Analysis and Reduction of Crosstalk on Coupled Microstrip Lines by Using FDTD Method

Pichaya Supanakoon[1], Monchai Chamchoy[1], Panarat Rawiwan[1], Prakit Tangtisanon[1]
Sathaporn Promwong[1], Teerasilpa Dumwipata[2] and Jun-ichi Takada[3]
[1]Department of Information Engineering, Faculty of Engineering,
King Mongkut's Institute of Technology Ladkrabang, Bangkok 10520, Thailand.
Phone: (+66)2737-2500-47 ext. 5140, E-mail: kspichay@kmitl.ac.th
[2]Department of Industrial Electrical Technology, Faculty of Engineering,
King Mongkut's Institute of Technology North Bangkok, Bangkok 10800, Thailand.
[3]Graduate School of Science and Engineering, Tokyo Institute of Technology,
Tokyo 152-8552, Japan. E-mail: takada@ide.titech.ac.jp

Abstract: The crosstalk among coupled microstrip lines is the major limiting factors of signal qualities in the high-speed digital and communication equipment. In this paper, a three-dimensional finite difference time domain (FDTD) method is applied to analyze the crosstalk between the coupled microstrip lines. The proposed structures of the coupled microstrip lines are investigated to reduce the coupling in a simple way by modifying their ground plane with an optimum gap. The examples of these structures with the different sizes of the gaps on their ground plane are studied. These structures are considered as the four-port network to evaluate transmission efficiency, near- and far-end crosstalk. Gaussian pulse is excited to evaluate the frequency characteristics from dc to 30 GHz. The transmission efficiency, near- and far-end crosstalk of each structure of the coupled microstrip lines are demonstrated. The numerical results of this study show that the majority of crosstalk is the far-end crosstalk. The usage of the optimum gap on the ground plane can reduce the far-end crosstalk of the coupled microstrip lines while the transmission efficiency is nearly equal.

1. Introduction

The coupling of microstrip lines is one of the most important topics in computer and microwave industries. The unwanted electromagnetic coupling between the coupled microstrip lines can seriously affect the signal qualities in analog and high-speed digital systems. The crosstalk between the coupled microstrip lines may result in the distortion of the analog signal and digital pulse, and may even cause the loss of digital information. Therefore, the reduction of the crosstalk among the coupled microstrip lines to a tolerable level is the important goal in a circuit design. The far-end crosstalk reduction in two coupled lines is considered in [1-2]. Many structures of the coupled microstrip lines are designed for the crosstalk reduction such as additional grounded PCB tracks between strip lines [3-4], a dual-plane microstrip interconnect [5], the coupled microstrip lines with a notch [6] and using via fences placed between the strip lines [7].

Usually, the problem of the crosstalk is analyzed by using a circuit approach. It is convenient for an engineer to understand EMI. However, as EMI is essentially a field problem, a field approach is more suitable to analyze this problem. The effect of the crosstalk reduction can be analyzed by using quasi-TEM and full-wave spectral domain approaches [8] and a field-oriented PCB layout CAD method [9]. The most popular approach for solving an electromagnetic in time domain is the finite difference time domain (FDTD) method. This technique, introduced by Yee in 1966 [10], is used to numerically solve Maxwell's equations in the time domain on a meshed computational domain. This method is simple to implement for complicated dielectric structures because arbitrary electrical parameters can be assigned to each lattice cell as well as boundary conditions is also simple to model.

In this paper, the three-dimensional FDTD method is applied to analyze the crosstalk between the coupled microstrip lines. The proposed structures of the coupled microstrip lines are investigated to reduce the coupling in a simple way by modifying their ground plane with an optimum gap. The examples of these structures with the different sizes of the gaps on their ground plane are studied. These structures are considered as the four-port network to evaluate transmission efficiency, near- and far-end crosstalk. Gaussian pulse is excited to evaluate the frequency characteristic form dc to 30 GHz. The transmission efficiency, near- and far-end crosstalk of each structure of the coupled microstrip lines are demonstrated. The numerical results of this study show that the majority of crosstalk is the far-end crosstalk. The usage of the optimum gap on the ground plane can reduce the far-end crosstalk of the coupled microstrip lines while the transmission efficiency is nearly equal.

2. FDTD Methodology

2.1 Finite Difference Equations

Finite difference equations are derived directly from Maxwell's curl equations in time domain. For the lossless coupled microstrip lines, the strip and the ground plane are made of a perfect conductor ($\sigma = \infty$) that has zero thickness, and substrate has a relative dielectric constant of ε_r. The Maxwell's curl equations can be written as

$$\mu\frac{\partial\vec{H}}{\partial t} = -\nabla\times\vec{E}, \quad (1)$$

$$\varepsilon\frac{\partial\vec{E}}{\partial t} = \nabla\times\vec{H}. \quad (2)$$

To obtain discrete approximation of the continuous partial differential equations, the centered difference approximation is used on both time and space first-order partial differences. The entire computation domain is the collection of all unit cells. The dimensions of the unit cell along $x-$, $y-$ and $z-$directions are Δx, Δy and Δz, respectively. The node with subscript indices i, j and k corresponds to node number in $x-$, $y-$ and $z-$directions. This notation implicitly assumes the $\pm 1/2$ space indices. The time steps are indicated with the superscript. After simple arrangement, the finite difference equations are described in [10]. The half time steps indicate that E and H are calculated alternately. In these equations, the values of permittivity and permeability are the function of locations. For the electric field components on the interface of two different dielectric materials, the average of the dielectric constant $(\varepsilon_1 + \varepsilon_2)/2$ is used as the permittivity. A maximum time step is limited by the stability restriction of the finite difference equation [11],

2.2 Voltage Source and Resistor Models

A voltage source is represented as the electric field E in the $z-$direction at the node i_s, j_s and k_s along $x-$, $y-$ and $z-$axis, respectively. If the resistance of the source is set to R_S then the usual FDTD electric field at the source location is given by [12]

$$E_z\big|^n_{i_s,j_s,k_s} = \frac{1}{\Delta z}V_S\big|^n + I_S\big|^{n-\frac{1}{2}}, \quad (3)$$

when the current through the source is given by

$$I_S\Big|^{n-\frac{1}{2}}_{i_s,j_s,k_s} = \frac{H_y\Big|^{n-\frac{1}{2}}_{i_s+\frac{1}{2},j_s,k_s} - H_y\Big|^{n-\frac{1}{2}}_{i_s-\frac{1}{2},j_s,k_s}}{\Delta x} - \frac{H_x\Big|^{n-\frac{1}{2}}_{i_s,j_s+\frac{1}{2},k_s} - H_x\Big|^{n-\frac{1}{2}}_{i_s,j_s-\frac{1}{2},k_s}}{\Delta y}. \quad (4)$$

For a resistor, the voltage source V_S is only set to zero.

2.3 Excitation Pulse

The wideband Gaussian pulse with dc constant is desirable as excitation because its frequency spectrum is also Gaussian. Therefore, it provides frequency domain information from dc to desired cutoff frequency by adjusting the width of the pulse. The wideband Gaussian pulse in time domain can be written as

$$V_S\big|^n = V_0\exp\left[-\left(\frac{n-n_0}{n_{decay}}\right)^2\right], \quad (5)$$

where n is the time step. The pulse has the maximum amplitude V_0 at the time step n_0 and has the $1/e$ characteristic decay of n_{decay} time step.

2.4 PML ABC Treatment

The tangential field components on the six mesh walls must be specified in such a way that outgoing waves are not reflected. The FDTD simulation in this paper used the perfectly matched layer absorbing boundary condition (PML ABC) [13].

The PML ABC can absorb propagation wave effectively by using nonphysical lossy media adjacent to the outer-grid boundaries backed by perfectly conducting walls. Based on the splitting of the field components into two subcomponents, electric loss (σ) and magnetic loss (σ^*) for the PML medium are specified by satisfying PML impedance-matching condition. After introduction of the electric and magnetic losses, the electromagnetic waves inside the PML medium will rapidly attenuate. The exponential time stepping of the field updating equations is used to replace the conventional FDTD algorithm.

The electric loss in the PML region is assumed to increase with depth from zero at $\rho = 0$ to a maximum value of σ_{max} at $\rho = \delta$ by

$$\sigma(x) = \sigma_{max}\left(\frac{\rho}{\delta}\right)^m. \quad (6)$$

The expression of the PML reflection coefficient at normal incident is

$$R(0) = \exp\left[\frac{-2\sigma_{max}\delta}{(m+1)\varepsilon c}\right]. \quad (7)$$

Polynomial scaling provides two parameters for a fixed PML thickness: σ_{max} and m. For large m, the conductivity distribution is relatively flat near the interface of the PML and the working volume. However, deeper into the PML, the conductivity increases more rapidly than for small m. In this region, the field amplitudes are substantially decayed and reflections due to the discretization error contribute less. Typically, m in the rage between 3 and 4 has been found to be suitable [14].

3. Numerical Results

In this section, the effects of the reducing crosstalk between the coupled microstrip lines using the optimum gap on the ground plane are investigated. The proposed structure of the coupled microstrip lines with the gap on ground plane is shown in Fig. 1. The dielectric substrate has the dielectric constant $\epsilon_r = 2.17$ and the thickness $h = 0.762$ mm. The strips and the ground plane are zero thickness and perfect conductor. The strips have the wide $w = 1.278$ mm. The lengths between the strips and from the edge to each strip are $W = 2.130$ mm. The strips have the length $L = 30$ mm. Four cases of the gap sizes are computed for this structure. For case a, the gap size is assigned as $g = 0$ mm. That

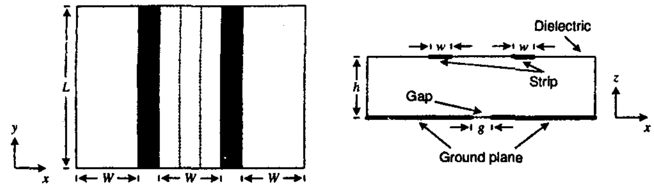

Figure 1. Geometry of the coupled microstrip lines structure with gap on the ground plane.

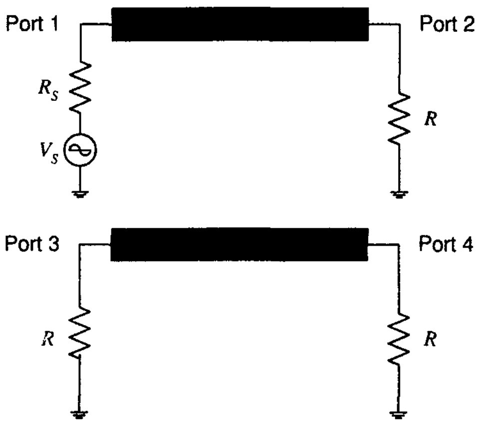

Figure 2. Equivalent four-port network of the coupled microstrip lines.

is the structure of the conventional coupled microstrip lines and its crosstalk effect is used as refer once value reduction. Subsequently, the gap sizes are assigned as g = 1.278, 1.704 and 2.130 mm for case b, c and d, respectively.

In the FDTD analysis, the appropriate selection of time and spatial parameters can bring about convergent numerical results. The spatial cell sizes in each region of the computational space are in general chosen to be less than one twentieth the wavelength which is determined by the maximum frequency and the electric permittivity of that region. The default three-dimensional meshes based upon the above rules are alluded to as a reference for determining the three-dimensional meshes of geometrically presented here. Therefore, the cells size of Δx = 0.213 mm, Δy = 0.300 mm and Δz = 0.254 mm are used. The PML ABC with 16 layers is employed to reduce the reflection error at the edge of the simulation boundary. The PML reflection coefficient at normal incident is set to $R(0) = 10^{-5}$. The PML scaling parameter is $m = 3$. Then, the total lattice has the dimension of 86 × 143 × 46 cells. The time step based on each cell is chosen to be the maximum time step that satisfies the Courant condition. The time step of Δt = 0.478 ps is used, resulting in n_{max} = 10,000 iterations for this simulation.

To evaluate transmission efficiency, near- and far-end crosstalk, the structures of the coupled microstrip lines are considered as four-port network, shown in Fig. 2. The voltage resistive source and resistor models are used. The source and load resistors are set to $R_S = R = 50\ \Omega$, which are matched to the characteristic impedance of the microstrip line at low frequency. The time waveform for the voltage source used in this simulation is the wideband Gaussian pulse peaking at the time step n = 57 with a value of 1.0. The pulse excitation is provided by the following equation

$$V_S|^n = \exp\left[-\left(\frac{n-57}{19}\right)^2\right], \qquad (8)$$

the frequency of this pulse ranges from dc to about 30 GHz.

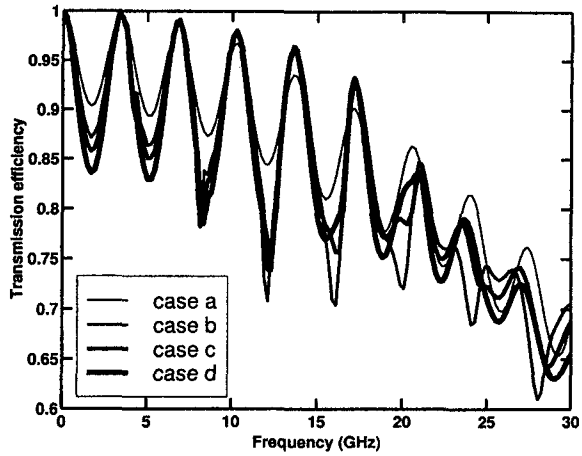

Figure 3. Transmission efficiency for each case of the coupled microstrip lines.

The transmission efficiency η is defined as a ratio of power and maximum available power of load at port 2. The computed transmission efficiency of each case of the coupled microstrip lines is shown in Fig. 3. From the figure we can see that the transmission efficiency of each case is nearly equal. The mean of the transmission efficiency of case a, b, c and d are 0.852, 0.832, 0.833 and 0.826, respectively.

The near- and far-end crosstalk is respectively defined as power of load at port 3 and 4 normalize by power of source. Fig 4 and 5 show respectively the near- and far-end crosstalk of each case of the coupled microstrip lines. From these figures, we can see that the majority of crosstalk is the far-end crosstalk. The mean of the near-end crosstalk of case a, b, c and d are −27.24, −27.09, −29.03 and −29.97 dB, respectively. In addition, The mean of the far-end crosstalk of case a, b, c and d are −16.47, −23.45, -23.29 and -23.24 dB, respectively. The far-end crosstalk can be reduced by the additional gap on ground plane about 7 dB.

4. Conclusions

In this paper, the FDTD method is applied to analyze the crosstalk between the coupled microstrip lines. The transmission efficiency, near- and far-end crosstalk of each structure of the coupled microstrip lines are demonstrated. From these results, they can be observed

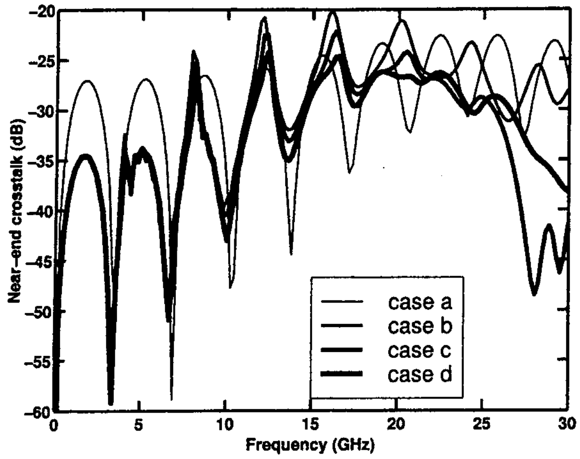

Figure 4. Near-end crosstalk for each case of the coupled microstrip lines.

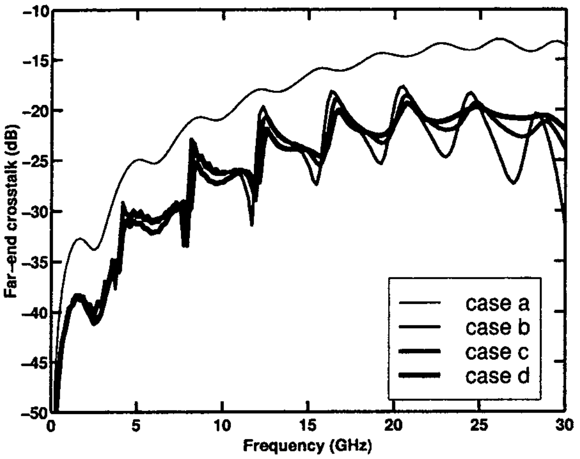

Figure 5. Far-end crosstalk for each case of the coupled microstrip lines.

that the majority of crosstalk is the far-end crosstalk. The usage of the optimum gap on the ground plane can reduce the far-end crosstalk of the coupled microstrip lines while the transmission efficiency is nearly equality. In addition, the FDTD method is flexible and convenient for evaluating the crosstalk on arbitrary coupled microstrip line structures.

References

[1] B. Eged, F. Mernyei, I. Novak and P. Bajor, "Reduction of far-end crosstalk on coupled microstrip PCB interconnects," *IEEE Instrumentation and Measurement Technology Conference*, vol. 1, pp. 287-290, 1994.

[2] Gazizov, "Far-end crosstalk reduction in doubled-layered dielectric interconnects," *IEEE Transaction on Electromagnetic Compatibility*, vol. 43, no. 4, pp. 566-572, 2001.

[3] I. Novak, B. Eged and L. Hatvani, "Measurement by vector-network analyzer and simulation of crosstalk reduction on printed circuit boards with additional center traces," *IEEE Instrumentation and Measurement Technology Conference*, pp. 269-274, 1993.

[4] I. Novak, B. Eged and L. Hatvani, "Measurement and simulation of crosstalk reduction by discrete discontinuities along coupled PCB traces," *IEEE Transaction on Instrumentation and Measurement*, vol. 43, no. 2, pp. 170-175, 1994.

[5] Y. Qian and E. Yamashita, "Characterization of a dual-plane microstrip interconnect with reduced pulse distortion and crosstalk," *IEEE MTT-S International Microwave Symposium Digest*, vol. 2, pp. 845-848, 1995.

[6] S. Dai, A. Z. Elsherbeni and C. E. Smith, "The analysis and reduction of crosstalk on coupled microstrip lines using nonuniform FDTD formulation," *IEEE MTT-S International Microwave Symposium Digest*, vol. 1, pp. 307-310, 1996.

[7] J. Gipprich and D. Stevens, "A new via fence structure for crosstalk reduction in high density stripline packages," *IEEE MTT-S International Microwave Symposium Digest*, vol. 3, pp. 1719-1722, 2001.

[8] J. C. Coetzee and J. Joubert, "Full-wave characterization of the crosstalk reduction effect of an additional grounded track introduced between two printed circuit tracks," *IEEE Transactions on Circuits and System-I: Fundamental Theory and Application*, vol. 43, no. 7, pp. 553-558, 1996.

[9] W. Xin, C. M. Lee, M. H. Pong, Z. Qian, "A novel approach base on electric field analysis to reduce crosstalk problem on PCB," *30th Annual IEEE Power Electronics Specialists Conference*, vol. 2, pp. 845-849, 1999.

[10] K. S. Yee, "Numerical solution of initial boundary value problems involving Maxwell's equations in isotropic media," *IEEE Transaction on Antennas and Propagation*, vol. 14, pp. 302-307, 1966.

[11] A. Taflove and M. E. Brodwin, "Numerical solution of steady-state electomagnetic scattering problems using the time-dependent Maxwell's equations," *IEEE Transaction on Microwave Theory and Techniques*, vol. 23, pp. 623-630, 1975.

[12] R. J. Luebbers and H. S. Langdon, "A simple feed model reduces time steps needed for FDTD antenna and microstrip calculations," *IEEE Transaction on Antennas and Propagation*, vol. 44, no. 7, pp. 1000-1005, 1996.

[13] J. P. Berenger, "A perfectly matched layer for the absorption of electromanetic waves," *J. Computational Physics*, vol. 114, pp. 185-200, 1994.

[14] J. P. Berenger, "Perfectly matched layer for the FDTD solution of wave-structure interaction problems," *IEEE Transaction on Antennas and Propagation*, vol. 51, pp. 110-117, 1996.