

# Broadband material model identification with GMS-parameters

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**Abstract**—Design and compliance analysis of PCB and packaging interconnects for 10-50 Gbps and higher data rates begins with the identification of broadband dielectric and conductor roughness models. Such models are not available from manufacturers and the model identification is the most important element of successful interconnect design. Overview of broadband dielectric and conductor roughness models for PCB and packaging interconnect problems is provided in the paper. Theory of broadband model identification with generalized modal S-parameters (GMS-parameters) is outlined and compared with the standardized Short Pulse Propagation (SPP) technique. Practical examples of successful dielectric and conductor roughness model identification up to 50 GHz are also provided.

## I. INTRODUCTION

The largest part of PCB and packaging interconnects can be formally defined and simulated as transmission line segments. Models for transmission lines are usually constructed with a static or electromagnetic field solver. Accuracy of such models is mostly defined by availability of causal frequency-continuous dielectric and conductor roughness models. Such models are needed because of PCB and packaging dielectric properties change substantially over bandwidth from DC to tens of GHz. In addition to dispersion, PCB dielectrics have layered structure and exhibit anisotropic behaviour with different in- and out-of-plane dielectric constant.

Manufacturers of PCB dielectrics usually provide out-of-plane dielectric constant and loss tangent identified with wide strip resonators at a few frequency points, typically below 10 GHz. These data may be sufficient to define wideband Debye model [1] that provides useful extension above 10 GHz. Multi-pole Debye model [1] may be also constructed with a few measurements of dielectric constant and loss tangent if all data points are consistent (measured with the same technique). Though, such model will have bandwidth restricted to approximately the last available frequency point. For accurate analysis of typical interconnect, the models defined with vendor spreadsheet data may require correction due to anisotropy. Narrow traces have electric fields with direction in- and out-of-plane and models require either effective value of dielectric constant or tensor description of dielectric.

Manufacturers of copper laminates typically do not have parameters usable to construct electrical roughness models at all. Parameters in datasheets are usable for mechanical purpose, but not for the electrical characterisation.

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 in context of interconnect analysis (for typical trace sizes). Availability of accurate broadband material models is the most important element for design success [2]. Overview of some dielectric parameters measurement methods and fitting with frequency continuous models is available in [1], [3]. Short Pulse Propagation (SPP) technique [3] was recently standardized by IPC [4]. Validation or identification of dielectric and conductor roughness models can be also done with generalized modal S-parameters (GMS-parameters) [5], [6]. Main steps of the identification with GMS-parameters are described in this paper and compared with SPP-technique. Possible methods for separation of dielectric and conductor roughness loss and dispersion effects are also discussed. Though, we start from material models description to facilitate further model use and portability for modelling software.

## II. BROADBAND DIELECTRIC AND CONDUCTOR MODELS

Wideband Debye (aka Djordjevic-Sarkar or Swensson-Dermer) and multi-pole Debye models [1] are dielectric models often used for broadband analysis of PCB and packaging interconnects.

Expression for complex permittivity of the wideband Debye model can be written as follows [1]:

$$\varepsilon(f) = \varepsilon(\infty) + \frac{\varepsilon_d}{(m2 - m1) \cdot \ln(10)} \cdot \ln \left[ \frac{10^{m2} + i \cdot f}{10^{m1} + i \cdot f} \right] \quad (1)$$

Where  $f$  is frequency. Values of dielectric constant at infinity  $\varepsilon(\infty)$  and parameter  $\varepsilon_d$  can be defined with dielectric constant and loss tangent at one frequency point. Values of  $m1$  and  $m2$  define position of the first and last pole in the continuous spectrum defined by the model. Those are typically set to very low and very high values outside of the frequency band of interest ( $m1=4$ ,  $m2=12$ ). Though, they may be also treated as variables in the identification process.

If model (1) does not capture observed dielectric dispersion, universal multi-pole Debye model can be used. Dielectric permittivity for the multi-pole Debye model can be written as follows [1]:

$$\varepsilon(f) = \varepsilon(\infty) + \sum_{n=1}^N \frac{\Delta \varepsilon_n}{1 + i \frac{f}{f'_n}} \quad (2)$$

Values of dielectric constant at infinity  $\epsilon(\infty)$  as well as pole frequencies  $f_n$  and residues  $\Delta\epsilon_n$  in general are not known for composite dielectrics and have to be identified. Multipole Debye model suitable for analysis of interconnects up to 50 GHz may require 5-10 poles that means from 11 to 21 variables to identify. Multipole models with complex poles (Lorentzian for instance) and Cole-Cole model may be also considered as the alternatives to the wideband and multi-pole Debye models (though, there is not physics behind that).

Conductor model includes bulk resistivity that can be identified with measurement of resistivity at DC as done in [3], [4]. To simulate effect of conductor roughness, modified Hammerstad [7] or Huray's snowball [8] roughness correction coefficients are used. They can be applied to either conductor surface impedance locally [7] or to the internal conductor impedance part of the global impedance per unit length of multiconductor line (if static field solver is used).

Modified Hammerstad correction coefficient can be expressed as follows [7]:

$$K_{rh} = 1 + \left( \frac{2}{\pi} \cdot \arctan \left[ 1.4 \left( \frac{\Delta}{\delta} \right)^2 \right] \right) \cdot (RF - 1) \quad (3)$$

It has two parameters:  $\Delta$  or surface roughness (SR) parameter (may be associated with rms peak to valley value for regular copper) and roughness factor  $RF$  (maximal possible increase of conductor losses due to roughness). Note that classical Hammerstad model has  $RF=2$  and just one parameter, but not very useful for characterisation of rough PCB copper [7].

Expression for the conductor surface impedance correction coefficient based on the Huray's snowball model can be simplified and written as follows [8]:

$$K_{rhu} = 1 + \sum_{n=1}^N B_n \cdot d_n^2 \cdot \left( 1 + \frac{2 \cdot \delta}{d_n} + \frac{2 \cdot \delta^2}{d_n^2} \right)^{-1} \quad (4)$$

This model has  $2N$  parameters:  $N$  ball diameters  $d_n$  and  $N$  coefficients  $B_n$ . They are not known for commonly used copper foils and must be identified.  $\delta$  in (3) and (4) is the frequency-dependent skin depth.

Other roughness correction coefficients can be also considered [7], but they all require the identification process. As the alternative to the correction coefficients, effective roughness dielectric (ERD) layer [9] can be used to simulate the roughness effect. Parameters of ERD layer can be computed from micro-photographs of the surface or identified similar to the roughness correction coefficients.

### III. MODEL IDENTIFICATION WITH GMS-PARAMETERS

Dielectric and conductor roughness models identification can be done by matching measured and computed generalized modal S-parameters (GMS-parameters) for a transmission line segment [5], [6]. GMS-parameters are S-parameters of a line segment in modal space normalized to characteristic impedances of t-line modes. GMS-parameters of N-conductor line segment can be written as follows:

$$S_G = \begin{bmatrix} 0 & \exp(-\Gamma \cdot L) \\ \exp(-\Gamma \cdot L) & 0 \end{bmatrix} \quad (5)$$

where  $\Gamma = \text{diag}(\Gamma_i, i=1, N)$  are complex propagation constants (Gammas) of  $N$  modes in line and  $L$  is the line segment length. All reflection and mode transformation parameters in (5) are exactly zeroes by definition. GMS-parameters (5) can be extracted from S-parameters measured for two line segments with length difference  $L=L_2-L_1$  through diagonalization of T-matrix [5], [10]. Gammas for comparison can be computed with an electromagnetic or quasi-static field solver. The material model identification is done by matching the measured and computed GMS-parameters as illustrated in Fig. 1.

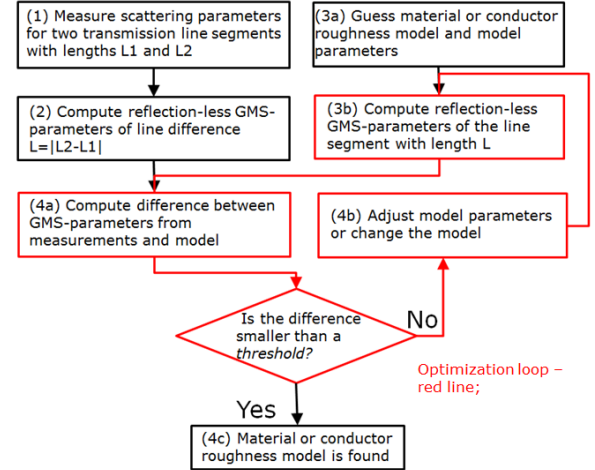


Fig. 1. Material model identification with GMS-parameters.

The two line segments should have substantially identical cross-sections and transitions to probes or connectors. Quality of measured transmission line S-parameters has to be estimated first and then TDR is used to verify consistency of the test fixtures [2], [3]. This procedure is implemented and automated in Simbeor software [11]. The key in this approach is availability of algorithms for analysis of transmission lines that supports the frequency-continuous material models (1)-(4) in step (3b) of the algorithm shown in Fig. 1.

It is known that the conductor roughness effect causes signal degradation (losses and dispersion) that are similar to the signal degradation caused by dielectrics [7]. Thus, it is important to separate properly the effects of losses and dispersion between the conductor roughness and dielectric models, or understand the consequences of not doing such separation. **There are four possible ways to build the conductor surface roughness model as outlined next.**

1) Optimize dielectric model to fit measured and modelled GMS-parameters following the procedure in Fig. 1 and do not use any additional conductor roughness model. The dielectric model will include effect of conductor surface roughness. Such model may be suitable for the analysis of a particular transmission line and has to be rebuilt if strip width or line type is changed. This combined model may be acceptable in cases of high-loss dielectrics when the effect of conductor roughness is minimal.

2) Define dielectric model with the data available from the dielectric manufacturer and then identify only the roughness

model with GMS-parameters following the procedure in Fig. 1. This approach works well if a manufacturer has reliable procedure to identify the dielectric properties (most of them do). This is the simplest way to identify the conductor roughness model.

3) If dielectric model is not available, identify dielectric and conductor roughness models separately. In addition to two line segments with rough copper, make two or more transmission line segments with flat rolled copper on the same board. First, use segments with flat copper to identify parameters in dielectric model following the procedure in Fig. 1. Then use the identified dielectric model for rough segments and identify the conductor roughness model following the same procedure Fig. 1, but for the roughness model.

4) If dielectric model is not available, identify dielectric and conductor roughness models simultaneously. It can be done with multiple line pairs with different widths of strips in each pair (narrow, regular and wide strips made of the same rough copper for instance). Dielectric model and conductor roughness model parameters can be optimized simultaneously following the procedure in Fig. 1, until differences of GMS-parameters for segments with all strip widths reach the stopping criteria. The resulting dielectric and roughness models will be usable for a given range of the strip widths. Though the procedure is the most complicated and may lead to multiple possibilities (ambiguity).

**Overall, the material model identification procedure with GMS-parameters is the simplest possible.** It needs measurements for two t-line segments with any geometry of cross-section and transitions. No extraction of propagation constants (Gamma) from measured data is required. The extraction of Gamma may be difficult and error-prone. Also, no de-embedding of connectors and launches is required. The approach needs the simplest numerical model - only propagation constants in (5) has to be computed for a given cross-section and with the material models to identify. No 3D electromagnetic models of the transitions are required. Procedure with GMS-parameters has minimal number of smooth complex functions to match during the identification process. Specifically, one complex parameter for single and two parameters for differential lines have to be matched. All reflection and modal transformation parameters are exactly zeroes always. Identified models are frequency-continuous and models (1), (3), (4) are not restricted to the frequency band used in the identification process. They are naturally extendable above the upper and below the lower frequencies.

#### IV. COMPARISON OF GMS AND SPP TECHNIQUES

It would be interesting to compare material model identification with GMS-parameters with Short Pulse Propagation (SPP) technique [3] standardized by IPC [4].

##### **Commonalities of GMS and SPP techniques:**

1) Exactly the same or similar test fixtures can be used for the identification - two segments of t-line with different length and nearly identical cross-section and transitions. Though, GMS technique does not have restrictions on the line lengths as well as on the line type and transparency of the transitions.

2) Both techniques use numerical transmission line models to identify the models for dielectric or conductor surface roughness. Though, reflection-less S-parameters of line segment are used in GMS technique and complex propagation constants (Gammas) are used in SPP technique. But, as we can see from equation (5) **the non-zero GMS-parameters are just exponential form of Gammas.** Thus, there is one to one transformation between the GMS-parameters and Gammas.

3) Resistance measurement at DC can be used to identify bulk resistivity in both techniques.

##### **Differences of GMS and SPP techniques:**

1) Measured S-parameters are used to extract GMS-parameters, but short pulse TDT measurements are used in SPP technique to extract complex propagation constants.

2) SPP uses measurements at 1 MHz to have low frequency asymptotes of dielectric constant. This step is not needed with the GMS-parameters if S-parameters are measured starting from sufficiently low frequency.

In summary, two techniques are very similar and Gamma can be extracted from the GMS-parameters and used to identify models exactly as defined in the SPP technique. **GMS-parameters can be considered as a modification of SPP technique that allows use of S-parameters instead of the short pulse TDT.**

#### V. EXAMPLES

Multiple examples of the material model identification with GMS-parameters up to 50 GHz were provided in [2], [7], [12]-[14]. As an example of material parameters identification up to 50 GHz (for 25-50 Gbps data channel) here we use measured data provided by Wild River Technology (<http://wildrivertech.com/>) for CMP-28 channel model platform validation board made with Isola FR-408HR dielectric. Detailed description of the board can be found in [14]. Five points for dielectric constant and loss tangent are available from the datasheet for the FR-408HR material. Though, the points are measured with different methods and the maximal frequency is 10 GHz only. Multi-pole Debye model (2) cannot be accurately defined up to 50 GHz with those data. No data for the conductor roughness was available from the board manufacturer. To identify or confirm the dielectric and conductor roughness model parameters, 2 and 8 inch single-ended strip and micro-strip lines and 2 and 6 inch differential strip and microstrip lines were used. For single-ended strip lines we can extract GMS-parameters of 6-inch line segment. Dielectric specifications of FR408HR show that it has dielectric constant (Dk) 3.66 and loss tangent (LT) is 0.0117 at 1 GHz (the other points are also consistent with that value and wideband Debye model). We can use that point to define the wideband Debye model (1). Though, Dk in the model has to be increased from 3.66 to 3.8 (3.3%), to match the measured group delay of GMS transmission parameter. That increase can be explained by the anisotropy of the dielectric. If we assume that the dielectric loss tangent from the manufacturer datasheet is actually accurate enough and attribute all observed excessive losses to the conductor

roughness model (section III, scenario 2), we end up with nearly perfect correspondence in GMS-parameters shown in Fig. 2. Modified Hammerstad model (3) with the roughness parameter  $SR=0.32$  and roughness factor  $RF=3.3$  is the result of the identification. The same result can be achieved by matching computed and measured complex propagation constants as illustrated in Fig. 3. Measured Gamma is extracted here directly from GMS-parameters and the real part is plotted as attenuation and imaginary as group delay per unit length.

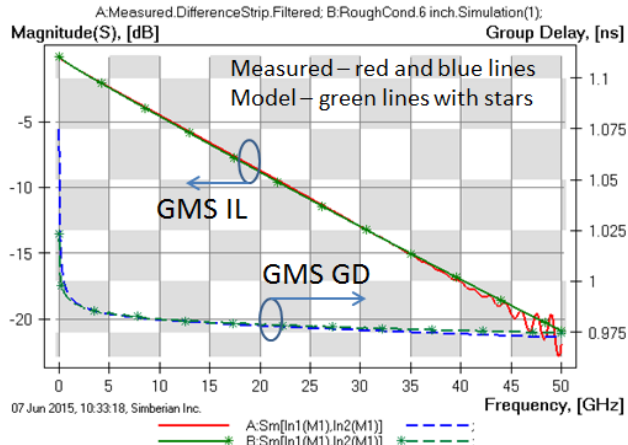


Fig. 2. Measured (red and blue curves) and computed (green curves) GMS insertion loss (left plot) and group delay (right plot) for 6 inch strip line segment with wideband Debye dielectric model with  $Dk=3.8$ ,  $LT=0.0117$  @ 1 GHz, modified Hammerstad model with  $SR=0.32$   $\mu m$ ,  $RF=3.3$  for copper roughness.

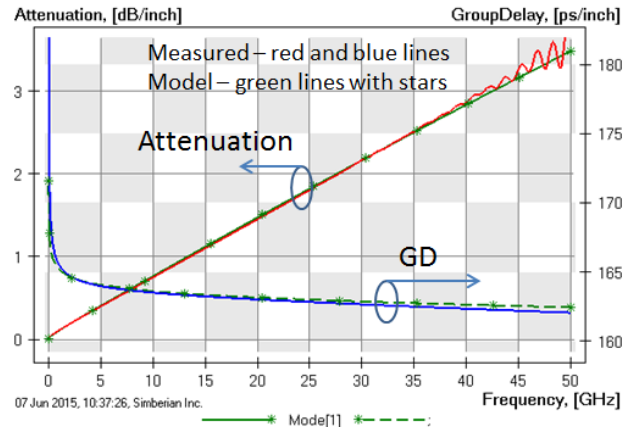


Fig. 3. Measured (red and blue curves) and computed (green curves) attenuation (left plot) and group delay (right plot) for strip line with wideband Debye dielectric model with  $Dk=3.8$ ,  $LT=0.0117$  @ 1 GHz and modified Hammerstad model with  $SR=0.32$   $\mu m$ ,  $RF=3.3$  for copper roughness.

Another interesting result of dielectric and conductor roughness models identification with GMS-parameters for multiple materials was recently reported in [12] and some data are provided here in Table 1 as another practical example.

Table 1. Results of model identification for some PCB materials.

Board Types	Model Parameters	WD Dielectric Constant @ 1 GHz	WD Loss Tangent @ 1 GHz	MH Roughness (SR, rms) ( $\mu m$ )	MH Roughness Factor (RF)
Megtron-6 with HVLP copper		3.64 (3.6)	0.002	0.38	3.15
Megtron-6 with RTF copper		3.72 (3.6)	0.002	0.37	4
Nelco N4000-13EPI with RTF copper		3.425 (3.4)	0.008	0.49	2.3

Dielectric constant and loss tangent values define wideband Debye model (1).  $Dk$  values in brackets and loss tangent (no changes in the model) are from datasheets, roughness parameters are for modified Hammerstad model (3). These models produced good correlation in the analysis of data links up to 50 GHz [12].

## VI. CONCLUSION

Overview of frequency-continuous dielectric and conductor roughness models for PCB and packaging interconnect analysis is provided in the paper. Practical procedure for identification of the model parameters with GMS-parameters has been described. Possible methods for separation of loss and dispersion effects between dielectric and conductor roughness models have been discussed. Model identification with GMS-parameters is compared with SPP technique. It is observed that two methods may be considered very similar. Instead of complex propagation constant extraction from pulse TDT response as in SPP techniques, it can be extracted from GMS-parameters and used for the material identification. It is demonstrated that matching of GMS-parameters and matching of complex propagation constants produces practically identical material models usable from DC up to 50 GHz.

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