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COUPLING MODEL FOR THE TWO ORTHOGONAL MICROSTRIP LINES IN TWO LAYER PCB BOARD (QUASI-TEM APPROACH)

M. Hashemi-Nasab and A. Cheldavi

College of Electrical Engineering
Iran University of Science and Technology
Tehran, Iran

Abstract—In the present paper a simple model has been given to simulate the signal propagation through two cross orthogonal microstrip lines in two different layers of the PCB board. First the structure has been analyzed using full wave software like HFSS, then a simple and suitable lumped equivalent circuit is proposed for the cross talk region and its parameters are obtained. Finally the s-parameters of this equivalent circuit compared with the results of full wave simulations. The results show good agreement up to some GHz.

1. INTRODUCTION

Now a day multilayer PCB boards are vastly used in the communications and specially RF circuits. Multilayer boards usually designed in such a way that transmission lines of one layer are orthogonal with transmission lines on the two adjacent layers. This procedure followed to decrease coupling phenomenon between two adjacent layers. There are a lot of classical model for the parallel multi conductor transmission line structure on the same layer of the PCB board or on the different layers [1–5]. But these methods can not be used for the orthogonal line structure, because in TEM modes, two orthogonal microstrip lines should not have any coupling to each other. In quasi-TEM mode longitudinal field, although very small, has a coupling effect to the orthogonal lines on the other layers.

To analyze this structure, with notice to electric and magnetic field distribution on the cross region, an equivalent simple lumped circuit is proposed as the first approximation. Then a simple method is proposed to obtain the optimized values for the equivalent circuit parameters.

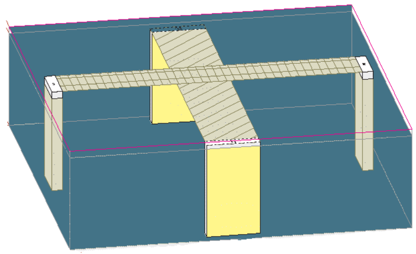


Figure 1. Two layer orthogonal microstrip transmission line.

The scattering matrix of the equivalent circuit is then calculated and compared with the scattering matrix obtained from the full wave analysis. Comparison of the equivalent model results and full wave analysis shows good agreement.

This paper is concentrated to the two strip lines in two layers but the results and this equivalent circuit can be generalized to more than two layers, and more lines on each layer.

2. EQUIVALENT CIRCUIT

Consider a two layer PCB structure with a microstrip line in each layer. In quasi-TEM approximation, for each line, the electric field lines are tied vertically to the ground plane. Magnetic field in this mode is closed around the strip. Propagation of the electric and magnetic fields can be modeled using distributed capacitance and inductance of the transmission line in form of lumped elements [6, 7]. In the first approximation losses are ignored. The coupling phenomenon is appeared in the middle of the lines where the lines crosses each other. We called this region "crosstalk region".

In the cross talk region electric field of the upper strip first closed to the down strip and then closed to the ground plane. This kind of field distribution causes coupling between two strips and forces the coupled signal to pass through down strip even when the down strip has no excitation . The upper microstrip signal is quasi-TEM and has longitudinal fields, so a leakage quasi-TEM signal propagates in the down strip. As the first approximation, this coupling phenomenon can be modeled by a mutual capacitor between two strips. Of course, when the upper strip has an excitation, the coupled signal from the upper strip to the down strip is more than the coupled signal from the down strip to the upper strip (when the down strip has an excitation and the upper strip has no excitation). In our first order approximation model we ignore this effect. Also, in a small region after crosstalk region the electric field of the upper strip instead of ground plane tied to the down strip. But in this research this effect also has been ignored.

To model the crosstalk region, first note that the region is too small, so the short length of the transmission lines in each layer can be modeled as two small inductors. To model the coupling phenomenon, as the first approximation, a mutual capacitance is considered between two transmission lines. Respect to these explanations, the equivalent circuit is proposed as shown in Fig. 2. In Fig. 2 the rectangular boxes represent the length of the transmission lines in four directions out of the crosstalk region. It is supposed that these 4 sections of transmission lines have no coupling effect to each other.

3. PARAMETERS OF THE EQUIVALENT CIRCUIT

The parameters of the equivalent circuit will be calculated when both dielectric layers have the same ε_r and same height. For the case of two different layers with different ε_r, the electric field distribution of the up and down strips will be more irregular, therefore calculating the structure parameters will be more complicated. First a rough estimation for the parameters of the model has been proposed. This estimation will be modified and optimized later using the exact scattering parameters of the crosstalk region obtained from HFSS.

3.1. Self-Capacitance

First the self capacitance between the down strip and the ground plane has been obtained. For microstrip line structure the per length capacitance can be obtained from [6]:

$$\begin{aligned}
\frac{w}{h} > 3 \Leftrightarrow C &= 4\varepsilon_r\varepsilon_0\left[\frac{w}{h}+\frac{2}{\pi}\ln 2\right]\frac{F}{m} \\
\frac{w}{h} < 3 \Leftrightarrow C &= 2\pi\varepsilon_r\varepsilon_0\left[\ln\left(\frac{8}{\pi}\cdot\frac{2h}{w}\right)+\frac{\pi^2}{48}\cdot\left[\frac{w}{2h}\right]^2\right]^{-1}\frac{F}{m}
\end{aligned} \tag{1}$$

The upper strip affects the electric field distribution between the down strip and the ground plane, and this effect decreases the self capacitance of the down strip. So, this value is the first approximation of the real value in this structure. Optimized value of self capacitance can be calculated from comparison of the scattering matrix parameters of this model and the results of full wave analysis.

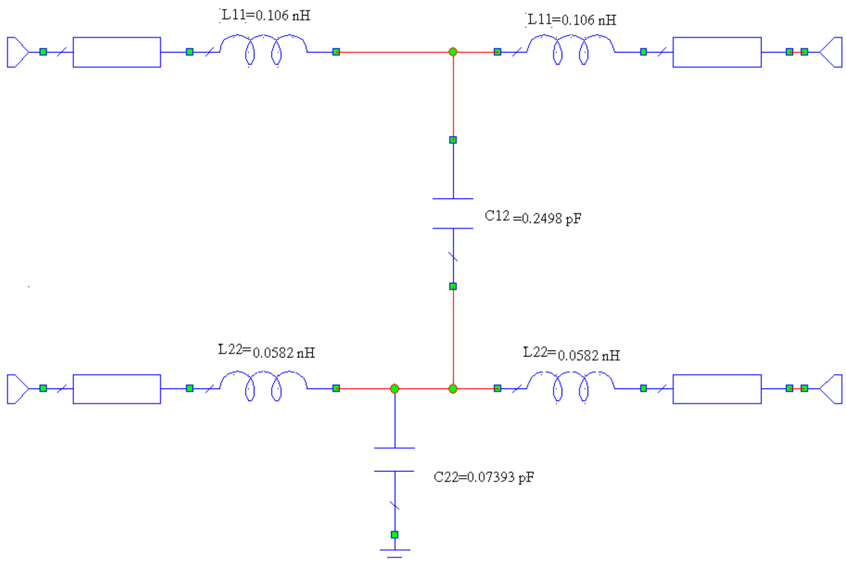


Figure 2. Equivalent circuit of two orthogonal microstrip lines in the cross talk region.

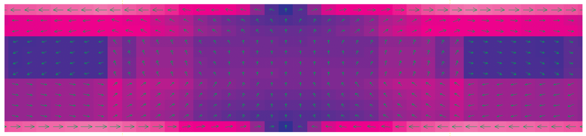
3.2. Inductance

For upper strip line, with a good approximation we can assume current distribution to be uniform. Both strips designed to be 50 ohm, so height of the upper strip is twice than the down strip, and its width is twice of the down strip. For both strips the width is much more than the thickness. Effects of the down strip to the upper strip current distribution are such low that can be ignored. Assuming current distribution of the upper strip to be uniform, one has [3]:

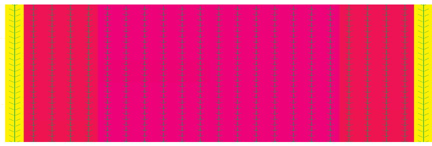
$$L_{plate} = \frac{\mu_0}{2\pi} \cdot l \cdot \left(\ln\left(u + \sqrt{u^2+1}\right) + u \cdot \ln\left(\frac{1}{u} + \sqrt{\left(\frac{1}{u}\right)^2 + 1}\right) + \frac{u^2}{3} + \frac{1}{3u} - \frac{1}{3u}(u+1)^{3/2} \right) F \tag{2}$$

When $u = \frac{l}{w}$ and $l = \frac{W_{down}}{2}$. For both L_{up} and L_{down} equation (2) is used.

Again this formula has been used as the first approximation and for optimized value an optimization method will be used.



(a)



(b)

Figure 3. Current distribution of two strip (a) down strip (b) upper strip.

With notice to Fig. 3 current distribution of the down strip in the cross region is completely non uniform. For increasing accuracy we separate cross region in two parts and calculate inductance for left part and right part separately.

In the other hand distributed model for cross talk region is a crosstalk current source (modeled using mutual capacitance) placed in the middle of the down microstrip in the cross region and power is distributed to the two side loads.

3.3. Mutual Capacitance

With notice of field distribution, as the first approximation, the cross region can be modeled by a plate capacitor with considering fringing field of the 4 sides. Ideal capacitance of a plate can be obtained as:

$$C = \varepsilon_r \varepsilon_0 \frac{W_1 W_2}{h} \tag{3}$$

Fringing fields of sides 1 and 2 considered by using W_{eff} in (3).

$$W_{eff} = \frac{2\pi h}{\ln\left[\frac{hF}{w} + \sqrt{1 + \left(\frac{2h}{w}\right)^2}\right]} \tag{4}$$

$$F = 6 + (2\pi - 6) \cdot e^{-\left(\frac{4\pi^2}{3}\right)\cdot\left(\frac{h}{w}\right)^{-\frac{3}{4}}} \tag{5}$$

Fringing fields of sides 3 and 4 will be considered by replacing

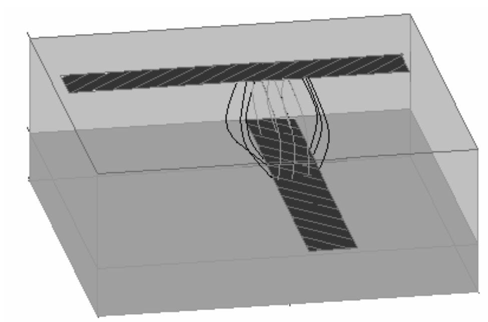


Figure 4. Cross section field distribution.

effective length instead of W_2.

$$\frac{\Delta l}{h} = 0.412 \cdot \frac{\varepsilon_{eff} + 0.3}{\varepsilon_{eff} - 0.258} \cdot \frac{\frac{w}{h} + 0.262}{\frac{w}{h} + 0.813} \tag{6}$$

Maximum error of Δl when compared with the practical data, is less than $0.05h$ for $\varepsilon_r = 1$ and less than $0.01h$ for $\varepsilon_r > 2.5$.

4. FREQUENCY DEPENDENCE OF THE MUTUAL CAPACITANCE

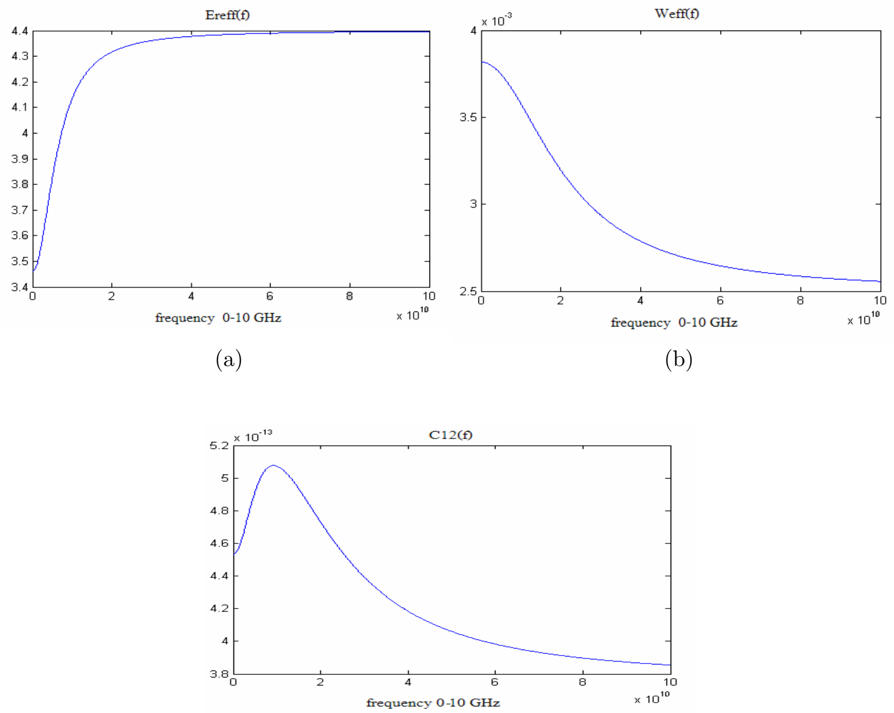
As frequency increases fields being more confined to the region below the strip and it causes ε_{reff} to increase up to the ε_r.

$$\varepsilon_{eff} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \cdot \left(1 + 10\frac{h}{w}\right)^{-0.5} \tag{7}$$

Getsingers in [10] obtained a relation for the frequency dependence of the ε_{reff} in the frequency range below 18 GHz.

$$\varepsilon_{eff}(f) = \varepsilon_r - \frac{\varepsilon_r - \varepsilon_{reff}}{1 + G \cdot \left(\frac{f}{f_p}\right)^2} \tag{8}$$

$$f_p = \frac{Z_0}{2\mu_0 h}, \qquad G = 0.6 + 0.009 \cdot Z_0$$



(a) (b)

(c)

Figure 5. Plot (a) ε_{reff} and (b) W_{eff} and (c) C_{12} variation in 1–10 GHz.

Owenze [11] obtained a relation for $W_{eff}(f)$ and then used it to estimate $Z(f)$:

$$W_{eff}(f) = W + \frac{W_{eff} - W}{1 + \left(\frac{f}{f_p}\right)^2}, \quad f_p = \frac{c}{2 \cdot W_{eff} \cdot \sqrt{\varepsilon_{reff}}}, \quad W_{eff} = \frac{h\eta}{Z_0 \cdot \sqrt{\varepsilon_{reff}}} \tag{9}$$

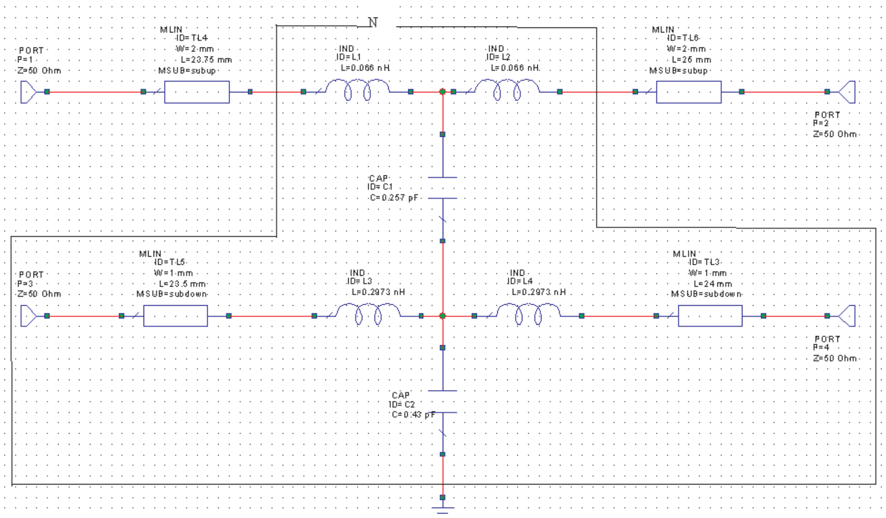


Figure 6. Equivalent two port network.

With using this relation, the mutual capacitor of the cross talk region becomes:

$$C_{12} = \frac{w_2}{h_{12}} \cdot \left(\varepsilon_r - \frac{\varepsilon_r - \varepsilon_{reff}}{1 + G \cdot \left(\frac{f}{f_p} \right)^2} \right) \left(W + \frac{W_{eff} - W}{1 + \left(\frac{f}{f_p} \right)^2} \right) \qquad (10)$$

Anyway it is not so important when the frequency range of our interest is not too broad. Also, we finally try to optimize the model and then get to the optimized values compared to the exact S-parameters.

5. OPTIMIZING EQUIVALENT CIRCUIT PARAMETERS

For optimizing model parameters we first terminate two ports of the down strip with their match loads. Considering cross region plus down strip, as a two port network, calculate ABCD parameters of this network and then with multiplying this transform matrixes, obtain ABCD and then scattering matrix between port 1 and 2 (see Fig. 6).

The elements of the crosstalk region ABCD matrix depends on C_{down}, L_{down}, C_{12}, L_{up} parameters therefore total ABCD matrix and also total scattering matrix depends on this parameters. In this structure because of symmetry, $S_{12} = S_{21}$ and $S_{11} = S_{22}$. To obtain optimized values of parameters, the exact S_{11} and S_{12} parameters are obtained from full wave simulation. This parameters are then compared to the s-parameters obtained from the model that depends on C_{down}, L_{down}, C_{12}, L_{up}. So, we obtain the optimized values of these parameters in a frequency band of 0.1–3 GHz using minimum least square error over 0.1–3 GHz frequency band.

It is very interesting to note that in a practical example the optimized values are very close to the main model parameters obtained using (1)–(3).

6. A PRACTICAL EXAMPLE

As an example consider two orthogonally crossed microstrip lines as shown in Fig. 1. The width and length of two microstrip lines are designed in such a way that impedance of both strip lines be 50 ohms. We used FR4 proxy fiber with $\varepsilon_r = 4.4$. In this fiber for 50 ohm lines w/h ratio should be 1.8. Height of two fibers are selected 0.635 mm and $W_{up} = 2.3495$ mm and $W_{down=1.174}$ mm. Length of the both microstrip lines is 1 cm. Fig. 7 shows the results of s-parameters obtained from HFSS simulation and the s-parameters obtained from the proposed lumped model in this paper.

For this structure the parameters of the lumped model are obtained from (1)–(3) as:

$$\begin{aligned} C_{down} &= 0.44\,\text{pF} & L_{down} &= 0.23\,\text{nH} \\ C_{coupling} &= 0.254\,\text{pF} & L_{up} &= 0.066\,\text{nH} \end{aligned}$$

The optimized parameters are obtained as:

$$\begin{aligned} C_{down} &= 0.44\,\text{pF} & L_{down} &= 0.21\,\text{nH} \\ C_{coupling} &= 0.250\,\text{pF} & L_{up} &= 0.05\,\text{nH} \end{aligned}$$

As it is clear the optimized values, with the goal function defined as the minimum least square error in the 0.1–3 GHz, are very close to that obtained from (1)–(3). To obtain better solution the goal function should be changed, and the frequency range should be reduced. For example for the frequency range 0.1–2 GHz the optimized values are given as

$$\begin{aligned} C_{down} &= 0.44\,\text{pF} & L_{down} &= 0.13\,\text{nH} \\ C_{coupling} &= 0.254\,\text{pF} & L_{up} &= 0.033\,\text{nH} \end{aligned}$$

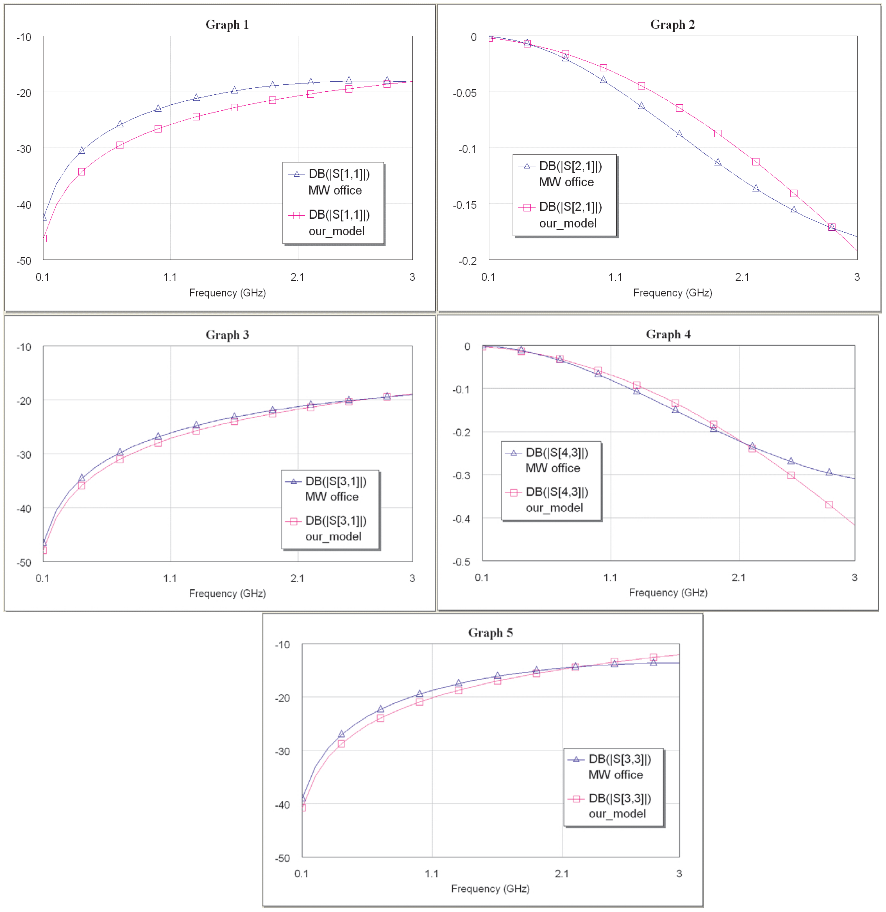


Figure 7. Comparing results of the equivalent circuit and physical structure.

7. CONCLUSION

Because of propagating of some non-TEM mode in microstrip structure, two orthogonal microstrip transmission line in different layer of a PCB has coupling effect to each other. The simulation results shows that, passing a signal through upper strips affects current distribution in the down strip in the region of approximately $3W_{down}$. An equivalent lumped model has been presented to simulate the coupling effect in cross talk region. This model as a four port network, can be replaced by the crosstalk region up to some GHz. The s-parameters of this lumped model have been compared with the results of HFSS simulation and shows good agreement.

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