Decompositional Electromagnetic Analysis of Digital Interconnects

Yuriy Shlepnev

Simberian Inc.
3030 S Torrey Pines Dr., Las Vegas, NV, 89146, USA
shlepnev@simberian.com

***Abstract* — Essential elements of decompositional electromagnetic signal integrity analysis of PCB and packaging interconnects are introduced in the paper. Digital interconnects can be formally divided into transmission line segments and discontinuities or transitions in lines such as via-holes and connectors. Multiport models of the components can be built separately with the electromagnetic analysis and then united into a complete channel model. This technique is known as decompositional electromagnetic analysis. Bandwidth and quality of multiport models, broadband material models, localization of all elements of a channel and systematic validation process are outlined in the paper as the key elements of the decompositional interconnect analysis that lead to successful analysis to measurement correlation from DC up to 50 GHz.**

I. 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. It imposes very special requirements on the interconnect modelling and design. No models or over-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 and discontinuities or transitions in lines such as via-holes and connectors. Multiport models of components are built separately with quasi-static or electromagnetic analysis, measurements or obtained from component vendors and then united into a complete channel model. This technique was originally developed for 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. Digital interconnects typically require analysis over much larger frequency band and may contain components that have not being used in microwave applications. **This paper outlines four essential elements of the de-compositional electromagnetic signal integrity analysis that guarantee analysis to measurement correlation up to 50 GHz and beyond.**

II. Quality of S-parameter Models

Any element of a linear time-invariant data channel can be modelled as a multiport described with S-parameter models. 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. 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 frequency and time domain and compliance analyses of interconnects. **Multiport models of interconnects must have sufficient bandwidth and have acceptable passivity, reciprocity and causality quality metrics. This is one of the key elements that lead to 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 accurate extrapolation, and have bandwidth 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 the total physical length of the system; c is speed of light and ε_{eff} is effective dielectric constant. The highest frequency in the sweep must be defined by a required resolution in time-domain or by the signal spectrum [1]. The highest frequency can be defined either with signal rise time t_r as $f_h > \frac{\alpha}{t_r}$ or with the first harmonic of the signal spectrum f_{s1} as $f_h > K \cdot f_{s1}$. α may be between 0.5 and 1, and K may range from 2 to 5, depending on the actual attenuation in the channel. All models for a channel interconnects must satisfy the target bandwidth requirement. Otherwise they have to be discarded and rebuilt.

Most of interconnect component S-parameter models are available as discrete or tabulated Touchstone models [2]. Interpolation or approximation of tabulated matrix elements may be necessary both for time and frequency domain analyses. Appropriate sampling is very important for discrete

Fourier transform and convolution-based time-domain analysis algorithms [1], but not so for algorithms based on rational approximation. For successful rational approximation there must be 4-5 frequency point per each resonance and the electrical length of a system should not change more than a quarter of wave-length between two consecutive frequency points. This condition can be expressed as the limit on the frequency step *df* as: $df < \frac{c}{4L \cdot \sqrt{\varepsilon_{eff}}}$. Under-sampling 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 artefacts that are not so easy to detect. To reveal such defects, S-parameters quality metrics have been recently introduced in [3], [4] and implemented in Simbeor software [5]. Metrics for passivity, reciprocity and causality computed for band-limited discrete models can be used for preliminary analysis of quality of S-parameters.

Passivity Quality Metric (*PQM*) can be defined as:

$$PQM = \max\left[\frac{100}{N}\left(N - \sum_{n=1}^{N} PW_n\right), 0\right]\% \tag{1}$$

$PW_n = 0 \text{ if } PM_n < A; \text{ otherwise } PW_n = (PM_n - A)/B$

Where $A = 1.00001,\ B = 0.1$ in Simbeor software; N is the total number of frequency points; $PM_n = \sqrt{\max\left[eigenvals\left(S^*(f_n) \cdot S(f_n)\right)\right]}$; $S(f_n)$ is S-parameter matrix at frequency f_n.

Reciprocity Quality Metric (*RQM*) can be defined as:

$$RQM = \max\left[\frac{100}{N}\left(N - \sum_{n=1}^{N} RW_n\right), 0\right]\% \tag{2}$$

$RW_n = 0 \text{ if } RM_n < C; \text{ otherwise } RW_n = (RM_n - C)/B$

Where $C = 10^{-6}$ in Simbeor software; $RM_n = \frac{1}{N_s}\sum_{i,j}\left|S_{i,j}(f_n) - S_{j,i}(f_n)\right|$, N_s is total number of subtracted off-diagonal elements of S-matrix.

Preliminary Causality Quality Metric (*CQM*) can be defined as a ratio of clockwise rotation measure to the total rotation measure in % (3-point rotation algorithm was introduced by V. Dmitriev-Zdorov).

All introduced metrics range from 0 to 100. Zero means big violations and 100 means no violations. Values above 99.9 are typical for good models. Ranges for acceptable [99, 99.9) and questionable models [80, 99) are defined on the base of analysis of thousands of models in [4]. Examples of preliminary Touchstone model quality estimation are also provided in [4]. Note that preliminary quality estimation is done for a discrete and band-limited data set and, thus, is incomplete. Though, it allows separation of models with unacceptable violations of passivity and reciprocity. If passivity or reciprocity metrics are too low (below 80 in general), the model has to be discarded and rebuilt. Large violations of preliminary causality metric (*CQM* below 50 for instance) for computed models may point at under-sampled data – such models have to be also rebuilt.

A model ranked as good with the preliminary metrics, may still have hidden defects and may not allow accurate interpolation or extrapolation for purpose of the time-domain analysis for instance. Rigorous estimation of passivity and causality can be done only for a frequency-continuous models defined from DC to infinity. Such models can be built with the rational approximation of the original tabulated data [6]. S-parameters approximated with the rational functions are causal by definition in case if passivity is ensured from DC to infinite frequencies. High-quality tabulated models can be accurately approximated with the passive rational macro-models. The final quality metric with the range from 0 to 100 can be constructed using root-mean square error (*RMSE*) of the passive rational approximation as follows [4]:

$$Q = 100 \cdot \max(1 - RMSE, 0)\%$$

$$RMSE = \max_{i,j}\left[\sqrt{\frac{1}{N}\sum_{n=1}^{N}\left|S_{ij}(n) - S_{ij}(f_n)\right|^2}\right],\ i,j = 1,...,N \tag{3}$$

Where $S_{ij}(n)$ is element of S-matrix from the original data set at frequency f_n and $S_{ij}(f_n)$ is the same S-parameter computed with passive and reciprocal rational macro-model.

Examples of the final model quality estimation with the rational macro-models are provided in [4]. Good models typically have quality metric above 90. Models with the final quality metric between 50 and 90 should be further inspected to evaluate the useful bandwidth. Final quality metric below 50 usually means that the tabulated model cannot be accurately interpolated and extrapolated with a causal model. It has to be discarded to avoid further problems. After model quality is ensured, rational macro-model can be either re-sampled and used as improved model or exported and used as rational or broad-band SPICE macro-model that guarantees consistent analyses both in frequency and time-domain in practically all tools.

III. Broadband Material 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 analysed 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 [7], [8]. Accuracy of transmission line models is mostly defined by availability of broadband dielectric and conductor roughness models. Wideband Debye (Djordjevic-Sarkar) and multi-pole Debye models [8] are examples of 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, Huray's snowball [9] and modified Hammerstadt [10] conductor roughness models can be effectively used. Parameters for such models are also not readily available from the PCB manufacturers. Manufacturers of dielectrics usually provide dielectric parameters at 1-3 points in the best cases. Those points may be acceptable to define the 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. **Availability of accurate broadband material models is the most important element for design success.** Validation or identification of dielectric and conductor models can be done with generalized modal S-parameters (GMS-parameters) [11]. 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 [5]. 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 with Nelco N4000-13EP dielectric and VLP copper [12].

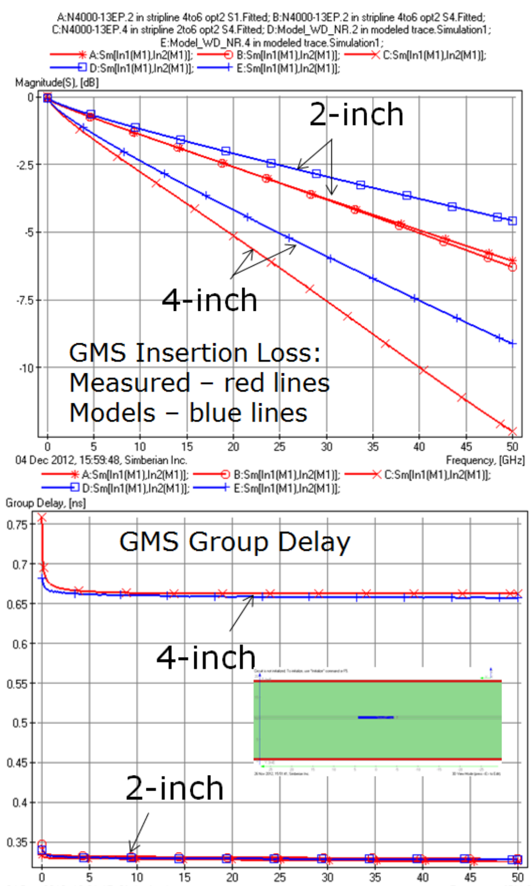

Fig. 1. Measured (red lines) and computed (blue lines) generalized modal insertion loss (top plot) and group delay (bottom plot) for 2 and 4 inch strip line segments (dielectric model from manufacturer and smooth conductor model).

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 shown as red lines in Fig. 1. 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. 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 (shown as blue lines in Fig. 1), the difference in the measured and computed group delay is small, but the difference in GMS insertion loss is huge as illustrated in Fig. 1. Dk in the model is slightly increased to match group delays – that increase can be explained by the layered structure and anisotropy of the dielectric. How to explain 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. 2, nearly perfect correspondence of measured and computed models can be achieved with the modified Hammerstadt model [10] with roughness parameter 0.27, roughness factor 4 and conductor resistivity adjusted to 1.1 (relative to resistivity of annealed copper).

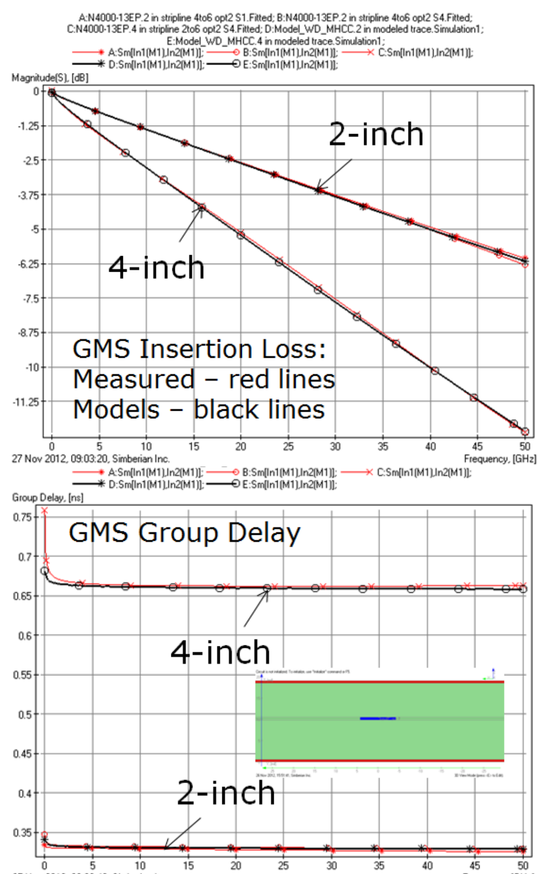

Fig. 2. Measured (red lines) and computed (black lines) generalized modal insertion loss (top plot) and group delay (bottom plot) for 2 and 4 inch strip line segments (dielectric model from manufacturer and rough conductor model with SR=0.27, SR=4).

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 effect will be less accurate, assuming that all additional losses are due to conductor roughness. For instance, model with increased LT and without roughness predicts up 40% smaller losses for differential strip with 4 mil wide strips with 4 mil distance at frequencies above 3-5 GHz. Model with the rough conductor produces more accurate insertion loss estimation for broader range of strip widths. This example illustrates typical situation and importance of the dielectric and conductor roughness model identification to have analysis to measurement correspondence for a particular board and t-lines up to 50 GHz. Note that the proper separation of loss and dispersion effects between dielectric and conductor models is very important, but not easy task. Though, if dielectric manufacturer used smooth conductors to identify dielectric parameters, that model can be used to identify parameters of the conductor roughness as it is done here. There are other ways to separate the effects, but this is outside of the scope of this paper. Another problem with the PCB materials is the layered structure and associated with that anisotropy. Difference between the vertical and horizontal components of the effective dielectric constant may be substantial and must be taken into account to have analysis to measurement correlation for transmission lines with different strip width and for vertical transitions.

IV. Modelling Discontinuities in Isolation

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, etc.) 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 built 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 [13]. Structures with the behaviour dependent on the board geometry are called not localizable and should not be used in 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 relatively low frequencies. **Only localizable transitions must be used to design predictable interconnects – this is one of the most important elements for design success.** 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 without changing phase reference planes and evaluate the differences in the computed S-parameters. If the difference is small, the structure is considered localizable and suitable for interconnect design. Note that practically all planar transmission lines and planar discontinuities used as PCB and packaging interconnects may be considered as localized structures that can be simulated in isolation. Though, it is possible only up to a certain frequency where high order modes turn from evanescent to propagating mode and these modes are not accounted for in the analysis. In addition, continuity of the reference plane(s) should be also preserved. Structures such as a line crossing a gap between two conductive planes cannot be simulated in isolation in general. Another typical example of non-localized structure on PCB is single via without or with electrically distant stitching vias. As illustrated in Fig. 3, change in simulation area size causes significant difference in computed reflection and transmission parameters for a single via on PCB. We cannot reliably predict behaviour of such via on a populated PCB without knowing models of all components connected to parallel planes. Even if we know the final board geometry and models for all components, the problem is typically too complicated to be accurately solved up to 50 GHz. Sufficient number of stitching vias connecting reference planes for top and bottom transmission lines placed close to the signal via localizes the problem as also illustrated in Fig. 3 (the transition is not optimal and can be further optimized if necessary). Such transition can be accurately simulated in isolation from the rest of the board in both pre-layout and post-layout analyses.

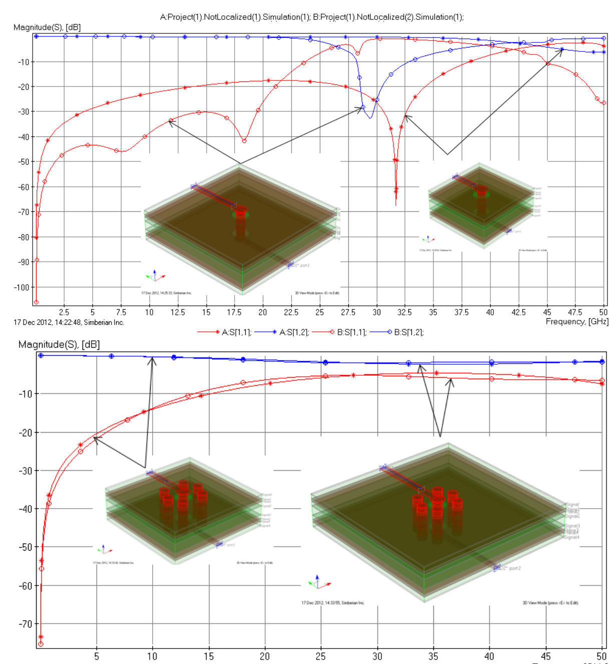

Fig. 3. Examples of non-localized structure (top plot – single via without stitching vias) and localized structure (single via with 6 stitching vias, bottom plot) on PCB. Simulated with de-embedded t-line ports and PEC boundary conditions at different distance. Transmission parameters – blue lines, reflection parameters – red lines.

Note that the number of stitching vias may be prohibitive for the isolation of single-ended transitions through multiple planes up to 50 GHz. Micro-via transitions that do not change reference planes or change one reference plane at a time may be a good alternative in this case. Also, differential vias can be used instead of single-ended. Though, only differential mode is localized in this case. Common mode behaviour is similar to the single-ended via and requires localization with the stitching vias for accurate modelling in isolation. Note that even a localized vertical transition on PCB or in package gradually loses this property and become dependent on PCB geometry with the increase of frequency as vias become electrically distant from each other.

Electromagnetic analysis of transmission line discontinuities in isolation is possible only with appropriately de-embedded transmission line ports. That procedure removes reflections due to parasitics of the model ports and is similar to de-embedding of DUT model from measured data. Quality of such numerical de-embedding defines the quality of the final interconnect model. A simple way to evaluate the de-embedding quality is to simulate a 90-degree segment of ideal 50-Ohm strip line as suggested in [14]. This 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 segment connection points. If substantial reflections observed in such numerical experiment, numerical models are not suitable for the decompositional analysis.

V. Validation and 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 the analysis results with measurements. **The benchmarking process is one of the most important elements for design success.** It reveals deficiencies of analysis, manufacturing and measurements. Validation board should include a set of structures to identify all dielectric and conductor roughness models. Ideally, there must be at least one pair of lines per one material model to identify separately models for solder mask or plating conductor, core and prepreg dielectrics or resin and glass, conductor roughness, and so on. Identification of two models at the same time may be not unique and lead to multiple possibilities and ambiguity as was pointed out at part III of this paper. Validation 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 (simple channels with single and differential vias for instance).

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 and models for vias for each structure should be strictly prohibited. Possible discrepancies reveal either limitations of a tool, incorrect material models, problems with board design or manufacturing defects that alter the expected behaviour. The source of the discrepancies must be investigated and revealed. 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 as a part of the channel. TRL-type de-embedding can be used for PCBs as was demonstrated in [15]. Though, it is difficult due to large PCB manufacturing variability and a number of additional structures are needed for the de-embedding. 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. TDR measurements have limitations because of every element of interconnect acts as a low-pass filter that reduced the resolution required to properly identify and compare interconnect discontinuities.

Examples of benchmarking boards developed and investigated up to 20-50 GHz have been provided in [10]-[12], [15]-[16]. One of the first benchmarking boards with 30 test structures have been systematically investigated up to 20 GHz and presented in [15]. Another board to investigate coupled structures on PCB up to 40 GHz is described in [16]. Both single-ended and coupled transmission line segments were used to identify material models with GMS-parameters. About 38 different structures on the board have been modelled with two different solvers and also experimentally investigated with VNA and TDNA. The results of this comprehensive investigation are reported in [16].

As an example of the model validation we will use simple interconnect structure shown on the insert in Fig. 4. The interconnect has about 6.4 inch strip line segment on signal layer S1 of 12-layer PCB made with Nelco N4000-13EP dielectric and VLP copper as described in [12]. Each signal layer on the board has 2 reference planes and layer S1 is closest to the connector side. There are two 2.4 mm connectors from Molex and two transitions from connectors to strip line (launches) on both ends of the segment. Models for dielectric and copper roughness for this board have been identified with GMS-parameters as described in part III of this paper. The board was designed and experimentally investigated by David Dunham from Molex. The structure is simulated with the decompositional electromagnetic analysis. Manufacturer provided S-parameter model for connector with good final quality metric. S-parameter models for launch and strip line were built with 3D electromagnetic analysis and then all models are concatenated. Modelled and measured S-parameters are compared in Fig. 4. Both magnitude and phase of transmission parameters S21 correlate well. Though, there are some substantial deviations in magnitude of the reflection parameters S11. It is typically very difficult to get ideal correlation in reflection parameters for low-reflective structures or at frequencies above 20 GHz.

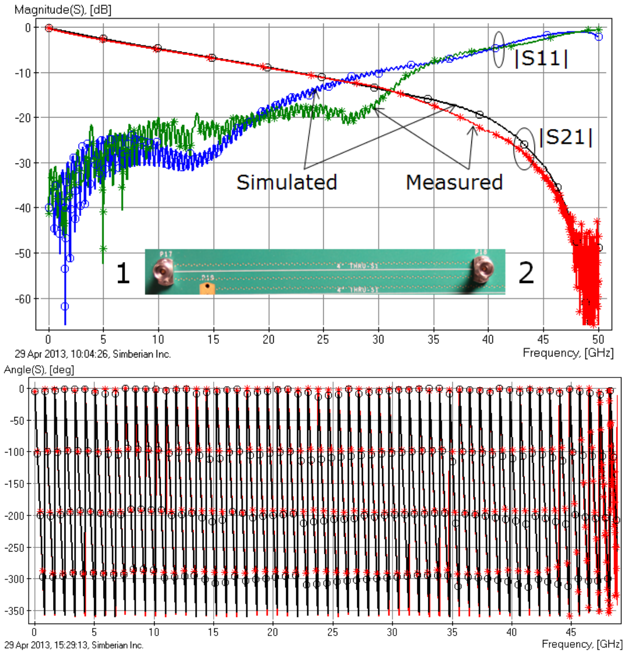

Fig. 4.Magnitudes of S-parameters(top graph, model – lines with circles, measured – lines with stars) and phase of transmission parameter (bottom graph, model – black line with circles, measured – red line with stars) for strip-line interconnect on layer S1.

There are multiple factors that may contribute to the discrepancy – manufacturing tolerances and actual inhomogeneity of the PCB dielectrics are usually the most probable reasons. Strip in signal layer S1 is connected to the connector pad with vias back-drilled from the other side of the board. Large manufacturing variability of such process may be a problem to have predictable reflection parameters up to 50 GHz. Though accurate estimation of transmission is still possible that is critical for the analysis of a data channel.

VI. Conclusion

Four elements essential for successful design of PCB and packaging interconnects up to 50 GHz and beyond have been outlined and illustrated in the paper. Bandwidth and quality of S-parameter models, broadband material characterization and identification, possibility to analyze all elements of a channel in isolation and systematic model validation and benchmarking process are equally important elements for design success. If even one element is neglected, it may compromise the whole project. It would be interesting to further investigate the impact of the elements on the interconnect performance evaluation. In particular, how to reduce the bandwidth requirements for material and component models without loss of accuracy if corresponding structures are relatively far from the signal source and high-frequency signal harmonics are filtered out by other structures. Though, this may be the subject for further investigation. Another interesting subject would be analysis of sensitivity of interconnect elements to the manufacturing variability and design and use of structures with small sensitivity on boards with larger manufacturing tolerances.

References

[1] S.L. Marple, Jr., Digital Spectral Analysis with Applications, Prentice-Hall, 1986.

[2] IBIS Open Forum Touchstone File Format Specification http://www.eda.org/ibis/

[3] Y. Shlepnev, Quality Metrics for S-parameter Models. IBIS Summit at DesignCon 2010, http://www.eda.org/ibis/summits/feb10/shlepnev.pdf

[4] Y. Shlepnev, Reflections on S-parameter Quality, IBIS Summit at DesignCon 2011, http://www.eda.org/ibis/summits/nov11b/shlepnev.pdf

[5] Simbeor Electromagnetic Signal Integrity Software, www.simberian.com

[6] B. Gustavsen, A. Semlyen, Rational approximation of frequency domain responses by vector fitting, *IEEE Trans. Power Del.*, vol. 14, no. 3, 1999, pp.1052 -1061.

[7] Y. Shlepnev, Modeling frequency-dependent conductor losses and dispersion in serial data channel interconnects, Simberian App. Note #2007_02, http://www.simberian.com/AppNotes.php

[8] Y. Shlepnev, Modeling frequency-dependent dielectric loss and dispersion for multi-gigabit data channels (with experimental validation), Simberian App. Note #2008_06, http://www.simberian.com/AppNotes.php

[9] P. G. Huray, O. Oluwafemi, J. Loyer, E. Bogatin, X. Ye, Impact of Copper Surface Texture on Loss: A Model that Works, DesignCon 2010.

[10] Y. Shlepnev, C. Nwachukwu, Roughness characterization for interconnect analysis. - Proc. of the 2011 IEEE International Symposium on Electromagnetic Compatibility, Long Beach, CA, USA, August, 2011, p. 518-523.

[11] Y. Shlepnev, A. Neves, T. Dagostino, S. McMorrow, Practical identification of dispersive dielectric models with generalized modal S-parameters for analysis of interconnects in 6-100 Gb/s applications. DesignCon 2010.

[12] D. Dunham, J. Lee, S. McMorrow, Y. Shlepnev, 2.4mm Design/Optimization with 50 GHz Material Characterization, DesignCon2011.

[13] Y. Shlepnev, Designing localizable minimal-reflection via-holes for multi-gigabit interconnects, Simberian App. Note #2009_05, http://www.simberian.com/AppNotes.php

[14] J.C. Rautio, An ultra-high precision benchmark for validation of planar electromagnetic analysis, IEEE on MTT, v.42, N11, 1994, p. 2046-2050.

[15] Y. Shlepnev, A. Neves, T. Dagostino, S. McMorrow, Measurement-Assisted Electromagnetic Extraction of Interconnect Parameters on Low-Cost FR-4 boards for 6-20 Gb/sec Applications, DesignCon 2009.

[16] J. Bell, S. McMorrow, M. Miller, A. P. Neves, Y. Shlepnev, Unified Methodology of 3D-EM/Channel Simulation/Robust Jitter Decomposition, DesignCon 2011.