# A Crosstalk Reduction Technique for Microstrip MTL Using Mode Velocity Equalization

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***Abstract*—This paper proposes a simple method to enhance the timing margin of high-speed digital signals for multiconductor transmission lines (MTLs) in inhomogeneous media. It uses interline capacitances to implement the capacitance matrix homogenizing the original inhomogeneous structure. As a result, it equalizes the velocities of all the modes involved in the MTL structure simultaneously. The design equations are derived and the limitation of the proposed scheme is explained. For the experiment, a 100-mm three-line coupled microstrip MTL (M-MTL) is built. The measured total peak-to-peak jitter including crosstalk and other noises was 91 ps in the test structure. When the M-MTL is loaded by the proposed values of lumped capacitors between the adjacent lines, the jitter is reduced to 12 ps, whereas the measurement using the even/odd mode approach shows 22 ps jitter. All the measurement results are in good agreement with the theoretical values, which proves the effectiveness of the proposed method.**

***Index Terms*—Coupled transmission lines, crosstalk, electromagnetic coupling, jitters, transmission line.**

## I. Introduction

HIGH data rates, complex wiring, and reduced area are common specifications for the current high-speed digital system. This requires the routing area for the chip-to-chip wired communication section in the high-speed system to be small, so that the density of the interconnected region has been increased and the line space between the transmission conductors has been reduced. Consequently, the crosstalk due to the coupling effect between the two transmission lines is expected to raise the primary limiting factor in the printed circuit board design to maintain the signal integrity. Moreover, in chip-to-chip communication, the microstrip multiconductor transmission line (M-MTL) has been popular in routing structures. Unfortunately, it can produce serious far-end crosstalk due to its inhomogeneity.

Many papers have researched equalizer circuits used to eliminate distortion at the receiver block, such as the feed forward equalizer and the decision feedback equalizer [1]–[3]. These equalizer schemes may be good solutions to the crosstalk noise problem; however, they are not suitable for low-power and low-cost systems. Also, many papers have discussed the reduction of crosstalk-induced jitter (CIJ) using passive elements. These papers have proposed to resolve CIJ by inserting capacitors between lines to equalize the velocities of the even/odd modes that exist in the two lines involved [4]–[6]. However, M-MTLs have $N$ modes in $N$ transmission lines, so that the equalization of the two modes between two lines may not be good enough.

In this paper, we suggest a method for obtaining more accurate values for the compensation capacitors that match the velocities of all the modes in M-MTL simultaneously, so that the timing margin is enhanced and the eye-diagram deterioration is minimized. This method can be applied to high-speed single-ended signaling through parallel links. We have assumed that M-MTL is lossless because it has less influence on the transmission properties than the coupling effect [7].

This paper is organized as follows. Section II introduces the fundamental theory of M-MTL structure and proposes the new mode velocity equalization method. In Section III, finding the optimal termination values for M-MTL is addressed. In Section IV, we present a limitation of a single lumped element compensation method. After that, our simulation and measurement results are shown in Section V. Conclusion is given in Section VI.

## II. Theory and Proposed Method

In this study, we consider interconnections with $N$ transmission conductors and a reference conductor in an inhomogeneous medium that provides $N$ transmission channels. Without a loss of generality, let us consider an M-MTL ($\varepsilon_{r1} = 1$) as shown in Fig. 1. There are $N$ different orthogonal modes that can propagate with different propagation constants. On the other hand, these modes degenerate and propagate with the same propagation constant provided that the MTL is in the homogeneous medium ($\varepsilon_{r1} = \varepsilon_{r2}$) in Fig. 1. This observation is the key to dealing with the crosstalk issue examined in this paper.

The basic parameters of M-MTL are inductance $\mathbf{L}$ and capacitance $\mathbf{C}$ matrices. These parameters can easily be obtained by a field solver. The solution to the wave equation gives $N$ orthogonal eigenmodes propagating in the M-MTL. This approach is based on the frequency-independent resistance, inductance, conductance, and capacitance (RLGC) model. However, most of the practical interconnection lines have the frequency-dependent RLGC parameters. In order to reduce the errors caused by using the frequency-independent RLGC model, the $\mathbf{L}$ and $\mathbf{C}$ matrices are extracted at the operating frequency of the system. The matrices $\mathbf{Z}$ and $\mathbf{Y}$ are obtained by (1) with $\mathbf{R}$ and $\mathbf{G}$ equal to zero

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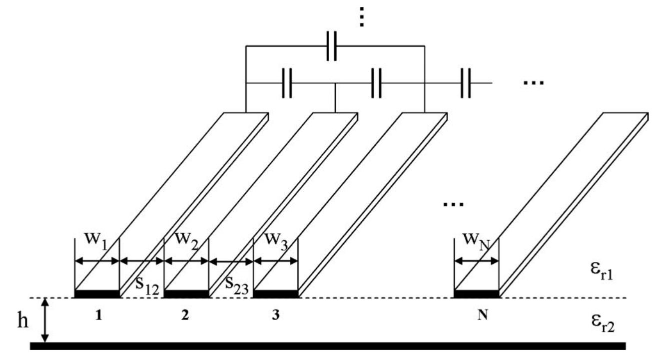



Fig. 1. Scheme for reducing crosstalk in M-MTL structure.

for lossless lines. The matrices **Z** and **Y** are symmetrical square matrices of order *N* and are frequency dependent. In this case, the propagation velocity of each mode can easily be obtained through a suitable diagonalization of the matrices **YZ** using the transformation matrix **S**, as in (2). The matrix **D** is a diagonal matrix of order *N* composed of the eigenvalues of **YZ**. Each of these eigenvalues is written as the square of the propagation constant $\gamma_\iota$ of the *i*th mode in (3) [7], [8]:

$$\mathbf{Z} = \mathbf{R} + j\omega\mathbf{L}, \quad \mathbf{Y} = \mathbf{G} + j\omega\mathbf{C} \tag{1}$$

$$\mathbf{S}^{-1}\mathbf{ZYS} = \mathbf{D} \tag{2}$$

$$\mathbf{D} = \mathrm{diag}_n(\gamma_1^2, \gamma_2^2, \ldots, \gamma_n^2). \tag{3}$$

In inhomogeneous media, the values of $\gamma_\iota$ generally differ from each other, so that the propagation velocities of the modes are different and large jitter results are at the far end. Therefore, it is necessary to minimize the difference between $\gamma_\iota$s in order to reduce the jitter.

The self-partial matrices that have all positive elements—$C_h$ and $L_h$ for M-MTL in homogeneous nonmagnetic media and $C_{ih}$ and $L_{ih}$ in inhomogeneous nonmagnetic media—are known to have an interesting relationship, as shown in

$$L_{ih} = L_h, \quad C_{ih} \neq C_h. \tag{4}$$

Notice that all the eigenvalues of **YZ** are the same for any capacitance matrix $C_{\mathrm{eq}}$ scaled from $C_h$ by $C_{\mathrm{eq}} = \alpha C_h$ even though the value itself differs from $C_h$. Therefore, we introduced a scaled capacitance matrix $C_{\mathrm{eq}}$ in (5), which is the capacitance matrix of the M-MTL in a homogeneous medium equivalent to the original inhomogeneous medium in the sense that the total self-capacitances are the same for these two structures. This is a concept similar to the effective dielectric constant used for microstrip line analysis:

$$C_{\mathrm{eq}} = \frac{\sum_{i=1}^{n} \mathrm{C}_{ih(ii)}}{\sum_{i=1}^{n} \mathrm{C}_{h(ii)}} C_h. \tag{5}$$

In (5), $C_{h(ii)}$ and $C_{ih(ii)}$ are self-capacitance elements (diagonal elements) of self-partial matrices of the *i*th conductor in a homogeneous and inhomogeneous medium, respectively. The equivalent homogeneous matrix $C_{\mathrm{eq}}$ equalizes the mode propagation velocities and becomes the target matrix for eliminating jitter in M-MTL. Fortunately, it was also found that each of the self-capacitances of $C_{\mathrm{eq}}$ is almost the same as those of $C_{ih}$. This result seems to be reasonable, considering the concept of effective dielectric constant for each line in M-MTL:

$$C_{eq(ii)} \cong C_{ih(ii)}, \qquad i = 1, \ldots, n. \tag{6}$$

The target matrix $C_{\mathrm{eq}}$ can be obtained by adding the interline capacitance matrix $C_{ic}$ to $C_{ih}$ in (7). The matrix $C_{ic}$ is actually a compensation matrix to make up the difference between $C_{\mathrm{eq}}$ and $C_{ih}$ and is shown in (8).

This suggests that all the modes in M-MTL modified properly by (8) have little difference in $\gamma_\iota$s, similar phase velocities, and a small CIJ at the far end:

$$C_{\mathrm{eq}} \cong C_{ih} + C_{ic} \tag{7}$$

$$C_{ic} \cong C_{\mathrm{eq}} - C_{ih} \cong \begin{bmatrix} 0 & C_{ic(12)} & \cdots & C_{ic(1n)} \\ C_{ic(21)} & 0 & \cdots & \cdots \\ \cdots & \cdots & \cdots & \cdots \\ C_{ic(n1)} & \cdots & \cdots & 0 \end{bmatrix}. \tag{8}$$

In order to realize the matrix $C_{ic}$, which is a distributed parameter based on per-unit-length, the lumped interline capacitor of the value $C_{ij,\mathrm{lumped}}$ given in (9) is attached between the *i*th and *j*th lines. Although the periodic mounting of many lumped capacitors with small values between two lines produces a better performance for the reduction of CIJ [4], this is not effective in a real system in terms of cost and area. Therefore, the additional capacitance of $C_{ij,\mathrm{lumped}}$ is loaded on the receiver chip as a lump element (as shown in Fig. 1)

$$C_{ij,\mathrm{lumped}} = C_{ic(ij)} \times \text{coupling length}. \tag{9}$$

The algorithm is based on the mimicking of a homogeneous MTL structure expressed by (5), where all the modes have the same mode velocity. It can be achieved by comparing the target capacitance matrix $C_{\mathrm{eq}}$ and the capacitance matrix $C_{ih}$ of the M-MTL, and augmenting proper capacitors compensating the differences between them, given by (8).

## III. Decision of Terminations Values

In the new M-MTL structure with interline capacitors from (9), the phase velocities are approximately the same for all modes. However, the characteristic impedances of modes need to be considered to estimate the reflection of the modes at the termination. This characteristic impedance matrix $\mathbf{Z_C}$ can be derived by

$$\mathbf{Z_C} = \mathbf{S\Gamma^{-1}S^{-1}Z} = \mathbf{S\Gamma S^{-1}Y^{-1}} \tag{10}$$

where

$$\mathbf{\Gamma} = \mathrm{diag}_n(\gamma_1, \gamma_2, \ldots, \gamma_n) \tag{11}$$

as **Z** and **Y** are given by (1) with $L_{ih} = L_h$ and $C_{\mathrm{eq}}$.

With this characteristic impedance matrix $\mathbf{Z}_C$, the reflection matrix **P** of the new M-MTL with interline capacitance is given by

$$\mathbf{P} = (\mathbf{Z_L} - \mathbf{Z_C}) \cdot (\mathbf{Z_L} + \mathbf{Z_C})^{-1} \tag{12}$$

for the given impedance matrix $\mathbf{Z_L}$ of th e termination. The element of the reflection matrix **P** is the voltage reflection coefficient $\rho_{ij}$.

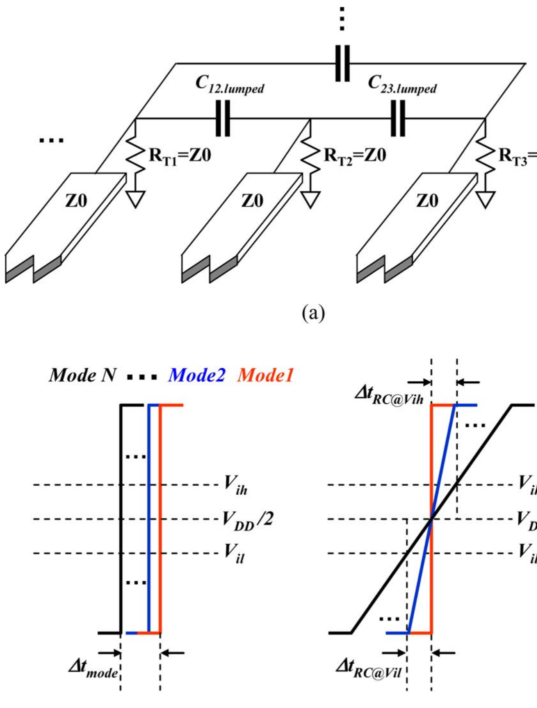



Fig. 2. (a) Termination structure of coupled M-MTL. (b) Timing notation.

The characteristic impedance matrix $\mathbf{Z_C}$ of the M-MTL and M-MTL with interline capacitance is not diagonal, whereas $\mathbf{Z_L}$ is diagonal since only the self-parallel resistors are used in receivers. To minimize the reflection noise, the proper value of termination $\mathbf{Z_L}$ is to be determined to minimize the maximum absolute column sum norm $\|\mathbf{P}\|$ obtained by (10)–(13) [8]:

$$\|\mathbf{P}\| = \max_j \sum_{i=1}^{n} |\rho_{ij(ii)}|\cdot \tag{13}$$

In this manner, the reflection noise of the M-MTL with interline capacitance can be obtained.

## IV. Single Lumped Element Compensation Method and Its Limitation

Although the proposed method shows how to calculate the compensation capacitors by (9), it has a limitation in its performance due to the fact that $C_{ij,\mathrm{lumped}}$s are lumped interline capacitors, whereas the derived compensation parameter $C_{ic(ij)}$ for equalized M-MTL is the distributed capacitance. In order to estimate the interconnection coupling length limitation of the proposed method, M-MTL with characteristic impedance $Z_0$ is terminated by $Z_0$ resistors for each line in Fig. 2(a), the supplied I/O voltage is $V_{DD}$, and a step pulse is applied as the excitation.

Since $N$-coupled M-MTL structure has $N$ modes and each mode has its own velocity, the difference in the propagation delay of these $N$ modes makes the jitter $\Delta t_{\mathrm{mode}}$, as shown on the left in Fig. 2(b). Also, notice that there exist the unwanted $RC$ delay $\Delta t_{RC}$ due to the finite rise/fall time in the mode voltage at the output due to the low-pass filter consisting of the lumped interline capacitor $C_{12,\mathrm{lumped}}$ and output load. This $RC$ delay, $\Delta t_{RC}$, is the required time for the signal to change from $V_{DD}/2$ to $V_{ih}$ or $V_{il}$ as shown on the right in Fig. 2(b). Therefore, the total jitter at the input high-voltage level $V_{ih}$ or the input low-voltage level $V_{il}$ is increased. Let $\Delta t_{\mathrm{target\,jitter}}$ be the target jitter determined by the system's timing margin. Then, these time factors should satisfy (14) for a successful operation:

$$\Delta \mathrm{t}_{\mathrm{CIJ@V}_{ih},\mathrm{V}_{il}} = |\Delta \mathrm{t}_{\mathrm{RC}} + \Delta \mathrm{t}_{\mathrm{mode}}| \le \Delta \mathrm{t}_{\mathrm{target\,jitter@V}_{ih},\mathrm{V}_{il}}. \tag{14}$$

To clarify the behavior of each mode affected by the lumped interline capacitor, the time-domain waveforms of every mode are to be investigated in the case of the M-MTL which is perfectly matched at the source side. When the load network is composed of interline capacitors and termination resistors at the end of transmission lines as shown in Fig. 2 (a), output voltage $\mathbf{V_L}$ can be obtained by (15) using the transmission coefficient $\mathbf{T_L}$ and incident voltage $V_{\mathbf{L}}^{+}$ in the Laplace s-domain:

$$\mathbf{V_L}(\mathbf{s}) = \mathbf{T_L}(\mathbf{s})\mathbf{V_L^+}(\mathbf{s}), \quad \mathbf{T_L} = \mathbf{P} + \mathbf{I} \tag{15}$$

where $\mathbf{P}$ is the reflection matrix of the incident voltage signal at the load given by (12) and $\mathbf{I}$ is a unity matrix.

For the diagonalization of (2), the voltage vector $\mathbf{V}$ is substituted by $\mathbf{S}\tilde{\mathbf{V}}$, where $\tilde{\mathbf{V}}$ is the transformed voltage. Hence, the transformed voltage output at the load can be represented by

$$\tilde{\mathbf{V}}_{\mathbf{L}}(\mathbf{s}) = \mathbf{S^{-1}T_L S}\tilde{\mathbf{V}}_{\mathbf{L}}^{+}(\mathbf{s}). \tag{16}$$

Note that $\mathbf{S}^{-1}\mathbf{T}_{L}\mathbf{S}$ in (16) is a nearly diagonalized matrix since the load network is obtained by lumping the distributed capacitance of the equalized transmission line into a lumped capacitor. Therefore, the $i$th mode, which is represented by a transformed voltage vector with a nonzero element only at the $i$th position, remains on the same mode without any conversion to other modes after passing through the load network.

In order to estimate the limitation of coupling length, the step function is used as an input since it gives the worst-case jitter of the structure [5], [6], [12]. The step input of the $i$th mode can be expressed by (17) in the s-domain:

$$\tilde{\mathbf{V}}_{\mathbf{L},i\mathbf{th}}^{+}(\mathbf{s}) = \left[0, \ldots, \frac{1}{s}, \ldots, 0\right]^t. \tag{17}$$

The $i$th mode output voltage in the time domain $\tilde{\mathbf{V}}_{\mathbf{L},i\mathbf{th}}(\mathbf{t})$ can be obtained by inverse Laplace transform of (16) with the input (17) as

$$\tilde{\mathbf{V}}_{\mathbf{L},i\mathbf{th}}(\mathbf{t}) = \mathcal{L}^{-1}[\tilde{\mathbf{V}}_{\mathbf{L},i\mathbf{th}}(\mathbf{s})]. \tag{18}$$

From (15) to (18), we can obtain the time-domain output waveforms of each mode and the waveforms can be used to find the coupling length limitation for a system. For example, the original 100-mm three-coupled M-MTL without any compensation capacitors has three modes with their own propagation velocities. These three different propagation velocities make a wide CIJ, 72 ps as shown in Fig. 3. If the distributed compensation capacitors are added between the lines, the difference in propagation velocities of each mode is reduced to $\Delta t_{\mathrm{mode}}$, 23 ps. On the other hand, the lumped interline capacitors at the end of the transmission lines increase the delay by $\Delta t_{\mathrm{RC}}$ at $V_{il}$ or

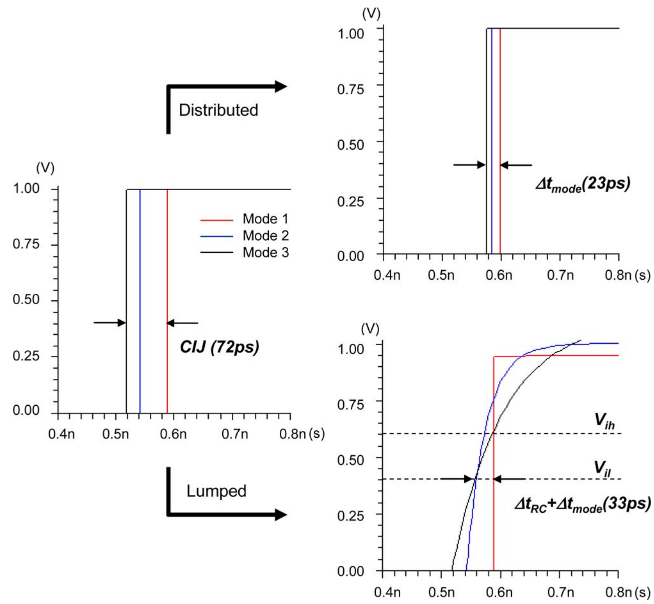
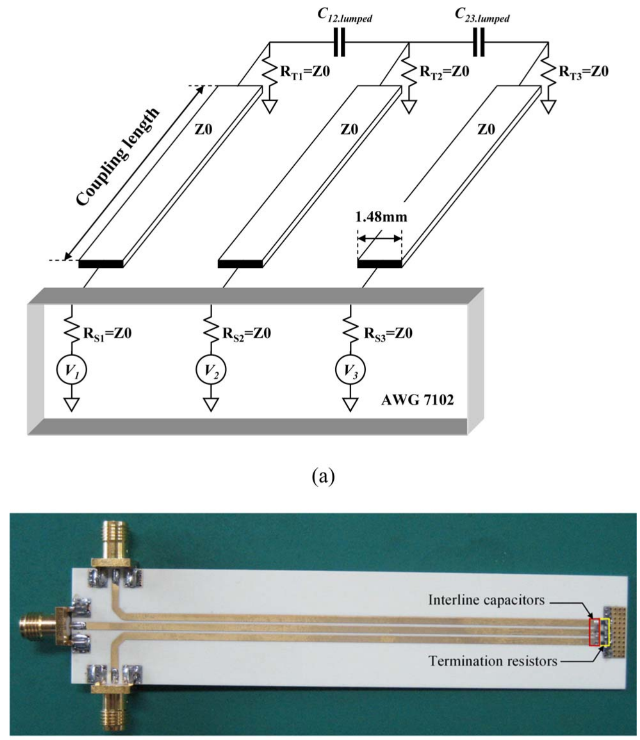



Fig. 3. Example of the jitter in the line with the distributed interline capacitance and with the lumped interline capacitance. The capacitor values are extracted by the proposed method.




(a)

(b)

Fig. 4. (a) Physical description of the test structure ($Z_0 = 50\ \Omega$). (b) Test fixture of three-line coupled transmission lines.

$V_{ih}$. We set $V_{il}$ and $V_{ih}$ to be $0.4V_{DD}$ and $0.6V_{DD}$, respectively, as shown in Fig. 3, where the effect of nonadjacent line coupling is neglected. In this way, we can determine the length limitation of the M-MTL interconnection system using (14) since $\Delta t_{RC}$ and $\Delta t_{\mathrm{mode}}$ depend on the length.

## V. Simulation and Measurement Results

To validate the proposed method, the simulation and measurements were performed for M-MTL with three 100-mm-length coupled 50 Ω lines on an RO4350B substrate with a $\varepsilon_r$, thickness, and width of 3.48, 0.762 mm, and 1.480 mm, respectively. In the test, pseudorandom-bit-sequence signals generated by utilizing 1.0-V $V_{DD}$ Tektronix's Arbitrary Waveform Generator 7102 were used. Tektronix's Digital Phosphor Oscilloscope 7254 and P6248 probes were also used for the test measurements as shown in Fig. 4.

The EM field solver program in Hspice was used to obtain the **L** and **C** matrices. From these matrices, the interline capacitance was obtained by using (1)–(8), as shown in Table I. The values of the interline capacitances obtained by the proposed method differ from those of the method based on the inductive ratio $L_{\mathrm{mutual}}/L_{\mathrm{self}} = C_{\mathrm{mutual}}/C_{\mathrm{total}}$ [4]–[6]. For the simulation and measurements, a 200-Mb/s operation data rate was used. But rise/fall time (20–80%) of the source (driver) was 60 ps, which is corresponding to that of a source over 2.0 Gb/s. To avoid unexpected noise from the refection or probe inductance and to show only the crosstalk noise effect, these source characteristics were chosen.

Fig. 5 shows the simulated output CIJ at the first line of the three coupled lines with various line-to-line spacing values (0.5–1.0 mm), when three lines are stimulated by pseudorandom bits in a synchronized transition. The interline capacitor between line 1 and line 3 does not affect CIJ of the center line (line 2) in

TABLE I
.Calculation Result of Interline Capacitance

| Line space(mm) | | Proposed interline capacitance (pF) | | | Inductive-ratio-based interline capacitance (pF) | | |
|---|---|---|---|---|---|---|---|
| line 1,2 | line 2,3 | line 1,2 | line 2,3 | line 1,3 | line 1,2 | line 2,3 | line 1,3 |
| 0.50 | 0.50 | 1.31 (1.30) | 1.31 (1.30) | 0.20 | 0.97 (1.00) | 0.97 (1.00) | 0.62 |
| 0.50 | 1.00 | 1.32 (1.30) | 0.91 (0.90) | 0.19 | 1.10 (1.10) | 0.74 (0.75) | 0.44 |
| 1.00 | 1.00 | 0.92 (0.90) | 0.92 (0.90) | 0.17 | 0.81 (0.80) | 0.81 (0.80) | 0.33 |
| 1.00 | 1.50 | 0.92 (0.90) | 0.69 (0.70) | 0.15 | 0.84 (0.85) | 0.62 (0.65) | 0.26 |

* ( ) values are used in measurement

the three-line coupled M-MTL. But, CIJs of lines 1 and 3 are different from that of line 2. The full interline compensation shows the best CIJ when the proposed interline capacitance values are used. The jitter is increased about 5–8 ps in this structure by using the adjacent line compensation capacitances only. However, it should be noticed that the adjacent line compensation by the proposed method shows better CIJ compared with the inductive-ratio-based method. Although the method given by [4] is good, the proposed method is more effective in reducing CIJ. This is explained by the fact that the inductive-ratio-based method considers only two modes between two lines involved, whereas the proposed method simultaneously considers all the modes involved in the entire line structure. All the signal propagation cannot be comprised of the combination of only two modes of propagation. Therefore, by (5)–(8), which gives the compensat-

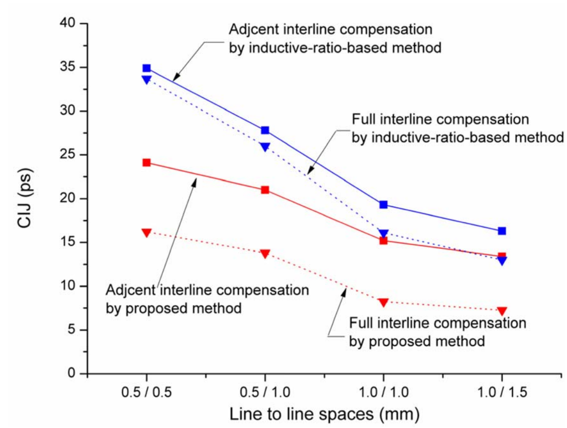



Fig. 5. Comparison simulation on adjacent and full interline capacitance effect at the first line of the three-line coupled MTL ($l = 100$ mm) by Hspice.

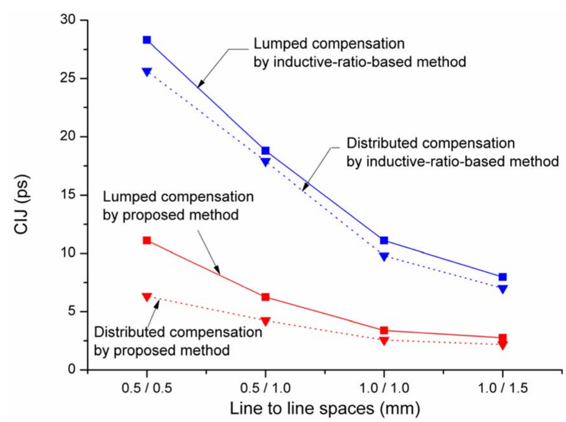
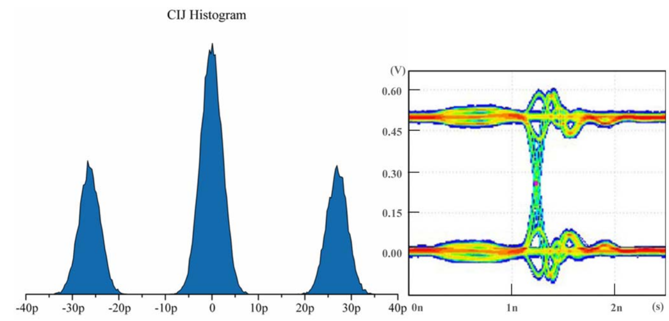



Fig. 6. Comparison simulation of CIJ using distributed and lumped interline capacitance at the center line in the three-line coupled M-MTL structure by Hspice.

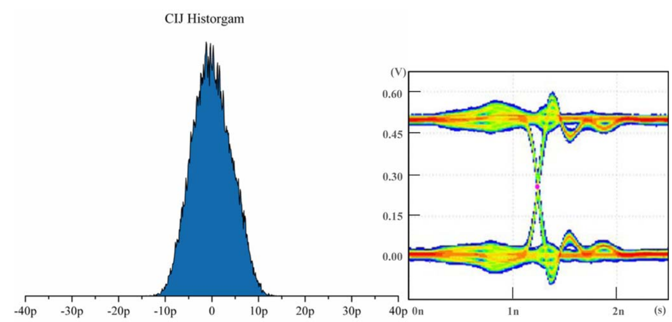
ing capacitances equalizing all the modes involved, we get more accurate values than the conventional method.

Fig. 6 shows the jitter difference between the lumped element capacitor $C_{ij \cdot \mathrm{lumped}}$ and distributed $C_{ij}$. As you can see, the error is shown to be less than 5 ps. Therefore, the lumped element compensation is an effective method of equalizing the mode velocities.

In Fig. 7, the eye diagrams and CIJ histograms were measured when the interline capacitor was mounted according to the values in Table I. Fig. 8 shows the simulation and test results obtained at the output of the center line (line 2). For the experiment, the interline capacitors were mounted only between the adjacent lines due to the dimensional limitations of lumped capacitors. Measurement data were close to the simulation results. In the case of three coupled lines with 0.5-mm line spaces without the interline capacitors, the CIJ was measured as 91 ps including crosstalk jitter and other jitters such as mismatch jitter. Although the exact values of the capacitors could not be used due to the limited availability of chip capacitor values, the measured total jitter with the interline capacitors was measured about 12 ps by the proposed method. It should be noted that the proposed method obtains an extra timing margin of 10 ps compared with that of the prior method. The results verify that the proposed method is an effective solution for the crosstalk reduction of M-MTL. The differences between the measured data and the simulation seem to come from line dispersion, termination mismatch, nonuniformity of transmission lines, and variation of the capacitors, which are not considered in the theory.




(a)




(b)

Fig. 7. (a) Measured eye diagram without interline capacitor. (b) Measured eye diagram with proposed interline capacitance values on the condition of CIJ at the first line of the three-line coupled MTL.

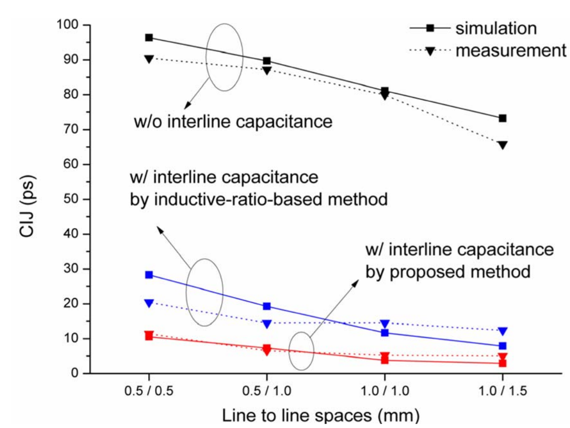



Fig. 8. Measurement result of CIJ at the center line of the three-line coupled MTL.

The proper value of the termination resistor $Z_L$ from (10) to (12) was found to be 50 Ω and $\|\mathbf{P}\|$ was 0.175 in the case of the 0.5 mm/0.5 mm line-to-line spaces M-MTL structure. But after adding the interline capacitors, the proper $Z_L$ is changed to 45 Ω and $\|\mathbf{P}\|$ increased to 0.220. When the termination resistor was fixed at 50 Ω, $\|\mathbf{P}\|$ changed to 0.263. These results mean that the reflection noise was increased by the passive equalization. Fortunately, the effect of reflection noise was small since the difference of $\|\mathbf{P}\|$ was below 0.09. In this case, the crosstalk noise was more dominant than the reflection noise.

To find the limitation of this proposed method, system specifications must be defined. Suppose the 2.0-Gb/s M-MTL system requires a maximum CIJ of 50 ps at $V_{ih}$ ($0.6V_{DD}$) or $V_{il}$ ($0.4V_{DD}$) on the condition that the driver/receiver has 50 $\Omega$ termination. Then, it is estimated that the maximum coupling length is restricted to about 70 mm in a conventional M-MTL system and the proposed structure with the interline capacitors increases the maximum coupling length to about 150 mm.

This proposed method is more effective in reducing the crosstalk compared with previous methods, even though its structural implementation is similar to those of prior methods.

## VI. Conclusion

In this paper, an effective method was proposed for the simultaneous equalization of the velocities of all modes in M-MTL. As a result, we obtained a better CIJ compared with the conventional even/odd mode compensation method. The optimized interline capacitors were calculated easily for any line spacing and coupling length. In addition, the performance was analyzed for the proposed technique and compared to other techniques using lumped capacitor loading.

It was shown that there was a coupling length limitation determined by *RC* delay, difference of the mode velocities, and the target jitter specifications. Due to the coupling length limitation, the proposed technique is more suitable for relatively short and dense channels having no guard trace like memory systems. The proposed crosstalk equalization technique offers low-cost, low-power implementation for high-speed single-ended multiple parallel channels.

Finally, the full interline compensation capacitors can be implemented by on-chip components using MIM or MOS capacitors in the CMOS receiver chip.

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