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Comparison of S-Parameter Concatenation to Full-Wave Simulation for High-Speed Interconnect Analysis

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Abstract

A standard method of high-speed digital channel simulation involves dividing the channel into discrete parts and analyzing each part independently to determine its individual S-parameters. These S-parameters are then combined or concatenated to determine the overall channel performance. Analyzed here is a portion of such a channel composed of a high-speed connector and it's footprint on a printed circuit board. For this structure, the paper compares the concatenation method to a single full-wave simulation of the entire structure. The first method is evaluated to determine its upper frequency limit thus showing when it is appropriate to use it over full-wave analysis, which is typically more complex and time consuming. Additionally, some structures that are not accounted for when using the concatenation method are investigated.

Authors' Biographies

Vittal Balasubramanian received his BS in Electrical Engineering from Delhi College of Engineering, University of Delhi in 2001 and his MS in Electrical Engineering from Penn State University in 2005. He was awarded the Doris Hughes Memorial Award at Penn State in recognition of his outstanding academic achievements and contribution to the college community. He has been working at FCI USA, Inc. since Jan 2005 where his responsibilities include the design and analysis of high-speed connectors. He also worked for Sapient Corporation Pvt. Ltd. from August 2001 to December 2002 as a technology associate and helped with software consulting. He is a member of the Institute of Electrical and Electronics Engineers since 1998.

Stephen B. Smith's current responsibilities include the design and analysis of high-speed connectors. He also engages in customer support activities working very closely with Marketing and Sales. Prior to coming to FCI in 2000, Smith worked at AMP Incorporated (now Tyco Electronics) for 10 years as a development engineer spending most of that time in the electromagnetics research group where he developed methods of modeling and simulating interconnection systems on projects spanning the frequency spectrum from power-frequency high-current utility connectors to high-speed and R.F. interconnects. Prior to that, he worked in acoustical research at Masland Industries (now Lear Corporation.) He has taught various courses at each of his jobs, and in the last couple of years, he has presented papers at the 2003 Conference on Information Science and Systems at The Johns Hopkins University and at the IEEE Holm Conference. At DesignCon 2005, he acted on behalf of the entire design team when he accepted the first annual DesignVision Award for the AirMax VS® Connector System. Smith has a B.S. in physics and an M.S.E.E., both from Penn State University.

Sedig S. Agili received his BS, MS, and Ph.D. in Electrical and Computer Engineering from Marquette University in 1986, 1989, and 1996, respectively. As a student, he was awarded fellowships from Marquette University and the U.S. Department of Education. Upon receiving his Ph.D., he joined the faculty at Marquette University where he taught several courses in electrical engineering and conducted research in the area of electro-optic devices, fiber optic communication and fiber optic sensors. In fall of 2001, he joined the electrical engineering and electrical engineering technology programs at Penn State University, Capital College. Currently he is teaching and conducting research in electronic communications, fiber optic communications, fiber optic sensors and signal processing. He has authored numerous articles published in journals and conference proceedings, and made presentations at many conferences and seminars. He also worked for Astronaut Corporation of America in Milwaukee, Wisconsin where he was involved in designing optical projection and heads-up display systems. He is a member of the Institute of Electrical and Electronic Engineers, American Society for Engineering Education, and Sigma Xi.

1. Introduction

The past several years have seen significant new developments in the design of connectors intended for high-speed transmission. Attributes of these new connectors include excellent impedance match, very low insertion loss, and very low crosstalk. The advent of these new connectors has facilitated the transmission of data at speeds exceeding 10 Gbit/s. At such high speeds, the vias in the connector footprint typically present the worst impedance discontinuity in the entire channel [1]. Hence, the footprints as well as the connector must be analyzed.

Treating the footprints and the connector as separate components permits the use of the concatenation method of simulation wherein the S-parameters of each component are separately calculated and then combined [2, 3] to give the overall S-parameters. Hence, it is only necessary to analyze the connector once, and its S-parameters can then be used in the simulation of the connector with various footprints. This is advantageous, because, in contrast to that of a footprint, the geometrical complexity of a connector typically makes it much more difficult to compute its S-parameters.

Concatenating S-parameters, although efficient, relies on the assumption that there are no non-transverse electromagnetic (non-TEM) interactions occurring at the interface between the connector and its footprints. Neglecting non-TEM effects is a reasonable practice when considering signals of low frequency. As the frequency increases however, these effects become increasingly important and perhaps, dominant. When this happens, the concatenation method would no longer work, and it would become necessary to use a full-wave solver to obtain an accurate characterization of the connector, footprints, and the interactions between them. A full-wave solution involves the analysis of a 3-D model of the footprints attached to the connector, and captures all of the three-dimensional field effects. Although the accuracy of full-wave solutions is desirable, it does involve considerable complexity and time. In addition, the entire connector must be included when solving each footprint. Hence, the concatenation method is frequently more practical and preferred, and it would be useful to find its limitations, in accuracy and frequency.

This paper compares the concatenation method to a full-wave solution of a connector and footprint combination in order to determine its accuracy and determine its upper frequency limit. Section 2 gives details of the analyses and compares the insertion loss, return loss, and differential impedance obtained using each method. Section 3 considers examples of structures typically ignored when using the concatenation method and determines if such assumptions are valid. Section 4 gives the conclusions.

2. Modeling of footprints and connector

To compare the concatenation method with modelling the entire structure using full-wave methods, it was necessary to obtain accurate S-parameters for each of the components so that the accuracy of the methods would be in question rather than the quality of the S-parameters. Hence, for consistency, all of the S-parameters used in the comparisons were obtained from full-wave solutions using the finite-element method (FEM) as implemented within Ansoft HFSS. Three structures were modelled: a footprint consisting of short traces connecting to vias, a connector-like structure, and a combination consisting of the connector-like structure with a footprint on each end.

2.1 Footprint analysis

The 3-D model of the footprint, shown in Figure 1, consisted of a single differential pair of two signal and two ground vias. The pc board had a thickness of 3.4104 mm and consisted of 10 layers, four of which were signal layers. The traces were routed on layer four (from the top) and had a width of 0.2032 mm (8 mils) and a separation of 0.2032 mm (8 mils). They also possessed a small amount of in-pair skew as might occur when routing to connectors whose differential pairs are arranged within a column. Hence, one trace had a length of 4.6791 mm and the other, a length of 3.1107 mm. For the initial simulation, press-fit pins of length 1.6 mm were included inside the vias. The vias themselves had a drill-hole diameter of 0.6 mm, a finished-hole diameter of 0.5 mm, pads of diameter 0.9 mm, and antipads of diameter 1.3 mm. The vias were spaced on 1.4 mm centerlines which is consistent with the footprint of a widely used backplane connector. The vias were backdrilled to a depth of 1.8279 mm thus allowing sufficient via barrel to support the press-fit pins. The board material was FR4 with a relative dielectric constant of 4. For this problem as well as all subsequent full-wave solutions, the geometry was meshed optimally for a frequency of 20 GHz, and an interpolating frequency sweep was performed from 50 MHz to 20.05 GHz. The problem was solved to determine the S-parameters of the two terminal or single-ended modes which were subsequently used to compute the S-parameters of the differential mode using equations found in [4.]

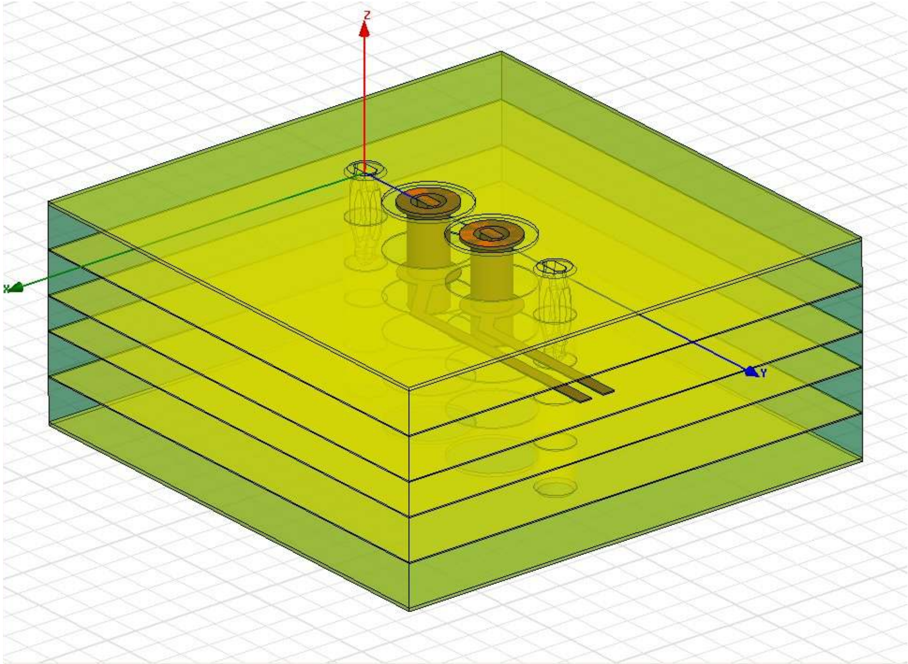


Figure 1: 3-D model of the footprint

In order to obtain reliable results from the full-wave solution, it was necessary to push the convergence by requiring a maximum delta S of 0.005 per adaptive pass as opposed to the default value of 0.02.

2.2 Connector analysis

Since the concatenation method is frequently used to analyze communication links, we chose to analyze a right-angled connector-like structure similar to those used on backplane systems. To simplify the analysis, the structure lacked a separable interface. The 3-D connector-like model, shown in Figure 2, consisted of two signal and two ground conductors forming a single differential pair. In the absence of ground planes or shields, the conductors were edge-coupled as in the AirMax VS® connector from FCI. As in the aforementioned connector, the conductors were spaced on centerlines of 1.4 mm. Each of the conductors was 0.35 mm thick and 0.6 mm wide and encapsulated in plastic with a relative dielectric constant of 3.5 to obtain a differential impedance of approximately 100 Ω.

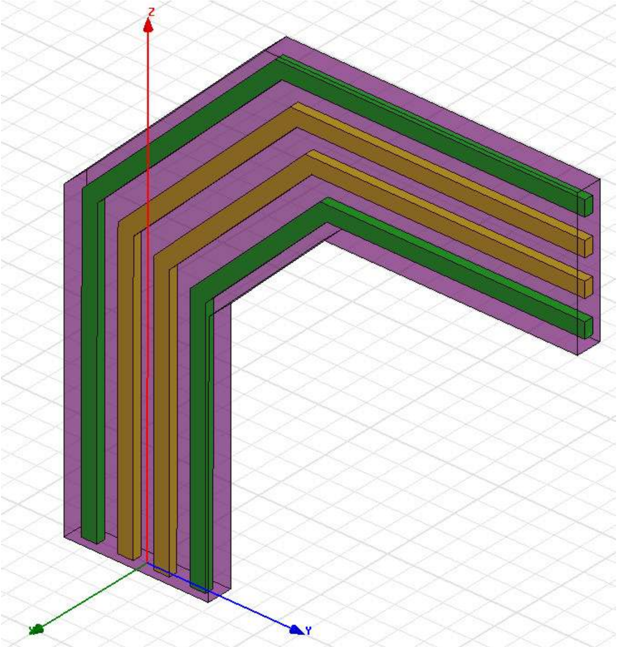


Figure 2: 3-D model of the connector-like structure

The structure was placed in an air box with absorbing boundary condition on all its faces. The simulation setup and convergence criteria were the same as those described in sub-section 2.1. In the following sub-section, the S-parameters obtained from this full-wave analysis were combined with those of two footprints, one on each end of the connector.

2.3 Concatenation of footprint and connector S-parameters

The process of joining multiple S-parameters together to obtain the S-parameters of any combination of S-parameters is known as concatenation, and is performed in MATLAB®. Application of this process to footprint-connector-footprint combination yields its S-parameters. To do this, the MATLAB® script uses an S-to-T matrix transformation [5] method as follows

$$S_{footprint} \rightarrow T_{footprint}$$

$$S_{connector} \rightarrow T_{connector}$$

where $S_{footprint}$ and $S_{connector}$ are the scattering matrices of the footprint and the connector respectively. The scattering matrices are transformed to the transfer scattering matrices $T_{footprint}$ and $T_{connector}$, and the concatenated T-parameters of the footprint-connector-footprint can then be represented as

$$T_{concat} = T_{footprint} T_{connector} T_{footprint} \, .$$

Finally, the scattering matrix of the footprint-connector-footprint combination, S_{concat}, is obtained as $T_{concat} \rightarrow S_{concat}$ using T-to-S matrix conversion [5]. The S-parameters of this combination are compared directly to the S-parameters of a complete full-wave model including two footprints as described in sub-section 2.4.

2.4 Comparison of S-parameter concatenation to full-wave simulation

To solve the combined footprint and connector-like structure using the full-wave solver, the 3-D models of sub-sections 2.1 and 2.2 were combined. As can be seen in Figure 3, the footprint was attached to each end of the connector-like structure and the entirety solved at once. The details of the problem set-up were identical to those of sub-section 2.1 and 2.2.

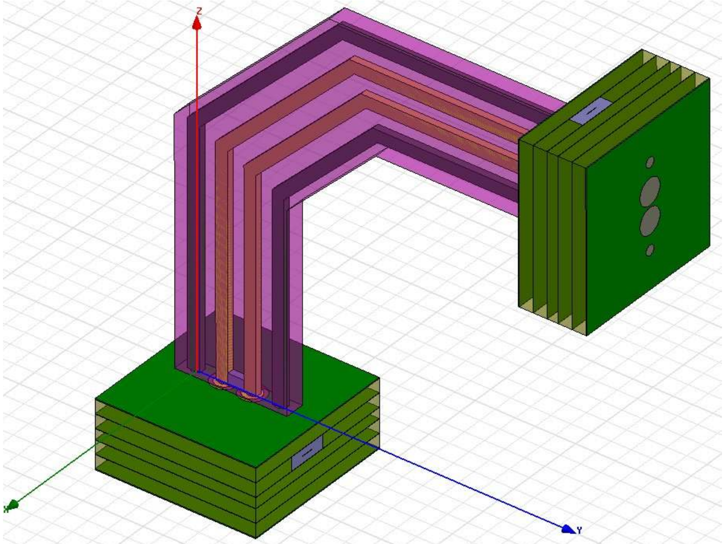
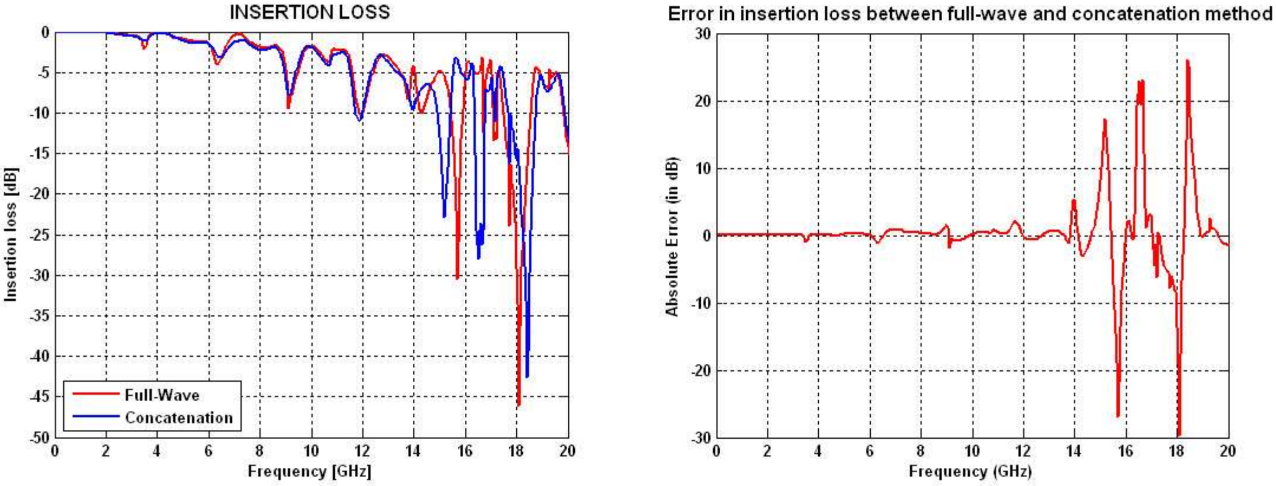


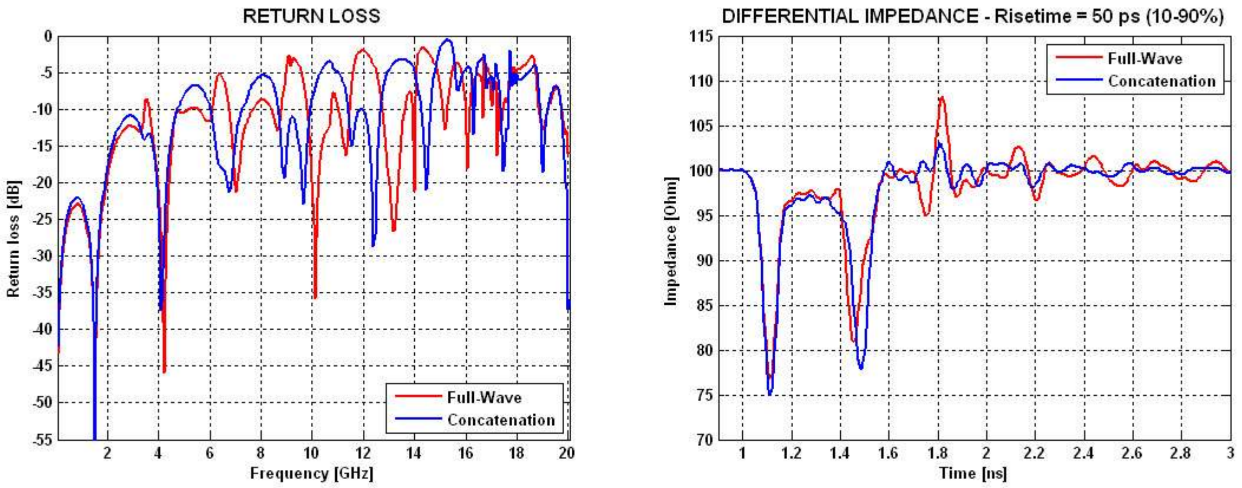
Figure 3: 3-D model of footprint-connector-footprint combination

Figures 4(a) shows the insertion loss obtained from the concatenated and the complete full-wave solutions plotted on the same graph. Figure 4(b) shows the results of subtracting the two different plots of insertion loss from Figure 4(a) and the two methods show excellent agreement within ±1 dB up to a frequency of 9 GHz and ±2 dB up to 14 GHz. The return loss shows good agreement up to about 5 GHz as shown in Figure 4(c). The two impedance profiles in Figure 4(d) are in close agreement up to about 1.6 ns which is where the second footprint in the footprint-connector-footprint combination ends. Interestingly, the impedance profile obtained from the full-wave analysis shows significantly more ringing than that obtained from the concatenation method and could contribute to the difference observed in the return losses.



a) Insertion loss

b) Error in insertion loss



c) Return loss d) Differential impedance

Figure 4: Comparative analysis of full-wave analysis and concatenation method

Return loss might be of interest when considering bidirectional channels. Most high-speed serial channels, however, are unidirectional, and in such cases, the importance of return loss is significantly reduced. Because of reflection artifacts in simulation, return loss is often difficult to correlate. Moreover, return loss can be notoriously difficult to measure repeatably, and so its impact on system analysis is not necessarily clear. In all channels, however, insertion loss is of primary concern. Given that the insertion losses of the two methods agree to within ±1 dB up to 9 GHz and ±2 dB up to 14 GHz, it follows that the concatenation method should be more than adequate for the characterization of systems at 10 Gbps (i.e. 5 GHz). Furthermore, if any non-TEM effects are occurring at the interface between the footprint and the connector, they clearly are not detrimentally affecting the performance up to 14 GHz. Beyond 14 GHz, the disparity between the two methods could be caused by non-TEM effects, inaccuracies in modeling, lack of adequate convergence, or any combination of these or other effects. In defense of the models, it should be pointed out that measurement of structures such as the ones considered here is extremely difficult at frequencies above 10 GHz.

While it was fairly straight forward to implement the concatenation method for the connector and footprint just described, there are other structures for which it poses a challenge. The following section presents some examples of such structures.

3. Examples of structures not accounted for in the concatenation method

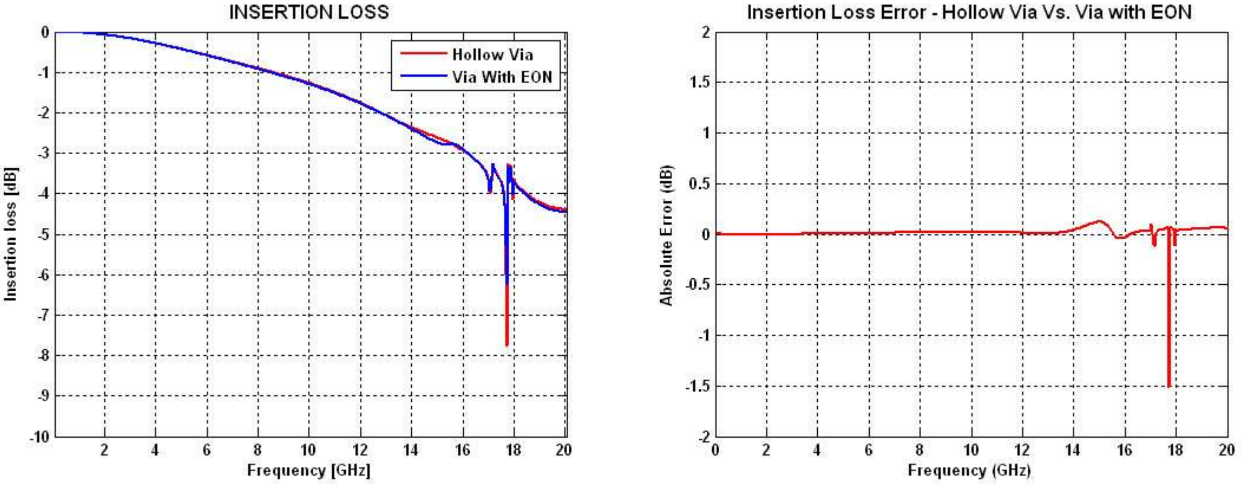
When using the concatenation method, it is necessary to select or sometimes create well-defined reference planes for the S-parameters of each individual component. Occasionally a question arises concerning small geometrical features and on which side of the reference plane they are located. For example, if one considers the top surface of a PCB to be a reference plane, it relegates the press-fit pins of a connector to be on the PCB (or footprint) side rather than the connector side of the reference plane. Frequently, after this is done, the presence of the press-fit pins is ignored, and the vias are modeled as either hollow or solid cylinders. The assumption, of course, is that the electromagnetic fields reside entirely outside of these cylinders and that the electrical performance of the footprint is entirely determined by the size and separation of the vias, and not by the press-fit pins. Any lack of accuracy resulting from the absence of the press-fit pins would represent a failing not of the concatenation method per se, but rather an inability to obtain accurate S-parameters to implement using the method. Such situations might still require the use of a full-wave analysis of the entire structure thus eliminating the ambiguity in the location of the reference plane.

In addition to the afore-mentioned press-fit pin geometry, the reference plane location problem appears in other connector footprint structures as does the possibility of non-TEM behavior. Several such structures are analyzed in this section to determine whether they possess sufficient field disruption to warrant the use of full-wave analysis or whether the disruption is inconsequential at the frequencies of interest. The results of the analyses of each of these example structures are described in the following sub-sections.

3.1. Point of contact between the press-fit pin and via barrel

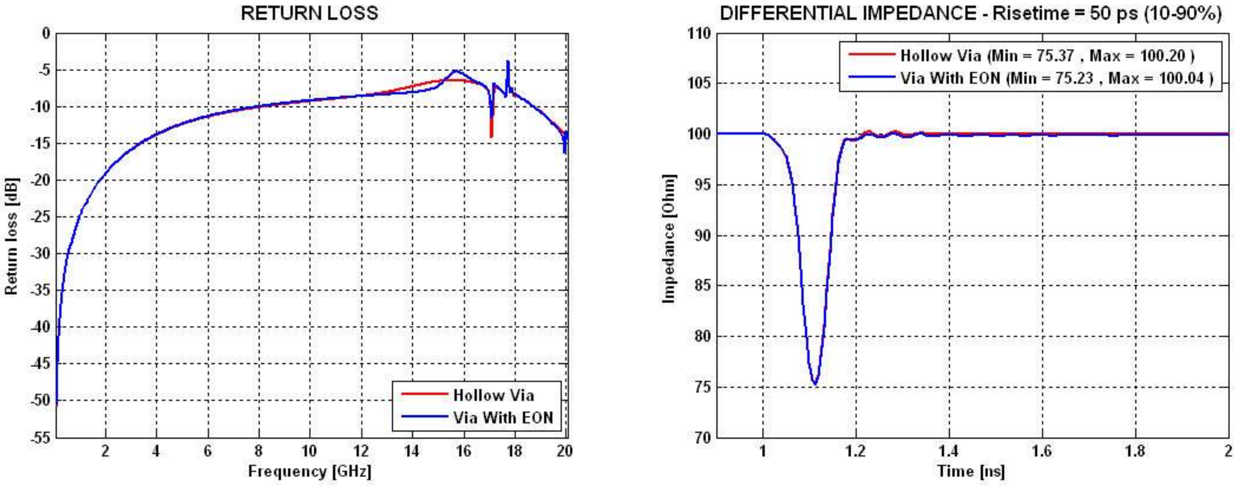
One example of the reference plane ambiguity problem would be the inclusion of the points of contact between an eye-of-the-needle (EON) press-fit pin and the inside of a via. The analysis of this particular structure answered the question of whether or not it is appropriate to simply analyze vias as if they were empty annular cylinders rather than hollow cylinders with press-fit pins inside. The 3-D models used for this analysis were similar to that shown in Figure 1, where, the EON was included in one model and not included in the other. As a result, the point of contact for launching the signal into the board was different. In the case where the EON was included in the model, the launch point was at the location where the EON first made contact with the inside of the via barrel (0.255 mm below the surface of the board). In the case without the EON, the signal was assumed to be launched at the via pad on the top layer of the board.

Figure 5 compares the insertion loss, return loss, and differential impedance for the cases with and without the EON in the via barrel. Clearly, Figure 5(b) shows negligible difference between the insertion losses of the two cases. That coupled with the similarities in the return loss and differential impedance as shown in Figures 5(c) and 5(d), respectively, suggests that it is not necessary to include the EON in footprint simulations. This has the potential to significantly reduce both the model complexity and solution time. Also the ambiguity in which side of the reference plane to place the EON goes away and there is no problem implementing the concatenation method with press-fit connectors.



a) Insertion loss

b) Error in insertion loss



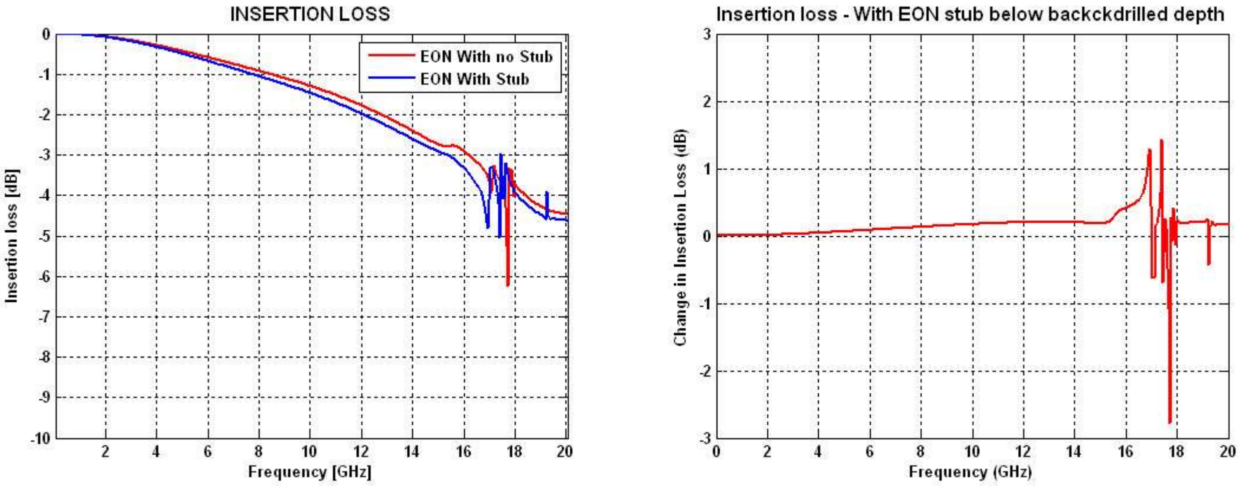
c) Return loss d) Differential impedance

Figure 5: Effect of point of contact between EON and via barrel

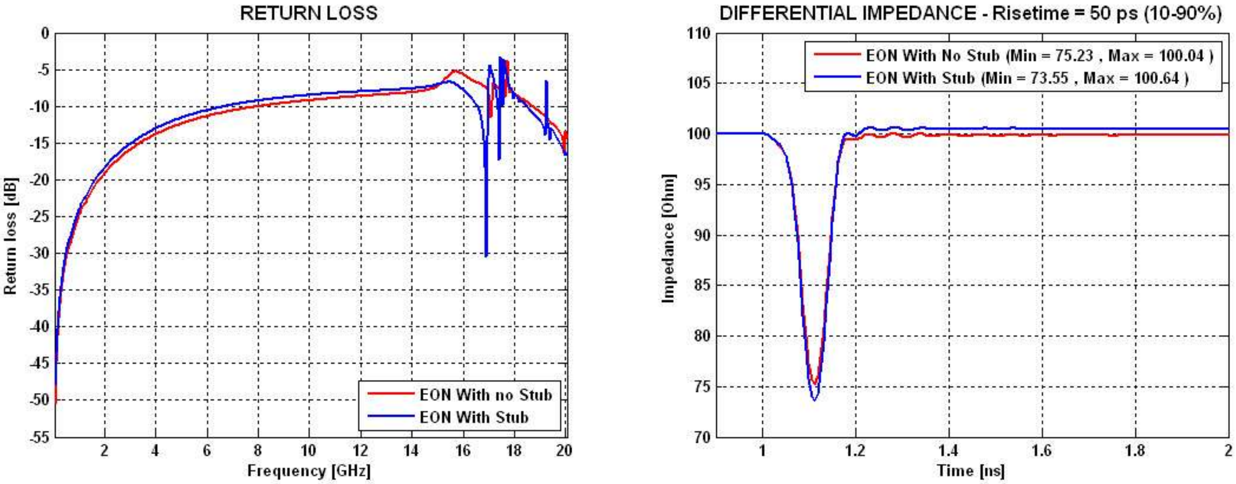
3.2. EON stub below backdrilled depth

Another situation that can present ambiguity in assigning features to a particular side of the reference plane is a press-fit pin that extends below the via into the backdrilled region of the board. The model of Figure 1 was solved again with the EON extending below the backdrilled depth by 0.92 mm. Figure 6 shows the difference in performance before and after extending the EON.

The minute performance change, insertion loss less than 0.25 dB and return loss less than 1 dB up to 15 GHz, can be explained by Figure 6(d), which shows that the longer EON exposes the signal to a slightly higher effective capacitance as would be expected. As in sub-section 3.1, the concatenation method is suitable despite ignoring the extra length of a press-fit pin.



a) Insertion loss b) Change in insertion loss

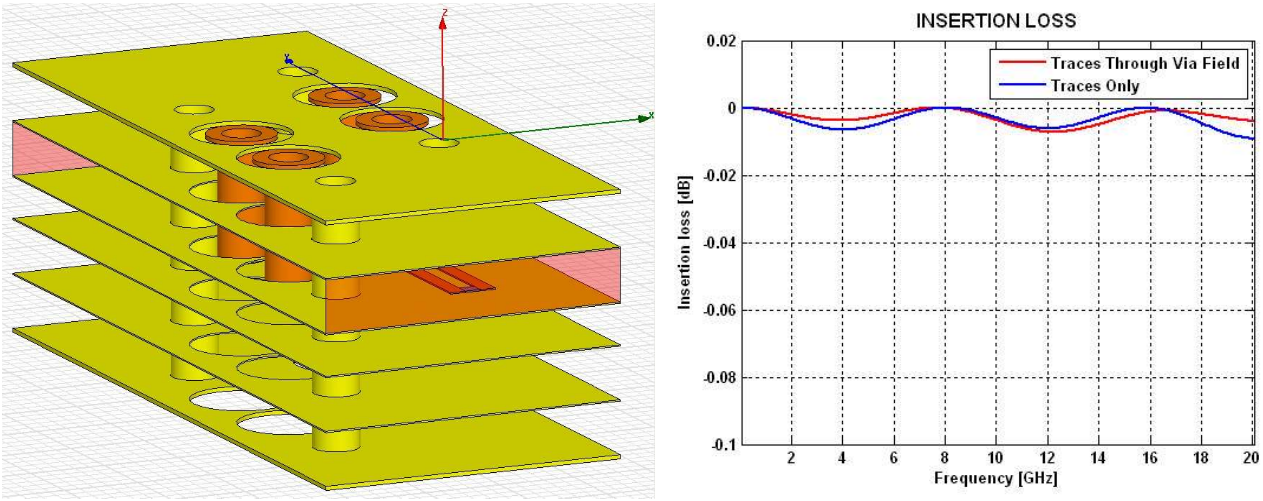


c) Return loss d) Differential impedance

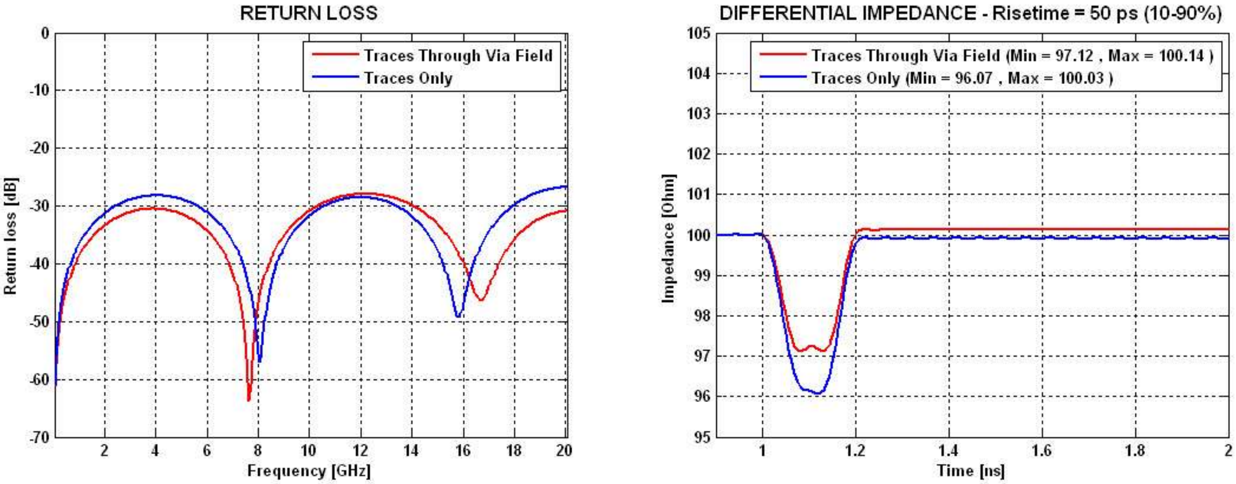
Figure 6: Effect of EON stub below backdrilled depth

3.3. Routing through a footprint

When using the concatenation method to characterize a channel, the S-parameters for lossy traces are typically obtained by modeling the thickness, width, separation, and length of the traces on a lossy substrate. The assumption is usually made that the traces pass through a homogeneous environment without encountering any irregularities or obstacles along the way. In reality, the traces usually bend and pass between geometrical features such as vias. Figure 7 shows the modeling and characterization of a differential pair of traces traveling between two sets of differential vias such as those encountered in a connector footprint.



a) 3D Model b) Insertion loss



c) Return Loss

d) Differential impedance

Figure 7: Effect of routing traces through a footprint

Because it appears in both models, the periodicity in the insertion and return losses is entirely a function of the electrical length of the traces and is not caused by the presence of the vias. Interestingly, routing through the via field slightly raises the impedance of the traces. The antipads seem to exert more influence than the vias. Since the antipads reduce the capacitance of the traces, they slightly raise the impedance. In this case, since the traces had an impedance lower than 100 ohms, the effects of the antipad slightly improve the impedance. Had the traces possessed an impedance of or higher than 100 ohms, the antipads, likely would have had a slightly detrimental effect on the return loss. Given that the change in the insertion loss is minute, this effect is probably negligible and ignoring it while using concatenation would be acceptable.

3.4. Trace-to-via crosstalk

Since the concatenation method ignores the effects of routing through a via field as discussed in sub-section 3.3, it logically follows that it would also ignore any crosstalk that might exist between traces and vias. Figure 8 shows differential crosstalk from the traces to the via as determined from the model in Figure 7(a).

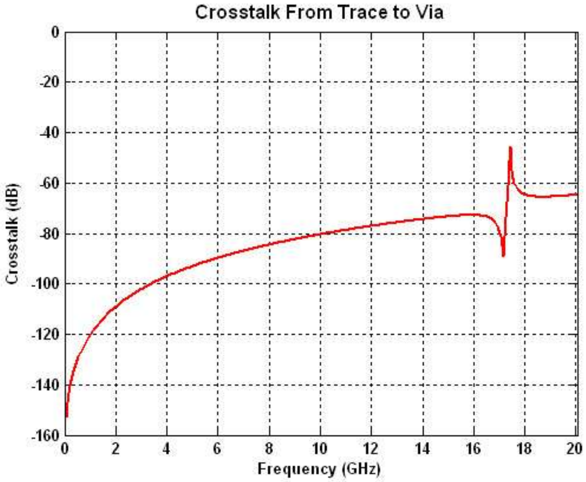
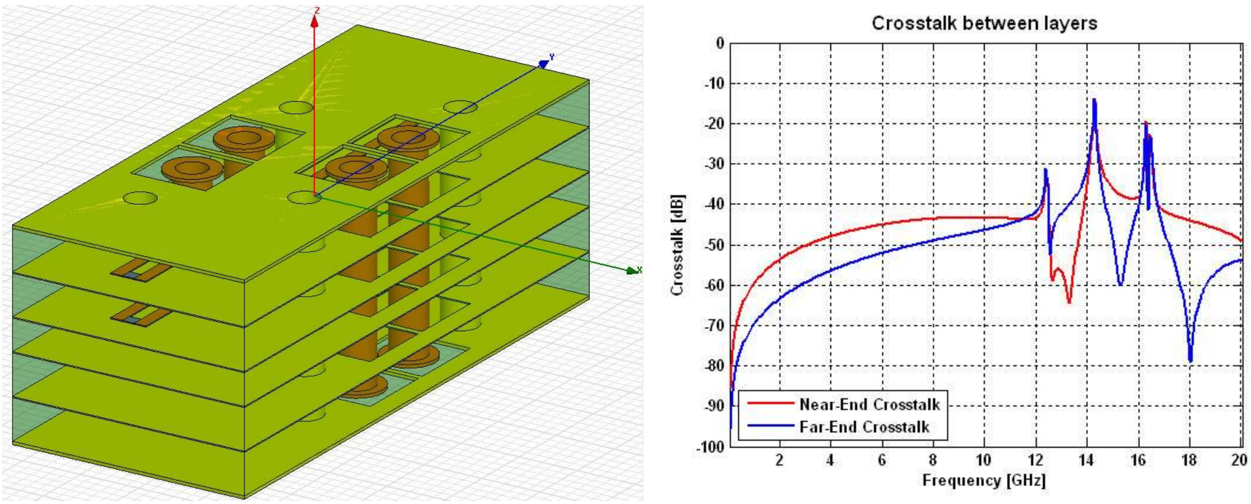


Figure 8: Differential crosstalk from trace to via

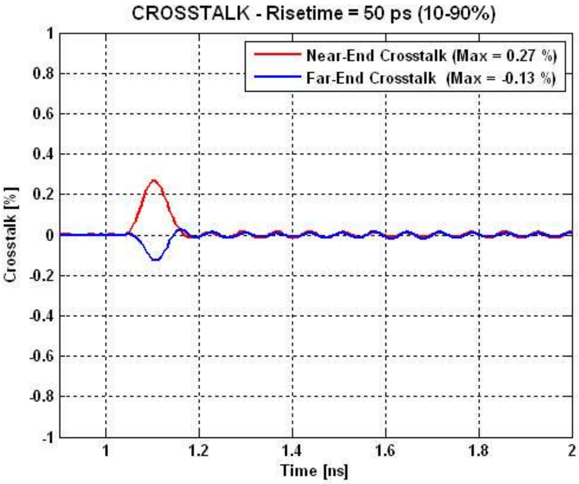
Across the frequency range, the figure shows that the crosstalk is negligible. This justifies ignoring trace-to-via crosstalk when using concatenation.

3.5. Trace crosstalk between board layers

Another effect ignored not only when using the concatenation method, but also when designing boards in general, is the possible existence of crosstalk between traces on different board layers. The model in Figure 9(a) was used to compute the crosstalk between traces on layers 2 and 4. Figures 9(b) and (c) show the near-end and far-end differential crosstalk in the frequency and time domains respectively.



a) Model used to compute crosstalk between layers b) Crosstalk between layers in frequency domain



c) Crosstalk between layers in time domain

Figure 9: Trace crosstalk between board layers

This crosstalk exists because of the presence of antipads along the transmission path. Initially, in Figure 9(a), the traces were flush with the edge of the antipads. One might do this when routing more than one differential pair on the same layer and in the same space between two sets of vias. The simulation was run a second time with the traces centered in the space between the vias and hence, not flush with the edge of the antipads. The crosstalk under this situation was essentially the same as that shown in Figures 9(b) and (c). Hence, there is not a penalty to pay in layer-to-layer crosstalk when routing traces flush with the edge of antipads.

In practical situations, the crosstalk is likely to be lower than that shown in Figure 9(c) because the signal risetime will probably have degraded to slower than 50 ps (10-90%) before encountering the antipads as the signal will have traversed through lossy trace and/or a connector. Hence, this effect can be neglected in practical situations and ignoring it while using concatenation would be acceptable.

4. Conclusions

This paper presented the analysis of a portion of a channel composed of a high-speed connector and it's footprint on a printed circuit board using two methods: the concatenation method and a single full-wave simulation of the entire structure. Comparison of the results obtained from the two methods showed excellent agreement in the insertion loss to within ±1 dB up to a frequency of 9 GHz and ±2 dB up to 14 GHz. Additionally, the differential impedance agreed to within 3 Ω over the entire electrical length of the structure. The similar results between the two methods coupled with the relative simplicity of the concatenation method suggested that it might be preferable to use the concatenation method for the characterization of systems up to and beyond 10 Gbps (i.e. 5 GHz.)

After the analysis of the connector and footprints, additional analyses were performed on other structures generally ignored when using the concatenation method. The results of each of these analyses showed that they have little impact on the accuracy of the concatenation method.

5. References

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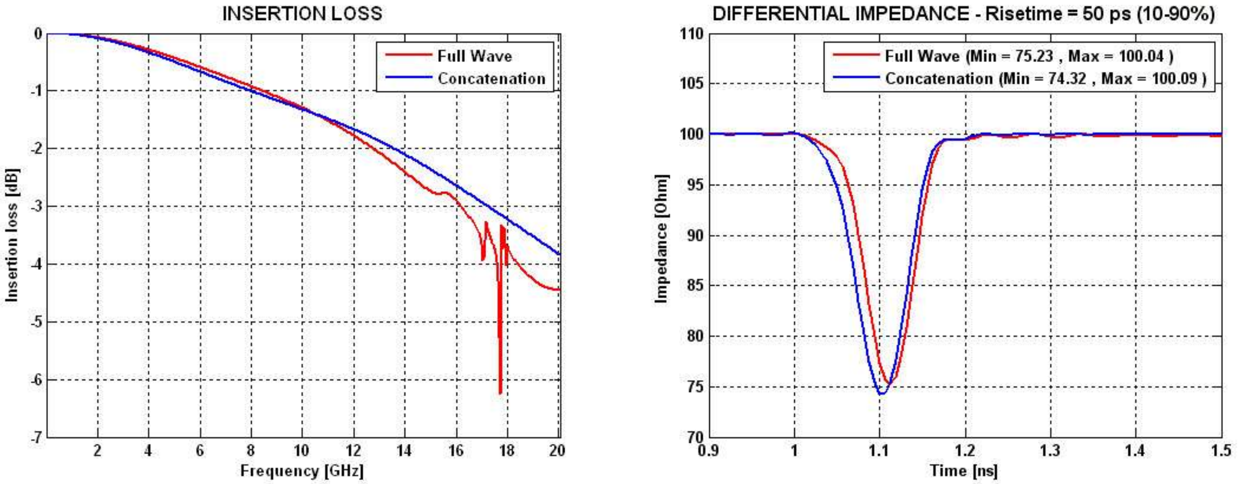
Appendix

Since the full-wave solution of the footprint in sub-section 2.1 involved significant complexity and time, the same analysis was attempted using a simpler proprietary quasi-static solver designed exclusively to analyze vias [6, 7]. The quasi-statically-derived via S-parameters were concatenated to those of traces for comparison to the full-wave solution. The results for insertion loss and impedance shown in Figures 10(a) and 10(b) respectively illustrated the adequacy of using a simpler quasi-static analysis to characterize vias.



a) Insertion Loss

b) Differential Impedance Comparison

Figure 10: Full-wave vs. quasi-static analysis

The insertion loss shows excellent agreement up to about 11 GHz and the impedances agree to within 1 Ω. Future use of a quasi-static solver could result in reduced complexity and shorter solve times without compromising accuracy.