

A Novel Methodology for Defining the Boundaries of Geometrical Discontinuities in Electronic Packages

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Abstract—For efficient RF/microwave modeling, measurement and design of signal paths in electronic packages (e.g., from chip to chip or chip to board), the boundaries of geometrical discontinuities along these paths must first be accurately defined. In this contribution, a novel methodology for defining the boundaries of all geometrical discontinuities in electronic packages and boards will be presented, illustrated and experimentally validated. Discontinuity is used in this work to describe any package/board component that distorts the electromagnetic (EM) field pattern on uniform transmission lines, e.g., flip chip interconnects, wirebonds, vias, bends on package/board traces, pins, solder balls, etc.

I. INTRODUCTION

As signal rise and fall times continuously decrease, causing the significant frequency content of digital signals to extend into the GHz range, parasitic effects arising from discontinuities along signal paths in electronic packages and boards may severely degrade the quality and timing of signals, leading to system mal-functioning or even failure. To prevent such occurrences, the effects of these discontinuities must be well investigated and compensated for, at the beginning of the design cycle. To reach this goal, thorough RF/microwave modeling and measurement is required. However, efficient modeling/measurement of any discontinuity at RF/microwave frequencies can only be guaranteed, if its boundaries are accurately defined, i.e., if the points away from the physical limits of the discontinuity where higher-order modes (HOMs), excited at the discontinuity, vanish or become insignificant, are known.

For more than half a century now, different methods have been used to study the effects of discontinuities. In 1955, *A. A. Oliner* [1] derived some analytical expressions to determine the equivalent circuit parameters of various discontinuities in a balanced strip transmission line. Quasi-static methods were used in, for example [2] – [4], for the extraction of the static capacitances and low-frequency inductances of vias and bends, respectively. Full-wave techniques were employed in, for example [5] – [7], to extract the frequency-dependent characteristics of these discontinuities. However, in all these works, the boundaries

of discontinuities were not defined. This implies, that the degree of EM field distortion in the vicinity of the discontinuities has been estimated so far only by guess work, because the distance away from the physical limit of the discontinuities where the HOMs reach wasn't defined.

In this work, a novel and comprehensive approach for defining the boundaries of all geometrical discontinuities in electronic packages and boards will be presented. Reasons for defining these boundaries will first be discussed in section II. This will then be followed by a presentation of the development, illustration and validation of the novel methodology in sections III, IV and V, respectively.

II. NEED FOR DEFINING THE BOUNDARIES OF GEOMETRICAL DISCONTINUITIES

Before outlining the need to define the boundaries of geometrical discontinuities, let's first take a closer look at what exactly happens at a discontinuity. Consider a signal excited at port 1, and propagating in a transverse electromagnetic (TEM) or quasi-TEM mode along a package/board trace with a discontinuity (e.g., a 90° bend in this case) towards port 2, as shown in Figure 1.

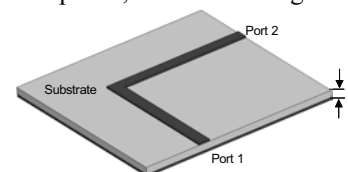


Figure 1: A microstrip trace with a 90° bend (h =substrate thickness).

As long as the cross-section seen by the signal at any point in time is the same as the cross-section for which the TEM or quasi-TEM mode was defined at port 1, the signal continues propagating and/or attenuating in the longitudinal direction without any change in the transverse direction. However, when the signal meets a discontinuity, a scattered field is excited, such that the total field, consisting of the original incident and the scattered field, matches all the necessary boundary conditions. In the vicinity of the discontinuity, the scattered field decomposes into various

modes of the interconnecting traces. One of these modes is the fundamental mode (i.e., TEM or quasi-TEM, depending on the configuration of the line) and the others are HOMs. Since the cutoff frequency of the fundamental mode is zero, the scattered TEM or quasi-TEM mode will be propagating. When viewed from port 1, it is seen as a wave reflected by the discontinuity. The excited HOMs are either propagating or evanescent, depending on frequency. If the operating frequency is lower than the cutoff frequency of the excited HOMs, then they will be evanescent or non-propagating. In this case, the fields of the HOMs will decay exponentially with distance, and at a sufficient distance away from the discontinuity, they will completely vanish, leaving the fundamental mode unperturbed. Such a distance marks the *end of the discontinuity effect* and is termed, the *boundary of the discontinuity*. These boundaries must be accurately defined because of the following reasons:

- In order to prevent/solve signal integrity problems in electronic packages/boards, efficient wideband models for complete signal paths within these subsystems are needed. The development of such models involves the use of multilumped modeling, whereby the signal paths are segmented, followed by an extraction of lumped models for each segment and then cascading the resulting models to increase bandwidth. However, since electronic packages/boards consist almost entirely of discontinuities, this segmentation can only be done at points away from each discontinuity where fields of HOMs have vanished or are considered insignificant, because lumped/distributed element modeling is based on TEM wave theory. *Consequently, defining the boundaries of each package/board discontinuity is a pre-requisite for efficient wideband modeling and analysis of complete signal paths in electronic packages and boards.*

- To predict/optimize the RF performance of electronic packages/boards, their discontinuities must be thoroughly characterized through 3D full-wave EM field computations and/or RF measurements at the beginning of the development cycle. To perform such computations with field solvers, the discontinuities are connected to ports using uniform transmission lines (e.g., microstrip, stripline...), which represent the actual configuration of traces in the package or board. *The length of the line used for this interconnection depends on the boundaries of the discontinuity under consideration, and it determines the accuracy, duration and memory requirements of the field computation.* If the lines are too short, then HOMs excited at the discontinuity will be captured at the ports, leading to erroneous and misleading results. A similar problem is also encountered when designing test structures for RF measurements. On the other hand, if the transmission lines are longer than required, then the field computation time and memory required will increase, leading either to unsuccessful computations due to insufficient memory or computations that may run for days, instead of just a few hours.

III. METHODOLOGY TO DEFINE THE BOUNDARIES OF GEOMETRICAL DISCONTINUITIES

In this section, the methodology we developed for defining the boundaries of geometrical discontinuities will be presented. However, due to space limitation, it will not be possible to give an in-depth analysis.

A. 2D EM Field Analysis of the Interconnecting Traces

Since the scattered field excited at the discontinuity decomposes into various modes of the interconnecting traces, the traces must first be separately analyzed, so as to be able to know when the HOMs they support become propagating, and how they attenuate with distance.

- **Computation of the Cut-off Frequency of the Lowest-order Higher-order Mode (Lo-HOM):** For this purpose, Maxwell's equations have to be completely solved. However, due to the complicated field structure in most planar transmission lines, it is generally not possible to obtain analytical solutions of Maxwell's equations for these structures, in which the total effects of all modes are considered. So, these lines are mostly approximated using waveguides, because analytical solutions of Maxwell's equations for most waveguides are feasible. For example, a parallel plate waveguide (PPWG) is used to model the propagation of HOMs in striplines [8]. A microstrip line can be approximated by a section of a PPWG, neglecting fringing, and the relationship between HOMs and the quasi-TEM mode in a microstrip line is analogous to the relationship of transverse electric (TE) and transverse magnetic (TM) modes to the TEM mode in a PPWG [9]. Based on an analytical approach, the cutoff frequency is given as

$$f_{TE_n, TM_n} = \frac{n}{2a\sqrt{\mu\epsilon}}, \quad (1)$$

where n designates the order of the mode. Therefore, for the Lo-HOM (TE₁ and TM₁), n will be replaced by 1 and the cutoff frequency will then be given as

$$f_{TE_1, TM_1} = \frac{1}{2a\sqrt{\mu\epsilon}} = \frac{v}{2a}. \quad (2)$$

From (1), it can be seen that the second Lo-HOM (e.g., TM₂) has a cutoff frequency equal to twice the value of TM₁. TM₃ has a cutoff frequency three times that of TM₁, and so on. This explains why it is sufficient to determine only the cutoff frequency of the Lo-HOM. If these modes are not propagating, then other HOMs can not also propagate.

Since a represents the substrate thickness, it can be deduced from (2) that the Lo-HOM can only propagate if a is greater than, or equal to half the wavelength of the propagating signal. This indicates that for miniaturized electronic packages used for RF/high-speed applications with very thin dielectric substrates, HOMs can not propagate.

For a more accurate computation of the cutoff frequency of any planar transmission line, an EM field solver must be used, so as to consider the fringing fields.

- **Computation of the Attenuation Constant, α , of Lo-HOM and Definition of the Range of Existence of its Fields Necessary for 3D EM Field Analysis:** All field components of HOMs have a general form given as:

$$X_{x,y \text{ or } z} = X_0 e^{-\gamma z} = X_0 e^{-\alpha z} e^{-j\beta z}, \quad (3)$$

where X_0 is the magnitude. Since $\gamma \approx \alpha$ for the Lo-HOM,

$$X_{x,y \text{ or } z} = X_0 e^{-\alpha z}. \quad (4)$$

From (4), the distance, l ($=z$), away from the point of excitation where the field components of a given mode attenuate to a particular value can easily be computed. Based on this equation, numerous investigations were performed and the range of existence of the fields of HOMs necessary for 3D EM analysis was defined as:

$$\frac{1}{10\alpha} \leq l \leq \frac{23}{5\alpha} \quad (5)$$

The exact boundaries of discontinuities lie within this range. To define these boundaries, a maximum of 12 data points, corresponding to 12 lengths of the interconnecting traces, have to be used for the 3D field analysis.

B. 3D EM Field Analysis of the Interconnecting Traces and Discontinuity: Examination of Poynting Theorem

Since an infinite number of HOMs are excited at a discontinuity, it is not possible to define the boundaries of discontinuities by defining the exact point where each mode vanishes. For this reason, the *methodology used to define the boundaries of discontinuities in this work is based on the fact that the HOMs disturb the fundamental mode. Hence the boundaries of a discontinuity are characterized by an end of this disturbance.*

To define the end of this disturbance, the variation of power within the range defined in (5) must first be examined. For this purpose, the Poynting theorem, given in (6), was used.

$$-\frac{1}{2} \int_V (\vec{E} \cdot \vec{J}_s^* + \vec{H}^* \cdot \vec{M}_s) dv = \frac{1}{2} \oint_V \vec{E} \times \vec{H}^* \cdot d\vec{s} + \frac{\sigma}{2} \int_V |\vec{E}|^2 dv + \frac{\omega}{2} \int_V (\epsilon |\vec{E}|^2 + \mu |\vec{H}|^2) dv + j \frac{\omega}{2} \int_V (\mu |\vec{H}|^2 - \epsilon |\vec{E}|^2) dv \quad (6)$$

In the immediate vicinity of the discontinuity, the imaginary part of (6) is very large due to the presence of the evanescent modes which store electric and magnetic energies, and the losses are negligible. As the distance away from the discontinuity increases, this imaginary part decrease exponentially due to the exponential decay of the fields of the HOMs. The boundaries of discontinuities are characterized by a very small and steady component of this imaginary part, because the current and field distributions approach those of an infinitely long transmission line where only the TEM mode is present. This variation in power can also be seen in S-parameters since, for example, $|S_{m}|^2$ is equal to the ratio of power reflected from, to power incident on, an n-port. For the computation of these S-parameters, a numerical technique that solves the complete wave equation must be used so as to take into consideration the effects of

all the modes. Only the fundamental mode has to be excited at the ports, since only the effects of HOMs on this fundamental mode are of interest. To prevent frequency-dependent losses, if any, from overshadowing the results, the S-parameters have to be computed only at a single frequency point – the highest frequency of interest.

C. Definition of the Boundaries of Discontinuities

Based on the extracted S-parameters, the exact boundaries of a discontinuity, represented by l_{bod} , can then be defined as follows:

$$\text{If } |S_{m}|_l - |S_{m}|_{l+\Delta l} \approx 0, \quad (7)$$

then $l = l_{bod}$.

Depending on the problem at hand, these boundaries can also be defined as the points where $|S_{m}|$ changes by less than a self defined limit, e.g., 5%. In this case,

$$\text{If } |S_{m}|_l - |S_{m}|_{l+\Delta l} \approx x, \quad (8)$$

and $\frac{x}{|S_{m}|_{l+\Delta l}} \leq 0.05$, then $l = l_{bod}$.

IV. ILLUSTRATION OF METHODOLOGY

For an illustration of the methodology developed for defining the boundaries of discontinuities, a through-hole via, whose cross-sectional view is shown in Figure 2, was used as an example. The via interconnects two uniform 50 Ω microstrip traces (each being 175 μm wide and 27 μm thick), on layers 1 and 4, in a four-layered FR4 substrate ($\epsilon_r=4.2$, $\tan \delta=0.007$). The inner layers are used as power/ground planes. Each of the pre-peg layers has a thickness of $h1=100 \mu\text{m}$, and the core is 150 μm thick. The diameters of the via hole and pad are 200 μm and 400 μm , respectively.

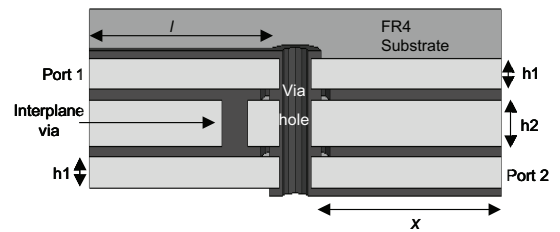


Figure 2: Cross-sectional view of via and interconnecting traces.

The cutoff frequency of the Lo-HOM was computed to be approximately 200 GHz. At 30 GHz (just an example), α was computed to be 1.48 E+03 Np/m. Using (5), the range of existence of fields necessary for 3D EM field analysis was defined. Within this range, 12 data points, corresponding to 12 different lengths were used to perform 3D EM field computations using HFSS (a full-wave EM field solver that employs the finite element method to solve Maxwell's equations). At the end of each computation, an S-matrix was extracted. $|S_{11}|$ at 30 GHz from each matrix was plotted

against the corresponding length and the graph shown in Figure 3 was obtained. As expected, $|S_{11}|$ decreases exponentially with distance away from the via pad. Therefore, according to (7),

$$\text{since } |S_{11}|_{l=\frac{14}{5\alpha}=1892\mu\text{m}} = 0.323$$

$$\text{and } |S_{11}|_{l+\Delta l\left(\frac{3}{5\alpha}\right)=\frac{17}{5\alpha}=2297\mu\text{m}} = 0.322 = \text{Constant}$$

$$\text{then } |S_{11}|_{l=\frac{14}{5\alpha}=1892\mu\text{m}} - |S_{11}|_{l+\Delta l\left(\frac{3}{5\alpha}\right)=\frac{17}{5\alpha}=2297\mu\text{m}} \approx 0,$$

Signifying that $l = l_{bod}$.

\Rightarrow The exact boundaries of the via are at a distance of

$$\frac{14}{5\alpha} = 1892\mu\text{m} \text{ away from the via pad.}$$

Considering the 5% limit:

$$\text{Since } |S_{11}|_{l=\frac{8}{5\alpha}=1081\mu\text{m}} = 0.334,$$

$$|S_{11}|_{l+\Delta l\left(\frac{9}{5\alpha}\right)=\frac{17}{5\alpha}=2297\mu\text{m}} = 0.322 = \text{Constant}$$

$$\text{and } |S_{11}|_{l=\frac{8}{5\alpha}=1081\mu\text{m}} - |S_{11}|_{l+\Delta l\left(\frac{9}{5\alpha}\right)=\frac{17}{5\alpha}=2297\mu\text{m}} = 12,$$

$$\text{then } \frac{12}{|S_{11}|_{l+\Delta l\left(\frac{9}{5\alpha}\right)=\frac{17}{5\alpha}=2297\mu\text{m}}} \approx 0.0373 < 0.05$$

Signifying that $l = l_{bod}$

\Rightarrow Considering the 5% limit, the boundaries of the via are

$$\text{at a distance of } \frac{8}{5\alpha} = 1081\mu\text{m} \text{ away from the via pad.}$$

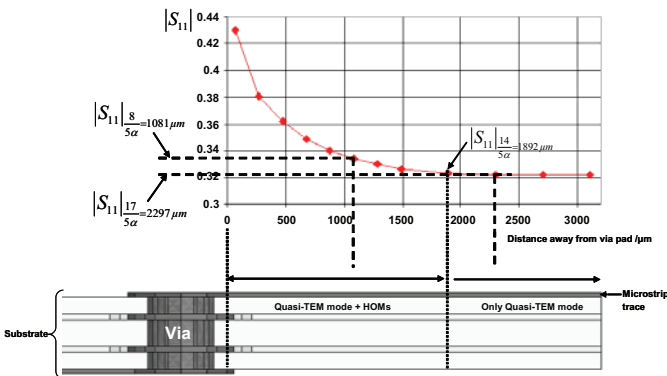


Figure 3: Variation of $|S_{11}|$ at 30 GHz with distance away from via pad.

V. EXPERIMENTAL VALIDATION

The proposed methodology was used to define boundaries of all geometrical discontinuities in a ball grid array package module. Test samples of these discontinuities were then designed on an FR4 substrate and measured using a vector network analyzer. Prior to the measurement, the LRRM method was used for the calibration. As can be seen in Figure 4, a good correlation was obtained between

measurement and simulation results, thus validating the proposed methodology. In this case, double bends on a package trace were considered as an example.

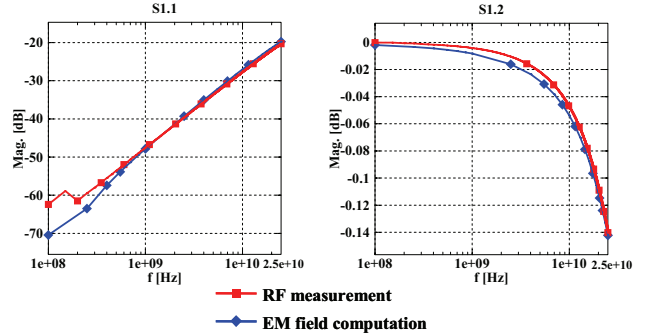


Figure 4: Comparison of measurement and EM field computation results for a microstrip trace with double bends.

VI. SUMMARY AND CONCLUSION

In this work, a novel methodology for defining the boundaries of discontinuities in electronic packages and boards was developed, illustrated and experimentally validated. With the help of this methodology, wideband models that account for the parasitic effects of all interconnecting components along complete signal paths in system-in-package modules were developed. These models were then used to track-down and solve signal integrity problems, before they even occur in the actual design.

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