Efficient and Accurate Signal Transient and Crosstalk Noise Determination of Frequency-Variant RLC Transmission Lines

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Abstract—A new signal transient and crosstalk noise determination method of frequency-variant RLC transmission lines is presented. In order to investigate the realistic transmission line characteristics, two coupled lines are experimentally characterized by using s-parameters up to 20GHz. Then the signal transients and crosstalk noises are directly determined with the measured s-parameters and compared with conventional SPICE simulation results. Thereby, it is shown that the conventional approach may lead to significant design error. In contrast, the proposed method is very accurate as well as efficient.

I. INTRODUCTION

As the clock frequency and the level of the integration of integrated circuits drastically increases, the interconnect lines play a pivotal role in the determination of circuit performance [1]-[3]. Moreover, in such high-performance circuits, the interconnect lines have to be treated as transmission lines in which the circuit model parameters are frequency-variant due to skin effect and proximity effect [4].

Note, since the frequency-variant RLC transmission line responses cannot be represented with the close forms of expressions, conventional quasi-static circuit-model-parameter-based circuit simulation techniques may have inherent limitations in their accuracy and efficiency. That is, conventional signal integrity verification using SPICE-like EDA tools for such frequency-variant RLC transmission lines may neither be cost-efficient nor accurate. Therefore, for the today’s integrated system designs, not only are experimental characterizations crucial but also efficient simulation method is highly required.

In this paper, a new signal integrity verification method for the frequency-variant RLC transmission lines is presented. Then it will be shown that the conventional EDA-tool-based simulation techniques may lead to significant design errors.

II. EXPERIMENTAL CHARACTERIZATION

In order to experimentally characterize transmission lines, test patterns as shown in Fig. 1-(a) are designed and fabricated by using a BGA package process. The cross-section of the lines is also shown in Fig. 1-(b). Since the transmission line characteristics are frequency-dependent, the 4-port s-parameters are determined up to 20GHz by using 2-port vector network analyzer (Agilent 8510C VNA) connected with two Cascade Microtech GSSG probe tips (150um pitch). The calibrations are performed by using standard SOLT (short-open-load-thru) calibration method and the s-parameters are based on 50 ohm reference impedance. Since the two coupled lines are symmetrical, the measured 4-port s-parameters can be represented by a 4 by 4 matrix as shown below.

\[
[S] = \begin{bmatrix}
S_{11} & S_{12} & S_{13} & S_{14} \\
S_{21} & S_{22} & S_{23} & S_{24} \\
S_{31} & S_{32} & S_{33} & S_{34} \\
S_{41} & S_{42} & S_{43} & S_{44}
\end{bmatrix} = \begin{bmatrix}
S_{11} & S_{12} & S_{13} & S_{14} \\
S_{21} & S_{22} & S_{23} & S_{24} \\
S_{31} & S_{32} & S_{33} & S_{34} \\
S_{41} & S_{42} & S_{43} & S_{44}
\end{bmatrix}
\] (1)

Since the two symmetrical lines have the even mode and odd mode, the modal s-parameters can be represented by [5]

\[
[S]_{even} = \begin{bmatrix} S_{11} + S_{12} & S_{13} + S_{14} \\
S_{31} + S_{32} & S_{33} + S_{34}
\end{bmatrix},
\] (2)

\[
[S]_{odd} = \begin{bmatrix} S_{11} - S_{12} & S_{13} - S_{14} \\
S_{31} - S_{32} & S_{33} - S_{34}
\end{bmatrix},
\] (3)

respectively. Since the coupled lines are decoupled into the two eigen mode s-parameters, the propagation constants (\(\gamma\)) and characteristic impedances (Z) of the coupled lines can be readily determined.
as in a single isolated line. The modal transmission line parameters of the single isolated line can then be directly determined from the measured s-parameters as shown below \[6\]

\[
e^{-j\gamma} = \frac{1 - S_{11} + S_{22}^2 \pm \sqrt{(S_{11} - S_{22}^2 + 1)^2 - (2S_{12})^2}}{2S_{21}}, \tag{4}
\]

\[
Z^2 = Z_0^2 \frac{(1 + S_{11})^2 - S_{22}^2}{(1 - S_{11})^2 - S_{22}^2}, \tag{5}
\]

where the measurement reference impedance \(Z_0\) is 50 Ohm. During the extraction of both complex parameters (i.e., \(\gamma(\omega)\) and \(Z(\omega)\)), the cyclically mapped phase outputs of the s-parameter have to be converted to the true radian measurement phase, which can be any real value. Then the transmission line parameters are determined. As shown in Fig. 2, the transmission line parameters are frequency-dependent due to the skin effect and proximity effect. That means the signal transient responses of the interconnect lines cannot be accurately determined without considering the frequency-dependent effects.

In addition, for the cross-section data of Fig. 1-(b), the transmission line parameters are determined by using commercial field-solver [7]. In Table I, the transmission line parameters that are determined at 100MHz are compared. It is noteworthy that the field-solver-based transmission line parameters show a significant difference from experimental data. Note that the field solver determines electro-magnetic flux intensity by using FEM method. Then the circuit model parameters are determined by exploiting two-terminal circuit parameter determination method. However, in reality, it may not be accurately applied to strongly coupled lines. Therefore, without experimental characterization, blind field-solver-based signal coupling (cross-talk noise) estimation may result in significant design error.

### TABLE II

**Comparison of the Calculated Transmission Line Parameters with Measurement Values at 100MHz**

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Calculated with Fig.1-(b)</th>
<th>Measured @ 100MHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>R [(\Omega/cm]]</td>
<td></td>
<td></td>
</tr>
<tr>
<td>R_{11}</td>
<td>0.44189</td>
<td>0.52169</td>
</tr>
<tr>
<td>R_{12}</td>
<td>0.05098</td>
<td>0.07747</td>
</tr>
<tr>
<td>L [nH/cm]</td>
<td></td>
<td></td>
</tr>
<tr>
<td>L_{11}</td>
<td>4.40909</td>
<td>4.43834</td>
</tr>
<tr>
<td>L_{12}</td>
<td>1.33780</td>
<td>0.69695</td>
</tr>
<tr>
<td>C [pF/cm]</td>
<td></td>
<td></td>
</tr>
<tr>
<td>C_{11}</td>
<td>0.81203</td>
<td>1.00910</td>
</tr>
<tr>
<td>C_{12}</td>
<td>0.36521</td>
<td>0.13602</td>
</tr>
<tr>
<td>C_{11+C_{12}}</td>
<td>1.17724</td>
<td>1.14512</td>
</tr>
</tbody>
</table>

III. **Efficient Simulation for RLC Lines**

Although simple RC model may permit the closed form of the models [8], it is not accurate as shown in Fig. 3. In contrast, the signal transient responses using RLC transmission line circuit models become complicated non-monotonic wave-shapes. Therefore, it may not be possible to get a closed form of an expression. Moreover, the SPICE simulation using the constant transmission line parameters may also be inaccurate as shown in Fig. 3.

In this work, signal transient responses of RLC transmission lines are determined by using TWA (traveling-wave-based waveform approximation) technique [9] which can be readily determine the signal transients of RLC transmission lines. Further, since the transmission lines are frequency-variant, the effects have to be taken into account. In reality, since the signal transient responses including the frequency-variant effects become more damped than those without the effects, more suitable transmission line parameters have to be determined. In this work, the transmission line parameters are tested by using the experimentally extracted transmission line parameters. Then suitable transmission line parameters at a specified high-frequency are selected for the signal transient determination of the frequency-variant transmission line.
In TWA, defining the effective time of flight \((t_{f0})\) and a time difference \((\delta)\) concerned with the wave reflection due to capacitive load \((C_l)\) below,

\[
 t_{f0} = \sqrt{(L \cdot L)(L \cdot C + C_l)},
\]

\[
 \delta = \sqrt{(L \cdot L)(L \cdot C + C_l)} - t \sqrt{L C},
\]

where \(l\) is the line length, the time domain wave-shapes are approximately determined as follows.

For \((2n-1)\tau_{f0} - \delta \leq t \leq (2n-1)\tau_{f0} + \delta\), the response can be modeled as a linear ramp shape,

\[
v_{r}(t) = \sum_{n=1}^{\infty} \left[ f_{n}(t) + g_{n}(t) \right] \cdot h_{n}(t),
\]

where

\[
f_{n}(t) = V_{j,\text{pole}} \left( (2n+1)\tau_{f0} - \delta \right),
\]

\[
g_{n}(t) = V_{j,\text{pole}} \left( 2(n-1)\tau_{f0} - \delta \right),
\]

\[
h_{n}(t) = u(t - 2(n-1)\tau_{f0} - \delta) - u(t - 2(n-1)\tau_{f0} + \delta).
\]

Note, \(n\) is the reflection count and \(C_l\) is the load capacitance of the line. The \(V_{j,\text{pole}}\) is the time-domain counter part of the frequency domain response that is determined by using the first three dominant poles. In contrast, for \((2n-1)\tau_{f0} + \delta \leq t \leq (2n+1)\tau_{f0} + \delta\), the response can be modeled as an RC-response-like shape

\[
v_{r}(t) = \sum_{n=1}^{\infty} \left[ f_{n}(t) + g_{n}(t) \right] \cdot h_{n}(t),
\]

where

\[
f_{n}(t) = 2V_{j,\text{pole}} \left( (2n-1)\tau_{f0} - \delta \right) - V_{j,\text{pole}} \left( 2(n+1)\tau_{f0} - \delta \right),
\]

\[
g_{n}(t) = u(t - 2(n-1)\tau_{f0} - \delta) - u(t - 2(n-1)\tau_{f0} + \delta),
\]

\[
o_{n} = V_{j,\text{pole}} \left( 2(n+1)\tau_{f0} - \delta \right) - 2V_{j,\text{pole}} \left( (n-1)\tau_{f0} \right),
\]

\[
\tau = (l \cdot R)(L \cdot C + C_l).
\]

Although the TWA technique is formulated with a step input, it can be modified for ramp input models which can be considered more general. A ramp input signal can be formulated by using delayed step signals as below

\[
\frac{V_{r}}{t_{f}} \left\{ h(t) - h(t-t_{r}) \right\} \approx \lim_{N \to \infty} \frac{V_{r}}{N} \sum_{k=1}^{N} \left( \frac{k t_{r}}{N} - \frac{t_{r}}{2} \right),
\]

\[
\frac{\sum_{n=1}^{\infty} f_{n}(t)}{t_{f}} = \sum_{n=1}^{\infty} \left[ \frac{f_{n}(t)}{t_{f}} - \frac{f_{n}(t-t_{r})}{t_{f}} \right] = \lim_{N \to \infty} \frac{V_{r}}{N} \sum_{k=1}^{N} \left( \frac{k t_{r}}{N} \right),
\]

where \(t_{r}\) is a rising time and \(N\) is the number of delayed step functions. In practice, \(N=5\) is accurate enough.

Note (8) and (9) can be applied to a single isolated line. However, since the two coupled lines can be decoupled into even mode \((\uparrow \uparrow)\) and odd mode \((\uparrow \downarrow)\) by using two fundamental switching modes, they can be directly exploited for the modal signals. Then the signal transient and crosstalk noise can be determined by combining the modal signal responses. For example, when the first line is switching from logic 0 to logic 1 and the second line remains in a quiet state, the signal transient at the first line can be determined,

\[
\begin{aligned}
\left( \uparrow 0 \right)_{\text{even}} &= \frac{\left( \uparrow \uparrow \right)_{\text{even}} + \left( \uparrow \downarrow \right)_{\text{even}}}{2} \\
\left( \uparrow 0 \right)_{\text{odd}} &= \frac{\left( \uparrow \uparrow \right)_{\text{odd}} - \left( \uparrow \downarrow \right)_{\text{odd}}}{2}.
\end{aligned}
\]

In contrast, the crosstalk can be determined at the second line for the same switching condition,

\[
\left( \uparrow 0 \right)_{\text{even}} = \frac{\left( \uparrow \uparrow \right)_{\text{even}} - \left( \uparrow \downarrow \right)_{\text{even}}}{2}.
\]

Note that the line of interest for the switching mode is marked with a square box.

Fig. 3. Signal transient and crosstalk noise of two coupled interconnect lines for \(\uparrow 0\) switching. \(t_{r}=50\) ps, \(l=1.7\) cm, \(R_{s}=50\), \(C_{L}=0.1\) pF.

IV. VERIFICATION OF THE PROPOSED TECHNIQUE

In order to verify the accuracy of the proposed technique, we compared three cases: (1) conventional SPICE simulation that employs the field-solver-based transmission line parameters, (2) TWA-based waveform determination that employs the transmission line parameters of a specified frequency (at 6GHz transmission line parameters), and (3) S-parameter-based direct signal transient determination [10]. As shown in Fig. 4, signal transients and crosstalk noise using the TWA-based compact models have excellent agreement with the s-parameter-based time-domain signal transient responses. In contrast, conventional SPICE simulation that employs the field-solver-based transmission line parameters shows significant deviation from the s-parameter-based time-domain signal transient responses. Further, since the TWA-based compact models are analytical expressions, they are much more efficient than that of SPICE simulation.
V. CONCLUSION

In this research, the signal transient and crosstalk noise determination of frequency-variant RLC transmission lines using compact CAD models was proposed. It was shown that the proposed technique is very accurate as well as efficient. The signal transient wave-shapes using the compact CAD models have excellent agreement with the s-parameter-based time-domain signal transient responses. In contrast, conventional SPICE simulation that employs the field-solver-based transmission line parameters may neither be accurate nor efficient.

REFERENCES