Finite Difference Analysis of Cylindrical Two Conductor Microstrip Transmission Line with Truncated Dielectrics

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ABSTRACT: The characteristics of the quasi-static TEM mode of a cylindrical microstrip transmission line are investigated using the finite difference technique. The transmission line consists of two perfectly conducting strips located between two different layers of dielectric materials, and a dielectric notch embedded in the substrate between two strips. The dielectric overlay and substrate are truncated for practical purposes. The formulation of the problem is based on the solution of Laplace's equation subject to appropriate boundary conditions and the use of Taylor's expansion to approximate the first and second order derivatives in Laplace's equation. To truncate the finite difference mesh, two artificial boundaries have been considered. The goal of this research is to study the effects of the parameters of the multi-layered cylindrical transmission line on the odd and even mode phase velocities, and to present several techniques to minimize the coupling and distortion between the two conductors. Copyright © 1996 Published by Elsevier Science Ltd

I. Introduction

Cylindrical microstrip transmission lines with multilayer dielectrics operating in the quasi-TEM mode have recently received much attention in the microwave literature (1–5). Using flexible dielectrics, it is possible to construct non-planar transmission lines that can be placed around conducting cylindrical surfaces. With the application of smaller and denser circuit dimensions, the coupling between circuit connections limits the performance of the cylindrical microstrip transmission lines (CMSTL). It is very important to find practical methods to accurately analyze and control the coupling between transmission lines.

In this paper, attention is focused on the problem of reducing coupling between two cylindrical microstrip transmission lines. One possible technique is to employ a dielectric notch embedded between the two conducting lines as shown in Fig. 1. The reduction of coupling can be achieved by properly selecting the size of the notch and the relative permittivities of the overlay, substrate and notch regions.

The CMSTL geometry shown in Fig. 1 is treated as a quasi-TEM mode problem.
FIG. 1. Geometry of a coupled cylindrical microstrip transmission line.

Laplace’s equation is solved for the potential distribution $V(\rho, \phi)$ in the various dielectric regions by applying the proper boundary conditions. The finite difference method (FD) is used to approximate the derivatives involved in Laplace’s equation. Then, the integral form of Gauss’s law is used to derive an expression for the scalar potential on the interfaces between different dielectric regions and to compute the total charge on the strips. Two approximate (asymptotic) boundary conditions (ABC) at the outer artificial boundary are used to truncate the finite difference mesh. The solution of Laplace’s equation in a cylindrical coordinate system has been described by Harrington and Pontoppidan (6). A method for the solution of the problem of a transmission line in one homogeneous dielectric medium is discussed by Joshi (7). The theory of quasi-TEM modes on coupled transmission lines in terms of voltage and current eigenvectors is developed by Kajfez (8, 9). It has been shown that the even and odd modes are possible only when the coupled transmission lines are of symmetric shape.

One of the goals of this study is to present an efficient FD solution to analyze and to control the coupling between the conducting lines through the use of dielectric notch and truncated dielectric substrate and superstrate. Here, an ideal model is considered, where the conductors are assumed to be perfectly conducting, the dielectric materials are lossless, isotropic and homogeneous, however the thickness of each conductor is not necessarily assumed to be zero.

The characteristic parameters of the CMSLTs shown in Figs 1 and 2, such as the self and mutual capacitances ($C_{11}, C_{21}$), coupling coefficient ($k_c$), even and odd modes
characteristic impedances \((Z_e, Z_o)\), normalized phase velocities \((v_e, v_o)\), and the effective permittivities \((\varepsilon_{\text{eff}})\) are studied through a set of numerical examples.

II. Formulation

2.1. Finite difference approximations

For the quasi-static 2-D problems shown in Figs 1 and 2, the potential, \(V\), everywhere can be described using Laplace's equation in the cylindrical coordinates as:

\[
\frac{\partial^2 V}{\partial \rho^2} + \frac{1}{\rho} \frac{\partial V}{\partial \rho} + \frac{1}{\rho^2} \frac{\partial^2 V}{\partial \phi^2} = 0. \tag{1}
\]

The finite difference technique is used to approximate the derivatives involved in Laplace's equation. This numerical approximation is based upon calculating the potential at a certain node as a function of neighboring grid points, or nodes. Consider that we have three consecutive nodes \(l, c, r\), as shown in Fig. 3. Taylor's expansion can be applied at node \(r\) and \(l\) in the \(\phi\) direction, i.e.

\[
V_r = V_c - (\phi_c - \phi_r) \frac{\partial V_c}{\partial \phi} + \frac{1}{2} (\phi_c - \phi_r)^2 \frac{\partial^2 V_c}{\partial \phi^2} \tag{2}
\]
FIG. 3. Finite difference mesh.

\[ V_i = V_c + (\phi_i - \phi_c) \frac{\partial V_c}{\partial \phi} + \frac{1}{2} (\phi_i - \phi_c)^2 \frac{\partial^2 V_c}{\partial \phi^2} \]  

(3)

where \( V_r, V_c, V_l \) represent the potential at node \( r, c \) and \( l \), respectively. Equations (2) and (3) can be re-arranged to solve for the first and second derivatives with respect to \( \phi \) as:

\[ \frac{\partial V_c}{\partial \phi} = V_c \frac{(\phi_i + \phi_l - 2\phi_c)}{(\phi_i - \phi_c)(\phi_l - \phi_c)} + V_l \frac{(\phi_c - \phi_i)}{(\phi_i - \phi_c)(\phi_l - \phi_c)} - V_r \frac{(\phi_l - \phi_c)}{(\phi_l - \phi_i)(\phi_c - \phi_i)} \]

(4)

\[ \frac{\partial^2 V_c}{\partial \phi^2} = V_c \frac{2}{(\phi_i - \phi_c)(\phi_l - \phi_c)} + V_l \frac{2}{(\phi_i - \phi_c)(\phi_l - \phi_c)} + V_r \frac{2}{(\phi_l - \phi_i)(\phi_c - \phi_i)}. \]

(5)

Similar expressions for the first and the second derivatives with respect to \( \rho \) can be obtained by applying Taylor’s expansion at nodes \( t \) and \( b \), in the \( \rho \) direction, as shown in Fig. 3, which are expressed as:

\[ \frac{\partial^2 V_c}{\partial \rho^2} = V_c \frac{2}{(\rho_i - \rho_c)(\rho_l - \rho_c)} + V_l \frac{2}{(\rho_c - \rho_b)(\rho_l - \rho_b)} - V_r \frac{2}{(\rho_l - \rho_c)(\rho_c - \rho_b)} \]

(6)

\[ \frac{\partial V_c}{\partial \rho} = V_c \frac{(\rho_i + \rho_l - 2\rho_c)}{(\rho_i - \rho_c)(\rho_l - \rho_c)} + V_l \frac{(\rho_c - \rho_b)}{(\rho_c - \rho_b)(\rho_l - \rho_c)} - V_r \frac{(\rho_l - \rho_c)}{(\rho_l - \rho_b)(\rho_c - \rho_b)}. \]

(7)

where \( V_l \) and \( V_b \) are the potentials at nodes \( t \) and \( b \), respectively.

The boundary conditions for the CMSTL shown in Figs 1 and 2 are: \( V = V_1 \) on strip 1, \( V = V_2 \) on strip 2, \( V = 0 \) on the surface of the circular core, and on a dielectric interface separating two regions, the normal component of the displacement vector \( D \)
must be continuous. This latter condition is based on Gauss’s law for the electric field, which can be described as:

$$\oint D \cdot ds = \oint \varepsilon \nabla V \cdot ds = g \oint \varepsilon \nabla V \cdot ndl = g \oint \varepsilon \frac{\partial V}{\partial n} dl = Q = 0 \quad (8)$$

where $Q$ is set to zero since no free charges exist on the dielectric boundary or in the dielectric regions. Since there is no dependence on $z$, the integral along the $z$-axis is a constant. The $\partial V/\partial n$ denotes the derivative of $V$ normal to the contour $L$, and $g$ is the length of the line along the $z$ direction. The surface integration in eq. (8) thus has been replaced by a contour integration.

Figure 3 shows the distribution of grid points in a homogeneous region, while Fig. 4 shows a node point $c$ (and its four immediate neighbors) on an interface between two dielectric regions and a node point $c$ on the artificial boundary. Applying Laplace’s eq. (1), to a point $c$ in a homogeneous region yields

$$V_c \left\{ \frac{2 \rho_e^2}{(\rho_i - \rho_e)(\rho_e - \rho_b)} + \frac{2}{(\phi_i - \phi_e)(\phi_e - \phi_c)} - \frac{\rho_e (\rho_i + \rho_b - 2 \rho_e)}{(\rho_i - \rho_e)(\rho_e - \rho_b)} \right\}$$

$$- V_i \left\{ \frac{2 \rho_e^2}{(\rho_i - \rho_e)(\rho_e - \rho_b)} + \frac{\rho_e (\rho_e - \rho_b)}{(\rho_i - \rho_e)(\rho_e - \rho_b)} \right\}$$

$$- V_b \left\{ \frac{2 \rho_e^2}{(\rho_e - \rho_b)(\rho_e - \rho_b)} - \frac{\rho_e (\rho_e - \rho_b)}{(\rho_i - \rho_e)(\rho_e - \rho_b)} \right\}$$

$$- V_i \left\{ \frac{2}{(\phi_i - \phi_e)(\phi_e - \phi_c)} \right\} - V_r \left\{ \frac{2}{(\phi_e - \phi_i)(\phi_i - \phi_r)} \right\} = 0. \quad (9)$$
It is important to note, at this point, that Eq. (9) applies to uniform as well as to nonuniform meshes in a homogeneous region. Other expressions are needed for points at the dielectric interfaces and at the artificial boundary. In order to develop such expressions, consider Fig. 5 which shows an expanded view of a special node point $c$ that is surrounded by four dielectric regions. Application of Gauss’s law, at point $c$ yields

$$
\oint_L \varepsilon \left( \frac{\partial V}{\partial \rho} \rho d\phi + \frac{\partial V}{\rho \partial \phi} d\rho \right) = 0.
$$

Equation (10) along with eqs (4) and (6) give a general expression for the node equation at point $c$ in the following form

$$
V_c \left\{ \frac{(\rho_i + \rho_c)}{4(\rho_i - \rho_c)} [\varepsilon_1(\phi_c - \phi_i) + \varepsilon_2(\phi_i - \phi_c)] + \frac{1}{2\rho_c(\phi_i - \phi_c)} [\varepsilon_2(\rho_i - \rho_c) + \varepsilon_3(\rho_c - \rho_b)] \right\}
+ \frac{(\rho_c + \rho_b)}{4(\rho_c - \rho_b)} [\varepsilon_3(\phi_i - \phi_c) + \varepsilon_4(\phi_c - \phi_i)] + \frac{1}{2\rho_c(\phi_i - \phi_c)} [\varepsilon_4(\rho_c - \rho_b) + \varepsilon_1(\rho_i - \rho_c)]
- V_i \left\{ \frac{(\rho_i + \rho_c)}{4(\rho_i - \rho_c)} [\varepsilon_1(\phi_c - \phi_i) + \varepsilon_2(\phi_i - \phi_c)] \right\}
- V_i \left\{ \frac{1}{2\rho_c(\phi_i - \phi_c)} [\varepsilon_2(\rho_i - \rho_c) + \varepsilon_3(\rho_c - \rho_b)] \right\}
$$
FIG. 6. Grid point c at a corner of dielectric notch.

\[-V_b \left\{ \frac{(\rho_c + \rho_b)}{4(\rho_c - \rho_b)} [\epsilon_3 (\phi_i - \phi_c) + \epsilon_4 (\phi_c - \phi_i)] \right\}
\]

\[-V_r \left\{ \frac{1}{2\rho_c (\phi_i - \phi_r)} [\epsilon_4 (\rho_c - \rho_b) + \epsilon_1 (\rho_c - \rho_r)] \right\} = 0 \]  

(11)

Special cases can be obtained from this general expression for a node at the corner of three different dielectric regions as shown in Fig. 6, or at the interface between two different dielectrics as shown in Fig. 4. Our computations indicate that the numerical data based on Eqns (9) and (11) are almost identical. Our data are therefore based on the general expression in Eq. (11) and the approximate expressions for the potential at the nodes on the artificial boundary as described in the next section.

2.2. First order approximate boundary condition (ABC1)

For the transmission lines shown in Figs 1 and 2, the outer artificial boundary can be shielded using a perfect electric conductor (PEC) and, hence, the potential at the boundary is set to zero. However, when the transmission line is not shielded by a PEC, an approximate boundary condition should be applied at an artificial boundary in order to truncate the FD mesh so that numerical analysis can be carried out. The first order boundary condition used here is proposed by Khebir et al. (10). Consider the series expansion for the electric potential defined as:

\[ V(\rho, \phi) = C_0 + A_0(\phi) \ln \rho + \sum_{n=1}^{\infty} \frac{A_n(\phi)}{\rho^n} . \]  

(12)
The constant term, $C_0$ is dropped because the potential at infinity is zero. The second term in Eq. (12) is also dropped because of the free charge enclosed by the mesh bounded by the artificial boundary. Therefore, Eq. (12) reduces to

$$V(\rho, \phi) = \frac{A_1(\phi)}{\rho} + \frac{A_2(\phi)}{\rho^2} + \frac{A_3(\phi)}{\rho^3} + \cdots$$

(13)

Differentiating Eq. (13) with respect to $\rho$ gives

$$\frac{\partial V(\rho, \phi)}{\partial \rho} = -\frac{A_1(\phi)}{\rho^2} - \frac{2A_2(\phi)}{\rho^3} - \frac{3A_3(\phi)}{\rho^4} - \cdots$$

(14)

Equation (13) is then multiplied by $1/\rho$ and added to Eq. (14), to yield

$$\frac{V}{\rho} + \frac{\partial V}{\partial \rho} = -\frac{A_2(\phi)}{\rho^3} - \frac{2A_3(\phi)}{\rho^4} - \frac{3A_4(\phi)}{\rho^5} - \cdots$$

(15)

As an approximation to Eq. (15), the right hand side will be assumed to be zero. This approximation leads to:

$$\frac{V}{\rho} + \frac{\partial V}{\partial \rho} = 0.$$  

(16)

Equation (16) will be implemented along the outer boundary of the mesh. The grid point system on the outer boundary of the mesh is shown in Fig. 4. The grid point $c$ resides on the outer boundary and $t$ is a point outside the outer boundary. Because point $t$ lies outside the mesh, it is not possible to enforce Eq. (9) or (11) at the node point $c$ in the usual manner. Instead, Eqs (16) and (6) are applied first to determine the potential at point $t$, in terms of $V_r, V_c, V_r, V_r$ and $V_b$. Once $V_r$ is known then Eq. (9) can be enforced. At point $c$, Eqs (6) and (16), can be arranged to obtain:

$$V_i = -V_c \left\{ \frac{(\rho_c + \rho_b - 2\rho_c)(\rho_c - \rho_b)}{(\rho_c - \rho_b)^2} + \frac{(\rho_c - \rho_c)(\rho_c - \rho_b)}{\rho_c(\rho_c - \rho_c)} \right\} + V_b \left( \frac{\rho_c - \rho_c}{(\rho_c - \rho_b)^2} \right).$$

(17)

Then by the substitution of (17) into Eq. (9), one obtains:

$$V_i = \left\{ \frac{2}{(\rho_c - \rho_c)(\rho_c - \rho_b)} + \frac{2}{\rho_c^2(\phi_i - \phi_i)(\phi_i - \phi_i)} - \frac{(\rho_c - \rho_b - 2\rho_c)}{\rho_c(\rho_c - \rho_b)(\rho_c - \rho_b)} \right\}$$

$$+ \left\{ \frac{(3\rho_c - \rho_c)(\rho_c - \rho_b - 2\rho_c)}{\rho_c(\rho_c - \rho_b)(\rho_c - \rho_b)} \right\}$$

$$- V_b \left\{ \frac{(3\rho_c - \rho_c)}{\rho_c(\rho_c - \rho_b)(\rho_c - \rho_b)} + \frac{(3\rho_c - \rho_c)(\rho_c - \rho_c)}{\rho_c(\rho_c - \rho_b)(\rho_c - \rho_b)^2} \right\}$$

$$- V_i \left\{ \frac{2}{\rho_c^2(\phi_i - \phi_i)(\phi_i - \phi_i)} \right\} - V_r \left\{ \frac{2}{\rho_c^2(\phi_i - \phi_i)(\phi_i - \phi_i)} \right\} = 0.$$  

(18)

2.3. Second order approximate boundary condition (ABC2)

In order to bring the artificial boundary closer to the strips, a higher order ABC needs to be developed. We first define:
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\[ U = \frac{\partial V}{\partial \rho} + \frac{V}{\rho}. \]  

(19)

Then the right hand side of Eq. (15) is replaced with \( U \) to obtain:

\[ U = -\frac{A_2}{\rho^3} \frac{2A_3}{\rho^4} - \frac{3A_4}{\rho^5} \ldots \]  

(20)

Equation (20) can be differentiated with respect to \( \rho \) to obtain:

\[ \frac{\partial U}{\partial \rho} = \frac{3A_2}{\rho^4} + \frac{8A_3}{\rho^5} + \frac{15A_4}{\rho^6} + \ldots \]  

(21)

Now, multiply Eq. (20) by \( 3/\rho \) and then add it to Eq. (21) to obtain:

\[ \frac{3}{\rho} U + \frac{\partial U}{\partial \rho} = -\frac{3A_2}{\rho^4} - \frac{6A_3}{\rho^5} + \frac{9A_4}{\rho^6} \ldots + \frac{3A_2}{\rho^4} + \frac{8A_3}{\rho^5} + \frac{15A_4}{\rho^6} + \ldots \]  

(22)

which can be rewritten as:

\[ \frac{\partial U}{\partial \rho} + \frac{3}{\rho} U = \frac{2A_3}{\rho^5} + \frac{6A_4}{\rho^6} + \frac{12A_5}{\rho^7} + \ldots \]  

(23)

If the definition of \( U \) is used and the right hand side of Eq. (23) is approximated as zero, one obtains:

\[ \left( \frac{\partial}{\partial \rho} + \frac{3}{\rho} \right) \left( \frac{\partial}{\partial \rho} + \frac{1}{\rho} \right) V = 0 \]  

(24)

which is equivalent to:

\[ \frac{\partial^2 V_c}{\partial \rho^2} + \frac{4}{\rho^c} \frac{\partial V_c}{\partial \rho} + \frac{2}{\rho^2} V_c = 0. \]  

(25)

The above expression can be used to obtain the voltage at point \( t \) for the grid point \( c \) on the outer boundary of the mesh as shown in Fig. 4. Substituting Eqns (6) and (7) into (25), we get:

\[ V_t = V_c \left\{ \frac{(\rho_i - \rho_b)}{\rho_i (\rho_i - \rho_b)(3\rho_c - 2\rho_b)} - \frac{2(\rho_i - \rho_b)(\rho_i + \rho_b - 2\rho_c)}{\rho_i^2 (\rho_i - \rho_b)(3\rho_c - 2\rho_b)} - \frac{(\rho_i - \rho_c)(\rho_c - \rho_b)}{\rho_i^3 (3\rho_c - 2\rho_b)} \right\} \]

\[ - V_b \left\{ \frac{(\rho_i - \rho_c)}{\rho_i (\rho_i - \rho_b)(3\rho_c - 2\rho_b)} - \frac{2(\rho_i - \rho_c)^2}{\rho_i^2 (\rho_i - \rho_b)(3\rho_c - 2\rho_b)} \right\}. \]  

(26)

Equation (26) can then be substituted into Eq. (9) to obtain:
\[ -V_b \left\{ \frac{3\rho_c - \rho_t}{\rho_c(\rho_c - \rho_b)(\rho_t - \rho_b)} - \frac{3\rho_c - \rho_b}{(\rho_t - \rho_b)(\rho_c - \rho_b)(3\rho_c - 2\rho_b)} \right\} \\
+ \frac{2(\rho_t - \rho_c)(3\rho_c - \rho_b)}{\rho_c(\rho_t - \rho_b)(\rho_c - \rho_b)(3\rho_c - 2\rho_b)} \right\} \\
- V_\ell \left\{ \frac{2}{\rho_c^2(\phi_\ell - \phi_\ell)(\phi_t - \phi_t)} \right\} - V_r \left\{ \frac{2}{\rho_c^2(\phi_r - \phi_r)(\phi_t - \phi_t)} \right\} = 0. \tag{27} \]

2.3. Banded matrix solution

From the equations describing the potential at the grid points, it is noted that the potential \(V\) at each node is related only to its four or three immediate neighboring nodes. Therefore, application of any of those equations to all grid points in the solution mesh leads to a set of simultaneous equations of the form \([A][X] = [B]\) where matrix \([A]\) is a sparse matrix, i.e. it contains many zero elements, \([X]\) is a column matrix containing all unknown potentials, \(V\), for the node points, and \([B]\) is a column matrix which contains the known potential at the fixed nodes. Matrix \([A]\) is banded in which its non-zero elements are clustered near the main diagonal. The solution of the matrix equation is efficiently obtained by using the Linpack (11) subroutines SGBCO and SGBSL. Once the solution of the unknown potentials is obtained, the characteristic parameters of the CMSTL can be determined as described in the next section.

III. Computation of the Cylindrical Transmission Line Characteristic Parameters

3.1. The charges on the strip lines

The computation of the characteristic impedance and phase velocity for a cylindrical transmission line with an inhomogeneous medium requires calculating the capacitances of the structure, with and without the dielectric material. Since the capacitance per unit length is directly related to the charge per unit length on the strips, the problem is reduced to finding the total charge per unit length on the strips. The microstrip transmission lines shown in Figs 1 and 2 have two conductors which are placed between the two different dielectrics, superstrate and substrate. It is assumed throughout the analysis of the problem that the conductors are perfectly conducting strips and the dielectric materials are lossless, isotropic, and homogeneous. Let \(Q_i\) denote the total charge on the \(i\)th conductor. If Gauss’s law is applied to a closed path \(L\) enclosing the \(i\)th conductor, the total charge per unit length on that conductor is then expressed as:

\[ \frac{Q_i}{\ell} = -\int_L \epsilon \frac{\partial V}{\partial n} dl = \int_{\text{right}} \epsilon(\rho) \frac{\partial V}{\partial \phi} d\rho + \int_{\text{top}} \epsilon(\phi) \frac{\partial V}{\partial \rho} \rho d\phi \\
+ \int_{\text{left}} \epsilon(\rho) \frac{\partial V}{\partial \phi} d\rho + \int_{\text{bottom}} \epsilon(\phi) \frac{\partial V}{\partial \rho} \rho d\phi \tag{28} \]

where the integral around the contour is the summation of the integrals at the top, bottom, left, and right sides of the path \(L\).
3.2. Computation of mutual and self-capacitances per unit length

The theory of multi-conductor transmission lines has been widely discussed. In general the equations relating voltages and charges to capacitances are given by

\[ Q_1 = C_{11} V_1 + C_{12} (V_1 - V_2), \quad Q_2 = C_{21} (V_2 - V_1) + C_{22} V_2 \]  

(29)

where the coefficients \( C_{ii} \) are called self-capacitances per unit length, and \( C_{ij} \) are mutual capacitances per unit length. For the odd mode of excitation, \( V_1 = -V_2 = V \) and hence the charges per unit length on strips 1 and 2 are denoted by \( Q_{1o} \) and \( Q_{2o} \), respectively. Whereas, when \( V_1 = V_2 = V \), the charges per unit length on strips 1 and 2 are denoted by \( Q_{1e} \) and \( Q_{2e} \), respectively. Applying these two types of excitations to the cylindrical microstrip line, and using Eq. (29), one gets

\[ C_{11} = \frac{Q_{1o} - Q_{1e}}{2}, \quad C_{12} = -\frac{Q_{2o} - Q_{2e}}{2}, \quad C_{21} = -\frac{Q_{2o} - Q_{2e}}{2}, \quad C_{22} = Q_{2e} \]  

(30)

3.3. Impedance, phase velocity and effective permittivity

For a symmetric transmission line, the total charge per unit length on the conductors will be equal in magnitude and sign for even mode, (i.e. \( Q_1 = Q_2 = Q_e \)), while for odd mode, the charges will be equal in magnitude but opposite in sign, (i.e. \( Q_1 = -Q_2 = Q_o \)). In such a case, the capacitances can be related to the odd and even total charges as:

\[ C_{11} = C_{22} = Q_e, \quad C_{12} = C_{21} = \frac{Q_o - Q_e}{2}. \]  

(31)

The effective permittivity of the line for even and odd modes are then defined, respectively by:

\[ \varepsilon_{re} = \frac{Q_e}{Q_{oe}}, \quad \varepsilon_{ro} = \frac{Q_o}{Q_{oo}} \]  

(32)

where the subscript \( a \) stands for the charge with the dielectric material replaced by free space (9). With known effective permittivities, the even and odd mode velocities of propagation (\( v_e \) and \( v_o \)) are given by:

\[ v_e = \frac{c}{\sqrt{\varepsilon_{re}}} = c \sqrt{\frac{Q_{qe}}{Q_e}}, \quad v_o = \frac{c}{\sqrt{\varepsilon_{ro}}} = c \sqrt{\frac{Q_{ao}}{Q_o}} \]  

(33)

where \( c \) is the velocity of light, and the even and odd impedances \( Z_{oe} \) and \( Z_{oo} \) are obtained from:

\[ Z_{oe} = \frac{1}{c \sqrt{Q_o Q_{oe}}}, \quad Z_{oo} = \frac{1}{c \sqrt{Q_o Q_{ao}}}. \]  

(34)

When this coupled transmission line is used as a coupling device, its characteristic impedance \( Z_o \) is usually given by \( \sqrt{Z_{oe} Z_{oo}} \).

Another important parameter that describes the electrical coupling factor between the strips is defined as \( k_e = V^2 / V', \) where \( V^2 \) is the voltage induced on strip 2 due to \( V' \) applied on strip 1 (12). In terms of the line capacitances, \( k_e \) reduces to
IV. Numerical Results

In all the numerical examples presented here, it is necessary to point out that all radial distances are normalized to the radius of the conducting core \( r \), while \( h_1, h_2, \) and \( h_3 \), are defined as \( (a-r)/r, (b-a)/r \) and \( (c-b)/r \), respectively, and \( t_s \) is the normalized thickness of the strips with respect to \( r \). All these parameters are shown in Figs 1 and 2. The truncation of the dielectric substrate and overlay, as shown in Fig. 2, can be specified by changing \( \alpha_1 \) and \( \alpha_n \). Non-truncated geometry is defined as \( \delta = 360^\circ \), the integer values \( N_1, N_1, N_2, N_3, \) and \( N_4 \) represent the number of the FD nodes on each strip, and the number of nodes along the radial direction along the substrate, strip, overlay, and air-layer, respectively. It is always assumed that the geometry of the CMSTL is symmetric around the \( y \)-axis.

4.1. Microstrip transmission line with a dielectric notch

4.1.1. Verification of computed data. In order to verify the numerical results generated by our computer program, a special case is selected to compare our results with published data in (3). The configuration parameters for this case are such that \( h_2 = 0.2, h_3 = 0.5, t_s = 0.001, \delta = 360^\circ, \delta_l = \delta_n = 2.0(a-r)/a, \varepsilon_{r_1} = 1.0, \varepsilon_{r_2} = \varepsilon_{r_3} = 9.6. \) The ratio of \( r/a \) is then varied from 0.5 to 0.9. Figure 7 shows the odd and even modes characteristic impedances and effective permittivities versus \( r/a \). As shown in the figure, the results obtained in this study and those in Figs 3 and 4 in (3) indicate good agreement with a maximum difference of 2.2%.

4.1.2. Effect of artificial boundary. It is very important to place the artificial boundary for the finite difference analysis at the proper position in order to obtain accurate results for the characteristics of the transmission line. It is also always desirable to place the artificial boundary as close to the strips as possible and yet not affect the accuracy of the numerical results. The closer the distance between the artificial boundary and the strips, the smaller the FD mesh size will be, and consequently the order of the resulting matrix equations is reduced and faster numerical solution is achievable. In order to study the effects of the position of the artificial boundary on the characteristics of the transmission line, the geometry shown in Fig. 1 is selected for this analysis, where, the distance \( h_3 \) is varied from 0.1 to 1.7. Other parameters selected are such that \( h_1 = 0.2, h_2 = 0.1, t_s = 0.001, \delta = 10^\circ, \delta_i = 10^\circ, \delta_n = 6^\circ, \) and \( \varepsilon_{r_1}, \varepsilon_{r_2}, \) and \( \varepsilon_{r_3} \) are set to 2.2, 4.7 and 9.6, respectively. Figure 8 shows the capacitances for even and odd modes versus \( h_3 \). It can be easily seen that using ABC2, the artificial boundary can be placed at a closer position to the strips than using the ABC1 or the PEC boundaries. This is because there are no significant changes in the CMSTL characteristics using ABC2 when the distance \( h_3 \) becomes greater than 0.5. The numerical results based on ABC1 and ABC2 are almost identical when \( h_3 \) is larger than 1, as expected. This behavior is clearly observed for the CMSTL with full or truncated dielectric material and with thin or thick strips. Similar results are obtained for truncated CMSTL in Fig. 2, although the results are not shown here (13). The second order approximate boundary
Fig. 7. Even and odd modes characteristic impedances and effective relative permittivities versus 
$r/a$. ($h_2 = 0.2, h_3 = 0.5, ts = 0.001, \delta_n = \delta_i = 2.0(a-r)/a, \delta = 360^\circ, \varepsilon_1 = 1.0, \varepsilon_2 = \varepsilon_3 = 9.6,$  
$N_2 = 1, N_3 = 4, N_4 = 5.$)

4.1.3. Convergence of numerical results. The convergence of our numerical solution 
can be observed in Fig. 9 for non-truncated and truncated CMSTLs. In the figure the 
even and odd mode characteristic impedances, $Z_e$ and $Z_o$, versus $N$ (the order of the 
matrix $[X]$) are shown. With $h_1 = 0.2, h_2 = 0.1, h_3 = 0.5, ts = 0.001, \delta_n = 10^\circ, \delta_i = 10^\circ,$  
$\delta = 6^\circ, \varepsilon_1 = 2.2, \varepsilon_2 = 4.7$ and $\varepsilon_3 = 9.6$. With fixed $h_3$, increasing $N$ improves the 
accuracy of the numerical results, however, it is clearly seen from the figure that 
convergence is achievable when $N$ is in the order of 2000.

4.1.4. Effect of the dielectric material of the overlay (superstrate). The effects of the 
dielectric material of the superstrate is studied with the configuration defined by 
$h_1 = 0.2, h_2 = 0.1, h_3 = 0.5, ts = 0.001, \delta_n = 10^\circ, \delta_i = 10^\circ, \delta_a = 6^\circ, \varepsilon_1 = 4.7$ and $\varepsilon_3$ is 
set to 1.0 and 9.6, respectively. The dielectric constant $\varepsilon_1$ is then varied from 1 to 16. 
The influences of various $\varepsilon_1$ and $\varepsilon_3$ on the characteristic parameters of the transmission 
line are tested for both truncated and non-truncated geometries as shown in Figs 1 and 
2. The size of the truncated geometry, $x$, and $x_{o}$, are set in this case equal to 72°, which 
means that the edges of the substrate and the overlay are 3° away from the conductor 
strips. Figure 10 shows the influences of $\varepsilon_1$ and $\varepsilon_3$ on the self and mutual capacitances
for non-truncated geometry, while Fig. 11 shows those influences for truncated geometry. As can be seen, the mutual capacitances are directly proportional to $\varepsilon_{r1}$, and it is noted that the self capacitance obtained by placing dielectric material into the notch exhibits a marked deviation from the result obtained by removing dielectric material from the notch. In this case, using dielectric material in the notch reduces the mutual capacitances, but not significantly as shown in Figs 10 and 11. These also show that the coupling coefficient increases with $\varepsilon_{r1}$. It is found that, for this geometry, using a dielectric notch between strips did not show significant decoupling between lines. Furthermore, the influences on the characteristic impedances, effective permittivities and normalized phase velocities are studied for even and odd modes of the transmission line with truncated and non-truncated geometries, respectively. It is noted from Fig. 10 that the even and odd normalized phase velocities, $v_e$ and $v_o$, are equalized at approximately $\varepsilon_{r1} = 4$ and $\varepsilon_{r1} = 6.5$, as $\varepsilon_{r3} = 1.0$ and $\varepsilon_{r3} = 9.6$, respectively, which means, for this specific case (non-truncated geometry), that the line distortion is eliminated. Similar behavior for truncated geometry can be observed in Fig. 11, where $v_e$ and $v_o$ are equalized at almost the same point, i.e. $\varepsilon_{r1} = 4.2$, for $\varepsilon_{r3} = 1.0$ and $\varepsilon_{r3} = 9.6$. These properties show that the use of truncated and non-truncated CMSTL with different dielectric materials in the notch allow for a distortionless line at different
values of the dielectric constant of the overlay. No significant changes in the behaviour characteristics of the characteristic impedance and the effective permittivities are observed when the truncation of the dielectric is applied in this example (13).

4.1.5. **Effect of the substrate dielectric material.** The configuration selected for investigating the effect of the dielectric constant in the substrate was such that \( h_1 = 0.2, h_2 = 0.2, h_3 = 0.5, t_s = 0.001, \delta_w = 10^\circ, \delta_i = 10^\circ, \delta_n = 6^\circ, \varepsilon_{r1} = 2.2, \varepsilon_{r2} = 4.7, \varepsilon_{r3} = 9.6, \delta = 360^\circ, N_w = 4 \sim 12, N_1 = 4, N_2 = 1, N_3 = 3, N_4 = 4.\)
the line is achievable using a dielectric notch in the substrate between strips for truncated or non-truncated geometries. To study the effect of removing the dielectric material from the notch, the same configuration and parameters are used in this case as those used for Fig. 12, but \( \varepsilon_2 \) is now set to 1, which represents an air-dielectric notch. The \( \varepsilon_2 \) is then varied from 1 to 16. The influences of \( \varepsilon_2 \) on the self and mutual capacitances between these two strips are shown in Fig. 13. It is noted that the influences of \( \varepsilon_2 \) on the self capacitance of the two strips do not show much differences as expected, compared with the case where a dielectric notch was used in the substrate. However, the influences of \( \varepsilon_2 \) on the mutual capacitance \( C_{12} \) (or \( C_{21} \)) for the truncated geometry is very different from the previous one. As can be seen in Fig. 13, the mutual capacitance (using truncated geometry) does not show significant percentage variations as \( \varepsilon_2 \) increases. Figure 13 also shows that the coupling coefficient decreases as \( \varepsilon_2 \) increases. It is also noted that the use of truncation geometry with an air-dielectric notch exhibits a better decoupling between lines, compared with the results obtained by using a dielectric notch. Figure 13 also shows that \( \nu_s \) and \( \nu_m \) are equalized at approximately \( \varepsilon_2 = 3 \). These properties show that a possible method for reducing mutual coupling between the transmission lines is by using a truncated geometry with an air-dielectric
Fig. 11. Effect of dielectric overlay on capacitances, coupling and phase velocity for truncated CMSTL. 

\[ \varepsilon_3 = 1.0 \quad \varepsilon_3 = 9.6 \]

4.1.6. Effect of the notch dielectric material and dimensions. Another case is selected to study the effect of the dielectric constant of the notch. The configuration selected is such that \( h_1 = 0.2, h_2 = 0.2, h_3 = 0.5, \delta_s = 0.001, \delta_\omega = 10^\circ, \delta_\lambda = 10^\circ, \delta_\eta = 6^\circ, \varepsilon_2 = 4.7, \varepsilon_3 = 9.6 \) and \( 1.0, x_s = 72^\circ, N_s = 10, N_1 = 4, N_2 = 1, N_3 = 3, N_4 = 4. \) The truncation is defined by \( \varepsilon_\omega = x_\omega = 72^\circ, \) and the dielectric constants, \( \varepsilon_1 \) and \( \varepsilon_2, \) are set to 2.2 and 4.7, respectively. The dielectric constant of the notch, \( \varepsilon_3, \) is then varied from 1 to 16. It is noticed that the self capacitance linearly increases with the increase of \( \varepsilon_3, \) and the coupling between the strip lines increases when \( \varepsilon_3 \) has a small value, and then decreases after \( \varepsilon_3 = 6. \) The truncated CMSTL shows less distortion for all values of \( \varepsilon_3 \) with respect to the non-truncated CMSTL, for this specific geometry.

In order to investigate the influences of the dielectric constant of the substrate on the characteristic parameters of the transmission line while \( \varepsilon_3 \) is varying, the dielectric constant \( \varepsilon_2, \) is reduced from 4.7 to 2.2, which means that \( \varepsilon_2 \) is now set equal to \( \varepsilon_1. \) Other configuration parameters are the same as used in the previous example. Compared with the above case, in which \( \varepsilon_2 \) is equal to 4.7, we found that both the self and mutual capacitances are reduced and the mutual capacitance \( (C_{21}) \) decreases after \( \varepsilon_3 \) is approximately 5. Figure 14 demonstrates that the coupling coefficient, \( k_c, \) decreases as \( \varepsilon_3 \)
increases. It is also noted that $v_e$ and $v_o$ are equalized at $\varepsilon_{r3} = 5.2$ for non-truncated geometry and at $\varepsilon_{r3} = 9$ for truncated geometry. This figure shows that CMSTL with truncated dielectric can be distortionless with less coupling.

Another case selected for study is the effect of the width of the notch on the characteristics of the CMSTL. The configuration is the same as the first case in this section, the separation between the strips, $\delta_s = 10^\circ$, the width of notch, $\delta_n$, is set to $2^\circ$ and $8^\circ$, respectively, and $\varepsilon_{r3}$ is then varying from 1 to 16. Figure 15 shows that there is an intersection between the $k_e$ curves when $\varepsilon_{r3} = 4.7$ (i.e. $\varepsilon_{r3} = \varepsilon_{r2}$), which means that there is no notch in the substrate. It also shows that the coupling coefficient is further reduced for a wider notch with no significant change in $v_e - v_o$. This property illustrates the advantage of using the widest notch possible for the best decoupling between the strips.

4.1.7. Effect of the strip width. In this investigation, the width of the strips, $\delta_w$, is varied from 2 to $20^\circ$ with a fixed spacing between the strips. Other configuration parameters selected are such that $h_1 = 0.2$, $h_2 = 0.2$, $h_3 = 0.5$, $ts = 0.001$, $\delta_s = 10^\circ$, $\delta_n = 6^\circ$, $\varepsilon_{r1} = 2.2$, $\varepsilon_{r3} = 9.6$, and the angles, $\alpha_s$, $\alpha_o$, are set equal to $65^\circ$. It is found that the self-capacitance increases as the width of strips gets wider. However, the mutual capacitance does not show a significant increase, and the coupling coefficient

Fig. 12. Effect of dielectric substrate on capacitances, coupling and phase velocities. ($h_1 = 0.2$, $h_2 = 0.2$, $h_3 = 0.5$, $ts = 0.001$, $\delta_s = 10^\circ$, $\delta_n = 6^\circ$, $\varepsilon_{r1} = 2.2$, $\varepsilon_{r3} = 9.6$, $\alpha_s = 72^\circ$, $N_w = 10$, $N_1 = 4$, $N_2 = 1$, $N_3 = 3$, $N_4 = 4$.)
is thus reduced as the strips have larger width. For the same geometry one notices that $v_e - v_o$ is very close to zero, which means, for this case, that the distortion on the transmission line is almost eliminated. It is also noted that the use of truncated geometry does not show significant changes on the values of the characteristic parameters of the transmission line (13).

4.1.8. Effect of the spacing between strips. The width of the strips, $\delta_w$, is set to $10^\circ$, and other configuration parameters are selected such that $h_1 = 0.2$, $h_2 = 0.2$, $h_3 = 0.5$, $ts = 0.001$, $\delta_n = 10^\circ$, $\delta_s = 6^\circ$, $k_s = 6.2$, $k_3 = 1.0$, $a_s = 72^\circ$, $N_1 = 10$, $N_1 = 4$, $N_2 = 1$, $N_1 = 3$, $N_4 = 4$.

Fig. 13. Effect of dielectric substrate on capacitances, on coupling and phase velocities. ($h_1 = 0.2$, $h_2 = 0.2$, $h_3 = 0.5$, $ts = 0.001$, $\delta_n = 10^\circ$, $\delta_s = 6^\circ$, $k_s = 6.2$, $k_3 = 1.0$, $a_s = 72^\circ$, $N_1 = 10$, $N_1 = 4$, $N_2 = 1$, $N_1 = 3$, $N_4 = 4$.)
previous example, in which a dielectric notch was used, however, the coupling coefficient does not show significant changes.

4.1.9. Effect of the substrate and overlay height. The heights of the substrate and the overlay are very important parameters to the characteristics of the transmission line. Usually, the heights of the substrate and overlay should be selected carefully in order to obtain the best decoupling and to equalize the even and odd phase velocities. For this purpose, the effect of the substrate height is first studied. The configuration selected for this case, is such that \( h_2 = 0.2, \ h_3 = 0.5, \ t_s = 0.001, \ \delta_w = 10^\circ, \ \delta_s = 10^\circ, \ \delta_n = 6^\circ, \ \varepsilon_r = 2.2, \ \varepsilon_r = 2.2, \ \alpha_s = 72^\circ, \ \alpha_w = 10^\circ, \ \alpha_s = 10^\circ, \ \alpha_n = 6^\circ, \ \varepsilon_r = 2.2, \ \varepsilon_r = 4.7, \ \varepsilon_r = 9.6, \ \alpha_s = \alpha_w = 72^\circ. \) The height of the substrate, \( h_1, \) is then varied from 0.1 to 0.9. Figure 16 shows that the self-capacitance decreases and mutual capacitance increases as \( h_1 \) increases, for both the truncated and non-truncated structures. With the use of truncated structure, \( v_e \) and \( v_o \) are equalized when \( h_1 \) is approximately 0.3, and the use of non-truncated structure does not exhibit this property. Note that the coupling is increased with varied \( h_1, \) which means that for the best decoupling, the height of the substrate should be kept as small as possible. For the same analysis with \( \varepsilon_r = 4.7 \) and \( \varepsilon_r = 2.2, \) Fig. 16 (case b) clearly shows that the coupling is further increased, \( v_e \) and \( v_o \) can not be equalized at any height of the substrate. Hence, in order to reduce the coupling between the strips, the dielectric constant of the overlay should be smaller than that of the substrate.
4.2. Performance of transmission line without a dielectric notch

In this section, the characteristics of the CMSTL without a dielectric notch is analyzed. To remove the dielectric notch from the substrate, the dielectric constant of the notch, \( \varepsilon_{r3} \), and the dielectric constant of the substrate, \( \varepsilon_{r2} \), are set to be equal to each other. The numerical results for several examples are discussed in the following subsections.

4.2.1. Effect of the dielectric constant. The effect of the dielectric constant of the overlay is first investigated. The configuration selected is such that \( h_1 = 0.2, h_2 = 0.2, \)
Fig. 16. Effect of thickness of substrate on coupling and phase velocities. a) $\varepsilon_r = 2.2$, $\varepsilon_r = 4.7$, b) $\varepsilon_r = 4.7$, $\varepsilon_r = 2.2$. ($h_2 = 0.2$, $h_3 = 0.5$, $t_s = 0.001$, $\delta_w = 10^\circ$, $\delta_s = 10^\circ$, $\delta_n = 6^\circ$, $\varepsilon_3 = 9.6$, $\alpha = 72^\circ$, $N_1 = 10$, $N_1 = 4$, $N_2 = 1$, $N_3 = 3$, $N_4 = 5$.)

Another analysis is made to study the effect of dielectric constant of the substrate. The same configuration is used for this case, but the dielectric constant of overlay, $\varepsilon_r$, is now set to 2.2, $\varepsilon_r$ and $\varepsilon_3$ are both varied from 1 to 16. The self and mutual capacitances are found to be both linearly increasing with $\varepsilon_r$, while, the coupling
coefficient decreases as $\varepsilon_{r2}$ and $\varepsilon_{r3}$ increase as shown in Fig. 18. In this case, however, the normalized $v_e$ and $v_o$ are equalized at approximately $\varepsilon_{r2} = \varepsilon_{r3} = 2.2$ for non-truncated structure, and at $\varepsilon_{r2} = \varepsilon_{r3} = 3.2$ for the truncated structure.

4.2.2. Effect of the spacing between the strips. The configuration selected in this case is such that $h_1 = 0.2, h_2 = 0.2, h_3 = 0.5, \delta_w = 10^\circ$, the dielectric constants of the substrate and the notch, $\varepsilon_{r2}$ and $\varepsilon_{r3}$, are both set to 4.7, the dielectric constant of the superstate, $\varepsilon_{r1}$, is equal to 2.2, $\delta_s$ is then varied from 6 to 24$^\circ$. In this investigation, strips with thickness are used to study their effects on the characteristic parameters of the transmission line. We noticed that the self-capacitance increases and the mutual capacitance decreases while the separation distance between the strips increases. It can be easily noticed that the use of thick strips increases both self- and mutual capacitances, compared with the use of thin strips as shown in Fig. 19. It is also noted that the coupling coefficient decreases with increased spacing between strips. All these figures clearly show that the use of thin strips leads to less coupling and distortion relative to thick strips.

4.3. Effect of the structure parameters

There are two major purposes for applying the truncated dielectric to CMSTL. One is for controlling the coupling between the lines and elimination of distortion, and the
other is to save dielectric materials if the use of the truncated geometry provides the
same characteristics as the non-truncated CMSTL. Since the size of truncation affects
the characteristics of the transmission line, the effect of some structure parameters such
as the size of truncation and the thickness of the strips are discussed in the following
sections.

4.3.1. Effect of the size of truncation. A case is selected to study the effect of the size
of the truncated dielectrics. The configuration selected is such that \( h_1 = 0.2, h_2 = 0.2, h_3 = 0.5, ts = 0.001, \delta_u = 10^\circ, \delta_s = 10^\circ, \delta_n = 6^\circ, \varepsilon_{r1} = 2.2, \alpha_x = 72^\circ, N_u = 10, N_1 = 4, N_2 = 1, N_3 = 3, N_4 = 4 \).}

Fig. 18. Effect of dielectric substrate on coupling and phase velocities. (\( h_1 = 0.2, h_2 = 0.1, h_3 = 0.5, ts = 0.001, \delta_u = 10^\circ, \delta_s = 10^\circ, \delta_n = 6^\circ, \varepsilon_{r1} = 2.2, \alpha_x = 72^\circ, N_u = 10, N_1 = 4, N_2 = 1, N_3 = 3, N_4 = 4 \).)
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Fig. 19. Effect of spacing between strips on capacitances of non-truncated CMSTL with no notch. 
\(h_1 = 0.2, h_2 = 0.2, h_3 = 0.5, t_s = 0.001\) and \(0.05, \delta_\sigma = 10^\circ, \delta_n = 4^\circ, \epsilon_{r2} = 2.2, \epsilon_{r3} = 4.7, \delta = 360^\circ, N_n = 10, N_1 = 4, N_2 = 1, N_3 = 3, N_4 = 5.\)

control the point where distortion on the line occurs. It is clearly noted from Fig. 20 that the larger the size of the truncation is the closer the results would be, compared with non-truncated structure.

4.3.2. Effect of the thickness of the strips. In order to investigate the effect of the thickness of the strips on the characteristic parameters of the transmission line, a structure is selected such that \(h_1 = 0.2, h_2 = 0.2, h_3 = 0.5, \delta_s = 10^\circ, \delta_n = 10^\circ, \delta_\sigma = 6^\circ, \epsilon_{r1} = 1.0\) and \(\epsilon_{r2} = 4.7\), and the angles, \(\alpha_s\) and \(\alpha_o\), are set to \(10^\circ\). The thickness of the strips, \(t_s\), is then varied from 0.001 to 0.28 with different \(\epsilon_{r3}\). The linear behavior can be observed in Fig. 21, where both self- and mutual capacitances are increased, as the thickness of the strips is increased. The coupling and distortion are found to be increasing with \(t_s\). It is clearly noted that thin strips should be used for reducing coupling and distortion.

V. Summary and Conclusions

A finite difference formulation for the analysis of the characteristics of a two-conducting strips cylindrical transmission line with a dielectric notch between the strips is presented. The dielectric materials are assumed to be lossless, isotopic, and
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Fig. 20. Effect of dielectric substrate on coupling and phase velocities with different truncations. 
\((h_1 = 0.2, h_2 = 0.2, h_3 = 0.5, ts = 0.001, \delta_w = 10^\circ, \delta_r = 10^\circ, \delta_p = 6^\circ, \varepsilon_{r1} = 2.2, \varepsilon_r = 9.6, N_w = 10, N_1 = 4, N_2 = 1, N_3 = 3, N_4 = 5)\)

homogeneous, while the thickness of the conducting strips are not necessarily assumed to be zero. Microstrip lines with full and truncated dielectric materials are used in the numerical analysis. The decoupling and distortion control between such two-conductor strip transmission line are investigated.

It has been shown here that the coupling between strip lines can be minimized by altering the dielectric constant of the notch material and use of proper truncation of the substrate material. A distortionless line is achieved when these parameters are selected properly. The truncation of the dielectric substrate and overlay provides a way to achieve zero distortion. With proper dielectric materials in different regions, it is found that the width of the strips should be as wide as possible and the thickness should be kept as thin as possible for the best decoupling and distortion control. The results also show that using dielectric overlay increases the coupling, but may reduce the distortion with proper values for the permittivity in the overlay. It is also found that transmission lines with truncated dielectric materials provides the same results obtained with non-truncated geometry.

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Fig. 21. Effect of thickness of strips on coupling and phase velocities for truncated CMSTL.

References


