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Anticipating EMI Using Transfer Functions and Signal Integrity Information

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Abstract

Discontinuities in a circuit can lead to signal integrity as well as EMI problems. A method, which efficiently combines full-wave tools and circuit simulators, is proposed herein to analyze the coupling at discontinuities. The proposed method may be applied to practical engineering designs.

Keywords

EMI, SI, radiation, circuit simulator, transfer function, full-wave.

INTRODUCTION

Simulations of EMC problems usually resort to full-wave tools due to the complexity of EMC phenomenon. While full-wave simulations can be accurate, they are time-consuming as well, which makes them difficult to apply to practical engineering designs. Circuit simulators, such as SPICE, are relatively faster and they can predict signal quantities fairly accurately.

The method developed in this work views the signal quantities, such as currents or voltages in a circuit, as the driving sources of an unintentional radiator and relates the signal quantities to the radiation levels with transfer functions. By integrating the transfer functions into the circuit simulators, the EMI levels can be predicted.

INTEGRATING EMI INTO CIRCUIT SIMULATORS

Some EMI problems can be generalized as shown in Figure 1. A signal is launched at a driver, propagating through a trace on a PCB. If the trace is long compared to the wavelength, it behaves as a transmission line. Often, the propagating signal encounters some discontinuities. The discontinuities can be gaps in the reference plane, via transitions, connectors etc. At the discontinuities, parasitic inductances and capacitances are introduced into the signal line, resulting in signal distortion. Meanwhile, the signal return current can drive the discontinuities, resulting in EMI problems. Figure 1 illustrates that signal integrity (SI) issues are closely related to EMI issues. Often, it is impossible to simulate the entire structure with full-wave methods because large difference in scale is involved in modeling the small features of the traces and the large features of the unintentional radiator.

However, if the interfaces between the circuit and the unintentional antenna can be identified at a pair of terminals (i.e., a port), the "EMI antenna" can be integrated into the circuit as a network or as an equivalent circuit. The network parameters of the "antenna" can be obtained from either full-wave simulations or measurements. Then, only circuit simulators are needed to estimate the radiation and the susceptibility of the circuit provided the "antenna" remains unchanged.

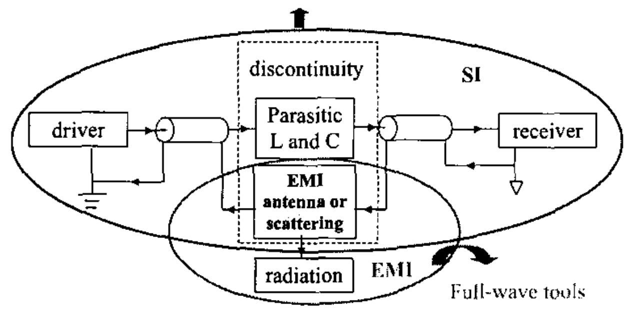


Figure 1. A schematic representation for incorporating EMI physics in circuit simulation.

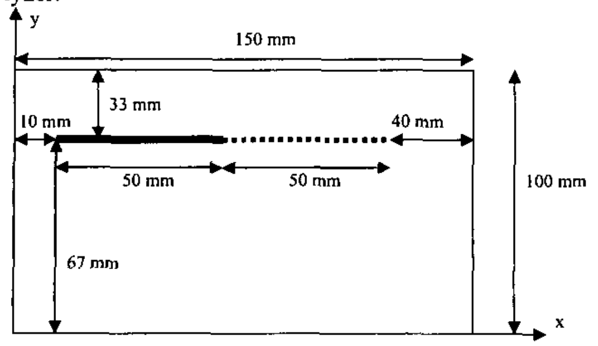
With this method, full-wave simulations are applied only to the "antenna" portion, instead of the entire circuit, which avoids the mixed-scale problems. Moreover, this method relates the EMI to the EMI sources through transfer functions. As long as the transfer functions remain unchanged, changes in the circuits can be analyzed with a circuit simulator, like SPICE, rather than requiring a new full-wave simulation.

APPLICATION EXAMPLES

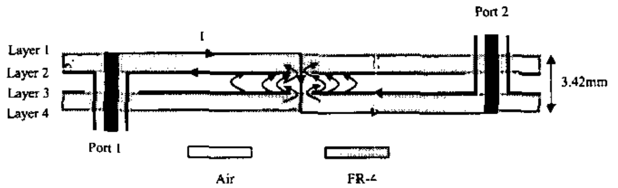
To demonstrate the method proposed in the previous section, two examples are considered below.

A. A via transition problem

The first example is a layer transition of a single-ended signal in a 4-layer board. The top and side views of the board geometry are shown in Figure 2. A 4-layer board consisting of two double-sided FR-4 boards was constructed by sandwiching the two pieces together. One double-sided board was stacked on the other with air filling in the space between them, as shown in Figure 2. Layer 2 and Layer 3 were metal layers, corresponding to a power-bus structure. A 10 mils wide trace was routed on Layer 1 for 5 cm, and went through a via to Layer 4 for another 50 cm. The characteristic impedance of the trace was approximately 120 Ω. Each end of the trace was connected to the center conductor of an SMA jack. The outer conductor of the SMA jacks was connected to the signal reference layer, Layer 2 or Layer 3. Two-port S-parameters of the single-ended trace were measured with a network analyzer.



(a) Top view.



(b) Side view.

Figure 2. Geometry of a four-layer board with via transitions.

The return current around the via is through displacement current. The displacement current will propagate through the power-bus structure, resulting in power-bus noise. Meanwhile, the input impedance looking into the power-bus at the via is present in the current return path of the circuit, distorting the propagating signal. Therefore, the unintentional antenna, namely the power-bus structure, needs to be included in the equivalent circuit to estimate the circuit and EMI quantities.

An equivalent circuit can be developed, as shown in Figure 3. The single-ended trace is modeled as two transmission lines W_1 and W_2, with W_1 corresponding to the part in Layer 1 and W_2 corresponding to the part in Layer 4. The per-unit-length inductance and capacitance of the transmission line were calculated using standard cross-section analysis. The through hole via is simply modeled as a short between W_1 and W_2. A more precise model should include the parasitic inductances and capacitances associated with the via. However, the parasitic inductances and capacitances were neglected since the focus of this work is to demonstrate the method of integrating the unintentional EMI antenna into the circuit. The equivalent circuit of the rectangular power-bus structure is shown in Figure 3 with $L = d/(ab\omega_{mn}\varepsilon)$, $C_0 = (ab\varepsilon)/d$, and $G_{mn} = C_0\omega_{mn}(\tan\delta + r/d)$, with the dielectric loss tangent $\tan\delta$ and the skin depth $r = \sqrt{2/(\omega\mu\sigma)}$, and $\omega_{mn} = \sqrt{(m\pi/a)^2 + (n\pi/b)^2}\Big/\sqrt{\varepsilon\mu}$ (mode indices m and n). The length and width of the power-bus structure are denoted as a and b, respectively, and d is the separation between the Layer 2 and Layer 3. The coefficients $N_{mni} = c_m c_n \cos(\frac{m\pi x_i}{a})\cos(\frac{n\pi y_i}{b})\operatorname{sinc}(\frac{m\pi W_{xi}}{2a})\operatorname{sinc}(\frac{n\pi W_{yi}}{2b})$ denote the ratio of ideal transformers, considering the location of the port (x_i, y_i) and the port widths W_{xi} and W_{yi} in the x and y directions, respectively. The constant $c_m = 0$ if $m = 0$, and $c_m = \sqrt{2}$ if $m \neq 0$, and $c_n = 0$ if $n = 0$, and $c_n = \sqrt{2}$ if $n \neq 0$ [2, 3, 4]. Port 3 in the equivalent circuit is an arbitrarily located port at an observation point on the power-bus and provides the open-circuit noise voltage.

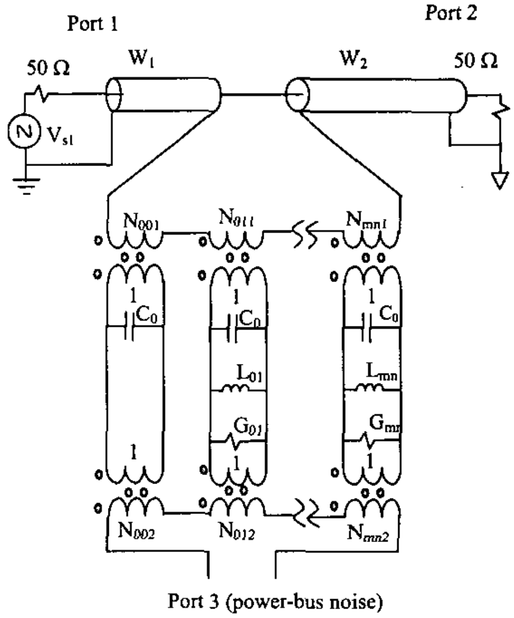


Figure 3. The equivalent circuit of the via transition in a 4-layer board.

The $|S_{11}|$ and $|S_{21}|$ of the equivalent circuit are shown in Figure 4. The large nulls in the $|S_{11}|$ are due to the resonance of the transmission line since the trace was not matched at the load and source ends. The small resonances are the power-bus resonances. The magnitude of the results from the proposed method agrees with the measurements within 3 dB over the frequency range from 50 MHz to 5 GHz. However, there are some shifts of the resonant frequencies. The shifts of the resonant frequencies probably is because the parasitic inductances and capacitances associated with the via and associated with the SMA jacks were not included in the equivalent circuit.

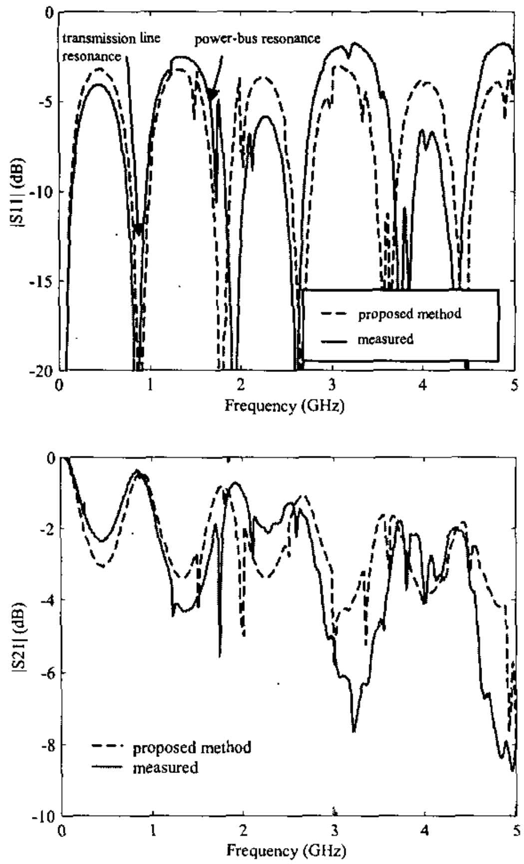
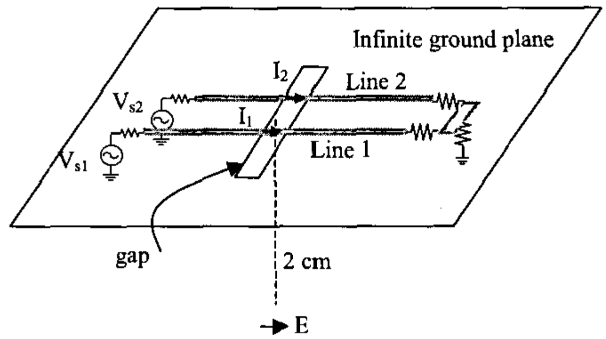


Figure 4. Comparisons of the results from the proposed method and from the measurements.

B. A differential line crossing a gap

The second example is to calculate the radiation from a differential line crossing a gap in an infinite ground plane. The geometry is shown in Figure 5. The unintentional radiator is the gap, and the sources of the radiator can be the currents on the two traces, I_1 and I_2. The radiation is represented by the E-field 2 cm away from the gap as shown. A better choice to define the sources is to define I_1 and I_2 as the line currents existing directly above the gap because in this way, the transfer functions from I_1 and I_2 to E will not change, even if other parts of the circuit change. Strictly speaking, the line currents driving the gap are distributed. However, if the gap width is small enough compared to the smallest wavelength, I_1 and I_2 can be approximated as two lumped sources.



trace width : 75 mils
trace spacing from edge to edge: 75 mils
trace length: 10 cm
spacing between trace and ground plane: 45 mils
gap width : 0.5 cm
gap length: 2.8 cm

Figure 5. Geometry of a differential line crossing a gap.

Since there are two antenna sources I_1 and I_2, two transfer functions are needed. Let

$$I_{cm} = I_1 + I_2\,, \tag{1}$$

and

$$I_{dm} = I_1 - I_2, \tag{2}$$

where I_{cm} and I_{dm} are the common-mode and differential-mode currents, respectively. Then, the radiation can be characterized in terms of a common-mode transfer function H_{cm} and a differential mode transfer function H_{dm}, with

$$H_{cm} = E/I_{cm}, \text{ when } I_{dm} = 0, \tag{3}$$

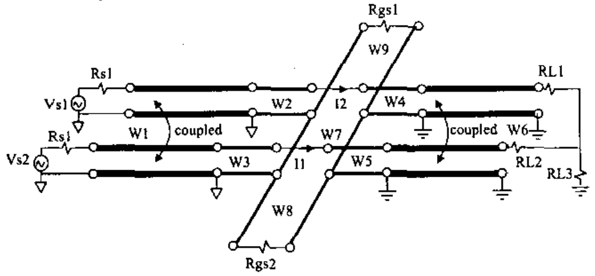
and

$$H_{dm} = E/I_{dm}, \text{ when } I_{cm} = 0. \tag{4}$$

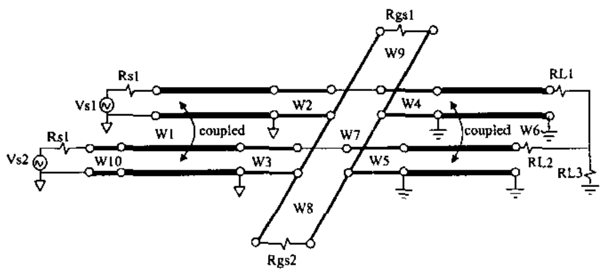
In order to obtain H_{cm} and H_{dm}, two FDTD simulations, one with $V_{s1} = V_{s2}$ and the other with $V_{s1} = -V_{s2}$, have been performed. Since the layout is balanced, $I_1 = I_2$ for the case of $V_{s1} = V_{s2}$ (corresponding to common-mode excitation), and $I_1 = -I_2$ for the case of $V_{s1} = -V_{s2}$ (corresponding to differential excitation). After the E field 2 cm away from the gap and I_1 and I_2 have been obtained from the FDTD simulations, the transfer functions H_{cm} and H_{dm} can be calculated using (3) and (4).

An equivalent circuit can be developed for the above problem, as shown in Figure 6 (a). The gap is modeled as a slotline and it consists of transmission lines W_7, W_8 and W_9. The resistors R_{gs1} and R_{gs2} are introduced to account for the radiation effects at the shorted ends of the slotline [1]. Transmission lines W_1 and W_6 are coupled transmission lines, corresponding to the differential line. The per-unit-length parameters of the transmission lines were calculated using cross-sectional analysis. Because the differential line is loosely coupled, there will be current on Line 1 and Line 2 as well as the ground plane for both common-mode and differential mode excitations. To force this physics in the circuit model, the parts of Line 1 and Line 2, within 0.5 cm away from the gap, are modeled as single-ended transmission lines W_2, W_3, W_4 and W_5. In this way, the currents in the equivalent circuit will return through the slotline for both common-mode and differential excitations.

Consequently, the slotline in the equivalent circuit can be excited by both modes.



(a) Balanced differential line crossing a gap.



(b) Unequal length differential line crossing a gap.

Figure 6. Equivalent circuit for the differential line crossing a gap.

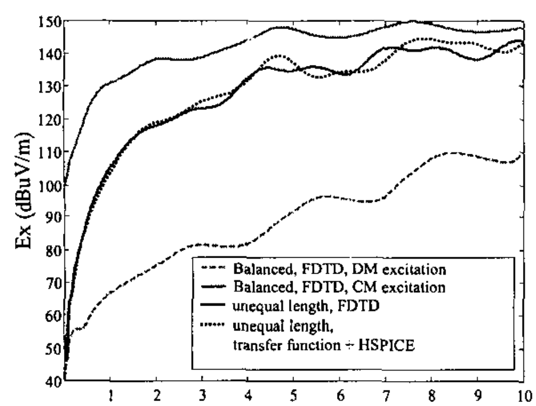


Figure 7. Comparison of the results from the proposed method to that from FDTD for the case of a differential line crossing a gap.

Imbalances can be introduced, for example, letting Line 1 be 5 mm longer than Line 2 at the feed ends. The impact on the radiation of this imbalance can be predicted using H_{dm} and H_{cm} without resorting to the FDTD simulations. The equivalent circuit for the case of unequal length is shown in Figure 6 (b). Transmission line W_{10} is introduced to account for the extra length. Inputing this circuit into HSPICE, I_1 and I_2 can be obtained. The E field can then be predicted using

$$E_{predict\_ul} = H_{cm} \times I_{cm\_ul} + H_{dm} \times I_{dm\_ul}, \qquad (5)$$

where the subscript "ul" represents the unequal length case. The results from the proposed method (the dark dotted curve) and that from the FDTD method (the dark solid curve) are shown in Figure 7. The agreement is within 5 dB over the frequency range from 100 MHz to 10 GHz. The discrepancy probably is because the modeling of the shorted termination of the gap with R_{gs1} and R_{gs2} is inadequate.

CONCLUSIONS

A method, which integrates the characteristics of an unintentional EMI "antenna" into the associated circuit, is proposed. The EMI "antenna" is characterized as a network with ports interfacing to the circuit, and the network parameters can be obtained from either measurements or full-wave simulations. Then, the "antenna" network is incorporated into the circuit, and only circuit simulations are needed to predict the EMI levels, and the susceptibility of the circuit. The proposed method is efficient compared to pure full-wave methods, and is applicable to practical engineering designs.

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