Abstract—The frequency-variant characteristics of a pair of package power and ground net are experimentally investigated and modeled by using frequency-variant grid-type equivalent circuits, where each cell of the grid is modeled with vertically stacked RLC-ladder circuits. Various test patterns are designed and fabricated by employing a ball grid array (BGA) package process. Then the test patterns are characterized by using impedance analyzer, time domain reflectometry and time domain transmission (TDR/TDT), and vector network analyzer (VNA). Based on the experimental characterizations, not only are material constants and circuit model parameters determined, but also the conductor roughness and other frequency-variant effects are modeled. It is shown that the SPICE-based circuit simulation using the proposed circuit model has an excellent agreement with the measurement data in the range of 100 MHz–15 GHz. Then, it is also shown that the various types of package performance evaluations can be very efficiently as well as accurately achieved with the proposed technique.

Index Terms—Ball grid array (BGA) package, frequency variant circuit model, power distribution net, S-parameters, skin effect, transmission line.

I. INTRODUCTION

As the operation frequencies for integrated circuits (ICs) and level of integration drastically increase, the power and ground reference potential fluctuations due to the simultaneous switching of many input/output (I/O) drivers and/or core logic gates can have a significant effect on the circuit performance. Since the simultaneous switching noise (SSN) during circuit operation can result in glitches, timing error, jitter, or even logic failure, the increase of the SSN implies significant reduction of timing or noise margin for a given circuit design. Unlike other electromagnetic effects such as crosstalk and timing delay, the SSN can be considered to be the most critical circuit design specification since it has a global effect on the overall circuit performance. In many high-performance ICs, it is commonly required for the power and ground reference potentials to be stabilized within the 5%–10% of the total voltage swing [1], [2]. It is well known that the major source of simultaneous switching noise (SSN) is due to the package interconnect paths which should be treated as a wave-based circuit model (i.e., a transmission line model), since the wave lengths of fundamental and harmonic frequencies are comparable with the physical lengths of the lines. Therefore, not only does the electrical design of the integrated circuit package can become a crucial circuit design issue, but also the package effect must be taken into account at the early stages of circuit design.

Previously, several methodologies for package performance evaluation and design have been proposed [3]–[11]. Some may be summarized as: 1) analysis based on the closed-form models of the SSN [3]–[5]; 2) package resonance effect analysis based on patch antenna models [6]; 3) package performance evaluation based on full circuit-modeling using lattice structures of many grid-type RLC-cells [8], [9]; 4) 2-D or 3-D numerical-analysis-based package performance evaluation [10], [11]. Based on the aforementioned evaluations, the power and ground reference potentials can be stabilized by utilizing the following design techniques: 1) insertion of extra-decoupling capacitances; 2) maximizing ground/power port assignment for effective package inductance reduction; 3) reduction of current loops (e.g., flip-chip packages); 4) slew rate (i.e., switching edge rate) control; 5) layout optimization with guard rings.

Nonetheless, it is formidable to meet today’s stringent package design specifications that require faster circuit switching speed as well as higher level of integration composed of mixed (e.g., digital/analog/radio frequency) signals. The aforementioned conventional approaches have the following fundamental limitations for further accurate package performance evaluation and design: 1) imprecise analytical closed-form modeling that may not reflect the realistic package effects (e.g., transmission line effects of package structure and complicated multiport switching effects), 2) 3-D field-solver-based numerical evaluation requiring heavy computation time and hardware resources, 3) constant transmission line circuit modeling in which the power and ground plane of a package are represented by lattice structures of many grid-type constant RLC-cells [8], and 4) large process variations and nonideal material characteristics (i.e., large dielectric constant variation due to mixed materials and metal roughness effect) which have a considerable effect on the circuit performance. It is to be noted that since the transmission line circuit model parameters are inherently frequency-variant due to skin, proximity, induced eddy current, and conductive substrate effect, the model may not be sufficient for high-frequency circuit design.
In addition, the relative dielectric constant for a nominal FR-4 may be in the range 3.4–6.11 since the intermetal dielectric material is composed of a mixture of glass fabric and epoxy resin [13] and the metal roughness effect is comparable to the skin depth of the ideal conductor. Thus, without considering the process variations, frequency-variant circuit model parameters, and nonideal material characteristics, all of the conventional approaches may inherently possess accuracy problems. Thus, a more accurate circuit model and model parameter determination based on experimental characterizations are essential.

In this research, a novel package circuit model design and a accurate parameter determination method which considers process variations, frequency-variant characteristics of model parameters, and nonideal material characteristics are developed, followed by package performance evaluation. Experimental characterization and model verification are performed with specified test patterns. The test patterns are fabricated using a ball grid array (BGA) package process. Model parameters are not only determined by using a commercial field-solver but also the experimental characterizations using time domain reflectometry and time domain transmission (TDR/TDT) and vector network analyzer (VNA). The frequency-variant circuit model is verified in the broad frequency band with experimental S-parameter measurements. Finally, it is shown that the package performances can be accurately as well as efficiently evaluated by employing the developed circuit models and experimental characterizations.

II. FREQUENCY-VARIANT CIRCUIT MODEL OF PACKAGE POWER AND GROUND PLANE

Lee [8] developed a generic high-frequency circuit model in which a pair of planes (power and ground planes) is divided into grid-type RLC circuit cells, as shown in Fig. 1. A cell size is determined with one tenth of the significant frequency wave length, thereby the transmission line effect can be taken into account. In the model, the edge parts of the plane are considered a half cell of an internal cell which is considered a full cell. Since a cell has a capacitance proportional to the cell area, the respective capacitances of a cell using the π model can be considered C/4 or C/2 [see Fig. 1(a)] [8], [11]. In contrast, letting the series resistance and inductance of a cell to be R and L, the respective resistance and inductance of a cell using the π model become 2R and 2L, respectively. Then finally, the power and ground plane can be modeled as a circuit, as shown in Fig. 1(b).

Such a circuit model is good enough to represent relatively low-frequency effects but it is not the case for those of the high-frequency since the frequency-variant characteristics such as skin-effect (i.e., frequency-variant resistance) and proximity effect (i.e., frequency-variant inductance) may not be accurately taken into account. In order to reflect these two frequency variant effects, a vertically stacked ladder circuit model as shown in Fig. 2 is usually employed [14]. The skin depth of a conductor is given by

$$\delta = \frac{1}{\sqrt{\pi f \mu \sigma}}$$  \hspace{1cm} (1)

where $f$, $\mu$, and $\sigma$ are a frequency, the permeability of the conductor, and the conductivity of the conductor, respectively [16].

Considering a cylindrical conductor, the skin effect can be approximately modeled with 4-RL (resistance–inductance) ladder circuits as shown in Fig. 2. At the beginning, letting the total dc resistance as $R_{dc}$, the first resistance may be approximately modeled as $R_1 = \alpha R_{dc}$, where the $\alpha R$ is a proportional constant between $R_1$ and $R_{dc}$. Similarly, assuming the resistance increase due to the skin effect in a band of frequency interval can be presented with a constant ratio ($\xi_R$)

$$\xi_R \equiv \frac{R_i}{R_{i+1}} (i = 1, 2, 3), \quad (\xi_R > 1)$$  \hspace{1cm} (2)

Then the constant ratio can be determined by solving

$$\xi_R^2 + \xi_R + \xi_R + 1 = \alpha R.$$  \hspace{1cm} (3)
Similarly, assuming the internal inductance in the given frequency interval also linearly varies, the proportional constant can be defined as

$$\xi_L \equiv \frac{L_i}{L_{i+1}} (i = 1, 2).$$  \hspace{1cm} (4)

Then the first inductance, \(L_1\), can be determined as [15]

$$L_1 = \frac{R_{dc}}{\omega \alpha_L} \left[ \frac{R_{hf} - R_{dc}(1 + \xi_R^2)}{R_{dc}(\xi_R^2 + \xi_R + 1) - R_{hf}} \right]$$  \hspace{1cm} (5)

where the subscript “hf” indicates the maximum frequency in a given frequency range. Thus, the \(R_{hf}\) is the resistance determined at the maximum frequency in the given frequency range. Note that the largest possible inductance of a structure is equal to the low frequency inductance while at the very high frequency, the internal inductance is negligibly small. Thus, the internal inductance can be represented by

$$L_{\text{internal}}^\text{hf} \approx \left( L_{\text{total}}^\text{hf} - L_{\text{external}}^\text{hf} \right)$$  \hspace{1cm} (6)

where the subscripts indicate low-frequency (lf) and high-frequency (hf), respectively. The superscript “internal,” “external,” and “total” indicate the internal inductance, external inductance, and total inductance, respectively. Assuming the first inductance can also be modeled with linear approximation, it can be defined as

$$L_1 \equiv \frac{L_{\text{internal}}^\text{hf}}{\alpha_L} = \frac{L_{\text{total}}^\text{hf} - L_{\text{external}}^\text{hf}}{\alpha_L}$$  \hspace{1cm} (7)

where the \(\alpha_L\) is a proportional constant. The \(\xi_L\) can be determined by [15]

$$\left( \frac{1}{\xi_L} \right)^2 + \left( \frac{1}{1 + \xi_R} \right)^2 \frac{1}{\xi_L} + \left( \left[ \frac{1}{\xi_R} \right]^2 + \frac{1}{\xi_R} + 1 \right)^2$$

$$- \alpha_L \left( \left[ \frac{1}{1 + \xi_R} \right] \left[ \frac{1}{\xi_R} \right]^2 + \frac{1}{\xi_R} + 1 \right)^2 = 0,$$  \hspace{1cm} (8)

$$\frac{R_{hf}}{R_{dc}} < \xi_R^2 + \xi_R + 1, \quad \frac{R_{hf}}{R_{dc}} < \xi_R^2 + \xi_R + 1.$$  \hspace{1cm} (9)

Therefore, \(\xi_R\) should be determined within the following inequality:

$$\xi_R^2 + 1 < \frac{R_{hf}}{R_{dc}} < \xi_R^3 + \xi_R + 1.$$  \hspace{1cm} (10)

Once the \(\xi_R\) is determined, \(R_3 - R_4\) and \(L_1 - L_3\) can be readily determined. However, note, although a realistic conductor may not be a cylindrical type, the \(\xi_R\) can be readily adjusted by using experimental data. Then, considering the return current path, the circuit model of a cell can be represented, as shown in Fig. 3.

Note, even though the circuit model is accurate enough for an ideal structure, it is not really the case for the practical package structure since there are the metal roughness effect and large effective dielectric constant deviation due to mixed materials. Thus, blind field-solver-based circuit model parameter determination based on nominal material parameters and physical dimensions may lead to the significant design error. It will be discussed in Section III in more detail.

### III. EXPERIMENTAL MODEL PARAMETER DETERMINATION

Since there are large process variations during the manufacturing process, the layout dimensions may have large deviations from the real physical dimensions. Thus, the physical dimensions are usually determined by using the microscope pictures of cross sections. However, in addition to the physical dimensions, more importantly, the electrical parameter variations due to material parameters and frequency-dependent transmission line effects have to be characterized. In this work, all the parameters are determined with the procedures, as summarized in Fig. 4. These experimental characterizations will be discussed in the ensuing subsections.

#### A. Experimental Dielectric Constant Determination

In order to electrically characterize the transmission line effects and material characteristics from low frequency to high frequency, various lengths of test patterns as shown in Fig. 5 are designed and fabricated by using a BGA package process. Since the dielectric constant is one of the important material constants which is strongly related with electric field variations during the circuit operation (i.e., signal delay, crosstalk, characteristic impedance, etc.), it needs to be carefully characterized. Thus, in this work, it is characterized by using various measurement equipments, i.e., 1) impedance analyzer, 2) TDR/TDT meter, and 3) vector network analyzer.

**Impedance-Analyzer (Agilent 4284A)-Based Characterization**: In order to determine the effective dielectric constant, capacitances between the signal line and ground plane are measured. However, since the data may not be reliable, they are not considered in this paper.

**TDR/TDT (Agilent 54754A)-Based Characterization**: The TDR/TDT equipment internally generates a time-domain step signal with 37.5 ps rise time as an input test wave and detects a reflected wave and transmitted wave. Thus,
the reflected waves of a pair of transmission lines which have different lengths are measured and the rise time difference between the two at 50% of the final value is selected, as shown in Fig. 5(d). Since the propagation delays of the lines are proportional to the line lengths, the propagation delay difference is directly related with line length difference. Thus, since

$$v_p = \frac{l}{t_f} = \frac{C}{\sqrt{\varepsilon_r}}$$  \(11\)

the dielectric constant can be determined by

$$\varepsilon_r = \left( \frac{t_f \cdot C}{l} \right)^2$$  \(12\)

where \(v_p, l, t_f, C,\) and \(\varepsilon_r\) are the phase velocity, line length, time of flight, and velocity of light, respectively. Note, since the pad parasites of the two lines can be considered identical, the de-embedding is not required for the measurement data. The propagation delay between the 17-mm-long line and the thru line is 0.109375 ns [see Fig. 5(d)]. Similarly, the propagation delay between the 19.7-mm-long line and the thru line is 0.127344 ns. Thus, the corresponding dielectric constants are 3.72 and 3.76, respectively.

**VNA (Agilent 8510C)-Based Characterization:** In this work, the VNA is calibrated with SOLT (short, open, load, and thru) calibration standards using the cascade-micro-tech impedance standard substrate (ISS). The parasitic pads are de-embedded with the well-known \(y\)-parameter-based de-embedding technique \[18, 19\]. The 2-port \(S\)-parameters of the test patterns are measured with a pair of GSG probes. Then the transmission line parameters (i.e., the propagation constant and characteristic impedance) can be determined, followed by the transmission line circuit model parameters (i.e., per-unit-length resistance, inductance, conductance and capacitance) \[18, 19\]. The \(S\)-parameter-based transmission line parameter determination method will be discussed in next subsections in more detail. Alternatively, the dielectric constant can be determined from phase difference of \(S_{21} [17]\). That is

$$\varepsilon_{\text{eff}} = \left( \frac{C}{\Delta l} \cdot \frac{\Delta \phi}{\omega} \right)^2$$  \(13\)

where \(C\) is the velocity of light and \(\Delta l\) is a length difference between the two lines. The \(\Delta \phi\) is the phase difference. Then, from (13), the effective dielectric constant depending upon the frequency can be determined.

**Consideration for the Measurement Data and Effective Dielectric Constant:** As described in the previous subsections, the dielectric constant is determined by using the impedance analyzer, TDR/TDT, and VNA. Among these techniques, the impedance-analyzer-based dielectric constant data show large discrepancy from the other ones. It is considered that there may be measurement errors due to equipment itself limitation or de-embedding problem (i.e., relatively large pad capacitance than those of lines), etc. Therefore, we discarded the impedance analyzer-based data. In contrast, as for both TDR/TDT and VNA-based data, all the data seem to be reasonable. Thus, as shown in
structure is composed of solder resist and intermetal dielectric materials, as shown in Fig. 5. Assuming the dielectric constant of solder resist is 3.4, the relationship between intermetal dielectric materials and the effective dielectric constant may be separately determined by using a commercial field solver. In this case, the dielectric constant of the bottom side of the mixed material corresponding to the previously measured effective dielectric constant (3.73) is 3.92.

B. Experimental Resistance Determination

During the real package process, in order to improve adherence between the conductor and dielectric material, the contact surfaces is not smooth but rough as shown in Fig. 6 (so called “metal roughness effect”). The rough surface thickness is approximately 1–3 μm which may be comparable to the skin depth. However, since the conventional skin effect model does not consider the metal roughness effect, S-parameters for the test patterns are measured and pad parasitics are de-embedded. Once such de-embedded S-parameters are determined, the transmission line characteristic parameters, i.e., the propagation constant and characteristic impedance, can be readily determined in terms of S-parameters [18]

\[
e^{-\gamma l} = \left\{\frac{1 - S_{21}^2 + S_{12}^2}{2S_{21}} \pm K\right\}^{-1} \quad (14)
\]

\[
K = \left\{(S_{11}^2 - S_{21}^2 + 1)^2 - 4(S_{11})^2\right\}^{1/2}
\]

\[
Z_c^2 = Z_0^2 \frac{(1 + S_{11})^2 - S_{21}^2}{(1 - S_{11})^2 - S_{21}^2} \quad (15)
\]

where the measurement reference impedance \(Z_0\) is 50 \(\Omega\). Thus, the transmission line parameters can be determined by

\[
R = Re\left\{\frac{\gamma Z_0}{\omega}\right\}, \quad L = Im\left\{\frac{\gamma Z_0}{\omega}\right\},
\]

\[
G = Re\left\{\frac{\gamma}{Z_0}\right\}, \quad C = Im\left\{\frac{\gamma}{Z_0}\right\}. \quad (16)
\]

The S-parameter-based frequency-variant resistances of test patterns [i.e., L3 (3.200 mm) and L6 (6.565 mm)] and the conventional skin-effect-model-based calculated data (note, the conductivity of the metal (copper)) is assumed as \(\sigma = 5.8 \times 10^7 \text{[S/m]}\) are plotted in Fig. 7. As expected, the line resistance has an excellent agreement with the S-parameter-based resistance in the frequency band of 0.1–3 GHz while it shows large deviation above 3 GHz. Note, since the effective skin depth is approximately 1.2 μm at 3 GHz from (1) and the metal roughness of the copper used for the BGA package process is known by 1–3 μm, it is considered that the metal roughness may have a significant modulation effect on the ideal skin effect model. Thus, taking the metal roughness effect \(\chi_{\text{roughness}}\) into account, the frequency-variant resistance can be modeled as

\[
R(f) = \frac{L}{\sigma \omega \delta(f)} \cdot \chi_{\text{roughness}}. \quad (17)
\]
where \( l, w, \) and \( \sigma \) are the line length, line width, and the conductivity of the line, respectively. Since the metal roughness effect (\( \chi_{\text{roughness}} \)) is considered to be inversely proportional to the skin depth, it can be approximately modeled as

\[
\chi_{\text{roughness}} \approx \frac{\delta_x}{\delta(f)} \tag{18}
\]

where \( \delta_x \) is the skin depth at a critical frequency, where the modulation effect occurs. Thus, the more practical frequency-variant resistance can be modeled as

\[
R(f) = \frac{l}{\sigma \omega \delta(f)} \approx R_{\text{RF-measured}} \quad (\text{for } f < f_s)
\]

\[
R(f) \approx \frac{l}{\sigma \omega \delta(f)} \cdot \left( \frac{\delta_x}{\delta(f)} \right) \quad (\text{for } f \geq f_s) \tag{19}
\]

where \( f_s \) is a frequency that the conventional skin-depth become smaller than the physical metal thickness \( t \). As shown in Fig. 7, the \( \delta_x \) of the tested structure can be approximately determined with 1.2 \( \mu m \) that is the skin depth at 3 GHz. The new model-based frequency-variant resistance is plotted in Fig. 7 by using long-dashed line. It shows good agreement with \( S \)-parameter-based measurement data.

IV. MODEL VERIFICATION AND PACKAGE PERFORMANCE EVALUATION

In order to investigate the model accuracy, two pairs of power and ground planes are designed and fabricated in the same test substrate as the transmission line patterns. One structure size is 10 mm \( \times \) 10 mm and the other is 37 mm \( \times \) 37 mm. The measurement ports are positioned at the center of the left- and right-side edges of the structure as shown in Fig. 8. Then, the \( S \)-parameters of two ports are measured by using VNA from 0.1–26.5 GHz. The cell size of the circuit model has to be less than one tenth \( (\lambda/10) \) of the wave length of the significant frequency. Thus, assuming the significant frequency is 26.5 GHz, a cell size is 0.5 mm that is determined by \( \lambda = c/(f \sqrt{\varepsilon_r}) \) and transmission line model parameters (i.e., \( R, L, \) and \( C \)) are determined by using the expressions [16], [21]

\[
R_{\text{RF}} = \frac{1}{\sigma} \cdot \left( \frac{t}{\delta_{\text{RF}}} \right)^2 \tag{20}
\]

\[
R_{\text{RF}} = \frac{1}{\sigma} \frac{1}{\delta_{\text{RF}}} \cdot \left( \frac{\delta_x}{\delta_{\text{RF}}} \right) \cdot 2 \tag{21}
\]

\[
L_{\text{RF}} = \mu \left( \frac{t_1 + t_2}{2} \cdot \frac{1}{6} \right), \quad L_{\text{RF}} = \mu d \tag{22}
\]

\[
C = \frac{\varepsilon_0 \cdot \varepsilon_r \cdot I \cdot v}{d} \tag{23}
\]

where \( \sigma \) is copper conductivity \((= 5.8 \times 10^7 \text{ S/m})\) and \( \varepsilon_r \) is the previously determined FR-4 epoxy dielectric constant \((= 3.92)\). Both \( t_1 \) and \( t_2 \) are the thickness of the ground plane and
that of the power plane, respectively. All the cross-sectional dimensions are the same, as shown in Fig. 8. The $\delta_{HF}$ is a skin depth at 26.5 GHz. Then the transmission line parameters are $R_{dx} = 0.0022989 \, \Omega$, $R_{dx} = 0.251 \, \Omega$, $L_{dx}^{\text{internal}} = 61.2 \, \text{pH}$, $L_{dx}^{\text{external}} = 57.8 \, \text{pH}$, and $C = 0.1886 \, \text{pF}$, respectively. As shown in Fig. 9 with these transmission line parameters in the vertically stacked four-ladder circuit model, the $\xi_{BR}$ can be approximately determined by 5.1. Thus, letting $\xi_{BR} = 5.1$, the circuit model parameters for both the half-cell and full-cell can be determined as shown in Fig. 10. Then finally a circuit model for a pair of the ground and power planes may be represented with a mesh-type (grid-type) circuit as shown in Fig. 10.

In order to investigate the accuracy of the model, the frequency-variant circuit model responses (HSPICE-based $S$-parameter data) are compared in terms of $S$-parameters that are determined by using a commercial full-wave-based field solver (i.e., HFSS) and the VNA-based measurements [20]. As shown in Figs. 11 and 12, the measurement-based frequency-variant circuit model responses have excellent agreement with VNA-based measurement data up to 15 GHz.

In contrast, the conventional transmission line model shows a substantial error. Note, since the material parameters and process variations have a significant effect on the high-frequency characteristics, all the input parameters for the numerical simulations such as metal roughness, resistivity, and physical dimensions are experimentally characterized. Otherwise, the conventional full-wave-based simulation (i.e., HFSS) using nominal material parameters and frequency-invariant model [8] may have large deviation from real measurement data as shown in Figs. 11 and 12. In addition to the experimental characterization of the input parameters for the numerical simulations, blind simulation may have significant inaccuracy problems. Consequently, it is noteworthy that, without careful experimental characterizations, blind numerical simulation or circuit model parameter determination using the commercial field-solvers may be very risky.

Once the accuracy of the circuit model is verified, the package performance for the various conditions can be readily evaluated by using a general purpose circuit simulator, e.g., HSPICE. In this work, a test package structure is assumed, as shown in Fig. 13, where the external power is supplied at the bottom side of the package. The power is distributed for die I/O circuits.
Fig. 12. Comparison of the proposed circuit-model-based S-parameters with measured S-parameter data for 37 mm × 37 mm power/ground plane pair. Note, 3-D-field-solver-based S-parameter data are determined by using both characterization data and nominal parameters. (a) S_{11} (magnitude[dB]). (b) S_{21} (magnitude[dB]).

Fig. 13. Schematic representation of power distribution network.

Fig. 14. Impedance variations of six-layered power delivery network (PDN) due to the extra decoupling capacitance insertion. (a) Schematic representation of power distribution network of six-layered package with decoupling capacitance. (b) Impedance variations of PDN with the decoupling capacitance insertion. Ceramic capacitances are connected on the bottom of the package substrate. (c) Impedance variations due to the inverse variation of the dielectric constant and dielectric thickness of PDN.

package material effects, these influences can be readily evaluated by combining the frequency-variant circuit model and the external circuit components such as die I/O circuits and decoupling capacitors.

As an example, such package performances can be accurately investigated with the proposed experiment-based circuit model for a test package, as shown in Fig. 14. Note, in the package, external ceramic capacitances are connected between the power and ground plane through vias. As shown in Fig. 14(b) and (c), the low-frequency impedance of the package can be significantly reduced by using the external multilayer ceramic capacitors (MLCCs) while the high-frequency impedance can be reduced by using the power/ground planes. Nonetheless, significant resonance effects due to the parasitic inductances and capacitances may not be reduced with the capacitances. Thus, in order to accurately control the resonance effects, the measurement-based circuit simulation for the integrated circuit packages
may be desirable. Possibly, small-scale "embedded capacitors" with much higher dielectric constant material (i.e., much higher per-unit-area capacitance) and much less parasitic inductance should be employed to meet future high-performance integrated circuit package or system in package (SIP) design requirements. Similarly, other circuit performances such as driver strength, signal integrity, jitter, intersymbol interference (ISI) noise, etc., can be efficiently as well as accurately evaluated with the proposed package circuit model.

V. CONCLUSION

For a real package, there exist numerous process variations. Moreover, the transmission line circuit model parameters of a package may be frequency-variant. Thus, since cross-section-based simple numerical simulation may greatly deviate from the real measurement data, experimental characterizations are considered crucial for high-performance package design.

In this research, a novel high-frequency measurement-based frequency-variant circuit model and model parameter determination methods were developed with specified test patterns which were fabricated with a BGA package process. The test patterns were experimentally characterized by using an impedance analyzer, TDT/TDT, and VNA. Not only were the material parameters such as dielectric constant and resistance determined, but also the transmission line parameters (i.e., the propagation constant and characteristic impedance) over a wide frequency band were determined. Thereby, practical circuit model parameter variations and frequency-variant characteristics of the transmission lines concerned with the package were accurately investigated and modeled. It was shown that the measurement-based frequency-variant circuit model for a pair of power/ground planes has excellent agreement up to 15 GHz with both the VNA-based measurements and the measurement-based full-wave simulation, whereas the conventional transmission line circuit model and the nominal-parameter-based full-wave numerical simulation show large discrepancy with the measurement data. The proposed technique can be usefully employed in industry for accurate simulation and design of high-performance integrated circuits and packages.

REFERENCES


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