototype structure has given better than 15 db per inch attenuation in the reverse direction over a bandwidth from 1.5 kMc to 6.0 kMc. The forward loss caused by the ferrite is about 0.2 db to 0.4 db over this range of operation.

I. INTRODUCTION

A GENERAL rule for developing microwave devices is—once it works, broad-band it. Isolators are no exception to this rule. The development of broad-band resonance isolators is characterized by the work of Weiss[1] and Duncan, Swern, Tomiyasu, and Hannwacker.[2] The latter authors describe a device which operates over an octave bandwidth beginning at 2 kMc. Recently, Jones, Matthaei and Cohn[3] described a broad-band isolator using the Faraday rotation effect in a ferrite.

In studying the propagation of TE waves between parallel plates, it is found that the RF magnetic field is elliptically polarized if the wave is slowed down by some smooth or fine-grained structure along the direction of propagation. In particular, the polarization may be made to approach arbitrarily close to circular polarization for sufficient slowing of the wave. The ratio of transverse to longitudinal components of the RF magnetic field is given by

$$\frac{H_x}{H_z} = \frac{-j}{\sqrt{1 - \frac{k^2}{\beta^2}}}, \tag{1}$$

where β is the propagation constant along the structure and k is the free-space propagation constant. For infinite parallel plates, this relation holds everywhere external to the slowing structure! Thus, for all frequencies for which $\beta \gg k$, we will have nearly perfect circular polarization.

Given circular polarization, how do we obtain broad-band resonance isolation? This problem is considered in Section III of this paper. Fortunately, it is possible to broaden inhomogeneously the ferrimagnetic resonance line of a ferrite in such a way that strong absorption takes place at locations in the structure where the various frequency RF components are strong.

II. THE SLOW WAVE STRUCTURE

Two possible slow wave structures have been considered for this isolator. They are shown in Fig. 1. In all cases, we will only be interested in the lowest order TE mode. All field quantities will be assumed to vary as $\exp\left[j(\omega t - \beta z) - \sqrt{\beta^2 - k^2}\,|x|\right]$. For the capacitively loaded line, we find

$$\beta \approx \frac{\omega^2 \mu_0 h C}{2}, \tag{2}$$

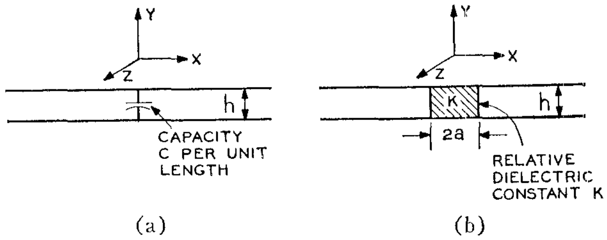

Fig. 1—Slow wave structures in parallel-plane transmission line.

where μ_0 is the permeability of the medium external to the capacity loading, h is the height of the guide, and C is the capacitance per unit length. Note that

$$\frac{\beta}{k} = \sqrt{\frac{\mu_0}{\epsilon_0}}\,\frac{\omega C h}{2}, \tag{3}$$

so that by (1), the circularity of the polarization improves with increasing frequency. However, by the x dependence of the field quantities, $\exp\left[-\sqrt{\beta^2 - k^2}\,|x|\right]$, it is seen that the higher frequency fields are bound more closely to the center of the line.

Several authors have derived expressions for the RF fields for the dielectric loaded line.[4,5] In this case, the expression for the propagation constant, β, involves the solution of the transcendental equation

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[1] M. T. Weiss, "Improved rectangular waveguide resonance isolators," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-4, pp. 240–243; October, 1956.

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

[3] E. M. T. Jones, G. L. Matthaei, and S. B. Cohn, "A nonreciprocal, TEM-mode structure for wide band gyrator and isolator applications," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-7, pp. 453–460; October, 1959.

[4] D. Fleri and G. Hanley, "Nonreciprocity in dielectric loaded TEM mode transmission lines," IRE TRANS. ON MICROWAVE AND TECHNIQUES, vol. MTT-7, pp. 23–27; January, 1959.

[5] M. Cohn, "Propagation in a dielectric-loaded parallel plane waveguide," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-7, pp. 202–208; April, 1959.

$$\tan \sqrt{K - \frac{\beta^2}{k^2}}\, ka = \sqrt{\frac{\frac{\beta^2}{k^2} - 1}{K - \frac{\beta^2}{k^2}}} \tag{4}$$

for the lowest-order TE mode. The ratio β/k is shown as a function of $\sqrt{K}\, ka$ with K as a parameter in Fig. 2. K is the relative dielectric constant of the center section of the guide. Higher-order TE modes occur for

$$\sqrt{K - 1}\, ka = \frac{n\pi}{2} \tag{5}$$

which reduces to

$$\sqrt{K}\, ka \approx \frac{n\pi}{2} \tag{6}$$

for $K \gg 1$. Hybrid modes may be removed to an arbitrarily high frequency by reducing the height of the guide.[5]

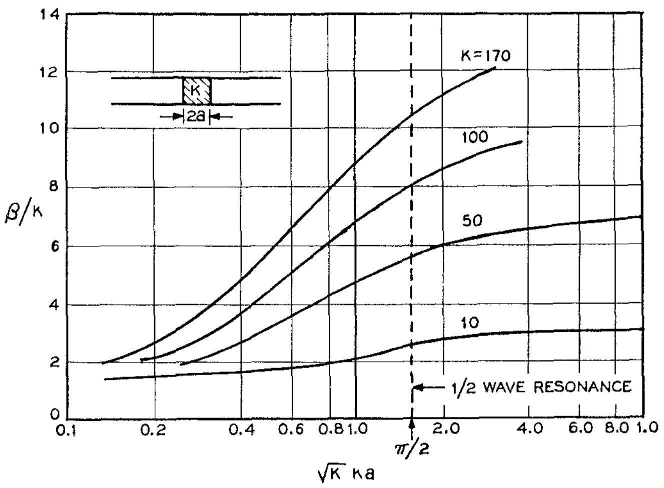

Fig. 2—Slowing factor β/k for various degrees of dielectric loading of parallel-plane transmission line.

For the case of $K=100$ (typical of TiO_2 in ceramic form), it is possible to obtain $\beta/k > 3$ over more than two octaves before problems of higher order modes come in. This type of moding problem does not exist for a guide loaded with ideal capacitors. As in the capacitively loaded guide, high frequency fields are bound closer to the dielectric than low frequency fields.

Expressions for the RF field components and the effects of finite guide width are given in the Appendix. The presence of open-circuit or short-circuit walls at the edge of the guide has little effect if

$$\exp\left[-\sqrt{\beta^2 - k^2}\, b\right] \ll 1,$$

where b is the halfwidth of the guide. Short-circuit walls introduce a low-frequency cutoff in the propagation characteristic.

An approximation to the capacitively loaded line may be made by capacitive posts. The heavily dielectric-loaded structures may use TiO_2 ($K \approx 100$), for the dielectric. This material may have a loss tangent of the order of 10^{-3} at microwave frequencies.[6]

III. The Ferrite Structure

The high-frequency RF fields are concentrated near the slowing structure while the lower frequency fields may extend to large distances from the center of the transmission line. It is possible to introduce ferrite slabs next to the slowing structure and bias them in such a way that high-frequency resonance absorption occurs near the center of the line and low-frequency resonance absorption occurs away from the center of the line.

Consider the magnetic pole configuration of Fig. 3.

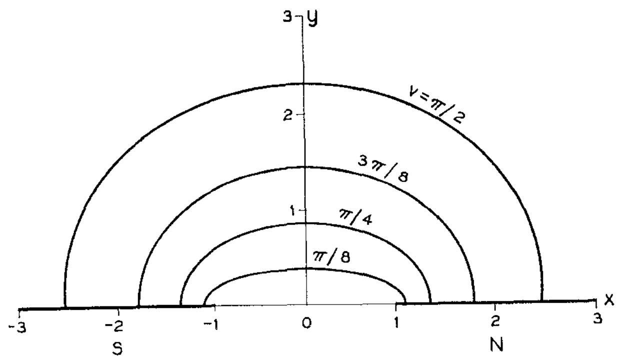

Fig. 3—Magnetic-pole shape and lines of constant magnetic field intensity.

The lines of flux, $v=$ constant, are given by[7]

$$\frac{x^2}{\cosh^2 v} + \frac{y^2}{\sinh^2 v} = 1. \tag{7}$$

At $y=0$ the magnetic field is perpendicular to the plane of the poles, and the relative magnetic field intensity as a function of x, the distance from the center, is given by

$$H_0(x) = \frac{M_0}{2 \times E}, \tag{8}$$

where M_0 is the pole strength and E is the complete elliptic integral

$$E = \int_0^{\pi/2} \sqrt{1 - \tanh^2 v \sin^2 \theta}\, d\theta \tag{9}$$

and

$$v = \cosh^{-1} x. \tag{10}$$

For a "thin" ferrite slab directly above the magnetic poles, the magnetic field will be perpendicular to the ferrite slab and of intensity given by (8). When the

[6] A. R. Von Hippel, "Dielectric Materials and Applications," John Wiley and Sons, Inc., New York, N. Y., pp. 304–305; 1954.

[7] S. Ramo and J. R. Whinnery, "Fields and Waves in Modern Radio," John Wiley and Sons, Inc., New York, N. Y., p. 136; 1953.

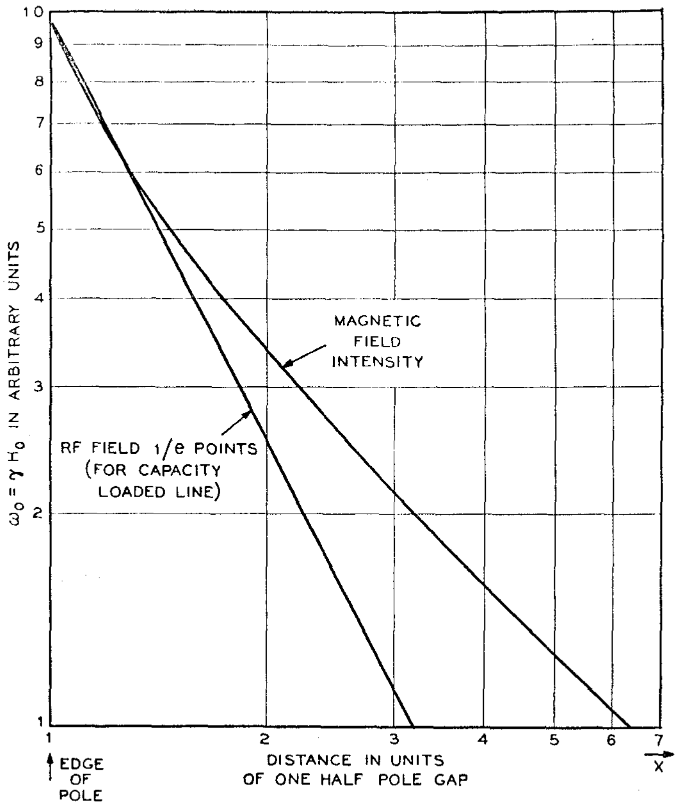

Fig. 4—Relative magnetic field intensity for pole configuration of Fig. 3.

demagnetizing factor for a flat plate[8] is included, the resonance condition is given by

$$\omega_0 = \gamma[H_0(x) - 4\pi M_s], \tag{11}$$

where $4\pi M_s$ is the saturation magnetization. A plot of H_0 as a function of distance from the edge of the magnetic pole gap is given in Fig. 4. Also, a line is shown representing the $1/e$ amplitude points of the RF field in the $\beta \gg k$ region for $\beta \sim \omega^2$.

IV. Experimental Results

Isolators using both capacitive and dielectric slowing have been constructed and tested. The resultant structures are shown in Fig. 5. The capacitively-loaded structure consisted of a parallel plate line 0.025 inch high with an 8 $\mu\mu$f post spaced every 0.065 inch which gave an upper cutoff frequency of about 7 kMc, as would be expected from (3). To obtain this high capacity, it was necessary to have a gap between the post and bottom plate of only 0.00025 inch. A thin sheet of mylar was used to give this spacing. The dielectric-loaded structure used a slab of ceramic rutile (TiO_2, $K \approx 100$), 0.075 inch wide by 0.025 inch high.

The ferrites used in these experiments were flat plates in the shape of isosceles trapezoids with bases of 1.25 inches and 2.0 inches, an altitude of 0.375 inch, and

[8] C. Kittel, "On the theory of ferromagnetic resonance absorption," *Phys. Rev.*, vol. 73, pp. 155–161; January, 1948.

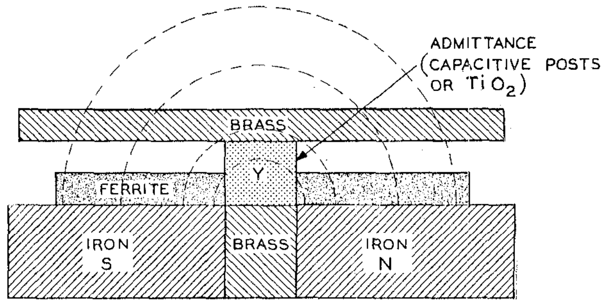

Fig. 5—Cross section of isolator.

a thickness of 0.020 inch. Three ferrite materials were used. All were polycrystalline and had a saturation magnetization, $4\pi M_s$, of the order of 500 gauss. The first sample was a magnesium aluminum-manganese ferrite ($Mg_{1.0}Al_{0.6}Fe_{1.35}Mn_{0.1}O_4$) with a linewidth of 500 oersteds. The other two samples were aluminum substituted yttrium-iron garnet ($3Y_2O_3 \cdot Al_2O_3 \cdot 4Fe_2O_3$). One YIG sample had a linewidth of 15 oersteads and the other a linewidth of 50 oersteds.

The behavior of all three ferrite materials was remarkably similar. The forward and reverse transmission through the structure as a function of magnetic field (arbitrary units) are shown in Fig. 6. This is data for the capacitive post structure with ferrite slabs on both sides of the posts. The ferrite in this case is the YIG sample with a 15-oersted linewidth. This material seemed to give slightly better results than the other materials. If we choose an operating magnetic field corresponding to the arrow in Fig. 6, an isolation vs frequency curve results, as shown in Fig. 7. Reverse attenuation curves for all three ferrites are shown in Fig. 8. For this last set of data the ferrite was placed on only one side of the structure, resulting in about one-half the isolation shown in Fig. 7.

The very low forward loss indicated in Figs. 7 and 9 is caused by the presence of the ferrite alone. This forward loss was determined by observing the change in transmitted power as the ferrite was removed from the structure. In all this work, it is important to realize that only the characteristics of the isolation mechanism are under consideration. Thus, a 0.3- to 0.5-db copper loss has been subtracted from the data, as well as input and output mismatch losses. We did not attempt to build a broad-band match into the test structures. At frequencies where high input and output VSWR's tended to mask the transmission characteristics of interest, tuning stubs were used to reduce input and output reflection coefficients.

The experiments to date with the dielectric-loaded isolator have yielded data similar to to that of the capacity-loaded structure. A typical set of data for this case is shown in Fig. 9. The YIG sample with a 15-oersted linewidth was used here. Short tapers were used to provide some impedance matching, but the reflections from the structure were still appreciable.

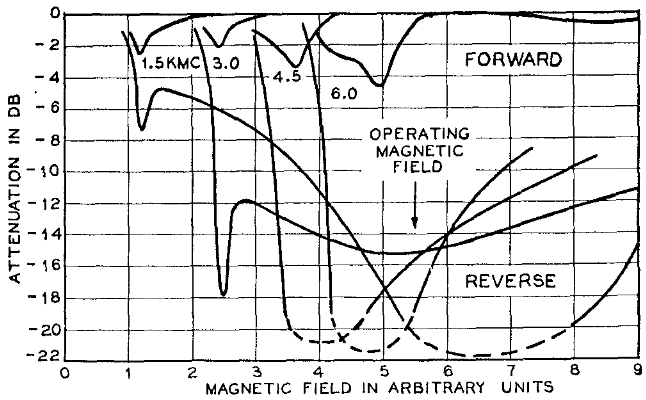

Fig. 6—Capacitively-loaded isolator characteristics as a function of magnetic field.

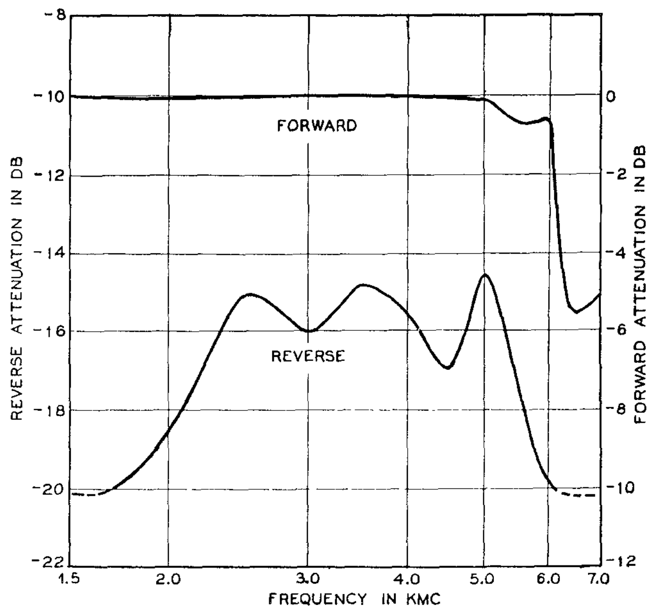

Fig. 7—Capacitively-loaded isolator characteristic for a fixed magnetic field.

V. Conclusion

A design technique for obtaining wide-band resonance isolation has been described. Experiments to date indicate that the concepts presented in Sections II and III of this paper are correct and useful.

Several interesting problems have yet to be solved. The first is to obtain a good broad-band match to either of the structures described. This probably can be done by a long taper into the slowing structure. The second problem is a thorough experimental investigation to find the optimum dc magnetic-field shape for broad-band isolation. Preliminary experiments indicate that the isolation-frequency characteristic is rather sensitive to dc magnetic-field shape at the edge of the ferrite next to the slowing structure. A third problem is associated with the second problem, namely, a theoretical treatment of wave propagation in an inhomogeneously magnetized ferrite medium.

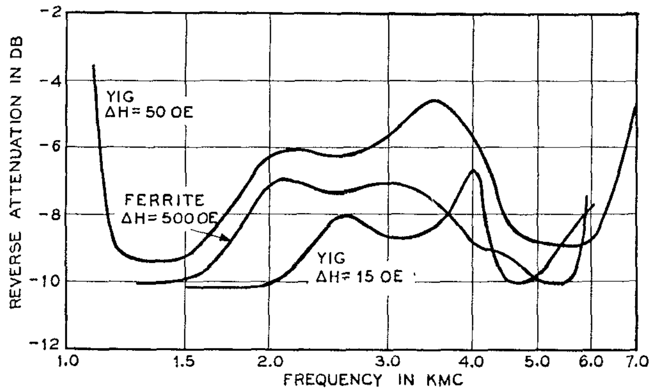

Fig. 8—Comparison of isolator characteristics for various ferrite materials.

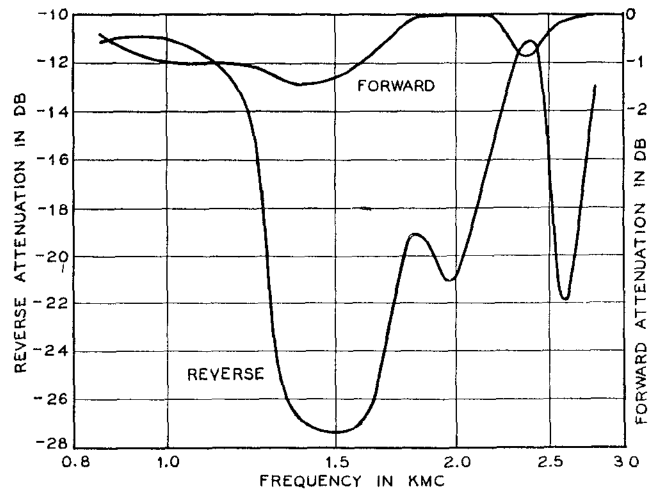

Fig. 9—Isolation characteristic of dielectric-loaded line.

Appendix

Fields in Slow Wave Structure

A. The Capacity-Loaded Guide

For a parallel-plate transmission line of infinite width, height h, and capacity C per unit length along the center of the line, we find

$$E_y = E_0 e^{j(\omega t - \beta z) - p|x|} \tag{12}$$

$$H_x = -\frac{\beta}{\omega\mu_0} E_y, \tag{13}$$

$$H_z = -\frac{jp}{\omega\mu_0} E_y. \tag{14}$$

The coordinate system used here is shown in Fig. 1. The transverse propagation constant p must satisfy the wave equation, giving the relation

$$-\beta^2 + p^2 = -k^2 \tag{15}$$

where k is the propagation constant of free space (or the material filling the volume of the guide). The boundary condition at $x=0$ determines the longitudinal propagation constant β through the equation

$$p = \sqrt{\beta^2 - k^2} = \frac{\omega^2 \mu_0 h C}{2}. \tag{16}$$

Eqs. (13)-(15) then give (1)

$$\frac{H_x}{H_z} = \frac{-j}{\sqrt{1 - \frac{k^2}{\beta^2}}}. \tag{1}$$

For the case of $\beta \gg k$, (16) reduces to (2)

$$\beta \approx \sqrt{\beta^2 - k^2} = \frac{\omega^2 \mu_0 h C}{2}. \tag{2}$$

If we add impedance side walls at $x = \pm b$, the RF fields are given by

$$E_y = E_0 e^{j(\omega t - \beta z)}(e^{-p|x|} + A e^{p|x|}), \tag{17}$$

$$H_x = -\frac{\beta}{\omega \mu_0} E_0^{j(\omega t - \beta z)}(e^{-p|x|} + A e^{p|x|}), \tag{18}$$

$$H_z = -\frac{jp}{\omega \mu_0} E_0 e^{j(\omega t - \beta z)}(e^{-p|x|} - A e^{p|x|}). \tag{19}$$

The presence of the side walls degrades the circular polarization of the RF magnetic field since we now have

$$\frac{H_x}{H_z} = \frac{-j}{\sqrt{1 - \frac{k^2}{\beta^2}}} \frac{e^{-p|x|} + A e^{p|x|}}{e^{-p|x|} - A e^{p|x|}}. \tag{20}$$

For open circuit walls at $x = \pm b$, the transverse propagation constant is determined by the equation

$$pb \tanh pb = \frac{\omega^2 \mu_0 C h b}{2}, \tag{21}$$

while for short circuit walls, the propagation constant p is found from

$$\frac{pb}{\tanh pb} = \frac{\omega^2 \mu_0 C h b}{2}. \tag{22}$$

In this later case, there is a low-frequency cutoff, given approximately by

$$\omega_c = \sqrt{\frac{2}{\mu_0 C h b}}. \tag{23}$$

To improve the circular polarization as given in (20), the quantity A should be as small as possible. For an open circuit wall at $x = \pm b$, the constant A in (20) is given by

$$A = e^{-2pb}, \tag{24}$$

while for short circuit walls at $x = \pm b$ we find

$$A = -e^{-2pb}. \tag{25}$$

B. The Dielectric Loaded Guide

Cohn[5] has given the complete expressions for the even and odd electric field modes in an infinite, parallel-plane, dielectric-loaded transmission line. We will only add the effects of open or short circuit walls at the edge of the transmission line. In the dielectric region, the electric fields are

$$E_y = E_0 \begin{pmatrix} \cos k_d x \\ \sin k_d x \end{pmatrix} e^{j(\omega t - \beta z)}, \tag{26}$$

where $k_d = k\sqrt{K - (\beta^2/k^2)}$. The even electric field modes are the cos $k_d x$ terms, and the odd field modes are the sin $k_d x$ terms. In the region external to the dielectric, the fields are

$$E_y = E_0 \begin{pmatrix} \cos k_d a \\ \sin k_d a \end{pmatrix} \frac{e^{-p|x|} + A e^{p|x|}}{e^{-pa} + A e^{pa}} e^{j(\omega t - \beta z)}. \tag{27}$$

The ratio of transverse-to-longitudinal magnetic field is then given by (20) and the constant A by (24) or (25).

The propagation constant β for the various mode and wall configurations is given by the following transcendental equations:

1) Even electric field, open circuit walls

$$\tan \sqrt{K - \frac{\beta^2}{k^2}}\, ka = \sqrt{\frac{\frac{\beta^2}{k^2} - 1}{K - \frac{\beta^2}{k^2}}} \tanh \sqrt{\frac{\beta^2}{k^2} - 1}\, k(b - a). \tag{28}$$

2) Even electric field, short circuit walls

$$\tan \sqrt{K - \frac{\beta^2}{k^2}}\, ka = \sqrt{\frac{\frac{\beta^2}{k^2} - 1}{K - \frac{\beta^2}{k^2}}} \coth \sqrt{\frac{\beta^2}{k^2} - 1}\, k(b - a). \tag{29}$$

3) Odd electric field, open circuit walls

$$-\cot \sqrt{K - \frac{\beta^2}{k^2}}\, ka = \sqrt{\frac{\frac{\beta^2}{k^2} - 1}{K - \frac{\beta^2}{k^2}}} \tanh \sqrt{\frac{\beta^2}{k^2} - 1}\, k(b - a). \tag{30}$$

4) Odd electric field, short circuit walls

$$-\cot \sqrt{K - \frac{\beta^2}{k^2}}\, ka = \sqrt{\frac{\frac{\beta^2}{k^2} - 1}{K - \frac{\beta^2}{k^2}}} \coth \sqrt{\frac{\beta^2}{k^2} - 1}\, k(b - a). \tag{31}$$