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ELECTROMAGNETIC COUPLING TO CIRCULANT SYMMETRIC MULTI-CONDUCTOR MICROSTRIP LINE

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Abstract—In this paper a method is introduced and applied to calculate the effects of an external field on a circular symmetric microstrip transmission line. The primary/secondary field idea is used for this purpose. The primary field is determined analytically for the cases of normal TM^z and TE^z incidence. The secondary field is determined using multi-conductor transmission line theory. The method is applied to a special structure and some useful results are obtained.

1. INTRODUCTION

Circular Symmetric Multi-conductor Microstrip Line (CSMSL) is a recently introduced multi-conductor structure useful for a variety of signal transmission tasks [1–3], especially in high frequency and microwave IC packaging, in which large number of high frequency connections and routings are to be made throughout the package. The effect of external field exposure on terminal voltages and currents of a transmission line is a phenomenon that should be well understood before making any judgment over EMC/EMI properties of the line. This turns out to be critical in high frequency integrated circuit design due to the importance of prediction and control of EMI between neighboring components.

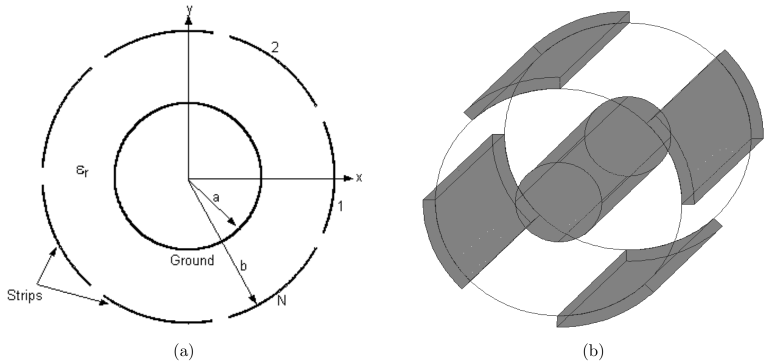

Figure 1. A CSMSL structure (a) The cross section (b) A three dimensional view.

Fig. 1 shows the so called CSMSL with N strips symmetrically surrounding the dielectric layer and a metallic core with a radius of "a". The z-axis is assumed to lay along the axis of the metallic core. Suppose that there are N comparatively thin strips surrounding the structure and each strip is located in the domain $\phi_i - \theta_i < \phi < \phi_i + \theta_i$ $(i = 1, 2, \ldots, N)$ on a cylinder with radius of "b". The space between the core and strips is occupied with a dielectric material with relative electric permittivity of ε_r. The strips and their relative currents and voltages are numbered (indexed) from the one at $\phi = 0$ in a counter clockwise fashion.

To develop a prime understanding about EMI in the CSMSL structure, a method first developed by Bernardi [4] earlier in 1990 is used. The total electric and magnetic field is divided into a primary and a secondary part. The primary part belongs to the field solution in the absence of metal strips and the secondary part belongs to the Quasi-TEM mode in the multi-conductor transmission line. In this paper, Bernardi's method is used for CSMSL structures. First the strips are removed from the CSMSL and a uniform TM^z and TE^z plane wave is exposed to determine the primary field. Then a transmission line formulation with distributed excitation sources is developed and solved. Finally a comprehensive example is presented.

2. DETERMINATION OF PRIMARY FIELD

This section deals with calculation of the so called "primary field". A far field plane wave is applied to a long CSMSL structure in which the metal strips are not present. For the sake of simplicity, it is assumed that the incident wave propagates towards positive x direction. Two different polarizations are possible, one the TM^z and other the TE^z. All relations are expressed in the frequency domain and our notion is compatible to that of [5].

2.1. TM^z Primary Field

Here a normally incident field is assumed such that its electric component is aligned in the $+z$ direction. Mathematically speaking this field is expressed as

$$\vec{E}^i = E_0 e^{-jk_0 x}\hat{x} = E_0\left(\sum_{m=-\infty}^{\infty} j^{-m} J_m(k_0\rho)e^{jm\phi}\right)\hat{z} \tag{1}$$

$$\begin{aligned}
\vec{H}^i &= -\frac{E_0}{\eta_0}e^{-jk_0 x}\hat{y} \\
&= -\frac{E_0}{\eta_0}(\sin(\phi)\hat{\rho}+\cos(\phi)\hat{\phi})\left(\sum_{m=-\infty}^{\infty} j^{-m} J_m(k_0\rho)e^{jm\phi}\right)\hat{y} \\
&= \frac{E_0}{\eta_0}\left(\sum_{m=-\infty}^{\infty}\left[\frac{j^{-(m+2)}mJ_m(k_0\rho)}{k_0\rho}\right]e^{jm\phi}\right)\hat{\rho} \\
&\quad +\frac{E_0}{\eta_0}\left(\sum_{m=-\infty}^{\infty}\left[j^{-(m+1)}J'_m(k_0\rho)\right]e^{jm\phi}\right)\hat{\phi}
\end{aligned} \tag{2}$$

in which $k_0 = \omega\sqrt{\mu_0\varepsilon_0}$ and $\eta_0 = \sqrt{\mu_0/\varepsilon_0}$ are the propagation coefficient and wave impedance in the free space. It is assumed that a penetrating field inside the dielectric layer and a scattered field outside the structure form up. These are denoted by the letters "d" and "s" for dielectric and scattered, respectively. Each of these components can be expressed using the electric and magnetic field potential vectors, as follows

$$\vec{E}^d = -j\omega E_0\left(\sum_{m=-\infty}^{\infty}\left[a_m J_m\left(k_0\sqrt{\varepsilon_r}\rho\right)+b_m Y_m\left(k_0\sqrt{\varepsilon_r}\rho\right)\right]e^{jm\phi}\right)\hat{z} \tag{3}$$

$$\vec{H}^d = \frac{E_0}{\mu_0\rho}\left(\sum_{m=-\infty}^{\infty} jm\left[a_m J_m\left(k_0\sqrt{\varepsilon_r}\rho\right)+b_m Y_m\left(k_0\sqrt{\varepsilon_r}\rho\right)\right]e^{jm\phi}\right)\hat{\rho}$$

$$-\frac{k_0\sqrt{\varepsilon_r}E_0}{\mu_0}\left(\sum_{m=-\infty}^{\infty}\left[a_m J'_m\left(k_0\sqrt{\varepsilon_r}\rho\right)+b_m Y'_m\left(k_0\sqrt{\varepsilon_r}\rho\right)\right]e^{jm\phi}\right)\hat{\phi} \tag{4}$$

$$\vec{E}^s = -j\omega E_0\left(\sum_{m=-\infty}^{\infty}\left[c_m H_m^{(2)}\left(k_0\rho\right)\right]e^{jm\phi}\right)\hat{z} \tag{5}$$

$$\begin{aligned}\vec{H}^s = & \frac{E_0}{\mu_0\rho}\left(\sum_{m=-\infty}^{\infty}jm\left[c_m H_m^{(2)}\left(k_0\rho\right)\right]e^{jm\phi}\right)\hat{\rho}\\ & -\frac{k_0E_0}{\mu_0}\left(\sum_{m=-\infty}^{\infty}\left[c_m H_m'^{(2)}\left(k_0\rho\right)\right]e^{jm\phi}\right)\hat{\phi}\end{aligned} \tag{6}$$

Primes in superscripts represent derivative of functions with respect to their arguments.

The unknown coefficients a_m, b_m and c_m in (3)–(6) are to be obtained such that boundary conditions for electric and magnetic fields are satisfied. Boundary conditions at $\rho = a$ and $\rho = b$ are respectively as follows

$$E_z^d = 0 \tag{7}$$

and

$$E_z^i + E_z^s = E_z^d \tag{8a}$$

$$H_\phi^i + H_\phi^s = H_\phi^d \tag{8b}$$

Using (1)–(8), the following relations are obtained.

$$a_m J_m\left(k_0\sqrt{\varepsilon_r}a\right) + b_m Y_m\left(k_0\sqrt{\varepsilon_r}a\right) = 0 \tag{9a}$$

$$a_m J_m(k_0\sqrt{\varepsilon_r}b)+b_m Y_m\left(k_0\sqrt{\varepsilon_r}b\right)-c_m H_m^{(2)}(k_0b) = -\frac{j^{-(m+1)}}{\omega}J_m(k_0b) \tag{9b}$$

$$a_m J'_m(k_0\sqrt{\varepsilon_r}b)+b_m Y_m(k_0\sqrt{\varepsilon_r}b)-c_m\frac{1}{\sqrt{\varepsilon_r}}H_m'^{(2)}(k_0b) = -\frac{j^{-(m+1)}}{\omega\sqrt{\varepsilon_r}}J'_m(k_0b) \tag{9c}$$

Relations in (9) form a linear set of algebraic equations to which a non-trivial solution is the answer to our primary field problem.

2.2. TE^z Primary Field Calculation

Here a normally incident field is assumed such that its electric component is aligned in the $-y$ direction. Mathematically speaking this is expressed as

$$\begin{aligned}\vec{E}^i &= -E_0 e^{-jk_0x}\hat{y}\\ &= -E_0\left(\sin(\phi)\hat{\rho}+\cos(\phi)\hat{\phi}\right)\left(\sum_{m=-\infty}^{\infty} j^{-m}J_m(k_0\rho)e^{jm\phi}\right)\hat{y}\\ &= E_0\left(\sum_{m=-\infty}^{\infty}\left[\frac{j^{-(m+2)}mJ_m(k_0\rho)}{k_0\rho}\right]e^{jm\phi}\right)\hat{\rho}\\ &\quad +E_0\left(\sum_{m=-\infty}^{\infty}\left[j^{-(m+1)}J'_m(k_0\rho)\right]e^{jm\phi}\right)\hat{\phi} \qquad (10)\end{aligned}$$

$$\vec{H}^i = -\frac{E_0}{\eta_0}e^{-jk_0x}\hat{z} = -\frac{E_0}{\eta_0}\left(\sum_{m=-\infty}^{\infty} j^{-m}J_m(k_0\rho)e^{jm\phi}\right)\hat{z} \qquad (11)$$

Here scattered and penetrating components of the electric field can be expressed as

$$\begin{aligned}\vec{E}^d = &-\frac{E_0}{\varepsilon_0\varepsilon_r\rho}\left(\sum_{m=-\infty}^{\infty} jm\left[a_mJ_m\left(k_0\sqrt{\varepsilon_r}\rho\right)+b_mY_m\left(k_0\sqrt{\varepsilon_r}\rho\right)\right]e^{jm\phi}\right)\hat{\rho}\\ &+\frac{k_0E_0}{\varepsilon_0\sqrt{\varepsilon_r}}\left(\sum_{m=-\infty}^{\infty}\left[a_mJ'_m\left(k_0\sqrt{\varepsilon_r}\rho\right)+b_mY'_m\left(k_0\sqrt{\varepsilon_r}\rho\right)\right]e^{jm\phi}\right)\hat{\phi} \qquad (12)\end{aligned}$$

$$\vec{H}^d = -j\omega E_0\left(\sum_{m=-\infty}^{\infty}\left[a_mJ_m\left(k_0\sqrt{\varepsilon_r}\rho\right)+b_mY_m\left(k_0\sqrt{\varepsilon_r}\rho\right)\right]e^{jm\phi}\right)\hat{z} \qquad (13)$$

$$\begin{aligned}\vec{E}^s = &-\frac{E_0}{\varepsilon_0\rho}\left(\sum_{m=-\infty}^{\infty} jm\left[c_mH_m^{(2)}\left(k_0\rho\right)\right]e^{jm\phi}\right)\hat{\rho}\\ &+\frac{k_0E_0}{\varepsilon_0}\left(\sum_{m=-\infty}^{\infty}\left[c_mH_m'^{(2)}\left(k_0\rho\right)\right]e^{jm\phi}\right)\hat{\phi} \qquad (14)\end{aligned}$$

$$\vec{H}^s = -j\omega E_0\left(\sum_{m=-\infty}^{\infty}\left[c_mH_m^{(2)}\left(k_0\rho\right)\right]e^{jm\phi}\right)\hat{z} \qquad (15)$$

The unknown coefficients a_m, b_m and c_m in (12)–(15) are to be obtained such that boundary conditions for electric and magnetic fields are satisfied. Boundary conditions at $\rho = a$ and $\rho = b$ are respectively as follows

$$E_\phi^d = 0 \qquad (16)$$

and

$$H_z^i + H_z^s = H_z^d \qquad (17a)$$

$$E_\phi^i + E_\phi^s = E_\phi^d \tag{17b}$$

The following relations are obtained, again from (10)–(17).

$$a_m J_m'\left(k_0\sqrt{\varepsilon_r}a\right) + b_m Y_m'\left(k_0\sqrt{\varepsilon_r}a\right) = 0 \tag{18a}$$

$$\begin{aligned} a_m J_m\left(k_0\sqrt{\varepsilon_r}b\right) + b_m Y_m\left(k_0\sqrt{\varepsilon_r}b\right) - c_m H_m^{(2)}\left(k_0 b\right) \\ = +\frac{j^{-(m+1)}}{\eta_0\omega} J_m(k_0 b) \end{aligned} \tag{18b}$$

$$\begin{aligned} a_m J_m'\left(k_0\sqrt{\varepsilon_r}b\right) + b_m Y_m'\left(k_0\sqrt{\varepsilon_r}b\right) - c_m\sqrt{\varepsilon_r} H_m'^{(2)}\left(k_0 b\right) \\ = +\frac{j^{-(m+1)}\sqrt{\varepsilon_r}}{\eta_0\omega} J_m'(k_0 b) \end{aligned} \tag{18c}$$

Here again, relations in (18) form a linear set of algebraic equations to which a nontrivial solution is the answer to our primary field problem.

3. TRANSMISSION LINE EQUATIONS AND THEIR SOLUTIONS

In this section, exposed multi-conductor transmission line equations and their solution are presented in frequency domain. Once the frequency domain voltage and currents obtained, FFT can be used to find the time domain responses.

Incorporating the definitions of capacitance and inductance matrices, using some algebraic manipulation and rewriting the resultant equations in matrix form, the so called "multi conductor transmission line equations with distributed forcing functions" are obtained as [6]

$$\frac{\partial}{\partial z}\boldsymbol{V}(\omega, z) + j\omega\boldsymbol{L}\boldsymbol{I}(\omega, z) = \boldsymbol{V}_F(\omega) \tag{19}$$

$$\frac{\partial}{\partial z}\boldsymbol{I}(\omega, z) + j\omega\boldsymbol{C}\boldsymbol{V}(\omega, z) = \boldsymbol{I}_F(\omega) \tag{20}$$

in which $\boldsymbol{V}$ and $\boldsymbol{I}$ are $N \times 1$ voltage and current vectors, respectively and $\boldsymbol{C}$ and $\boldsymbol{L}$ are capacitance and inductance matrices of the lines, respectively that can be obtained using a similar method to that of [1] or [7] for CSMSL structure. Also, $\boldsymbol{V}_F$ and $\boldsymbol{I}_F$ are $N \times 1$ distributed forcing sources and regarding to [6] can be obtained as follows.

$$\boldsymbol{V}_F(\omega) = j\omega\mu_0 \begin{bmatrix} \vdots \\ \int\limits_{\rho=a}^{\rho=b} H_\phi^p(\varphi=\varphi_i)d\rho \\ \vdots \end{bmatrix} \tag{21a}$$

$$\boldsymbol{I}_F(\omega) = -j\omega\boldsymbol{C} \begin{bmatrix} \vdots \\ \int\limits_{\rho=a}^{\rho=b} E_\rho^p(\varphi=\varphi_i)d\rho \\ \vdots \end{bmatrix} \tag{21b}$$

Defining the state vector $\boldsymbol{X} = [\boldsymbol{V}\ \boldsymbol{I}]^T$ and combining (19) and (20), we get

$$\frac{\partial}{\partial z}\boldsymbol{X}(\omega,z) + \boldsymbol{A}\boldsymbol{X}(\omega,z) = \begin{bmatrix} \boldsymbol{V}_F(\omega) \\ \boldsymbol{I}_F(\omega) \end{bmatrix} \tag{22}$$

in which

$$\boldsymbol{A} = \begin{pmatrix} 0 & j\omega\boldsymbol{L} \\ j\omega\boldsymbol{C} & 0 \end{pmatrix} \tag{23}$$

The solution of (22) can be written as

$$\boldsymbol{X}(\omega,z) = \boldsymbol{e}^{-\boldsymbol{A}z}\left(\boldsymbol{X}(\omega,0) + \int\limits_0^z \boldsymbol{e}^{\boldsymbol{A}z'}\begin{bmatrix} \boldsymbol{V}_F(\omega) \\ \boldsymbol{I}_F(\omega) \end{bmatrix} dz'\right) \tag{24}$$

in which

$$\boldsymbol{e}^{\mp\boldsymbol{A}z} = \boldsymbol{T}\begin{bmatrix} e^{\mp\lambda_1 z} & 0 & \cdots & 0 & 0 \\ 0 & e^{\mp\lambda_2 z} & \cdots & 0 & 0 \\ \vdots & \vdots & \ddots & \vdots & \vdots \\ 0 & 0 & \cdots & e^{\mp\lambda_{N-1} z} & 0 \\ 0 & 0 & \cdots & 0 & e^{\mp\lambda_N z} \end{bmatrix}\boldsymbol{T}^{-1} \tag{25}$$

In (25) $\boldsymbol{T}$ is a $N \times N$ matrix formed by putting the eigen-vectors of $\boldsymbol{A}$ together in columns and λ_1, λ_2, λ_N represent eigen-values of $\boldsymbol{A}$. The solution (24) comes like

$$\begin{bmatrix} \boldsymbol{V}(\omega,z) \\ \boldsymbol{I}(\omega,z) \end{bmatrix} = \begin{bmatrix} \boldsymbol{P}(z) & \boldsymbol{Q}(z) \\ \boldsymbol{R}(z) & \boldsymbol{S}(z) \end{bmatrix}\left(\begin{bmatrix} \boldsymbol{V}(\omega,0) \\ \boldsymbol{I}(\omega,0) \end{bmatrix} + \begin{bmatrix} \boldsymbol{F}_V(\omega,z) \\ \boldsymbol{F}_I(\omega,z) \end{bmatrix}\right) \tag{26}$$

in which

$$\begin{bmatrix} \boldsymbol{F}_V(\omega, z) \\ \boldsymbol{F}_I(\omega, z) \end{bmatrix} = \boldsymbol{T} \begin{bmatrix} (e^{+\lambda_1 z}-1)/\lambda_1 & 0 & \cdots & 0 & 0 \\ 0 & (e^{+\lambda_2 z}-1)/\lambda_2 & \cdots & 0 & 0 \\ \vdots & \vdots & \ddots & \vdots & \vdots \\ 0 & 0 & \cdots & (e^{+\lambda_{N-1} z}-1)/\lambda_{N-1} & 0 \\ 0 & 0 & \cdots & 0 & (e^{+\lambda_N z}-1)/\lambda_N \end{bmatrix} \boldsymbol{T}^{-1} \begin{bmatrix} \boldsymbol{V}_F(\omega) \\ \boldsymbol{I}_F(\omega) \end{bmatrix} \tag{27}$$

and

$$\begin{bmatrix} \boldsymbol{P}(z) & \boldsymbol{Q}(z) \\ \boldsymbol{R}(z) & \boldsymbol{S}(z) \end{bmatrix} = \boldsymbol{e}^{-\boldsymbol{A}z} \tag{28}$$

Also, the terminal conditions at $z = 0$ and $z = d$, are

$$\boldsymbol{V}(\omega, 0) = \boldsymbol{V}_L - \boldsymbol{Z}_L \boldsymbol{I}(\omega, 0) \tag{29a}$$

$$\boldsymbol{V}(\omega, d) = \boldsymbol{V}_R + \boldsymbol{Z}_R \boldsymbol{I}(\omega, d) \tag{29b}$$

in which sub-indices L and R stand for 'Left End' and 'Right End', respectively. Using (26) at $z = d$ and (29), one can obtain the left end and the right end terminals' voltages and currents.

$$\begin{aligned} \boldsymbol{I}(\omega, 0) = {} & (\boldsymbol{Z}_R(\boldsymbol{S}(d) - \boldsymbol{R}(d)\boldsymbol{Z}_L) + \boldsymbol{P}(d)\boldsymbol{Z}_L - \boldsymbol{Q}(d))^{-1} \\ & \cdot((\boldsymbol{P}(d) - \boldsymbol{Z}_R\boldsymbol{R}(d))(\boldsymbol{V}_L + \boldsymbol{F}_V(\omega, d)) \\ & -\boldsymbol{V}_R + (\boldsymbol{Q}(d) - \boldsymbol{Z}_R\boldsymbol{S}(d))\boldsymbol{F}_I(\omega, d)) \end{aligned} \tag{30a}$$

$$\boldsymbol{V}(\omega, 0) = \boldsymbol{V}_L - \boldsymbol{Z}_L \boldsymbol{I}(\omega, 0) \tag{30b}$$

Once this is achieved, (26) is applicable to calculate the voltage and current vectors at any point along the z-axis

4. EXAMPLES AND RESULTS

As an example a CSMSL with four equally width strips covering 20% of the surface of a dielectric rod, which is $2b = 3.0$ mm thick in diameter and contains an $a = 1.0$ mm metallic core, is exposed to both TM^z and TE^z polarizations of a plane wave radiation with an electric field strength of 1.0 V/m and in the direction of $+x$. Relative dielectric constant of the line or ε_r, is either 2.54 or 10.0. It is assumed that termination source voltage vectors, i.e., $\boldsymbol{V}_L$ and $\boldsymbol{V}_R$, are zero and that both left and right end of each line is connected to a 50 Ω load.

The excitation is assumed to occur at a frequency of 3.0 GHz and the line is $d = 20$ cm long (twice the wavelength in free space). Primary field components and force functions are calculated using a program written in *Maple*. Then a simple *Matlab* program is used to calculate the resulting termination voltage and current vectors.

Table 1. The distributed force voltage and current functions of CSMSL with $\varepsilon_r = 2.54$ and normal incidence.

ε_r=2.54	**TM^z**				**TE^z**			
	$\|V_F\|$ [mV]	$<V_F$ [deg.]	$\|I_F\|$ [μA]	$<I_F$ [deg.]	$\|V_F\|$ [mV]	$<V_F$ [deg.]	$\|I_F\|$ [μA]	$<I_F$ [deg.]
Line 1	107.24	3.84	0	-	0	-	0	-
Line 2	125.28	28.20	0	-	0	-	818.48	89.70
Line 3	154.12	46.19	0	-	0	-	0	-
Line 4	125.28	28.20	0	-	0	-	818.48	-90.30

Table 2. The distributed force voltage and current functions of CSMSL with $\varepsilon_r = 10.0$ and normal incidence.

ε_r=10.0	**TM^z**				**TE^z**			
	$\|V_F\|$ [mV]	$<V_F$ [deg.]	$\|I_F\|$ [μA]	$<I_F$ [deg.]	$\|V_F\|$ [mV]	$<V_F$ [deg.]	$\|I_F\|$ [μA]	$<I_F$ [deg.]
Line 1	107.54	3.85	0	-	0	-	0	-
Line 2	125.61	28.19	0	-	0	-	796.64	89.63
Line 3	154.50	46.18	0	-	0	-	0	-
Line 4	125.61	28.19	0	-	0	-	796.64	-90.37

Tables 1 and 2 show the distributed force voltage and current functions for $\varepsilon_r = 2.54$ and $\varepsilon_r = 10.0$, respectively. Figs. 2 and 3 depict the magnitude of voltage and current along the CSMSL length, for $\varepsilon_r = 2.54$ and $\varepsilon_r = 10.0$, respectively. Also, Fig. 4 depicts the power delivered to the loads on the right or left end of CSMSL versus load values again for both $\varepsilon_r = 2.54$ and $\varepsilon_r = 10.0$.

From Figs. 2–4 and Tables 1–2, it is understood that in the TE^z normal incidence only side strips No. 2 and 4 are excited, while in the TM^z normal incidence all four strips are excited and excitation in the TM^z polarization is much stronger than that of in the TE^z polarization. Also, the power(s) delivered to the line(s) behind the metallic core is smaller than that of other lines. This is a special property of circulant structures, which is not observed in planar ones. Also, as seen in Fig. 4, one can state that the powers delivered to the loads become maximum corresponding to a certain values of the loads and these powers decrease as the permittivity of the dielectric increases. So, one can conclude that it is better to choose the permittivity of the substrate as large as possible.

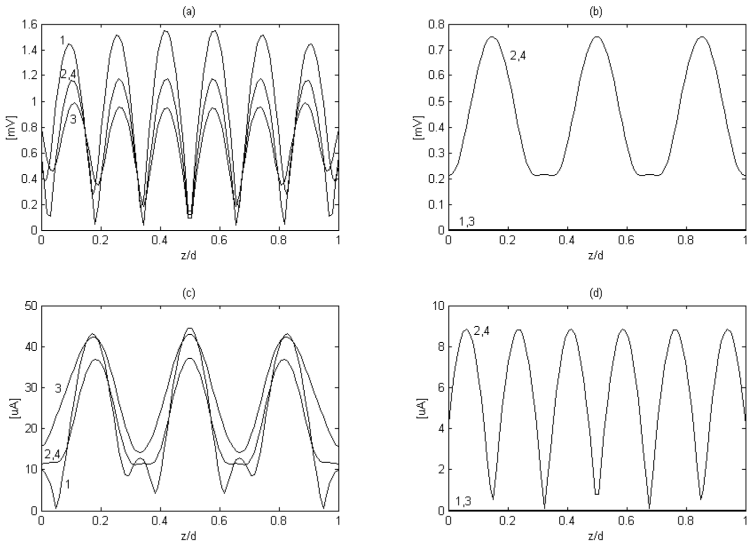

Figure 2. Magnitude of voltage and current along the CSMSL length with $50\,\Omega$ loads, $\varepsilon_r = 2.54$ and normal incidence a) $|\boldsymbol{V}|$ for TM^z b) $|\boldsymbol{V}|$ for TE^z c) $|\boldsymbol{I}|$ for TM^z d) $|\boldsymbol{I}|$ for TE^z.

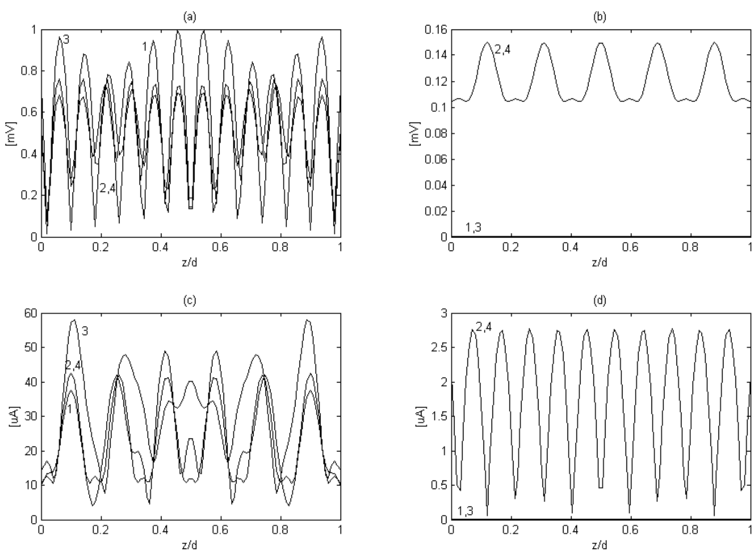

Figure 3. Magnitude of voltage and current along the CSMSL length with $50\,\Omega$ loads, $\varepsilon_r = 10.0$ and normal incidence a) $|\boldsymbol{V}|$ for TM^z b) $|\boldsymbol{V}|$ for TE^z c) $|\boldsymbol{I}|$ for TM^z d) $|\boldsymbol{I}|$ for TE^z.

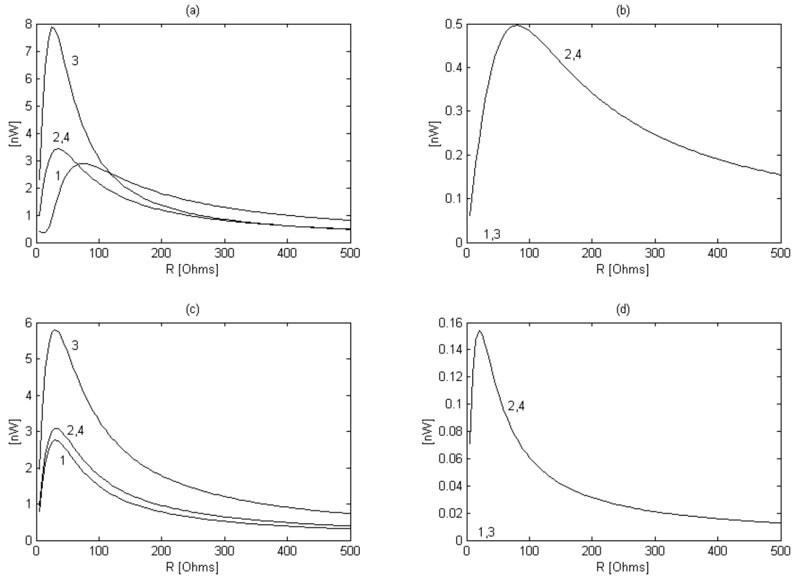

Figure 4. The power delivered to the loads of right or left end of CSMSL exposed by normal incidence, versus the load values a) $\varepsilon_r = 2.54$ and TM^z b) $\varepsilon_r = 2.54$ and TE^z c) $\varepsilon_r = 10.0$ and TM^z d) $\varepsilon_r = 10.0$ and TE^z.

5. CONCLUSION

In this paper a primary/secondary field regime is used to calculate the effect of external field illumination on a Circulant Symmetric Multi-conductor Microstrip Line or CSMSL structure. For both possible TE^z and TM^z polarizations, primary field components are calculated analytically. Then the net effect of excitation on line voltage and current is obtained. The method is examined to obtain the effect of external illumination on a four strip CSMSL. It has been concluded that the powers delivered to the line(s) behind or in front of the metallic core are negligible in the case of TE^z incidence and the power delivered to the line(s) behind the metallic core is smaller than that of the other lines in both polarizations. Also, the delivered powers decrease as the permittivity of the dielectric rise. Although the method has been applied to a CSMSL, it is worthy of mention that symmetry of the strips is not a necessary or restricting condition to the applicability of the method.

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