Modeling and Measurement of Mode-Conversion and Frequency Dependent Loss in High-Speed Differential Interconnections on Multilayer PCB

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**SUMMARY** We propose an accurate and efficient model of having an unbalanced differential line structure, where mode-conversion and frequency dependent loss effects are considered above the GHz frequency range. To extract model parameters of the proposed unbalanced differential line model, we measured s-parameters of test patterns using a 2-port VNA and defined a new type of mixed-mode s-parameter. The model parameters were obtained and are described for various types of the unbalanced differential line structures. Finally, the validity of the proposed model and the model parameters were successfully confirmed by a lattice diagram analysis.

**key words:** mode-conversion, unbalanced differential line, mixed-mode s-parameter, serial interconnection, frequency dependent loss

1. Introduction

As I/O data rate continues to increase, differential signaling schemes are expected to replace most of single-ended signaling schemes on PCBs for chip-to-chip links, because the differential signaling scheme has significant advantages in terms of signal integrity, power integrity, and electromagnetic interference (EMI) concerns [1]. First, the differential signal scheme rejects common-mode noise, injected or coupled from electromagnetic radiation and crosstalk [2]. In addition, a return current path is always guaranteed in the differential line structure, where mode-conversion and frequency dependent loss effects are considered in the GHz frequency range. The proposed differential line model consists of two simplified separate SPICE-type distributed RLGC circuit models. Because the two traces in a differential line are unbalanced, the model parameters of each trace are slightly different according to the trace imbalance. We also have defined a new type of mixed-mode s-parameter to enable convenient parameter processing and conversion.

The validity of the proposed model and the model parameters were successfully confirmed by a lattice diagram analysis, and by a series of time-domain measurements, including time domain reflectometry (TDR) measurement and time domain transmission (TDT) measurement.

2. Modified Mixed-Mode S-Parameter

The concept of the s-parameter was extended to differential interconnections and mixed-mode s-parameter was proposed in 1995 to describe wave propagations and reflections at the differential interconnections. The mixed-mode s-parameter includes propagations and reflections of differential-mode, common-mode and mode-conversion response, where mode-conversion consists of differential to common (DTC) mode-conversion and common to differential (CTD) mode-conversion. It is obtained by the relationship between incident waves and reflected waves of the normalized differential-mode waves and common-mode waves, as described in Eq. (1) [8],

\[
\begin{bmatrix}
    b_{dl1} \\
    b_{dl2} \\
    b_{cm1} \\
    b_{cm2}
\end{bmatrix} =
\begin{bmatrix}
    S_{dd11} & S_{dd12} & S_{dc11} & S_{dc12} \\
    S_{dd21} & S_{dd22} & S_{dc21} & S_{dc22} \\
    S_{cd11} & S_{cd12} & S_{cc11} & S_{cc12} \\
    S_{cd21} & S_{cd22} & S_{cc21} & S_{cc22}
\end{bmatrix} \begin{bmatrix}
    a_{dl1} \\
    a_{dl2} \\
    a_{cm1} \\
    a_{cm2}
\end{bmatrix}
\]

where \(a_{dl1}\) and \(a_{dl2}\) are the normalized differential-mode in-
are the normalized common-mode incident and reflected waves, respectively while \(a_{cmi}\) and \(b_{cmi}\)
are the normalized common-mode incident and reflected waves, respectively as shown in Fig. 1.

To create equivalent circuit model of an unbalanced differential line considering the mode-conversion, each trace of the unbalanced differential line is modeled separately. The existing mixed-mode s-parameter of Eq. (1) represents differential-mode, common-mode, and mode-conversion s-parameters, respectively. However, in a real case, when a differential-mode signal is applied to the differential line, the differential-mode and the common-mode signals are returned and transmitted simultaneously by the mode-conversion. Accordingly, we have modified the mixed-mode s-parameters to describe the differential-mode, the common-mode, and the mode-conversion simultaneously. As a consequence, the modified mixed-mode s-parameters represent the relation between the incident waves of the differential-mode \((a_{dmi}, a_{dm2})\), the common-mode \((a_{cm1}, a_{cm2})\) and the reflected waves at each port \((b_1, b_2, b_3\) and \(b_4\)).

The transformation can be developed by considering the relationships between the single-ended and mixed-mode reflected waves, which can be written as Eq. (2).

\[
\begin{bmatrix}
  b_1 \\
  b_2 \\
  b_3 \\
  b_4
\end{bmatrix}
= \frac{1}{\sqrt{2}}
\begin{bmatrix}
  1 & 0 & 1 & 0 \\
  -1 & 0 & 1 & 0 \\
  0 & 1 & 0 & 1 \\
  0 & -1 & 0 & 1
\end{bmatrix}
\begin{bmatrix}
  b_{dmi} \\
  b_{dm2} \\
  b_{cm1} \\
  b_{cm2}
\end{bmatrix}
\]

We can obtain the modified mixed-mode s-parameters if the transformation, Eq. (2), is put in the mixed-mode s-parameter form, such as in Eq. (3). The modified mixed-mode s-parameter is shown in Eq. (4):

\[
\begin{bmatrix}
  S_{d1d1} & S_{d1d2} & S_{d1c1} & S_{d1c2} \\
  S_{d2d1} & S_{d2d2} & S_{d2c1} & S_{d2c2} \\
  S_{cd1d1} & S_{cd1d2} & S_{cd1c1} & S_{cd1c2} \\
  S_{cd2d1} & S_{cd2d2} & S_{cd2c1} & S_{cd2c2}
\end{bmatrix}
= \begin{bmatrix}
  a_{dmi} \\
  a_{dm2} \\
  a_{cm1} \\
  a_{cm2}
\end{bmatrix}
\]

Each partition of the modified mixed-mode s-parameter represents a two-by-four s-parameter submatrix. The partition labeled \(S_{ij}\) is the relation between the differential-mode incident waves and the reflected waves to each ports of Fig. 1, where \(i\) is the single ended port number, and \(j\) denotes the differential-mode or common-mode port number. For example, \(S_{2d1}\) denotes the relation between the incident wave to the differential-mode port 1 and the reflected wave to the single ended port 2. \(S_{cc1}\) is the relation between common-mode incident waves and reflected waves to each port. The modified mixed-mode s-parameter can consider differential-mode and common-mode simultaneously. For example, if \(S_{3d1}\) and \(S_{4d1}\) are written as Eq. (5), they means that the common-mode signal as well as the differential-mode signal are transmitted to the single-ended ports 3 and 4, even though a pure differential-mode signal is applied to the differential-mode port 1 of Fig. 1(a).

\[
S_{3d1} = \frac{1}{\sqrt{2}}(S_{dd21} + S_{cd21})
S_{4d1} = \frac{1}{\sqrt{2}}(-S_{dd21} + S_{cd21})
\]

As shown in Fig. 1(a), the differential-mode ports are connected to single-ended ports 2 and 4 with a 180 degree out of phase with each other. Therefore, the phase difference appears in \(b_2/\alpha_{dmi}, b_3/\alpha_{dm2}\), \(b_2/\alpha_{dm1}\) and \(b_3/\alpha_{dm2}\). Finally, \(S_{2d1}, S_{2d2}, S_{4d1}\) and \(S_{4d2}\) of Eq. (4) should be multiplied by \(-1\) in order to remove the phase difference. For example, \(S_{4d1}\) of Eq. (5) must be changed into Eq. (6).

\[
S_{4d1} = \frac{1}{\sqrt{2}}(S_{dd21} - S_{cd21})
\]

Using the modified mixed-mode s-parameters, each trace of the differential line can be modeled separately as explained in the following section.

3. Proposed Modeling Procedure

The proposed modeling procedure is based on a 2-port s-parameter measurement using a 2-port VNA. Using the
modified mixed-mode s-parameters as explained in the previous chapter, the proposed differential line model can consider mode-conversion and frequency dependent loss. The proposed modeling procedure is shown in Fig. 2.

First, we measure open and short-circuited differential line patterns using the 2-port VNA and a microprobe in order to perform accurate calibration and minimize undesirable parasitic elements for connectors and connector mount pads. It is difficult to embody perfect open and short-circuit patterns of the PCB due to the parasitic capacitance of the fringing field at the open end point and the parasitic inductance of the via for short-circuit pattern. Therefore, the parasitic capacitance and the parasitic inductance must be de-embedded. In order to de-embed the parasitic capacitance and inductance, two test patterns of open and short-circuited differential lines are measured, as shown in Fig. 3. The two open-ended differential line patterns have different line lengths, $l_1$ and $l_2$, respectively. Moreover, the two short-circuited differential line patterns also have different line lengths, $l_1$ and $l_2$.

In the following, the proposed modeling procedure is explained step-by-step. First, four kinds of 2-port s-parameters ($\begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix}$) of open and short-circuited differential line pattern, the lengths of which are $l_1$ and $l_2$, respectively. In the same way, the labeled $S_{11}$ and $S_{22}$ represent the short-circuited differential line patterns, the lengths of which are $l_1$ and $l_2$.

Because $S_{dd11}$ and $S_{dd12}$ of the modified mixed-mode s-parameters can be represented by Eq. (7), an equation to represent the relation between the mixed-mode s-parameter and the standard s-parameter [8], each measured 2-port s-parameter is converted into differential-mode s-parameters ($S_{dd11,0,J1}$, $S_{dd11,0,J2}$, $S_{dd11,0,J1}$ and $S_{dd11,0,J2}$) and DTC mode-conversion s-parameters ($S_{cd11,0,J1}$, $S_{cd11,0,J2}$, $S_{cd11,0,J1}$ and $S_{cd11,0,J2}$) by Eq. (7).

\[
S_{dd11} = \frac{1}{2} (S_{11} - S_{21} - S_{12} + S_{22})
\]

\[
S_{dd12} = \frac{1}{2} (S_{11} - S_{21} - S_{12} + S_{22})
\]

Each mixed-mode s-parameter converted from the measured 2-port s-parameter can be used to induce the modified mixed-mode s-parameter, ($S_{11l,0,J1}$, $S_{11l,0,J2}$, $S_{11l,0,J1}$ and $S_{11l,0,J2}$) and ($S_{21l,0,J1}$, $S_{21l,0,J2}$, $S_{21l,0,J1}$ and $S_{21l,0,J2}$), by using Eq. (8) for $S_{11l}$ and $S_{21l}$, which is derived from Eqs. (3) and (4).

\[
S_{11l} = \frac{1}{\sqrt{2}} (S_{dd11} + S_{dd12})
\]

\[
S_{21l} = \frac{1}{\sqrt{2}} (S_{dd11} - S_{dd12})
\]

As a result, if $S_{11l}$ and $S_{21l}$ of Eq. (8) are inserted into Eq. (9), [9] the input impedance of each trace of the open and short-circuited differential line patterns can also be acquired, as shown in Figs. 4 and 5. ($Z_{in,T1,0,J1}$, $Z_{in,T2,0,J1}$, $Z_{in,T1,0,J2}$, $Z_{in,T2,0,J2}$), ($Z_{in,T1,0,J1}$, $Z_{in,T2,0,J1}$, $Z_{in,T1,0,J2}$, $Z_{in,T2,0,J2}$), and $Z_{in,T1}$ and $Z_{in,T2}$ are the input impedance of traces 1 and 2 when the differential-mode signal is applied to the differential line formed by traces 1 and 2.

\[
Z_{in,T1} = \frac{Z_{in,T1} + S_{11l} + Z_{0,T1} \tanh \gamma_{T1}(l_1 - l_2)}{1 - S_{11l}}
\]

\[
Z_{in,T2} = \frac{Z_{in,T2} + S_{11l} + Z_{0,T2} \tanh \gamma_{T2}(l_1 - l_2)}{1 - S_{21l}}
\]

As shown in Fig. 4, because the input impedance looking into each trace of the differential line at a distance $l_2$ with the open load is $Z_{in,T1,0,J2}$ and $Z_{in,T2,0,J2}$, the open-ended differential line with a line length of $l_1$ is treated as a differential line with a line length of $l_1 - l_2$. Then, the input impedance of the open-ended differential line with a line length of $l_2$ can be treated as load impedance.

\[
Z_{in,T1,0,J1} = \frac{Z_{in,T1,0,J1} + Z_{0,T1} \tanh \gamma_{T1}(l_1 - l_2)}{Z_{0,T1} + Z_{in,T1,0,J2} \tanh \gamma_{T2}(l_1 - l_2)}
\]

\[
Z_{in,T2,0,J1} = \frac{Z_{in,T2,0,J1} + Z_{0,T2} \tanh \gamma_{T2}(l_1 - l_2)}{Z_{0,T2} + Z_{in,T2,0,J2} \tanh \gamma_{T2}(l_1 - l_2)}
\]
From Eqs. (10) and (12), we obtained the characteristic impedance and the propagation constant of each unbalanced trace measured 2-port s-parameters, the characteristic impedance 0 and the propagation constant, \( \gamma \) for each trace of the differential line. In addition, \( Z_{0, T1} \) and \( \gamma_{T1} \) for the trace 1 of the differential line. As shown in Eqs. (14) and (15), the equivalent circuit model parameters of trace 1 and the trace 2 are obtained by using the characteristic impedance and propagation constant of each traces [9].

\[
R_{T1} = \text{Re}[\gamma_{T1}Z_{0,T1}]
\]
\[
L_{T1} = \text{Im}[\gamma_{T1}Z_{0,T1}]/2\pi f
\]
\[
G_{T1} = \text{Re}[\gamma_{T1}/Z_{0,T1}]
\]
\[
C_{T1} = \text{Im}[\gamma_{T1}/Z_{0,T1}]/2\pi f
\]

Then, we extract the equivalent circuit model parameters, resistance, conductance, inductance and capacitance per unit length of each trace in a differential line. From Eqs. (10) and (12), we obtained the characteristic impedance, \( Z_{0, T1} \) and the propagation constant, \( \gamma_{T1} \) for the trace 1 of the differential line. In addition, \( Z_{0, T2} \) and \( \gamma_{T2} \) for the trace 2 of the differential line can also be derived from Eqs. (11) and (13).

\[
Z_{m,T1,S,T1} = Z_{0,T1}Z_{m,T1,S,T1} + Z_{0,T1} \tan \gamma_{T1}(l_1 - l_2)
\]
\[
Z_{m,T2,S,T2} = Z_{0,T2}Z_{m,T2,S,T2} + Z_{0,T2} \tan \gamma_{T2}(l_1 - l_2)
\]

Then, we extract the equivalent circuit models of the differential line using the extracted equivalent circuit model parameters for each trace of the differential line. Finally, Fig. 6 presents the equivalent circuit model of the differential line considering the mode-conversion and the frequency dependent loss. As shown in Eqs. (14) and (15), the equivalent circuit model parameters of trace 1 and the trace 2 are obtained by using the characteristic impedance and propagation constant of each traces [9].

\[
Z_{m,T2,S,T1} = Z_{0,T2}Z_{m,T2,S,T1} + Z_{0,T2} \tan \gamma_{T2}(l_1 - l_2)
\]

We have created the equivalent circuit models of the differential line using the extracted equivalent circuit model parameters for each trace of the differential line. Finally, Fig. 6 presents the equivalent circuit model of the differential line considering the mode-conversion and the frequency dependent loss, which represents one section of a distributed SPICE-Type RLGC W-element. Because we have modeled the two traces of the differential line to reflect the structural imbalance of the differential line, the mode-conversion can be calculated. Moreover, because the resistance and the conductance of traces 1 and 2 are extracted by equations for the characteristic impedance and propagation constant, which are functions of frequency, we can include the frequency dependent loss effect.

The proposed model in Fig. 6 is applicable for the case of differential-mode propagation and DTC mode-conversion. In order to extract the equivalent circuit model for the case of common-mode and CTD mode-conversion, \( S_{c11} \) and \( S_{d11} \) are used instead of \( S_{dd11} \) and \( S_{cd11} \) in the second step of the proposed modeling procedure as shown in Fig. 2. \( S_{c11} \) and \( S_{d11} \) of the modified mixed-mode s-parameter are obtained from \( S_{cc11} \) and \( S_{dc11} \) by using Eq. (16).
The modeling procedure is then repeated, and the equivalent circuit model taking into account the common-mode and the CTD mode-conversion is extracted. Moreover, if the equivalent circuit model, as shown in Fig. 6 is combined with the equivalent circuit model considering common-mode and CTD mode-conversion, the complete equivalent circuit model can be extracted.

4. Extracted Model Parameters

To demonstrate the proposed modeling procedure and equivalent circuit model, we fabricated unbalanced differential lines. Figure 7 shows the selected unbalanced differential line structures for the modeling and measurements. The first unbalanced structure (Test Vehicle I) has an unbalanced trace width. $W_1$ is 250 $\mu$m and $W_2$ is 270 $\mu$m, 310 $\mu$m and 350 $\mu$m, respectively. In the second unbalanced structure (Test Vehicle II), the ground plane is located adjacent to a trace of the differential line. The gap between the signal trace and ground plane, $G$, is 100 $\mu$m, 200 $\mu$m, and 300 $\mu$m respectively. The ground plane underneath the third selected structure (Test Vehicle III) is partly removed and the distance between the center of the differential line and the edge of underneath ground, $CE$, is 0 $\mu$m, 100 $\mu$m, and 200 $\mu$m.

We extracted the equivalent circuit model parameters for three types of unbalanced differential lines based on the proposed modeling procedure, as shown in Table 1.

In the Test Vehicles I and II, the inductance of trace 2 is decreased and the capacitance ($C_{T2}$) of trace 2 is increased in the circuit model of Fig. 6 because of the increased width ($W_2$) and the decreased gap ($G$) between trace 2 and the ground located near trace 2 in Fig. 7. On the other hand, for Test Vehicle III, the inductance of trace 2 is increased and the capacitance of trace 2 is decreased because of the decreased distance ($CE$) between the center of the differential line and the edge of the underneath ground. The resistance ($R_{T1}$, $R_{T2}$) is fitted by a function proportional to the square root of the frequency and the conductance ($G_{T1}$, $G_{T2}$) is fitted by a function proportional to the frequency. The coefficients of the fitted functions are shown in table 1, where we assume that the DC loss of the traces is zero.

5. Verification of Extracted Model Parameters

We verified the extracted model parameters using a lattice diagram analysis and TDR and TDT measurements. In particular, we measured the common-mode reflected and transmitted waveforms when a differential-mode voltage step pulse was applied to the unbalanced differential lines to substantiate DTC mode-conversion. When the differential-mode signal was applied to the differential-mode port 1 in Fig. 1, the common-mode TDR waveform was measured as the sum of the reflected waves to single ended ports 1 and 2. Moreover, the common-mode TDT waveform was measured as the sum of the transmitted waves to single ended ports 3 and 4. The differential-mode voltage step pulse had a 35 ps rising time and a 480 mV differential-mode input voltage between trace 1 and trace 2. Before measuring the common-mode TDR and TDT converted from the differential-mode signal, we predicted the waveforms using a lattice diagram. The lattice diagram is a technique used to solve multiple reflections on a transmission line in a visual display. In practice, it is hard to predict the mode-conversion effect because the two traces in the unbalanced differential line have different characteristic impedance and different flight time. However, it was found that the common-mode TDR and TDT waveforms converted from the differential-mode signal can be estimated by the lattice diagram.

To predict the experiment results, we selected the unbalanced differential line shown in Fig. 8(a), where the characteristic impedance of each trace is $Z_0_{T1}$ and $Z_0_{T2}$, and the flight time from point A to point B of each trace is $t_{d,T1}$ and $t_{d,T2}$.

\[
S_{1c1} = \frac{1}{\sqrt{2}}(S_{cc1} + S_{dc1}) \quad S_{2c1} = \frac{1}{\sqrt{2}}(S_{cc1} - S_{dc1})
\]

\[
W_1 = 250 \mu m \quad W_2 = 270 \mu m, 310 \mu m \quad 350 \mu m
\]

\[
G = 100 \mu m, 200 \mu m, 300 \mu m
\]

\[
CE = 0 \mu m, 100 \mu m, 200 \mu m
\]
We assumed that $t_{d,T2}$ is smaller than $t_{d,T1}$ since the two traces of the unbalanced differential line have different flight times. The difference between $t_{d,T2}$ and $t_{d,T1}$ is $\Delta t_d$, as shown in Eq. (17).

$$\Delta t_d = t_{d,T2} - t_{d,T1} \quad (17)$$

The reflection coefficients at load are $\Gamma_{L,T1}$ and $\Gamma_{L,T2}$ and those at source are $\Gamma_{S,T1}$ and $\Gamma_{S,T2}$. Each reflection coefficient was calculated using the equations shown in the box in Fig. 8(b). When the source voltage changes from $0V$ to $+V$ and $-V$, the initial voltages on each trace, $V_{i,T1}$ and $V_{i,T2}$, are determined by the voltage dividers $V_{i,T1} = -VZ_{0,T1}/(Z_s + Z_{0,T1})$ and $V_{i,T2} = VZ_{0,T2}/(Z_s + Z_{0,T2})$. At time $t_{d,T1}$, the incident voltage $V_{i,T1}$ arrives at a load $Z_L$. At this time a reflected component is generated with a magnitude of $V_{i,T1}\Gamma_{L,T1}$, which is added to the incident voltage $V_{i,T1}$, creating a total voltage at a load of $V_{i,T1} + V_{i,T1}\Gamma_{L,T1}$. The reflected portion of the wave, $V_{r,T1}\Gamma_{L,T1}$ then travels back to the source. After time $\Delta t_d = t_{d,T2} - t_{d,T1}$, the incident voltage $V_{i,T2}$ arrives at the load $Z_L$. Similarly, a reflected component is generated with a magnitude of $V_{i,T2}\Gamma_{L,T2}$, this portion travels back to the source. At time $2t_{d,T1}$, the voltage $V_{i,T1}\Gamma_{L,T1}$ arrives at the source $Z_S$ and the voltage $V_{i,T2}\Gamma_{L,T2}$ also arrives at the source $Z_S$ after time $2\Delta t_d$.

Using the lattice diagram of Fig. 8(b), the common-mode TDR and TDT waveforms converted from the differential-mode signal are calculated as shown in Figs. 9 and 10. TDR waveforms at point A of Fig. 8 of traces 1 and 2 are shown in Fig. 9(a) and the common-mode TDR waveform is obtained by the sum of two TDR waveforms of traces 1 and 2 in Fig. 9(b). The common-mode TDR waveform has an initial voltage level of $a$, because the incident voltages ($V_{i,T1}$, $V_{i,T2}$) of traces 1 and 2 are different. Moreover, the voltage level of $b$ is generated for $2\Delta t_d$, because the difference of arrival times to point A of traces 1 and 2 in Fig. 8(a) is $2\Delta t_d$.

The common-mode TDT waveform is also acquired by the sum of two TDT waveforms of traces 1 and 2 in Fig. 10(a). As shown in Fig. 10(b), the common-mode TDT waveform has a voltage level of $a'$ for $\Delta t_d$ because the arrival times at point B of traces 1 and 2 have a difference of...
Fig. 11 Measured and simulated common-mode voltage caused by DTC mode-conversion for Test Vehicle I. (a) Simulated common-mode TDR waveform using the proposed model. (b) Measured common-mode TDR waveform. (c) Simulated common-mode TDT waveform using the proposed model. (d) Measured common-mode TDT waveform.

Fig. 12 Measured and simulated common-mode voltages caused by DTC mode-conversion for Test Vehicle II. (a) Simulated common-mode TDR waveform using the proposed model. (b) Measured common-mode TDR waveform. (c) Simulated common-mode TDT waveform using the proposed model. (d) Measured common-mode TDT waveform.

$\Delta t_d$.

Finally, we measured the common-mode TDR and TDT waveforms when the differential-mode signal is applied to unbalanced differential lines (Test Vehicle I, II and III) in order to verify the proposed model and the lattice diagram analysis in Fig. 9 and Fig. 10. Figure 11 shows the common-mode TDR and TDT waveforms of the unbalanced differential line that has different trace widths in Test Vehicle I. We obtained test results similar to the expected common-mode TDR and TDT waveforms from the lattice diagram. As the width of trace 2 increases, the imbalance of the differential line becomes higher and the amplitude of the common-mode TDR and TDT waveform also increases. In addition, common-mode TDR and TDT waveforms have a positive voltage because the characteristic impedance of trace 1 is larger than that of trace 2 and the velocity of trace 1 is greater than that of trace 2.

Figures 11(a) and (c) are the common-mode TDR and TDT waveforms using the proposed model and Figs. 11(b) and (d) are the measured common-mode TDR and TDT waveforms. The equivalent circuit model used for the simulation consists of multi-section of a SPICE-Type RLGC model extracted at Sect. 4. This means that the mode-conversion is generated uniformly along the differential line. However, it is not uniform on the test vehicles used for measurement because of the process variation. Although the common-mode TDR and TDT waveforms are not exactly identical, the maximum amplitude of the mode-conversion TDR and TDT waveforms simulated by the proposed model agrees well with the measured results. The maximum amplitude of common-mode TDR is 23.9 mV and that of common-mode TDT is 75.4 mV, when the width of trace 2 is 350 µm. This means that 4.98% and 15.71% of the 480 mV differential-mode input voltage is converted to the common-mode voltage. From these results, we can confirm that the common-mode voltage level suddenly increases because of the mode-conversion effect when the differential-mode signal arrives at the receiver. If it exceeds the common-mode sensing range of the receiver, the voltage margin decreases and the original signal can not be recovered. The proposed model predicts these effects and agrees well with the measurement result.

Figure 12 shows the common-mode TDR and TDT waveforms for the unbalanced differential line of Test Vehicle II, which has a ground adjacent the trace 2. As the gap between the ground and trace 2 ($G$ of Fig. 7) narrows, the characteristic impedance of trace 1 is increased and that of trace 2 is decreased because of the capacitance that exists in the gap between trace 2 and the ground. Moreover, the velocity of trace 1 is greater than that of trace 2. From these results, the amplitude of the common-mode TDR and TDT waveform increases and the voltage of the common-mode TDR and TDT waveforms is positive. 3.19% and 16.44% is changed into the common-mode maximally for Test Vehicle II, when the gap between trace 2 and the ground is 100 µm.

We also measured the common-mode TDR and TDT for Test Vehicle III of which the underneath ground plane is partly removed, as shown in Fig. 13. In this case, because a part of ground of trace 2 disappears, the increase of inductance of trace 2 results in an increase of the characteristic impedance and a reduction of velocity. From these results, we obtained negative voltage TDR and TDT waveforms for Test Vehicle III, which are the inverse of the results of Test Vehicles I and II, because the characteristic impedance of trace 1 is larger than that of trace 2, and the flight time of trace 1 is greater than that of trace 2. In case of Test Vehicle III, the characteristic impedance of trace 1 is smaller than that of trace 2, and flight time of trace 1 is lower than that of...
trace 2.0 \( \mu \text{m} \) CE, \(-67.4\) mV and \(-110.5\) mV common-mode voltages are reflected and transmitted maximally; these are 14.04% and 23.02% of the input differential-mode voltage. We know that Test Vehicle III generates the largest mode-conversion.

From these results, it is verified that the amplitude of the transmitted and reflected differential- to common-mode conversion voltage is increased as the imbalance of the differential line increases. The proposed model predicts these effects and agrees well with the measurement results. Moreover, the proposed model successfully estimates the mode-conversion effects.

6. Conclusion

Using the proposed modeling method, we have successfully modeled, measured, and studied three types of unbalanced differential line structures representing typical cases structures in practical PCB designs. The unbalanced structures for this study include trace width mismatch and ground proximity mismatch. From the measurements and the analysis, it is verified that the line imbalance and the loss are well reflected by the proposed model. The proposed model and the model parameters were successfully verified by a lattice diagram analysis, and by DTC conversion TDR and TDT measurements.

The proposed model can be effectively used to estimate the impact on signal integrity by the line imbalance resulting from unwanted process uniformity variation or unavoidable design circumstances. In addition, it can also be used to minimize the risk at an initial PCB design. Furthermore, the model is helpful to understand and solve the source of signal integrity problems and to provide proper design solutions.

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References


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