ulator HP-Momentum.[1] Fig. 4 compares simulated and modeled S-parameters of the structure of Fig. 3. Two models are considered. The first one is a low-frequency quasi-static model, which consists of cascaded coupled transmission-line sections from which the circuit parameters are determined with a quasi-static EM solver [11]. The quasi-static model does not include field effects at the location of the discontinuities. The second model is obtained with the described algorithm starting from the simulated S-parameters. The algorithm is applied with a smooth step signal with rise time $t_r = 100$ ps. Fig. 4 shows that the derived model is valid up to 7 GHz and that the field effects of the discontinuities must be considered in the equivalent-circuit model.

The second example is shown in Fig. 5. The DUT consists of two connector pin's of a multipin's backplane connector placed on a component board and a backplane board. Striplines on both printed circuit boards are making contact with the connector pin's. In order to connect the striplines with the measurement instrument, the component board is provided with planar contacts for coplanar probes, while the backplane board has SMA connectors. Measurements are performed with the HP8510 network analyzer from 50 MHz to 20.05 GHz using a SOLT calibration. The obtained circuit model for the connector via holes and the connector pin's is shown in Fig. 6. The connector via holes are modeled with a transmission line, while the connector pin's are modeled by three coupled transmission lines. Fig. 7 compares measured and modeled reflection, transmission, backward, and forward crosstalk for an incident step signal with a rise time of 150 ps. As can be concluded from these pictures, not only reflection, but also transmission (rise-time degradation) and crosstalk are accurately modeled.

IV. Conclusion

An algorithm is presented for the circuit modeling of coupled interconnection structures. Based on the principle of causality, a hybrid circuit-model equivalent consisting of lumped elements, transmission lines, and coupled elements is derived for the structure under test. The simulation and measurement examples show that reflection, transmission, and crosstalk properties of the interconnection are accurately modeled.

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Coax Via—A Technique to Reduce Crosstalk and Enhance Impedance Match at Vias in High-Frequency Multilayer Packages Verified by FDTD and MoM Modeling

Edward R. Pillai

***Abstract*—Large-scale crosstalk at vias and poor via electrical performance are major drawbacks in state-of-the-art high-frequency multilayer first- or second-level integrated-circuit/monolithic-microwave integrated-circuit (IC/MMIC) packages. The coax via design modeled in this paper breaks new ground in achieving more than 30-dB ultrawide-band crosstalk reduction and providing an enhanced impedance match.**

I. Introduction

The multilayer integrated-circuit (IC) package known as a single-chip module (SCM) or multichip module (MCM) used in high input/output (I/O) digital applications—whether organic, alumina, or glass/ceramic based—is now being aggressively sought after as the circuit carrier of choice in integrating microwave, RF wireless, and high-speed digital circuits driven by the underlying cost factors associated with those markets. Although manufacturing such a high-frequency package with good material and dimensional tolerances is proving to be less difficult at frequencies well into the gigahertz range, it is a real challenge to be able to design the package electrically against distortion—namely, crosstalk. Constant-impedance TEM stripline transmission used in the inner layers is extremely effective in achieving negligible crosstalk [1]. Quasi-TEM microstripline transmission used on the top layer causes significant crosstalk [2], but can be circumvented by either placing shielding grounded vias (see Fig. 1) between lines or going directly from device/chip pad into an inner layer. The third and only other (yet most problematic) crosstalk hot spot occurs at signal via-hole transitions. Here, the physical discontinuity causes several unwanted modes that cause severe coupling to other vias and lines. No amount of optimization of the via dimensions, materials, or dielectric constants

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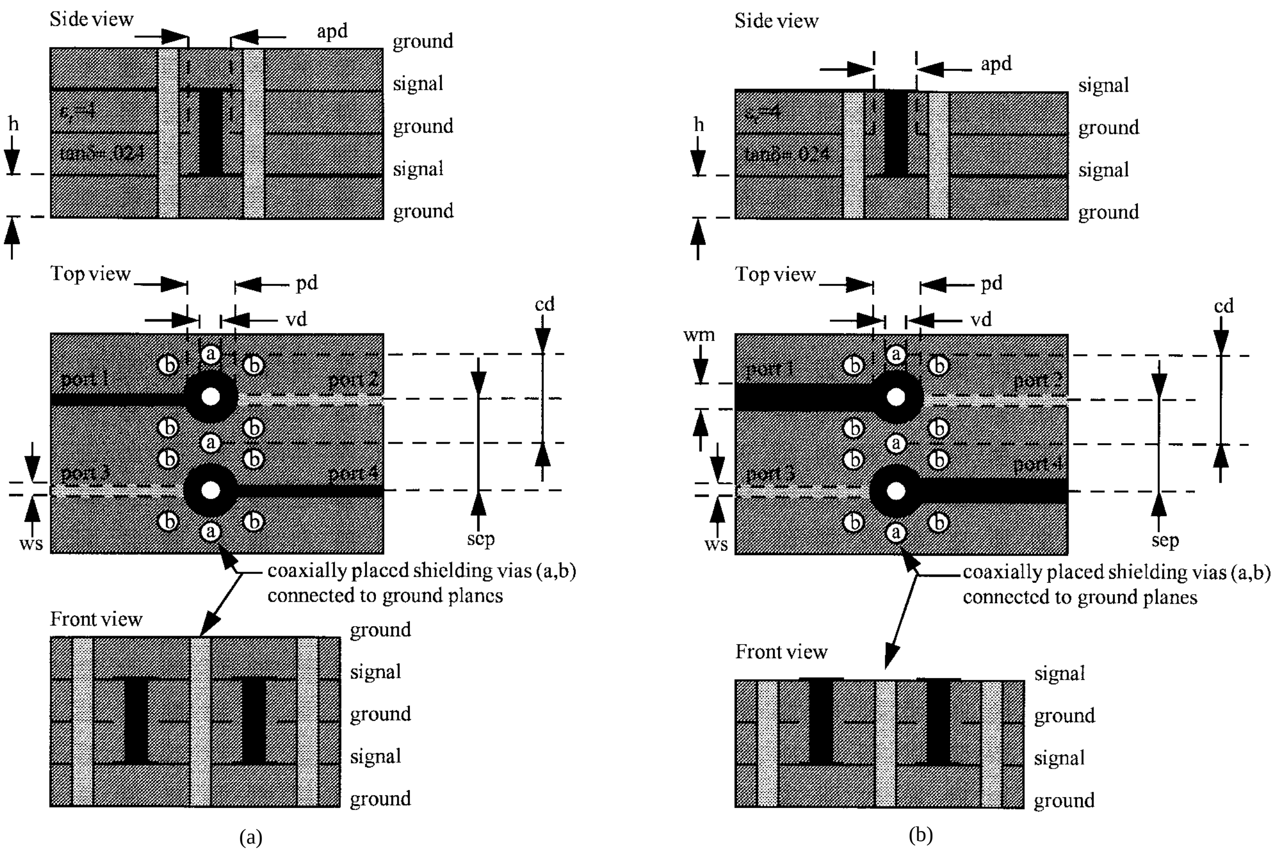


Fig. 1. (a) Coupled stripline-to-stripline coax vias and dimensions. (b) Coupled microstrip-to-stripline coax vias and dimensions.

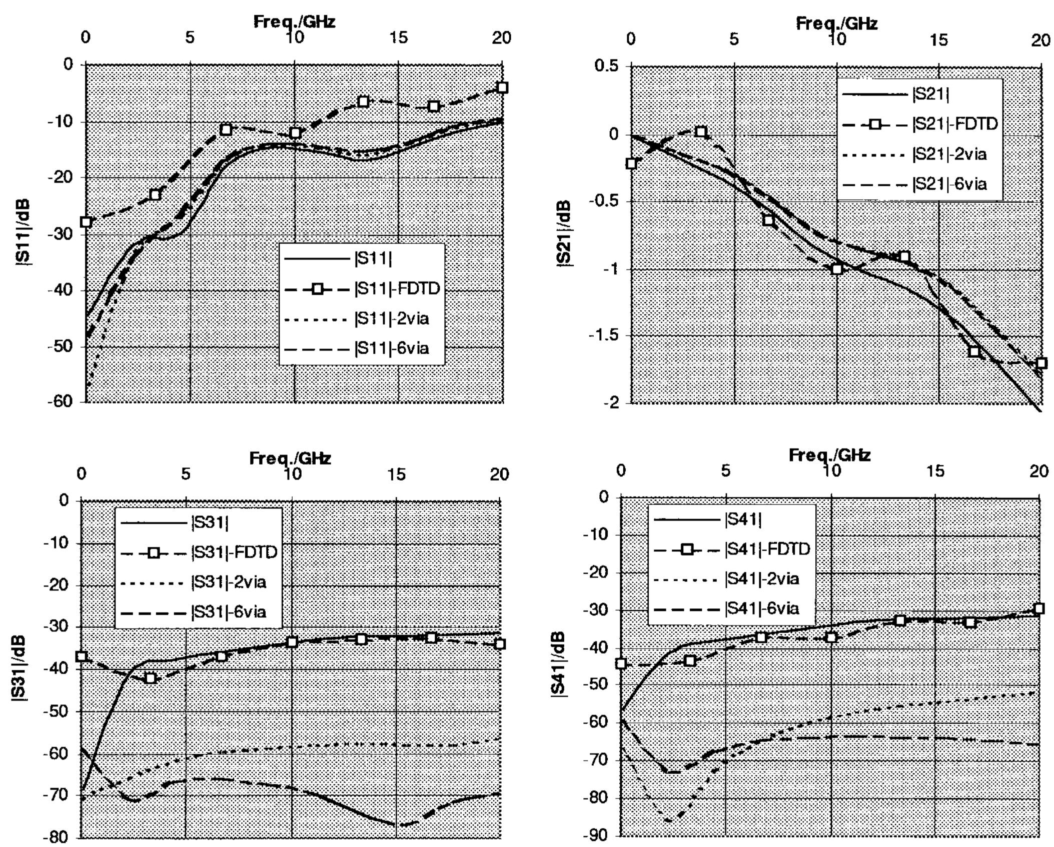
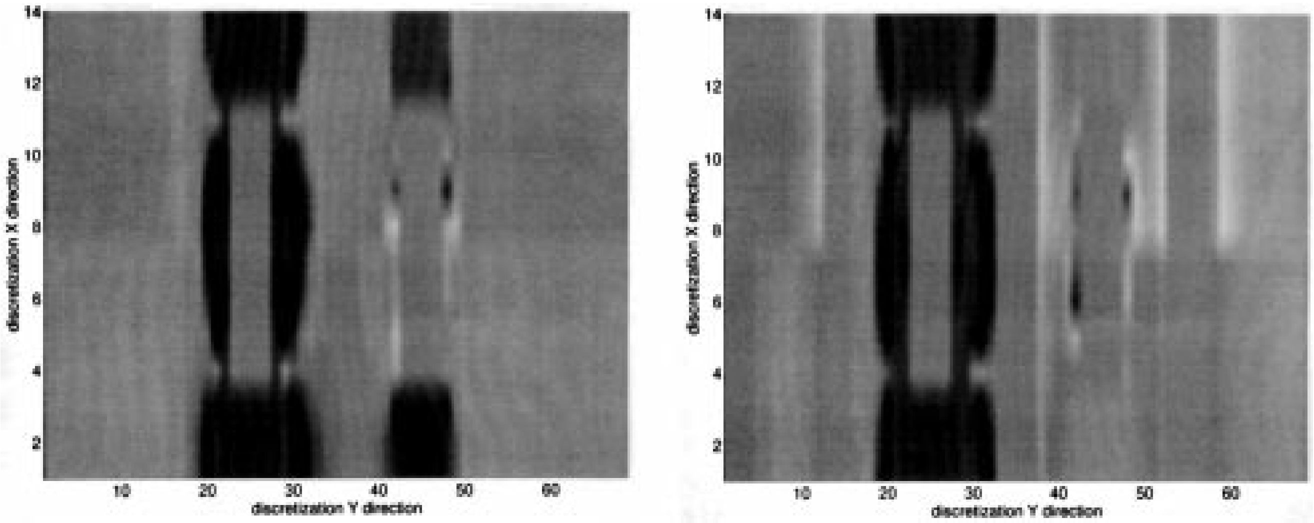


Fig. 2. Scattering-parameter results calculated with the MoM and FDTD method for the coupled stripline-to-stripline vias with zero, two, and six coax shielding vias.



(a) (b)

Fig. 3. (a) Time-harmonic vertical electric-field component in decibels at 10 GHz for cross-section "front view" in Fig. 1(a) without shielding vias. (b) Time-harmonic vertical electric-field component in decibels at 10 GHz for cross-section "front view" in Fig. 1(a) with two shielding vias per signal via.

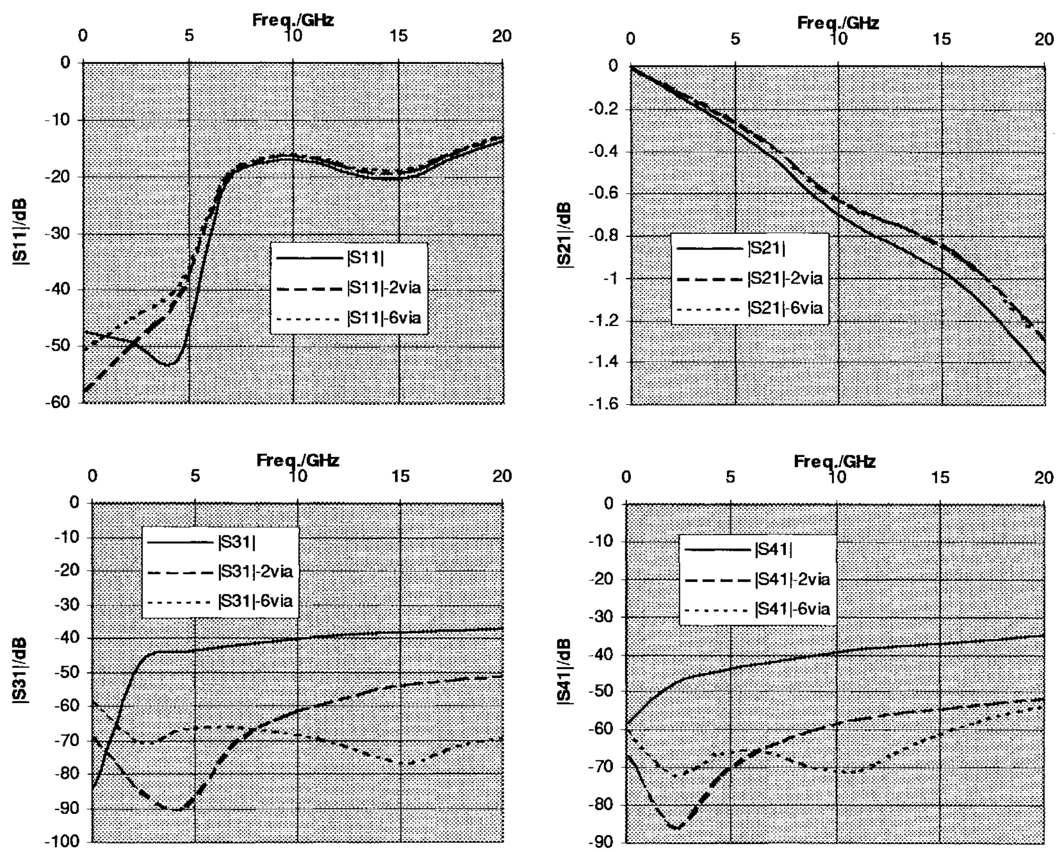


Fig. 4. Scattering-parameter results calculated with the MoM for the coupled microstrip-to-stripline vias with zero, two, and six coax shielding vias.

will significantly reduce crosstalk. Hence, this idea of the coax via is introduced. The ball-grid array (BGA) area-array interconnection used in the SCM and MCM lends itself excellently to implementing the coax via, not possible in many other types of packages.

The finite-difference time-domain (FDTD) method [3] and method of moments (MoM) [4], [5]—well-established electromagnetic (EM) full-wave tools—are used to analyze the coax via. The three-dimensional (3-D) FDTD is used for verification, while the two-and-one-half dimensional (2.5-D) MoM is used as the primary analysis tool, as it is better suited to planar structures, requiring less discretization (basis functions), and hence, lower computational overhead.

II. Coupled Stripline-to-Stripline Coax Vias

The structure in Fig. 1(a) with $\varepsilon_r = 4$, $\tan\delta = 0.024$, $\mathrm{apd} = 0.4$ mm, $\mathrm{pd} = 0.3$ mm, $\mathrm{vd} = 0.2$ mm, $h = 0.11$ mm, $\mathrm{cd} = 1$ mm, $\mathrm{sep} = 1$ mm, and $\mathrm{ws} = 0.111$ mm (50 Ω) is initially computed without the use of the surrounding shielding vias. Although

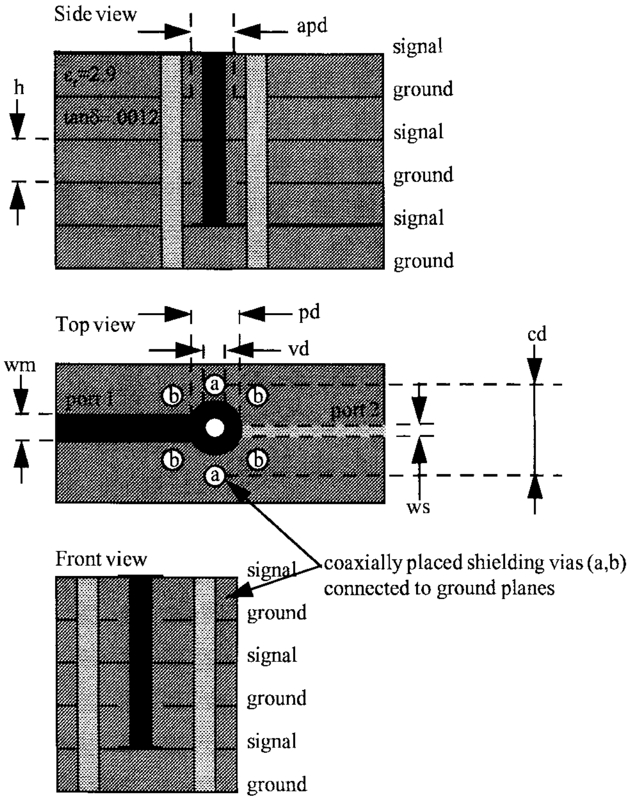


Fig. 5. Single multilayer microstrip-to-stripline coax via.

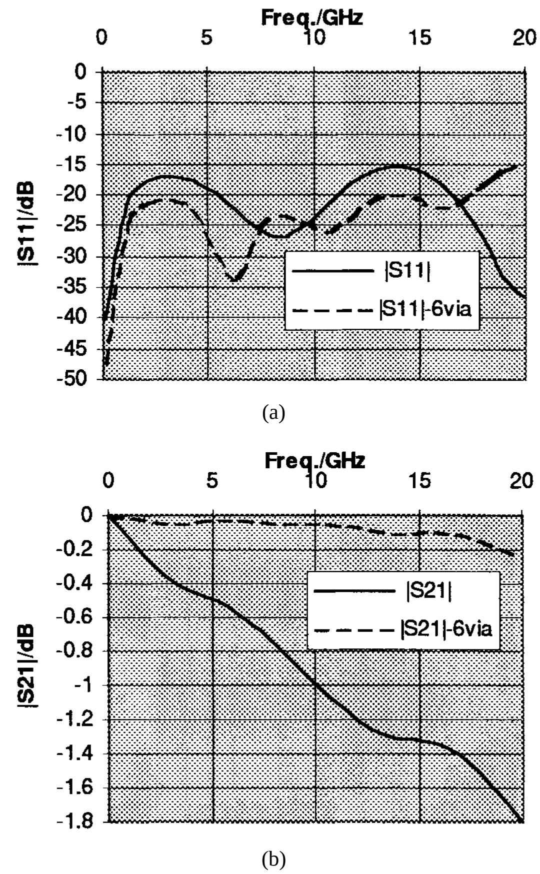


Fig. 6. (a) Scattering parameter $|S11|$ results for a multilayer microstrip-to-stripline via transition with and without shielding. (b) Scattering parameter $|S21|$ results for a multilayer microstrip-to-stripline via transition with and without shielding.

a reasonable match $|S11|$ is seen in Fig. 2(a), the transmission $|S21|$ in Fig. 2(b) is seen to be lossy due to the use of dri-clad having a high loss tangent (0.024). However, for the 1-mm spacing, the near-end coupling $|S31|$ in Fig. 2(c) and far-end coupling $|S41|$ in Fig. 2(d) show approximately -30-dB down values up to 20 GHz, which may seem acceptable for some RF applications, but one must keep in mind that in reality, when effects of coupling at line bends and line coupling are also taken into account, this figure will worsen to approximately -15 dB. This is unacceptable for most high-frequency applications. This result is verified through the FDTD modeling of the same structure (without shielding vias) shown in Fig. 2(a)–(d). The $|S11|$ result is in good agreement, variation being due only to port calibration differences. The $|S21|$ result has some ripple due to the Gibb's effect in the fast Fourier transform (FFT), but the center of the ripple coincides with the MoM curve, proving excellent agreement. Of greater importance is the excellent agreement observed for the $|S31|$ and $|S41|$ values.

The next case computed is conducted with only the coax shielding vias labeled a in Fig. 1(a). It is of interest to observe any improvement based on adding only two shielding vias per signal via, as package/board manufacturing costs are tightly linked to the number of vias used. The diametric positions of the a and b vias are calculated using the simple coaxial transmission equation [6], in this case $\mathrm{cd} = 1$ mm for 50 Ω. The corresponding results in Fig. 2(a)–(d) indicate a good match $|S11|$, and hence, about 0.25 dB less loss $|S21|$ can be obtained, but more significantly, almost 25-dB improvement in coupling as seen by $|S31|$ and $|S41|$ up to 20 GHz. The third case calculated utilizes all six shielding vias marked a and b in Fig. 1(a). Here again, as can be seen by $|S31|$ and $|S41|$, almost an extra 10 dB of wide-band isolation is achieved. The time-harmonic vertical electric-field component at 10 GHz is computed using the FDTD with the perfectly matched layer (PML) boundary condition [7] for the "front view" cross section of Fig. 1(a), and the result is shown in Fig. 3. Fig. 3(a) shows the case without shielding vias and Fig. 3(b) with two shielding vias per signal via. In this setup, the left signal via is excited at the lower conductor. The right signal via is passive. To avoid line-to-line coupling, the connecting conductors are placed on opposing sides of the dividing ground plane. The passive via clearly receives more crosstalk without the shielding vias.

III. Coupled Microstrip-to-Stripline Coax Vias

The structure in Fig. 1(b) is calculated with the MoM and with material parameters, with dimensions of $\varepsilon_r = 4$, $\tan\delta = 0.024$, $\mathrm{apd} = 0.4$ mm, $\mathrm{pd} = 0.3$ mm, $\mathrm{vd} = 0.2$ mm, $h = 0.11$ mm, $\mathrm{cd} = 1$ mm, $\mathrm{sep} = 1$ mm, $\mathrm{ws} = 0.111$ mm (50 Ω), and $\mathrm{wm} = 0.226$ mm (50 Ω). Results similar to Section II are seen in Fig. 4(a)–(b). For no shielding vias, the coupling is about -30 dB up to 20 GHz, as seen in $|S31|$ and $|S41|$. By introducing only the shielding vias a at a coaxial distance [as shown in Fig. 1(b)], a good match $|S11|$ with less loss $|S21|$ can again be obtained, as well as an almost 25-dB wide-band improvement in coupling $|S31|$ and $|S41|$ is observed. The addition of all shielding vias a and b, interestingly provides several additional decibel improvement in $|S31|$ and $|S41|$, especially at higher frequencies.

IV. Multilayer Microstrip-to-Stripline Via with and Without Shielding

In order to show that the coax-via concept is effective in reducing crosstalk and improving impedance match also with a several layer structure, the via in Fig. 5 is calculated. Here, a Rogers 6002 material with $\varepsilon_r = 2.94$, and $\tan\delta = 0.0012$ (very low loss) is used. The dimensions pertaining to Fig. 5 are $\mathrm{apd} = 1.27$ mm, $\mathrm{pd} = 0.635$ mm, $\mathrm{vd} = 0.3$ mm, $h = 0.254$ mm, $\mathrm{cd} = 2.54$ mm, $\mathrm{ws} = 0.312\,42$ mm (50 Ω), and $\mathrm{wm} = 0.635$ mm (50 Ω). The result in Fig. 6(a) for $|S11|$ shows improvement up to 18 GHz when using the six shielding vias. More significantly, $|S21|$ in Fig. 6(b) shows ten orders of magnitude in decibel improvement up to 20 GHz, indicating an excellent impedance match at the transmission line/via junctions and less energy escaping to produce crosstalk. This means that although for a longer several-dielectric-layer-thick via structure, crosstalk worsens when the via is unshielded, i.e., noncoaxial, the coax via's coupling remains at its low level, independent of the number of layers it traverses. This is explained through the fact that once a coax-like mode is successfully launched at the beginning of the via, it can propagate through several layers undisturbed.

V. Conclusion

It is well known that via-hole transitions excite parallel-plate modes in stripline structures. Proper inclusion of ground vias does indeed suppress these modes, thereby reducing coupling to other vias or lines. In addition, one would expect to see resonances in the S-parameter results presented, but were not evident. The package dimensions were small enough that the resonances did not occur in the bandwidth. This miniaturization is the key strength offered by the SCM and MCM technologies.

The use of a coax via is powerful in reducing crosstalk and, therefore, improving package performance. The shielding ground-plane connected via can be used at other points in the package to drastically lessen coupling between lines, as the coupling mechanism is through the dielectric.

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