# Building interconnect models with analysis to measurement correlation up to 50 GHz and beyond

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**Abstract:** Four essential elements of electromagnetic signal integrity analysis that guarantee successful design of PCB and packaging interconnects up to 50 GHz and beyond are introduced in the paper. Bandwidth and quality of S-parameter models, broadband material characterization and identification, localization of all elements of a channel and systematic benchmarking process are described in the paper as the elements of design flow that lead to success. Neglecting or missing even one of the elements may compromise the whole project.

**Introduction:** Faster data rates drive the need for accurate models for data channels and specifically for PCB and packaging interconnects. 10 gigabit Ethernet is practically the mainstream now and 25 gigabit is coming out. Spectrum of signals in such channels ranges from DC or MHz frequencies up to 20-50 GHz and beyond (into centimeter and millimeter wavelengths) and imposes very special requirements on the interconnect modeling and design. No models or simplified models may result in complete failure of such channels and require multiple iterations to fix and may be not possible at all. What is the best way to model such high-speed interconnects? It obviously depends on a problem to solve. For the signal integrity analysis, interconnects can be formally divided into transmission line segments (planar or cables) and discontinuities or transitions in lines such as via-holes and connectors as schematically shown in Fig. 1.

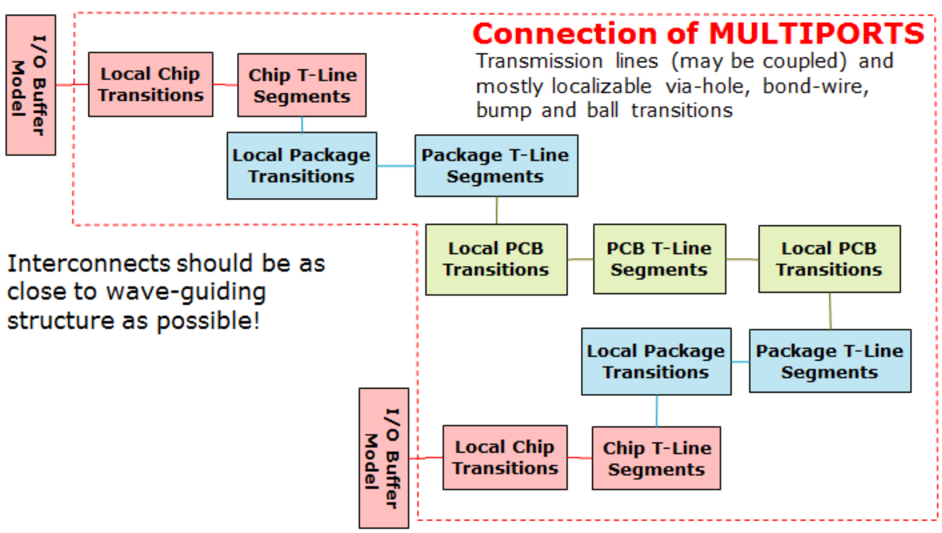


*Fig. 1. Interconnect model as connection of multiports. In general, the goal of interconnect design is to make it as close to a localized wave-guiding structure as possible and thus predictable with the analysis.*

Multiport models of components are built separately with static or electromagnetic analysis, measurements or obtained from component vendors and then united into a complete channel model. This technique was developed from microwave application and is known as decompositional electromagnetic analysis (also known as divide and concur or segmentation technique). It is also widely used in signal integrity analysis tools for digital applications. Though, the limitations of this technique and key elements that lead to success in case of digital interconnects is a subject of ongoing research (see latest DesignCon papers for instance). Interconnects typically require analysis over much larger frequency band and may contain components that have not being used in microwave application. This paper explains four essential elements of the de-compositional electromagnetic signal integrity analysis that guarantee analysis to measurement correlation up to 50 GHz and beyond:

*1) Quality of S-parameter models of interconnects (bandwidth, sampling, passivity and causality);*
*2) Broadband dielectric and conductor roughness models (important for analysis of transmission lines);*
*3) Localization property and de-embedding of discontinuities (possibility to be analyzed in isolation);*
*4) Procedure to validate models with measurements on a set of standard test structures (benchmarking).*

**Quality of S-parameter models of interconnects:** Any element of a linear time-invariant data channel can be modeled as a multiport described with S-parameter models [1], [2]. Multiport is a natural and scalable black-box description of linear structures smaller, comparable with or larger than wavelength. In decompositional analysis multiport parameters of transmission lines, via-holes and other components are united and then simulated with models of transmitter and receiver as schematically shown in Fig. 1. Multiports are often described with S-parameter models produced by circuit and electromagnetic simulators, VNAs and TDNAs. Very often such models have issues and may be not suitable for consistent broadband frequency and time domain and compliance analyses of interconnects. **Analysis to measurement correspondence is possible only with multiport models that have sufficient bandwidth and pass quality control. This is one of the key elements in the design success.**

S-parameter models are usually band-limited due to the limited capabilities of solvers and measurement equipment. Model should include DC point or allow extrapolation, and high frequencies defined by the signal spectrum. If a model does not contain DC point, the lowest frequency in the sweep should be below the transition to skin-effect (1-50 MHz for PCB applications), or below the first possible resonance in the system defined as $f_l < \frac{c}{4L \cdot \sqrt{\varepsilon_{eff}}}$, to allow extrapolation to DC. Here L is total physical length of the system, c is speed of light and $\varepsilon_{eff}$ is effective dielectric constant. The highest frequency in the sweep must be defined by the required resolution in time-domain or by spectrum of the signal [4]. The highest frequency can be defined either with signal rise time tr as $f_h > \frac{1}{2t_r}$ or with the main harmonic fs1 as $f_h > K \cdot f_{s1}$. Here K may range from 2 to 5, depending on the actual attenuation in the channel. **All models for a channel interconnects should satisfy the target bandwidth requirement. Otherwise they have to be discarded and rebuilt.**

In addition to the band-limitedness, most of interconnect component models comes in form of Touchstone models [3]. Touchstone models are just S-parameters defined at a set of frequencies. Interpolation or approximation of tabulated matrix elements may be necessary both for time and frequency domain analyses. Appropriate sampling is very important for DFT and convolution-based time-domain analysis algorithms [4], but not so for algorithms based on rational approximation. In general, there must be 4-5 frequency point per each resonance. In addition, the electrical length of a system should not change more than quarter of wave-length between two consecutive frequency points $df < \frac{c}{4L \cdot \sqrt{\varepsilon_{eff}}}$. **Under-sampling is typically occurs at lower frequencies and may lead to defects both in frequency and time domain analyses. Such models have to be discarded and rebuilt.**

In addition to the band-limitedness and possible under-sampling, models can be distorted with the measurement or simulation artifacts that are not so easy to detect. For instance, model convergence issues, fast frequency sweeps, non-conservativeness of ports, un-accounted high-order modes in electromagnetic analysis may cause model distortions. Measurement noise, calibration and measurement equipment problems can also lead to defective models. In general, those are human mistakes of tools developers and users. **How to estimate quality of S-parameter models to make sure that the models are suitable for analysis?** S-parameters quality metrics have been recently introduced in [5], [6] to simplify the task. Metrics for passivity, reciprocity and causality computed for band-limited discrete models can be used for preliminary analysis of quality of S-parameters. The metrics range is from 0 to 100. Zero means bad, 100 is good. Ranges for acceptable and questionable models are defined on the base of analysis of thousands of models. Example of preliminary analysis of a set of models in Simbeor® Touchstone Analyzer™ tool [7] designed for automation of S-parameters quality assurance is shown on Fig. 2.

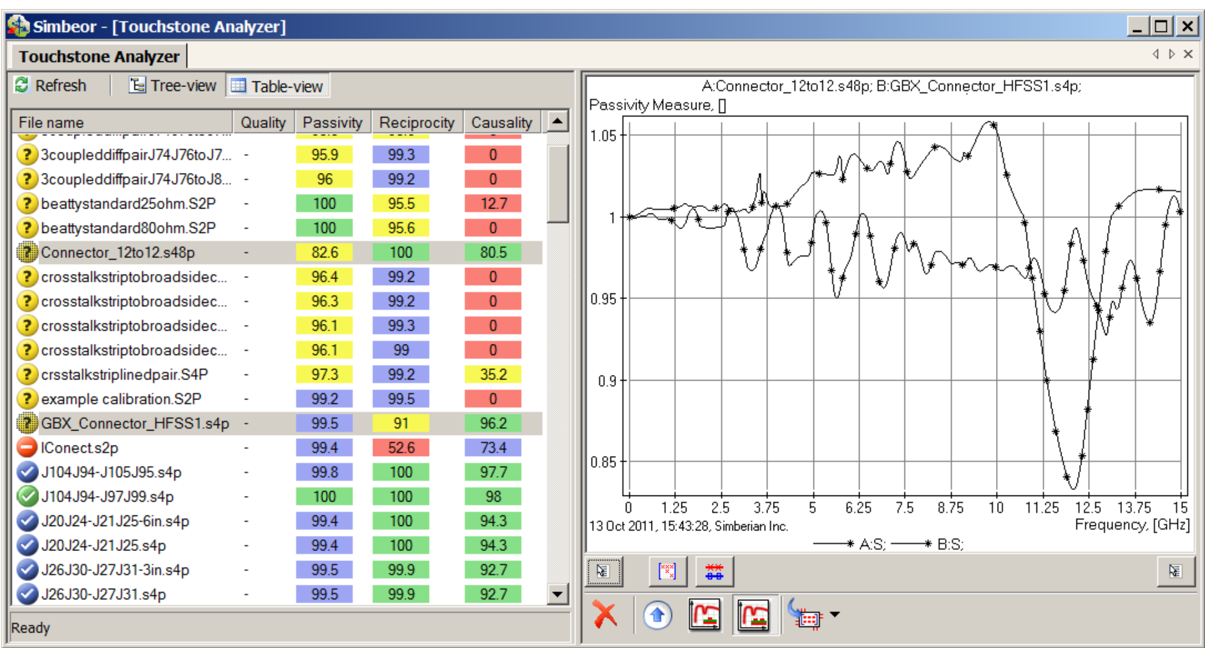


*Fig. 2. Example of preliminary analysis of Touchstone models quality in Simbeor Touchstone Analyzer [7]. Good metrics are shown in green; acceptable in blue; questionable in yellow; and bad is red. Graph on the right is adjusted for passivity measure for 2 selected models – passivity violated if the measure is above unit.*

Note that preliminary quality estimation is done for a discrete and band-limited data set and, thus, is incomplete. Though, it allows detecting unacceptable violation of passivity and reciprocity. **If passivity or reciprocity metrics are close to zero, the model has to be discarded and rebuilt.** Large violations of preliminary causality metric in computed models point at under-sampled data – such models have to be also rebuilt. Large causality violation in measured data may occur because of measurement noise and can be fixed with the rational approximation. A model ranked as good with preliminary metrics, may still have hidden defects and may not allow accurate interpolation or extrapolation for purpose of time-domain analysis for instance. Rigorous estimation of passivity and causality can be done only for a frequency-continuous model defined from DC to infinity. Such models can be built with the rational approximation of the original tabulate data. Rational macro-models are frequency-continuous models defined from DC to infinite frequency and can be used for the final quality estimation. **High-quality tabulated models can be accurately approximated with passive rational macro-models.** The final quality metric with the range from 0 to 100 can be constructed using the root-mean square error (RMSE) of the passive rational approximation as $Q = 100 \cdot \max\left(1 - RMSE, 0\right)\%$ [6]. S-parameters approximated with the rational functions are causal by definition in case if passivity is ensured from DC to infinite frequencies. Example of the final model quality estimation with the rational macro-models is shown in Fig. 3. **Low quality metric means that the tabulated model cannot be accurately interpolated and extrapolated with a causal passive model and, thus, has to be discarded.** After model quality is ensured, it can be further exported and used as broad-band SPICE macro-models (BB SPICE) or improved by re-sampling. Note that BB SPICE models are frequency-continuous and contain extrapolation to DC and infinity and, thus, guarantee consistent analyses both in frequency and time-domain in practically all tools.

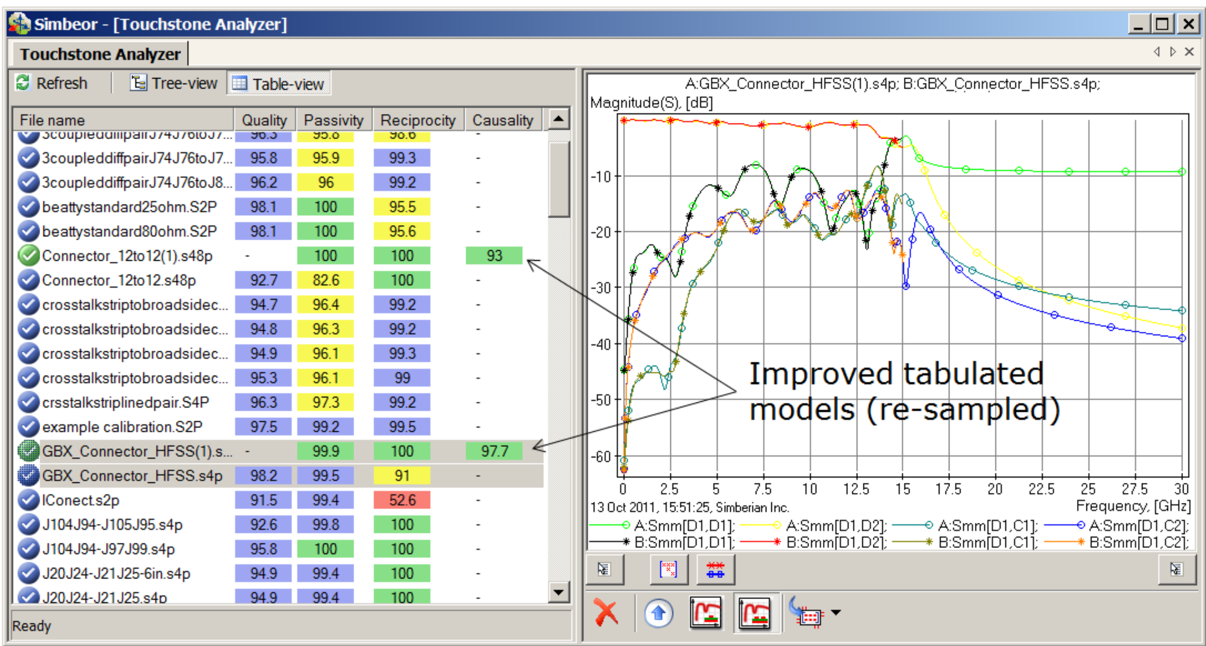


*Fig. 3. Example of the final model quality analysis in Simbeor Touchstone Analyzer [7]. Model icons and column Quality illustrate the model quality estimated with the passive rational approximation (good –green, blue - acceptable). Columns passivity and reciprocity show quality of the original tabulated data. Graph on the right is adjusted to compare magnitudes of the original and "improved" models.*

**Broadband dielectric and conductor roughness models:** The largest part of interconnects can be formally defined and simulated as transmission line segments. Models for transmission lines are usually constructed with a static or electromagnetic field solvers. Transmission lines with homogeneous dielectrics (strip lines) can be effectively analyzed with quasi-static field solvers and lines with inhomogeneous dielectric may require analysis with a full-wave solver to account for the high-frequency dispersion [8], [9]. Accuracy of transmission line models is mostly defined by availability of broadband dielectric and conductor roughness models. Wideband and multi-pole Debye models [9] are examples of widely used dielectric models suitable for accurate analysis of PCB and packaging interconnects. Parameters for such models are usually not available from manufacturers and have to be identified. To simulate effect of conductor roughness, modified Hammerstadt [10] and Huray's snowball conductor roughness models [11] can be effectively used, but parameters for such models are not readily available from a manufacturer. Manufacturers of dielectrics provide dielectric parameters at 1-3 points in the best cases – those points may be acceptable to define wideband Debye model. Manufacturers of copper laminates typically do not have parameters for the electrical roughness models. Thus, meaningful interconnect design and compliance analysis must start with the identification or validation of dielectric and conductor roughness models over the frequency band of interest. Even electromagnetic analysis of interconnects without such models may be simply not accurate and useless. **Accurate material characterization up to a target frequency is the most important element for design success.** The simplest procedure for such validation or identification is based on generalized modal S-parameters (GMS-parameters) [12]. S-parameters are measured for two line segments with substantially identical transitions and cross-sections, converted into reflection-less GMS-parameters and material models are then identified by matching computed and measured GMS-parameters. The procedure is automated in Simbeor software [7]. As an example of material parameters identification up to 50 GHz (for 25-30 Gbps data channel) we use measured data provided by David Dunham from Molex for one of the material characterization boards made from Nelco N4000-13EP dielectric and VLP copper [13]. A set of 2, 4 and 6-inch strip line segments was used to extract reflection-less GMS-parameters for 2 and 4 inch line segments as shown in Fig. 4. The dielectric specifications show that this dielectric may have dielectric constant (Dk) from 3.6 to 3.7 and loss tangent (LT) from 0.008 to 0.009.

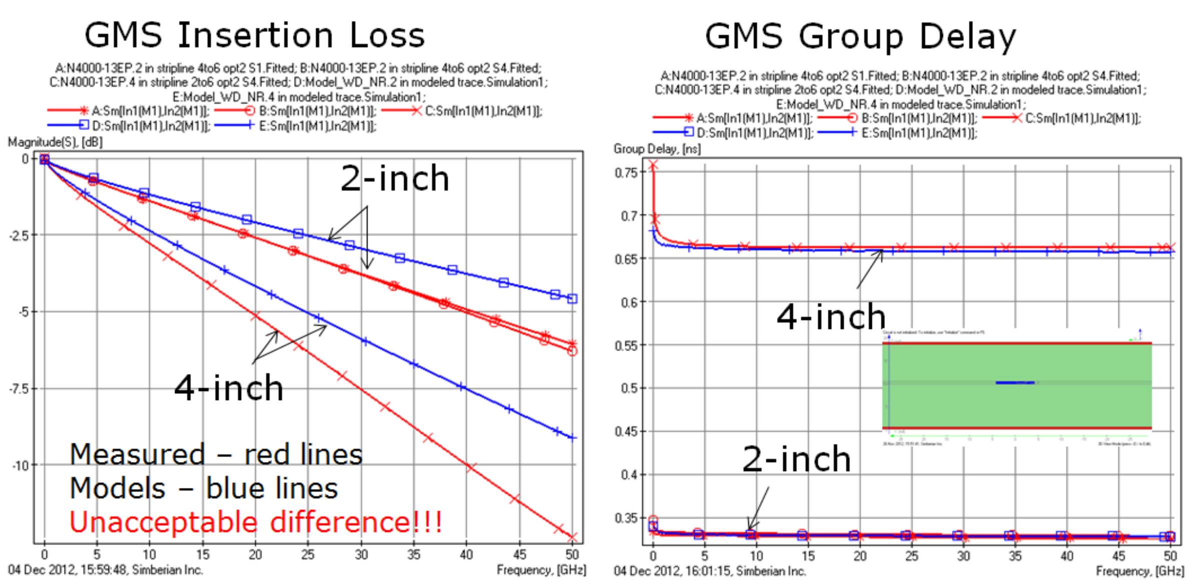


*Fig. 4. Measured (red lines) and computed (blue lines) generalized modal insertion loss (left plot) and group delay (right plot) for 2 and 4 inch strip line segments (dielectric model from manufacturer and smooth conductor model).*

If we compute GMS-parameters for 2 and 4 inch segments with the electromagnetic analysis with wideband Debye model and Dk=3.8 and LT=0.008 defined at 10 GHz, the difference in the measured and computed group delay is very small, but the difference in GM insertion loss is huge as illustrated in Fig. 4. Dk in the model is slightly increased to match group delays – that increase can be explained by anisotropy of the dielectric. Horizontal component of the Dk for layered glass-resin dielectrics can be up to 20% larger [14] and manufacturer probably used wider strips to identify dielectric. Wider strips have less energy in horizontal component of electric field and predict smaller Dk. What about the loss tangent – **how to explain such huge difference in the predicted and measured IL?** Typically this situation is explained as wrong data from the manufacturer. In this case LT should be increased to 0.0112 to have acceptable match for the insertion loss. Another option is to assume that the dielectric data from the manufacturer are actually correct, and attribute all observed excessive losses to the conductor roughness. As shown in Fig. 5, nearly perfect correspondence of measured and computed models can be achieved with the modified Hammerstadt model with roughness parameter 0.27, roughness factor 4 and conductor resistivity adjusted to 1.1 (relative to resistivity of annealed copper).

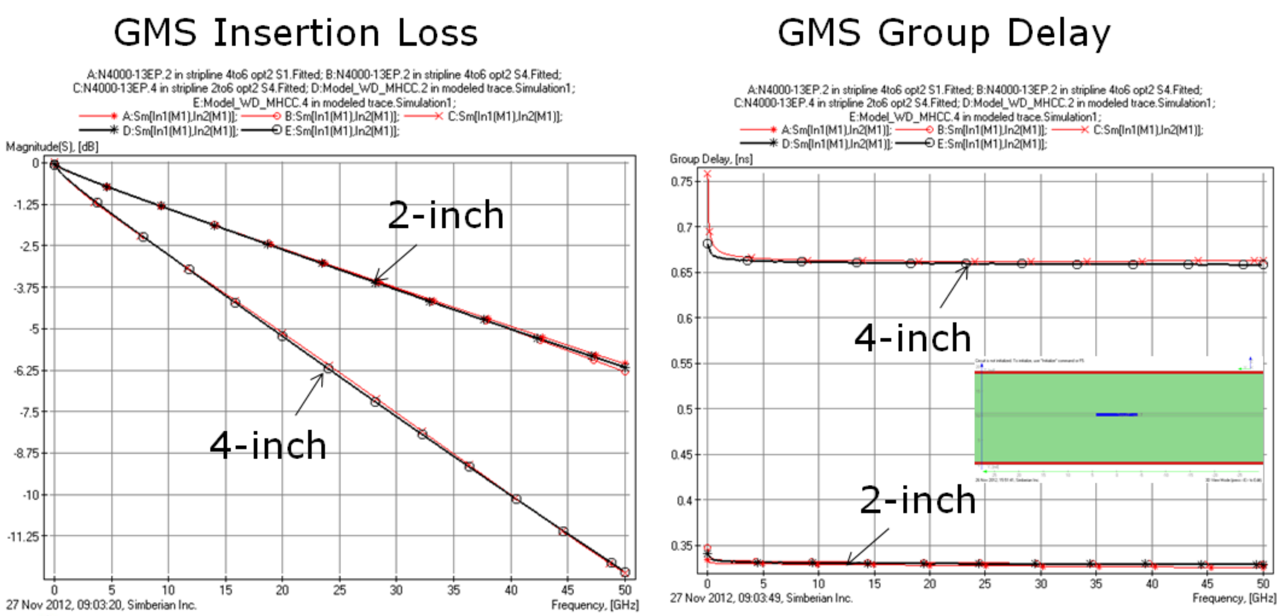


*Fig. 5. Measured (red lines) and computed (black lines) generalized modal insertion loss (left plot) and group delay (right plot) for 2 and 4 inch strip line segments (dielectric model from manufacturer and rough conductor model with SR=0.27, SR=4).*

As the result of this simple example we ended up with two models – with LT=0.0112 and no roughness and with LT=0.008 (as in the specs) and additional model for conductor roughness. Which one is correct? Both models are suitable for the analysis of the 8.5 mil strip line on that board. However, if strips with different width are used, the model without roughness model will be much less accurate. For instance, model without roughness predicts up 40% smaller losses for differential strip with 4 mil wide strips and 4 mil distance. **Model with the rough conductor is expected to produce more accurate insertion loss estimation for a broad range of strip widths.** This example illustrates typical situation and importance of dielectric and conductor roughness model identification to have analysis to measurement correspondence for a set of transmission lines on a particular board for a target frequency range.

**Localization property and de-embedding of discontinuities:** Ideally, all interconnects should look like uniform transmission lines with specified characteristic impedance. In reality, a channel is typically composed with transmission lines of different types (micro-strip, strip, coplanar, coaxial,…) and transitions between them such as vias, connectors, breakouts and so on. Even if we maintain the same impedance for the lines of different types, the transitions may be still reflective due to physical differences in cross-sections of the connected lines (coaxial and micro-strip for instance). The reflections cause additional losses and resonances and, thus, unwanted signal degradation. The effect of the transitions can be accounted for with models build with a full-wave 3D analysis. If such analysis is possible in isolation from the rest of the board up to a target frequency, the structure is called localizable [15]. Structures with the behavior dependent on the board geometry are called not localizable and should not be used for multi-gigabit interconnects in general. Analysis of such structures is possible only at the post-layout stage with substantial simplifications that degrade accuracy of the model at frequencies above 3-5 GHz. In other words, **only localizable transitions and discontinuities must be used to design predictable interconnects at frequencies above 3-5 GHz – this is one of the most important elements of successful design.** How to estimate the localization property of a transition? The simplest way is to run electromagnetic analysis of the structure with different boundary conditions or simply change simulation area size and evaluate the differences. Example of the localization evaluation for a single via with 6 stitching vias is shown in Fig. 6. The structure can be considered as localizable up to 50 GHz. The via in this example is localizable, but not optimal – the reflection loss may be not acceptable if multiple vias are used in a channel. Further via optimization can be done in Simbeor Via Analyzer and SiTune optimization tools [7].

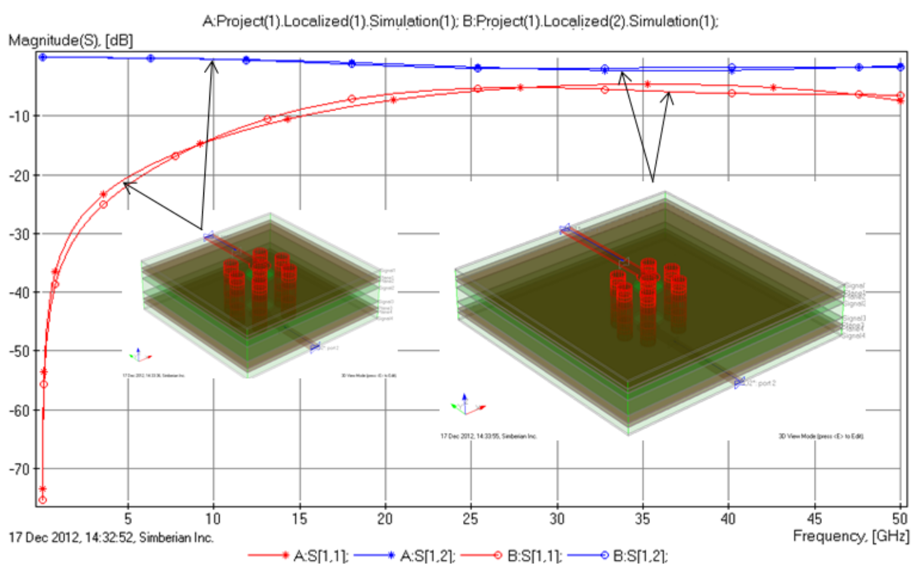


*Fig. 6. Localization property evaluation for a single via with 6 close stitching vias. Increase of simulation area size just slightly changed computed reflection (red lines) and insertion loss (blue lines).*

In this comparison of S-parameters of two vias simulated in different area, de-embedded transmission line ports are used to reduce the numerical reflection and to shift the phase reference plane closer to vias and to have S-parameters describing exactly the same portion of the geometry. **Quality of such numerical de-embedding defines the quality of the final interconnect model.** The simplest way to evaluate the de-embedding quality is to simulate a 90-degree segment of ideal 50-Ohm strip line as suggested in [16]. This simple test allows rigorous estimation of de-embedding accuracy and dynamic range. Another way is to simulate a t-line segment and concatenate the models into a longer segment – there should not be reflections observed at the connection points. If substantial reflections observed in such numerical experiment, the models are not suitable for the decompositional analysis.

**Analysis to measurement validation process (benchmarking):** Finally, how to make sure that the interconnect analysis works up to the target frequency and what is the problem if it does not? The best way to evaluate the accuracy of analysis is to build a validation or benchmarking board and compare analysis results with measurements. **The benchmarking is one of the most important elements of design success.** First of all, such board should include a set of structures to identify all dielectrics and conductor roughness models. There must be at least one pair of lines per one material model to identify separately models for solder mask, core and prepreg dielectrics or resin and glass, conductor roughness, plating material and so on. Identification of two models at the same time may lead to multiple possibilities and is problematic, as was pointed out at the dielectric and roughness model identification section. Benchmarking board should also include a set of structures to identify accuracy for transmission line models with possible coupling, resonant structures (Beatty standards or other type of planar resonators for instance) and typical discontinuities (channels with single and differential vias for instance). Examples of benchmarking boards developed and investigated up to 50 GHz are shown in Fig. 7 (see papers [12], [17]-[20] for details).

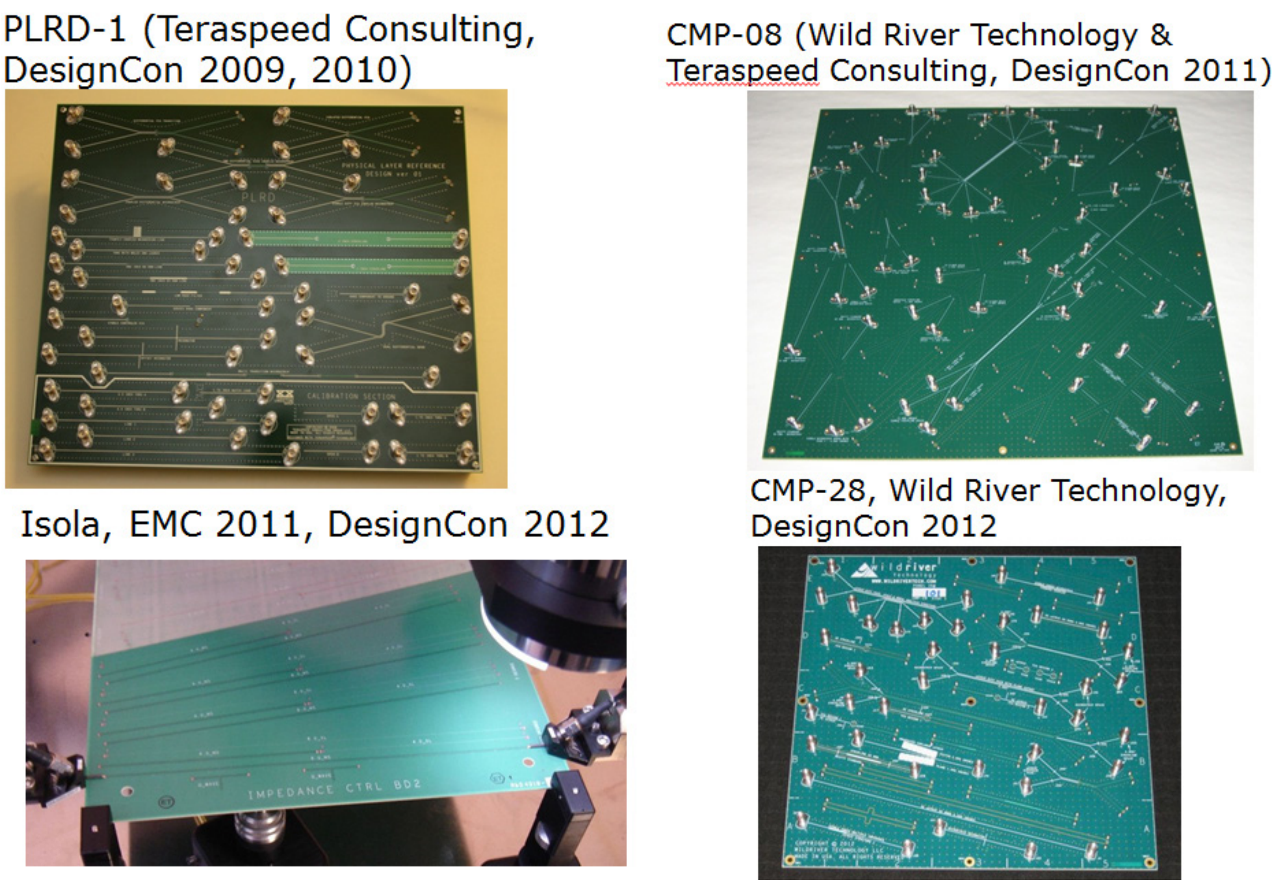


*Fig. 7. Examples of benchmarking boards features in [12],[17] (top lefs), [18] (top right), [19] (bottom left), [20] (bottom right).*

Considering the benchmarking process, identified material models must be consistently used for the analysis of all structures on the board. Tweaking dielectric or conductor roughness models for each structure for instance should be strictly prohibited. Possible discrepancies reveal either limitations of a tool or board design or manufacturing defects that altered the expected behavior. The source of the discrepancy must be investigated. The best choice for benchmarking measurement is VNA with the target frequency bandwidth. Either coaxial connectors or probe launches can be used. Probe launches are easier to model, but the measurement have to be done with a probe station – handheld probes are simply not suitable at microwave frequency range. In case of connectors with launches, they can be de-embedded or simulated too (more difficult due to unknown material properties). TRL-type de-embedding can be used for PCBs [17], but additional structures may be needed for the de-embedding (they can be also used for the material parameters identification). Note that simple T-matrix de-embedding does not simply work for PCB applications due to inhomogeneity of dielectrics and large manufacturing variations. Finally, measured and computed magnitudes and phases or group delays for all S-parameters have to be compared. Just insertion loss comparison is incomplete and may be misleading. Always compare phase or group delay and also reflection parameters.

**Conclusion:** Four essential elements that guarantee successful design of PCB and packaging interconnects up to 50 GHz and beyond have been outlined in the paper first time. Bandwidth and quality of S-parameter models, broadband material characterization and identification, localization of all elements of a channel and systematic benchmarking process are equally important elements of successful design. If even one element is missing or neglected, it may compromise the whole project.

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