DESIGNCON® 2013

JANUARY 28-31, 2013
SANTA CLARA CONVENTION CENTER

Elements of Decompositional Electromagnetic Analysis of Interconnects

Dr. Yuriy Shlepnev, Simberian Inc.

Simberian
Electromagnetic Solutions

UBM
Tech

Outline

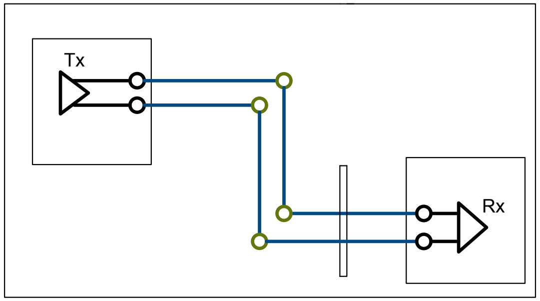
- Introduction in decompositional analysis of interconnects
- Multiport theory for interconnect analysis
  - Basics of S-parameters
  - Frequency-domain analysis with S-parameter models
  - Time-domain analysis with S-parameter models
  - Quality of S-parameter models
- Basics of signal propagation in interconnects
  - Signal degradation factors
  - Modeling transmission lines
  - Modeling via-holes and other discontinuities
- Broadband material models and model identification
- Validation of analysis with measurements (benchmarking)
- Conclusion
- Contacts and resources

Introduction

- Faster data rates drive the need for accurate models for data channels and specifically for interconnects
  - 10G Ethernet is practically mainstream now, 25-100G is coming…
- No models or over-simplified models may result in complete failure of multi-gigabit channel
- Without the accurate modeling of interconnects, a design may require
  - Test boards, experimental verification, …
  - Multiple iterations to fix or improve performance…
  - May be not possible to fix (whack-a-mole game)
- What is the best way to analyze interconnects and how to validate such analysis?
  - It depends on the problem to solve…

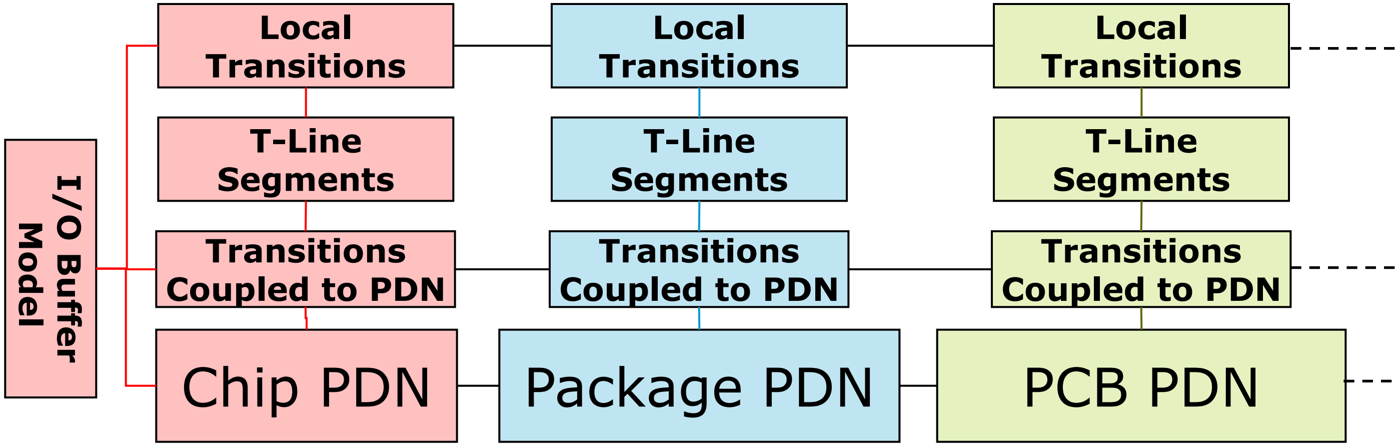
Possible ways to analyze interconnects

- Static or quasi-static analysis as a whole
  - Suitable for electrically small problems
  - Not suitable for PCB applications due to electrically large problem size
- 3D full-wave analysis as a whole
  - Suitable for EMI/EMC analysis (radiation/coupling)
  - Prohibitive simulation time or low accuracy for PCB SI analysis
- Decompositional electromagnetic analysis (diakoptics)
  - Divide, build or find models for elements and unite the models
  - Accurate and fast analysis of signal integrity (depends on accuracy of the component models)
  - May include coupling between nets and to parallel planes



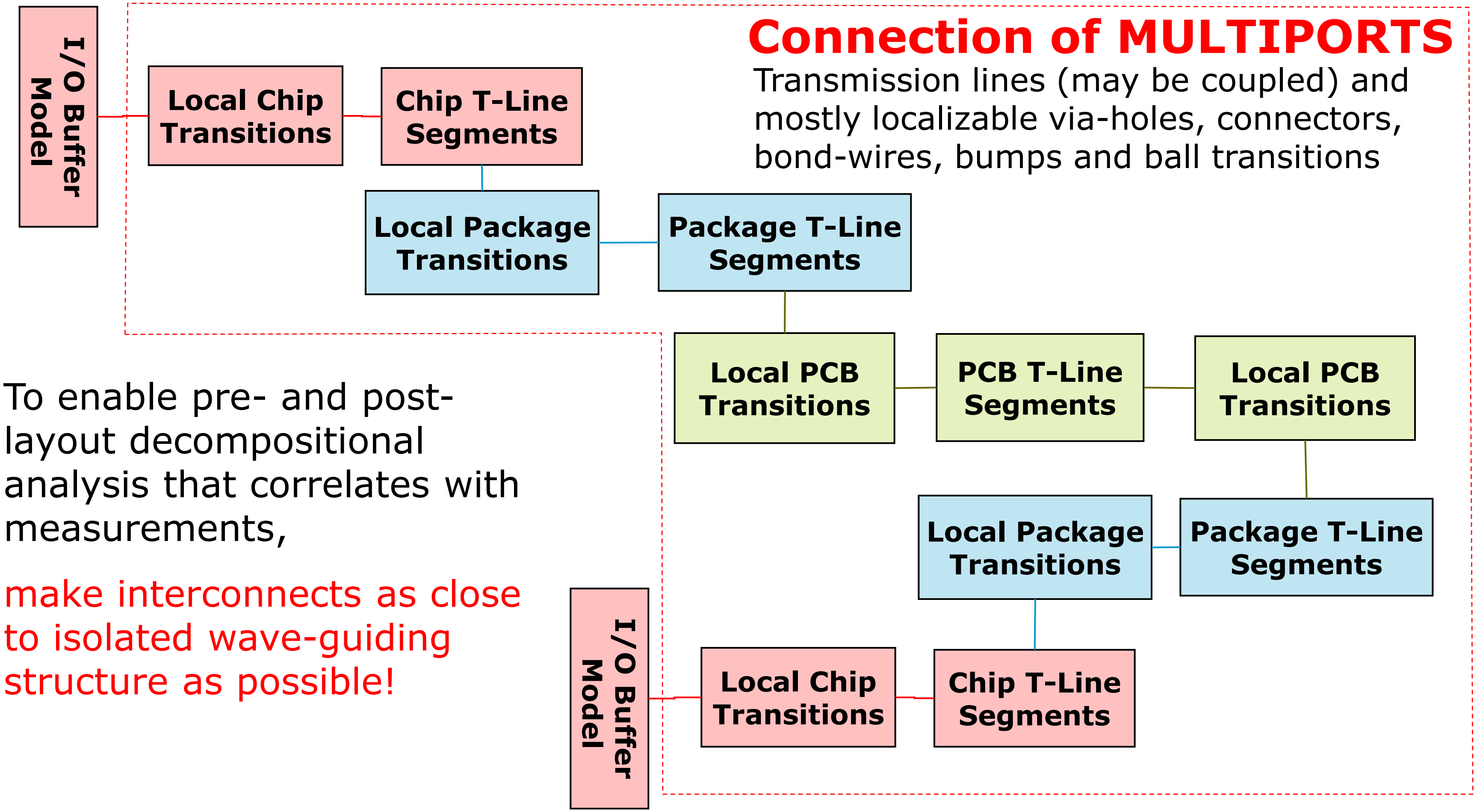
Elements of system decomposition

1) T-Line Segments: Segments of multi-conductor strip or micro-strip lines, periodic structures, CPWs, SIWs, slot-lines …
2) Local Transitions: Planar discontinuities, embedded passives, vias, non-uniform interconnects…
3) Transitions Coupled to PDN: Vias with returns through PNDs, discontinuities with changes in referencing, decoupling vias…
4) PDN: Transmission planes, strip lines,…



All things may be coupled…

Localized channel model



Connection of MULTIPORTS

Transmission lines (may be coupled) and mostly localizable via-holes, connectors, bond-wires, bumps and ball transitions

To enable pre- and post-layout decompositional analysis that correlates with measurements,

make interconnects as close to isolated wave-guiding structure as possible!

Essential elements of analysis for successful interconnect design

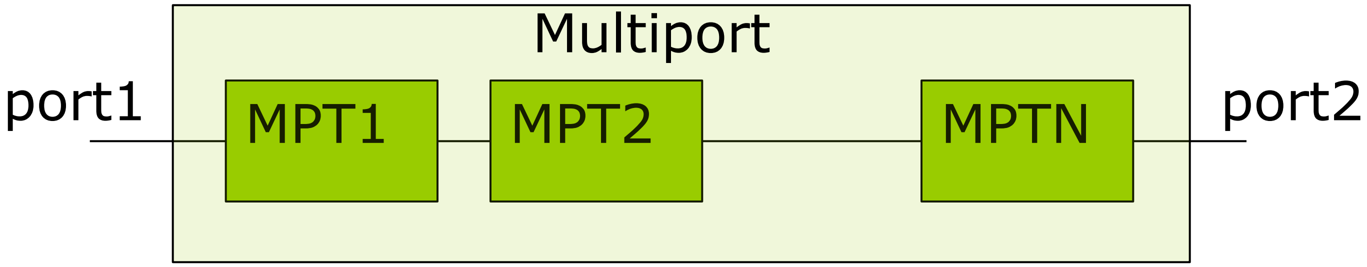
1. Quality of S-parameter models of components (bandwidth, sampling and causality)
2. Broadband dielectric and conductor roughness models (important for analysis of transmission lines)
3. Localization property (vias) and de-embedding of discontinuities (possibility to be analyzed in isolation)
4. Procedure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

Multiport theory for interconnect analysis

Basics of S-parameters

What is multiport?

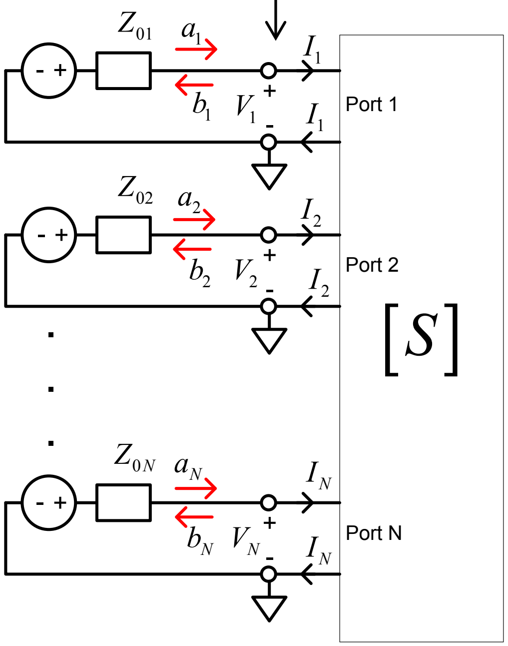
- Multiport is a natural and scalable black-box description of linear time-invariant system (suitable for non-linear too)
  - Reduces a system description to a simple input-output relationship irrespective of possible complicated internal structure
  - Suitable for systems smaller that, comparable with or larger than wavelength (literally DC to daylight)
  - Suitable for internal (wave-guiding) as well as external (antenna) systems



Multiport descriptors: Z, Y and S-parameters

Terminal or wave-ports:

$$\begin{cases} \overline{V} = \left(V_1, V_2, ..., V_N\right)^t & \text{- vector of port voltages} \\ \overline{I} = \left(I_1, I_2, ..., I_N\right)^t & \text{- vector of port currents} \end{cases}$$



Impedance (open-circuit) and Admittance (short-circuit) Matrices:

$$\overline{V} = Z \cdot \overline{I} \qquad \overline{I} = Y \cdot \overline{V}$$

$$Z_0 = diag\{Z_{0i}, i = 1, ..., N\} \in C^{N \times N}$$ normalization impedances

$$\overline{a} = \frac{1}{2} Z_0^{-1/2} \cdot \left(\overline{V} + Z_0 \cdot \overline{I}\right)$$ - vector of incident waves

$$\overline{b} = \frac{1}{2} Z_0^{-1/2} \cdot \left(\overline{V} - Z_0 \cdot \overline{I}\right)$$ - vector of reflected waves

Scattering matrix (exists always):

$$\overline{b} = S \cdot \overline{a}, \qquad S \in C^{N \times N}, \qquad S_{i,j} = \left.\frac{b_i}{a_j}\right|_{a_k = 0\ k \neq j}$$

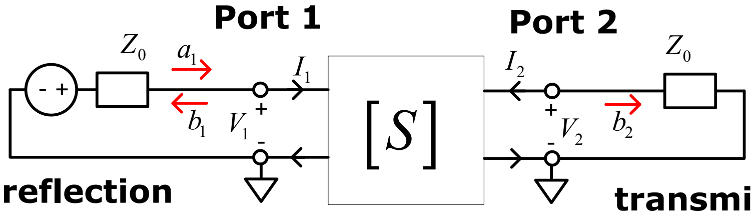
Frequency Domain (FD)

Reflected wave at port i with unit incident wave at port j defines scattering parameter S[i,j]

More in D.M. Pozar, Microwave engineering, John Wiley & Sons, 1998.

S-parameters for 2-port structure

$$\begin{bmatrix} b_1 \\ b_2 \end{bmatrix} = \begin{bmatrix} S_{1,1} & S_{1,2} \\ S_{2,1} & S_{2,2} \end{bmatrix} \cdot \begin{bmatrix} a_1 \\ a_2 \end{bmatrix}$$



reflection parameter

$$S_{1,1} = \left.\frac{b_1}{a_1}\right|_{a_2=0}$$

transmission parameter

$$S_{2,1} = \left.\frac{b_2}{a_1}\right|_{a_2=0}$$

$$v_i^+ = \sqrt{Z_0} \cdot a_i$$ **voltage of incident wave**

$$v_i^- = \sqrt{Z_0} \cdot b_i$$ **voltage of reflected wave**

$$V_i = v_i^+ + v_i^-$$ **total voltage**

$$I_i = \frac{1}{Z_0}\left(v_i^+ - v_i^-\right)$$ **total current**

$$P_i^+ = \left|a_i\right|^2$$ **power of incident wave**

$$P_i^- = \left|b_i\right|^2$$ **power of reflected wave**

$$\left|S_{1,1}\right|^2 = \frac{\left|b_1\right|^2}{\left|a_1\right|^2} = \frac{P_1^-}{P_1^+} \qquad \left|S_{2,1}\right|^2 = \frac{\left|b_2\right|^2}{\left|a_1\right|^2} = \frac{P_2^-}{P_1^+}$$

Magnitude is limited by 1 for passive systems!

FD:

$$\left|S_{i,j}\right| = \sqrt{\mathrm{Re}\left(S_{i,j}\right)^2 + \mathrm{Im}\left(S_{i,j}\right)^2}$$ **magnitude**

$$\angle S_{i,j} = \arctan\left(\mathrm{Im}\left(S_{i,j}\right) / \mathrm{Re}\left(S_{i,j}\right)\right)$$ **phase or angle**

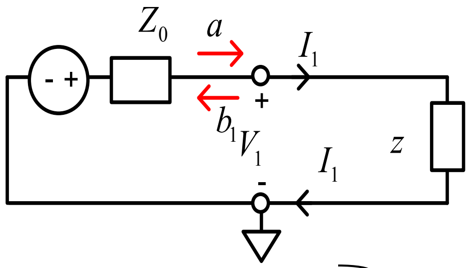
$$\left|S_{i,j}\right|_{dB} = 20 \cdot \log\left(\left|S_{i,j}\right|\right)$$ **magnitude in dB**

$$i = 1,2; \quad j = 1,2;$$

S-parameters are available in 2 forms

- Analytical models (equations)
  - Circuit with lumped elements (N poles)
  - Distributed circuits (Inf. poles or continuous spectrum)
  - Rational macro-models (N poles or Inf. poles with delay)
- Tabulated Touchstone models (discrete) produced by:
  - SPICE simulators
  - Microwave analysis software
  - Electromagnetic analysis software
  - Measurements (VNA or TDNA)
- All models may have reciprocity, passivity and causality violations
  - Quality of such models must be verified and assured!

Example: Terminator, one-port



z is a complex impedance $S \in C^{1\times 1}$

$$\left.\begin{array}{l} a_1 = \dfrac{1}{2\sqrt{Z_0}}\left(V_1 + Z_0 \cdot I_1\right) \\ b_1 = \dfrac{1}{2\sqrt{Z_0}}\left(V_1 - Z_0 \cdot I_1\right) \\ \\ V_1 = z \cdot I_1 \end{array}\right\}$$

$$b_1 = \frac{z - Z_0}{z + Z_0} \cdot a_1$$

$$S_{1,1} = \frac{z - Z_0}{z + Z_0}$$

Reflection parameter is equal to the reflection coefficient

Short-circuit:

$$z = 0 \Rightarrow S_{1,1} = -1$$

Open-circuit:

$$z = \infty \Rightarrow S_{1,1} = 1$$

Alternatively we can transform Z into S with

$$S = \left(Z_N - U\right)\cdot\left(U + Z_N\right)^{-1}, \quad Z_N = Z_0^{-1/2} \cdot Z \cdot Z_0^{-1/2}$$

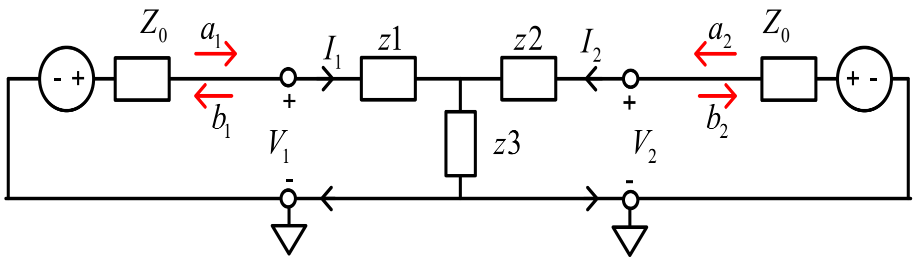
Passivity:

$$\left|S_{1,1}\right| \le 1$$

For real normalization impedance $\operatorname{Re}\left(z\right) \ge 0$

Always satisfied for nets composed of passive elements

Example: T-circuit, two-port



z1, z2, z3 are complex impedances

$$S \in C^{2\times 2}$$

We just use known Z and transform it to S

$$Z = \begin{bmatrix} z1+z3 & z3 \\ z3 & z2+z3 \end{bmatrix} \qquad Z_N = \frac{1}{Z_0}\begin{bmatrix} z1+z3 & z3 \\ z3 & z2+z3 \end{bmatrix}$$

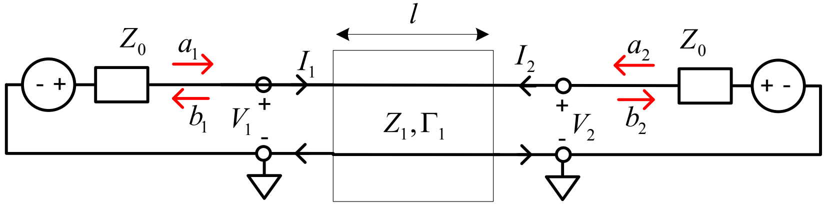
$$S = (Z_N - U)\cdot(U + Z_N)^{-1} = \frac{1}{A}\begin{bmatrix} -Z_0^2 + (z1-z2)\cdot Z_0 + B & 2\cdot z3\cdot Z_0 \\ 2\cdot z3\cdot Z_0 & -Z_0^2 - (z1-z2)\cdot Z_0 + B \end{bmatrix}$$

$$A = Z_0^2 + (z1+z2+2\cdot z3)\cdot Z_0 + B \qquad B = z1\cdot z2 + z2\cdot z3 + z1\cdot z3$$

S is always symmetric (reciprocal system) and non-singular

Passivity: $\left|eigenvals[S]\right| \le 1$ Always satisfied for nets composed of passive elements

One-conductor line segment



Passivity: Characteristic impedance and propagation constant must be causal and positive-real

$$Y(\omega,l) = \frac{1}{Z_1}\begin{bmatrix} cth(\Gamma_1 \cdot l) & -csh(\Gamma_1 \cdot l) \\ -csh(\Gamma_1 \cdot l) & cth(\Gamma_1 \cdot l) \end{bmatrix} \Rightarrow Y_N = \frac{Z_0}{Z_1}\begin{bmatrix} cth(\Gamma_1 \cdot l) & -csh(\Gamma_1 \cdot l) \\ -csh(\Gamma_1 \cdot l) & cth(\Gamma_1 \cdot l) \end{bmatrix}$$

$$S(\omega,l) = (U - Y_N)\cdot(U + Y_N)^{-1} \Rightarrow S(\omega,l) = \frac{1}{D}\begin{bmatrix} Z_1^2 - Z_0^2 & 2\cdot Z_1 \cdot Z_0 \cdot csh(\Gamma_1 \cdot l) \\ 2\cdot Z_1 \cdot Z_0 \cdot csh(\Gamma_1 \cdot l) & Z_1^2 - Z_0^2 \end{bmatrix}$$

$$S \in C^{2\times 2}$$

$$D = Z_1^2 + Z_0^2 + 2\cdot Z_1 \cdot Z_0 \cdot cth(\Gamma_1 \cdot l)$$

S-matrix is symmetric (S[1,2]=S[2,1]) and skew-symmetric (S[1,1]=S[2,2])

If normalization impedance is equal to the characteristic impedance of the mode, we get generalized modal S-matrix:

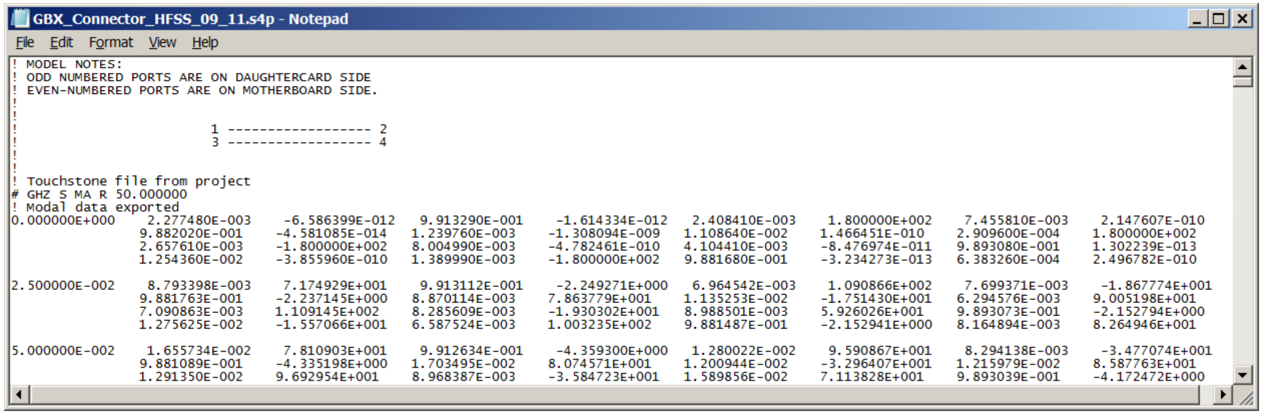
$$Z_0 = Z_1 \Rightarrow S(\omega,l) = \begin{bmatrix} 0 & \exp(-\Gamma_1 \cdot l) \\ \exp(-\Gamma_1 \cdot l) & 0 \end{bmatrix}$$

(anti-diagonal matrix)

Generalized modal S-parameters are useful for material parameters identification

Example of a discrete S-parameter model

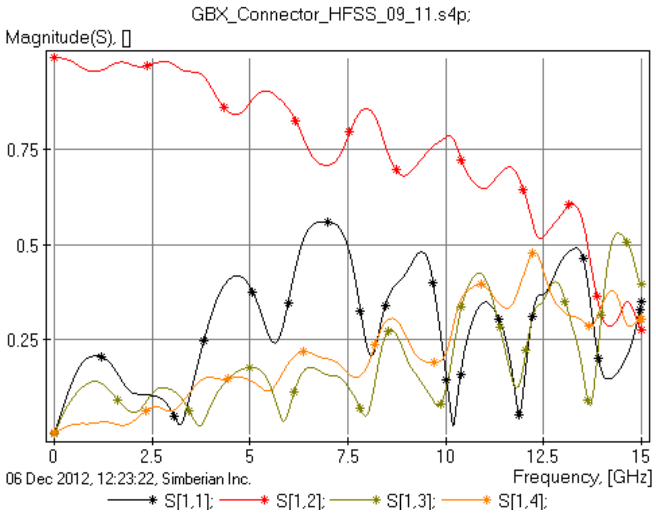
- Typical Touchstone model (see EIA/IBIS forum specs)

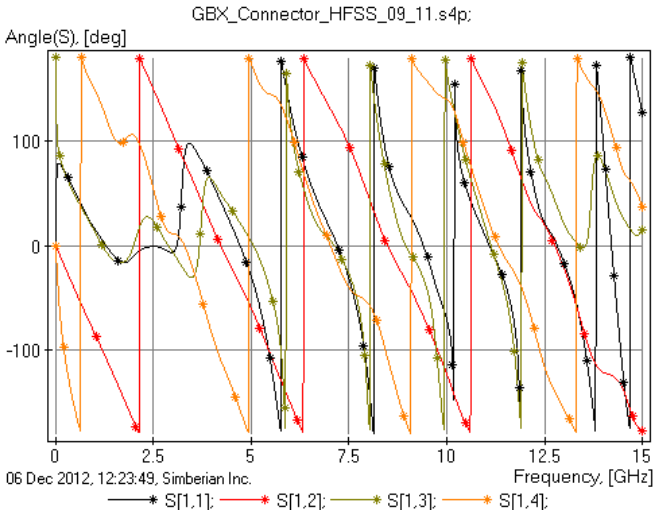


```
! MODEL NOTES:
! ODD NUMBERED PORTS ARE ON DAUGHTERCARD SIDE
! EVEN-NUMBERED PORTS ARE ON MOTHERBOARD SIDE.
!
!
!                    1 ------------------ 2
!                    3 ------------------ 4
!
!
! Touchstone file from project
# GHZ S MA R 50.000000
! Modal data exported
0.000000E+000    2.277480E-003    -6.586399E-012    9.913290E-001    -1.614334E-012    2.408410E-003    1.800000E+002    7.455810E-003    2.147607E-010
                9.882020E-001   -4.581085E-014   1.239760E-003   -1.308094E-009   1.108640E-002   1.466451E-010   2.909600E-004   1.800000E+002
                2.657610E-003   -1.800000E+002   8.004990E-003   -4.782461E-010   4.104410E-003   -8.476974E-011   9.893080E-001   1.302239E-013
                1.254360E-002   -3.855960E-010   1.389990E-003   -1.800000E+002   9.881680E-001   -3.234273E-013   6.383260E-004   2.496782E-010

2.500000E-002    8.793398E-003    7.174929E+001    9.913112E-001    -2.249271E+000    6.964542E-003    1.090866E+002    7.699371E-003    -1.867774E+001
                9.881763E-001   -2.237145E+000   8.870114E-003   7.863779E+001   1.135253E-002   -1.751430E+001   6.294576E-003   9.005198E+001
                7.090863E-003   1.109145E+002   8.285609E-003   -1.930302E+001   8.988501E-003   5.926026E+001   9.893073E-001   -2.152794E+000
                1.275625E-002   -1.557066E+001   6.587524E-003   1.003235E+002   9.881487E-001   -2.152941E+000   8.164894E-003   8.264946E+001

5.000000E-002    1.655734E-002    7.810903E+001    9.912634E-001    -4.359300E+000    1.280022E-002    9.590867E+001    8.294138E-003    -3.477074E+001
                9.881089E-001   -4.335198E+000   1.703495E-002   8.074571E+001   1.200944E-002   -3.296407E+001   1.215979E-002   8.587763E+001
                1.291350E-002   9.692954E+001   8.968387E-003   -3.584723E+001   1.589856E-002   7.113828E+001   9.893039E-001   -4.172472E+000
```

4-port structure (all ports have separate reference terminal)

Common defects: discreteness, bandwidth deficiency, passivity & causality



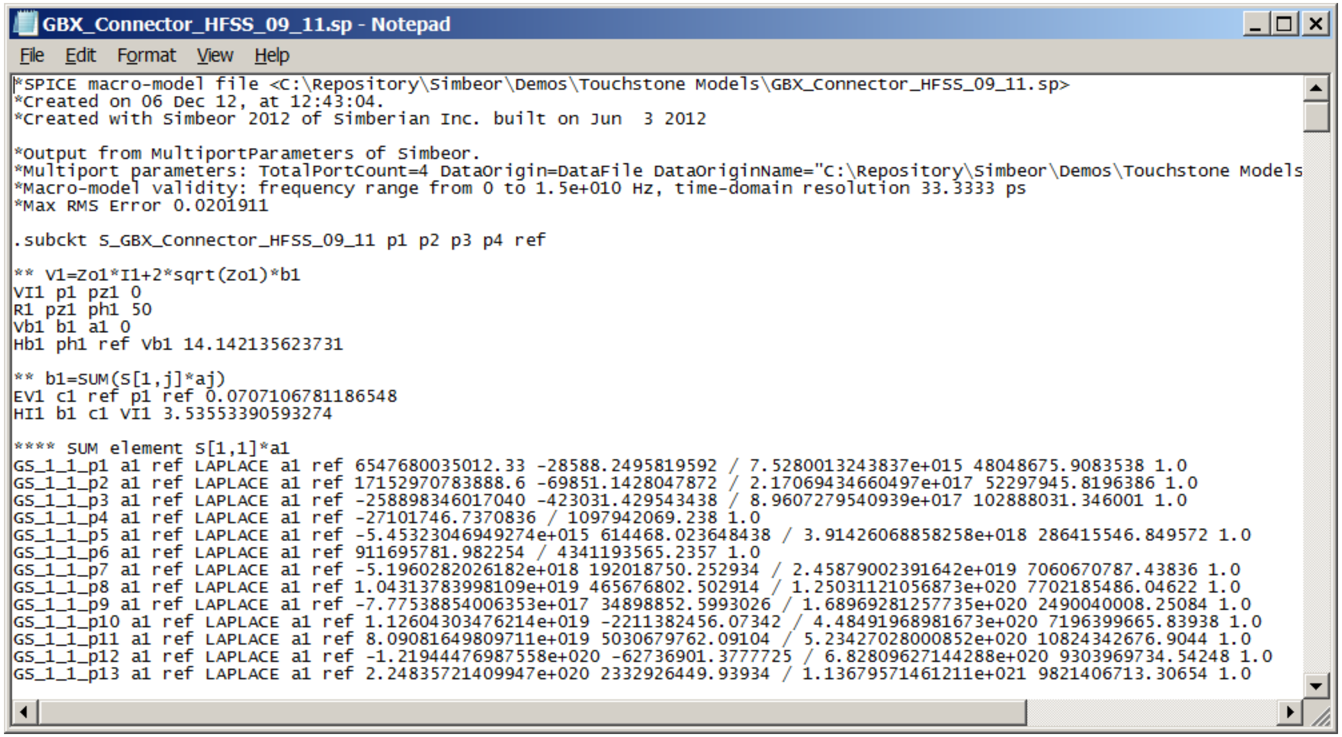


Example of BB SPICE S-parameter model

- BB SPICE model in HSPICE format

4-port sub-circuit with LAPLACE elements

Possible defects: bandwidth deficiency, passivity



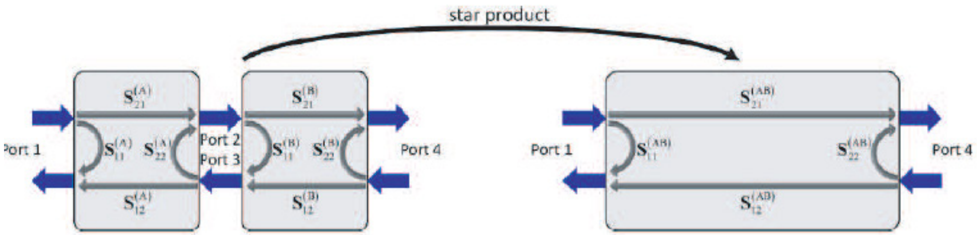
```
*SPICE macro-model file <C:\Repository\Simbeor\Demos\Touchstone Models\GBX_Connector_HFSS_09_11.sp>
*Created on 06 Dec 12, at 12:43:04.
*Created with Simbeor 2012 of Simberian Inc. built on Jun  3 2012

*Output from MultiportParameters of Simbeor.
*Multiport parameters: TotalPortCount=4 DataOrigin=DataFile DataOriginName="C:\Repository\Simbeor\Demos\Touchstone Models
*Macro-model validity: frequency range from 0 to 1.5e+010 Hz, time-domain resolution 33.3333 ps
*Max RMS Error 0.0201911

.subckt S_GBX_Connector_HFSS_09_11 p1 p2 p3 p4 ref

** V1=Zo1*I1+2*sqrt(Zo1)*b1
VI1 p1 pz1 0
R1 pz1 ph1 50
Vb1 b1 a1 0
Hb1 ph1 ref Vb1 14.142135623731

** b1=SUM(S[1,j]*aj)
EV1 c1 ref p1 ref 0.0707106781186548
HI1 b1 c1 VI1 3.53553390593274

**** SUM element S[1,1]*a1
GS_1_1_p1 a1 ref LAPLACE a1 ref 6547680035012.33 -28588.2495819592 / 7.5280013243837e+015 48048675.9083538 1.0
GS_1_1_p2 a1 ref LAPLACE a1 ref 17152970783888.6 -69851.1428047872 / 2.17069434660497e+017 52297945.8196386 1.0
GS_1_1_p3 a1 ref LAPLACE a1 ref -258898346017040 -423031.429543438 / 8.9607279540939e+017 102888031.346001 1.0
GS_1_1_p4 a1 ref LAPLACE a1 ref -27101746.7370836 / 1097942069.238 1.0
GS_1_1_p5 a1 ref LAPLACE a1 ref -5.45323046949274e+015 614468.023648438 / 3.91426068858258e+018 286415546.849572 1.0
GS_1_1_p6 a1 ref LAPLACE a1 ref 911695781.982254 / 4341193565.2357 1.0
GS_1_1_p7 a1 ref LAPLACE a1 ref -5.1960282026182e+018 192018750.252934 / 2.45879002391642e+019 7060670787.43836 1.0
GS_1_1_p8 a1 ref LAPLACE a1 ref 1.04313783998109e+019 465676802.502914 / 1.25031121056873e+020 7702185486.04622 1.0
GS_1_1_p9 a1 ref LAPLACE a1 ref -7.77538854006353e+017 34898852.5993026 / 1.68969281257735e+020 2490040008.25084 1.0
GS_1_1_p10 a1 ref LAPLACE a1 ref 1.12604303476214e+019 -2211382456.07342 / 4.48491968981673e+020 7196399665.83938 1.0
GS_1_1_p11 a1 ref LAPLACE a1 ref 8.09081649809711e+019 5030679762.09104 / 5.23427028000852e+020 10824342676.9044 1.0
GS_1_1_p12 a1 ref LAPLACE a1 ref -1.21944476987558e+020 -62736901.3777725 / 6.82809627144288e+020 9303969734.54248 1.0
GS_1_1_p13 a1 ref LAPLACE a1 ref 2.24835721409947e+020 2332926449.93934 / 1.13679571461211e+021 9821406713.30654 1.0
```

Multiport theory for interconnect analysis

Frequency-domain analysis with S-parameter models

Channel analysis with S-parameter models in frequency domain (FD)

1. Convert S-parameters of components into admittance (Y) or impedance (Z) parameters and use MNA (sparse matrices) $Y \cdot \overline{v} = \overline{i}$
2. Convert S-parameters of components into scattering T-parameters (or ABCD) and multiply T-matrices (concatenation) $T = T_1 \cdot T_2 \cdot T_2 \cdot \ldots$
3. Sparse S-matrix reduction techniques (Monaco/Tiberio, 1970…) $\left(S - \Gamma\right) \cdot \overline{a} = \overline{c}$
4. Unite S-parameters of components using Redheffer star product



$$\mathbf{S}_{11}^{(AB)} = \mathbf{S}_{11}^{(A)} + \mathbf{S}_{12}^{(A)} \left[\mathbf{I} - \mathbf{S}_{11}^{(B)} \mathbf{S}_{22}^{(A)}\right]^{-1} \mathbf{S}_{11}^{(B)} \mathbf{S}_{21}^{(A)}$$
$$\mathbf{S}_{12}^{(AB)} = \mathbf{S}_{12}^{(A)} \left[\mathbf{I} - \mathbf{S}_{11}^{(B)} \mathbf{S}_{22}^{(A)}\right]^{-1} \mathbf{S}_{12}^{(B)}$$
$$\mathbf{S}_{21}^{(AB)} = \mathbf{S}_{21}^{(B)} \left[\mathbf{I} - \mathbf{S}_{22}^{(A)} \mathbf{S}_{11}^{(B)}\right]^{-1} \mathbf{S}_{21}^{(A)}$$
$$\mathbf{S}_{22}^{(AB)} = \mathbf{S}_{22}^{(B)} + \mathbf{S}_{21}^{(B)} \left[\mathbf{I} - \mathbf{S}_{22}^{(A)} \mathbf{S}_{11}^{(B)}\right]^{-1} \mathbf{S}_{22}^{(A)} \mathbf{S}_{12}^{(B)}$$

Redheffer, R., Difference equations and functional equations in transmission-line theory," Modern Mathematics for the Engineer,E. F. Beckenbach, ed., Vol. 12, 282{337, McGraw-Hill, New York, 1961.

Interpolation and extrapolation may be needed for tabulated S-parameters of components

Interconnect simulation results in frequency domain

- Single-ended (terminal) and mixed-mode S-parameters
  - Reflection parameters (S[i,i], S[Di,Di])
  - Transmission and coupling parameters (S[i,j], S[Di,Dj], i!=j)
  - Mode conversion parameters (S[Di,Cj])
- Compliance metrics (interconnects only)
  - Insertion Loss (IL=-20log(|Sij|)), and Return Loss (RL=-20log(|Sii|))
  - Fitted Insertion Loss or Fitted Attenuation
  - Insertion Loss Deviation, Multiple Reflection
  - Power Sum Crosstalk (PSXT)
  - Near End PSXT (PSNEXT)
  - Far End PSXT (PSFEXT)
  - Common to Differential PSXT (CDPSXT)
  - Insertion Loss to PSXT Ratio (ICR)
  - Transmitter and Receiver Mismatch

IEEE 802.03ap – Annex 69B: Interconnect characteristics
SFF-8431 Specifications for Enhanced Small Form Factor Pluggable Module SFP+ - Appendix A: SFI Channel Recommendations
S. Sercu, V. Balasubramanian, J. De Geest, S. Smith, Compliance Testing of Passive Interconnects, DesignCon 2010

IEEE 802.03ap – Annex 69B: Interconnect characteristics (5 compliance metrics)

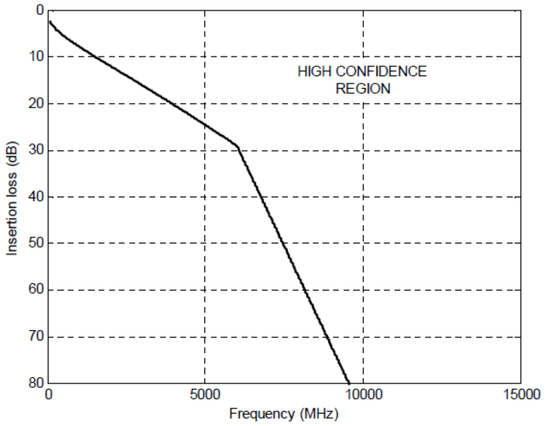


Figure 69B–5—Insertion loss limit for 10GBASE-KR

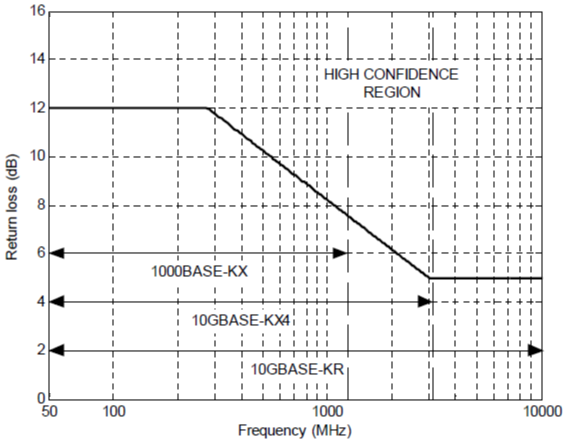


Figure 69B–7—Return loss limit

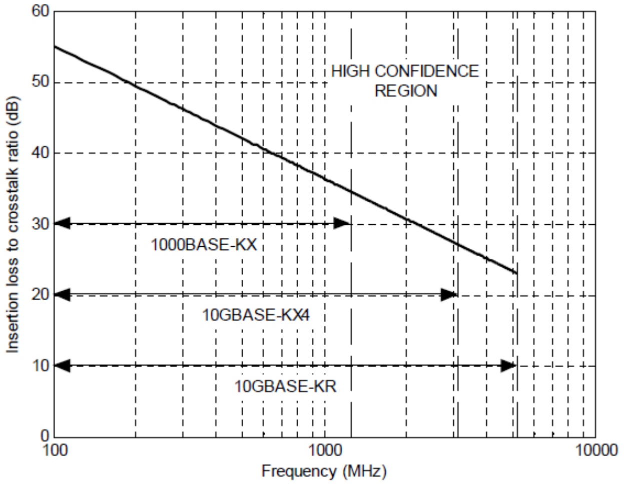


Figure 69B–8—Insertion loss to crosstalk ratio limit

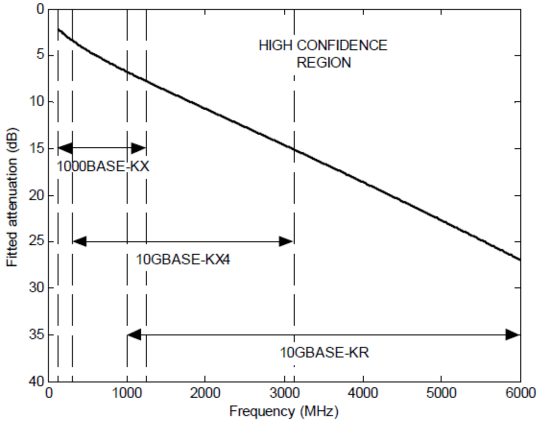


Figure 69B–2—Fitted attenuation limit

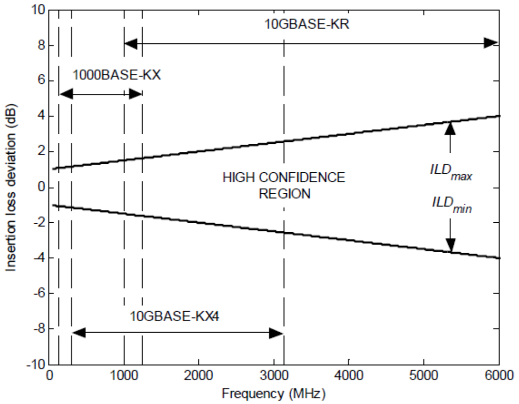
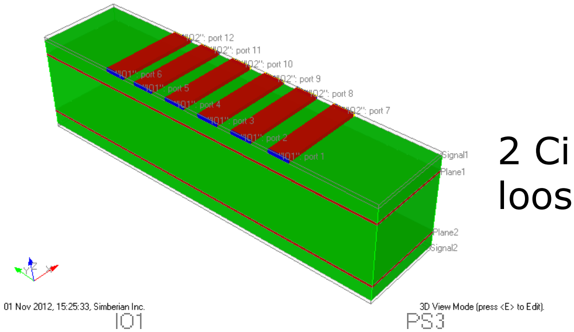


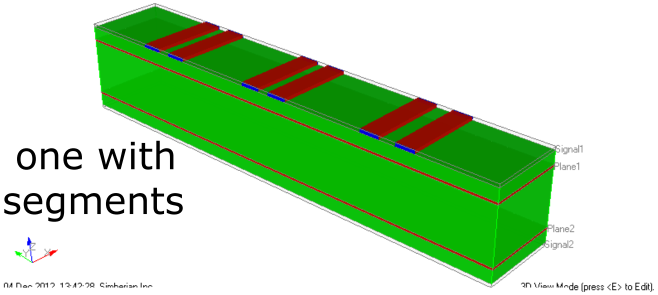
Figure 69B–6—Insertion loss deviation limits

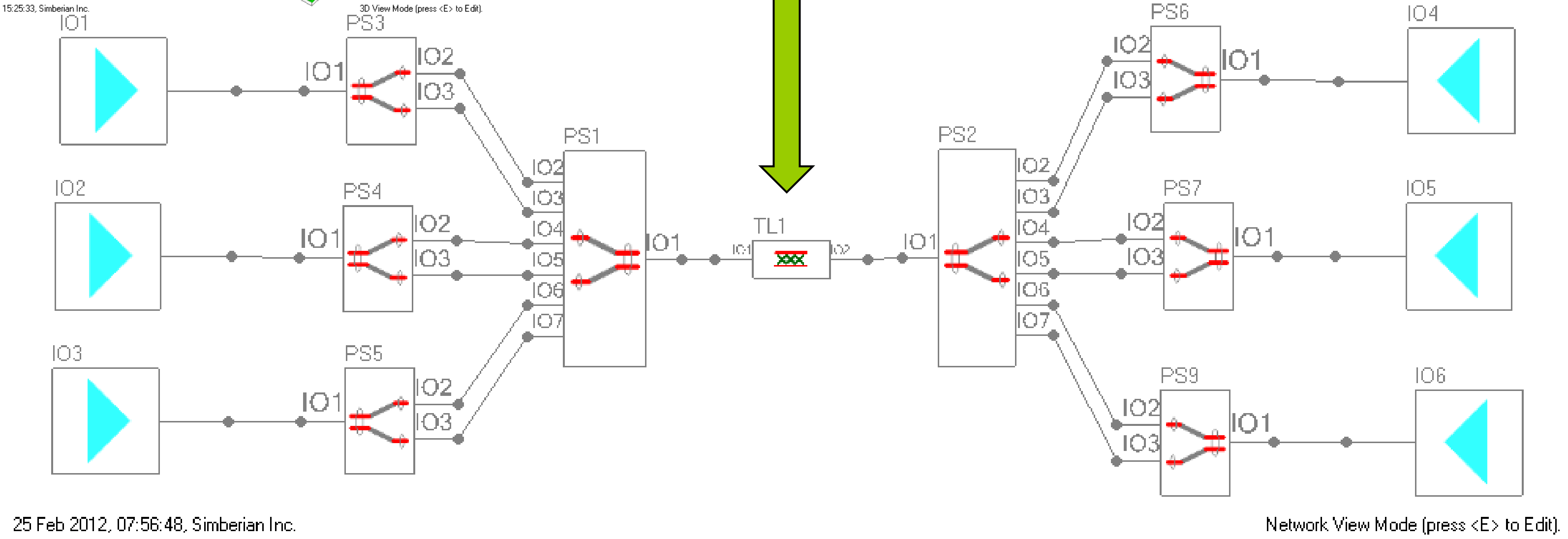
IL, Fitted Attenuation, RL and ILD may be plotted with inversed sign for consistency with S-parameters plotted in dB

Example of channel compliance evaluation



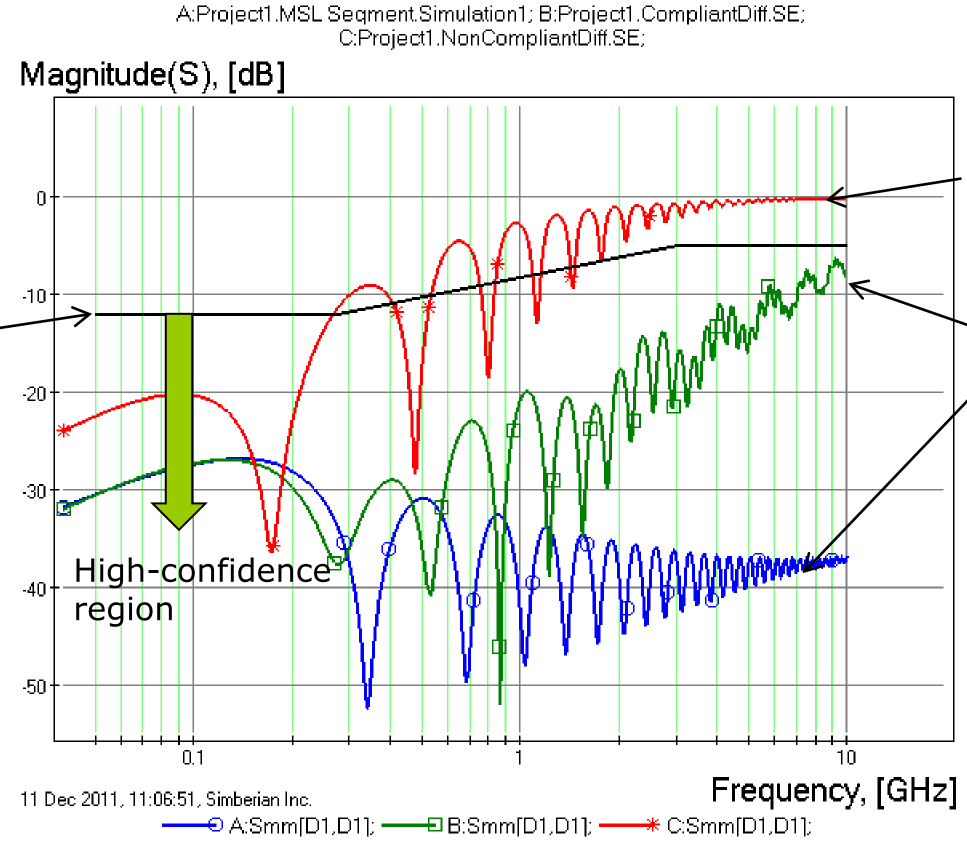
2 Circuits: one with tightly and one with loosely coupled microstrip line segments





Return or Reflection Loss (RL)

- Compliance with IEEE 802.03ap – Annex 69B



Failed

Passed

Compliance limit line

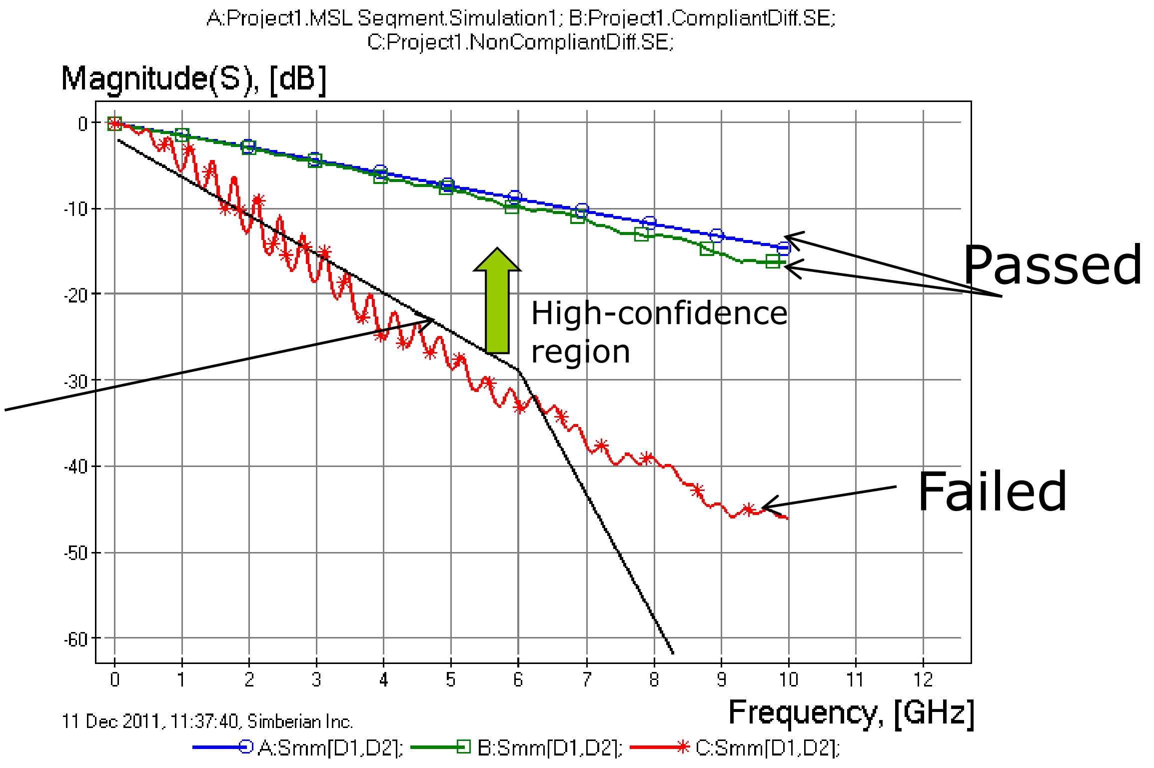
Differential reflection coefficients

Example #113 at http://kb.simberian.com/SimbeorExamples.php

Insertion Loss (IL)

- Compliance with IEEE 802.03ap – Annex 69B

Example #113 at http://kb.simberian.com/SimbeorExamples.php



Passed

High-confidence region

Compliance limit line

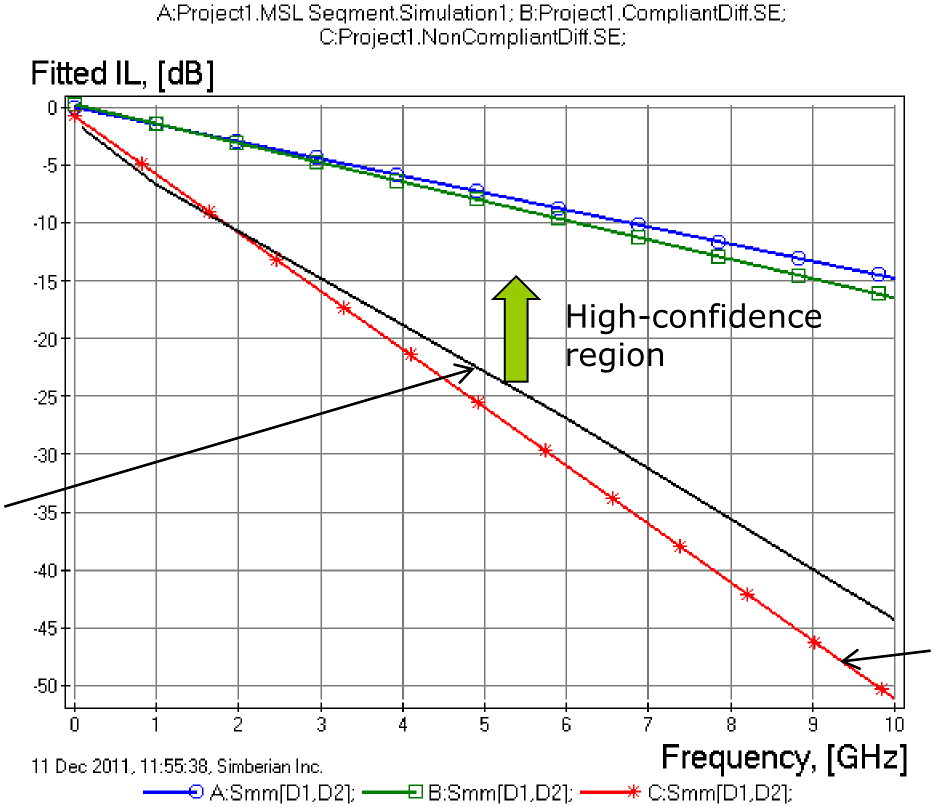
Failed

Differential transmission coefficients

Fitted Insertion Loss (Fitted IL) – same as Fitted Attenuation

- Compliance with IEEE 802.03ap – Annex 69B

$$FittedIL_{i,j} = a + c \cdot f$$



High-confidence region

Passed

Failed

Compliance limit line

Differential transmission coefficients

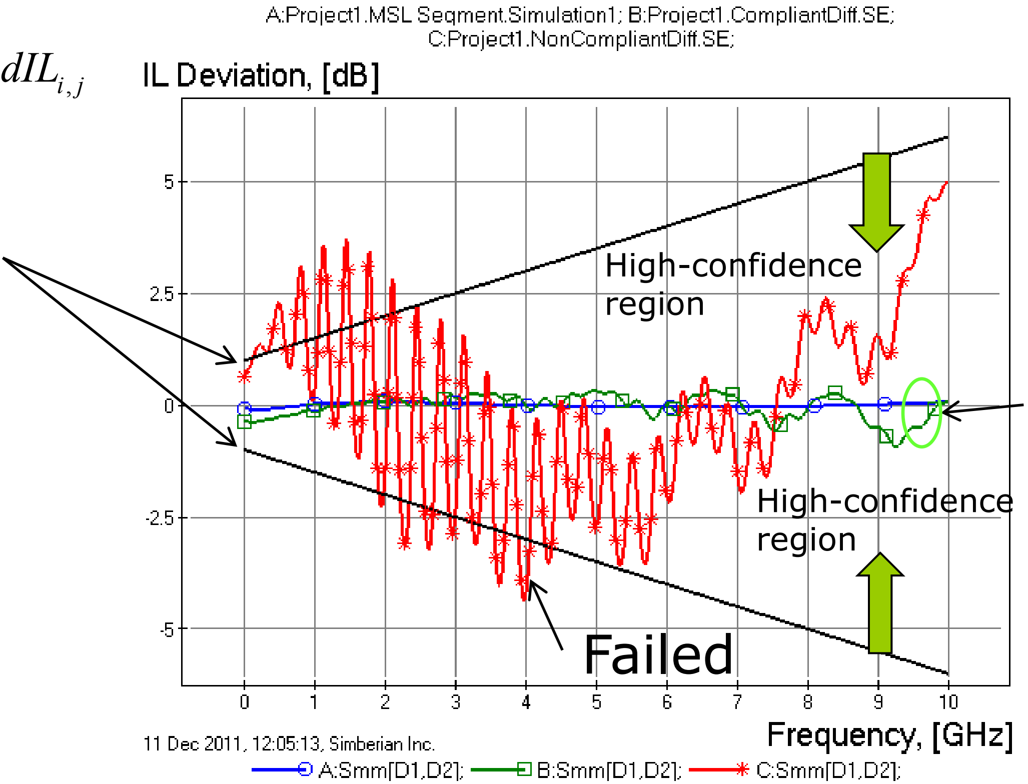
Example #113 at http://kb.simberian.com/SimbeorExamples.php

Insertion Loss Deviation (ILD)

- Compliance with IEEE 802.03ap – Annex 69B

$$ILD_{i,j} = IL_{i,j} - FittedIL_{i,j}$$

Compliance limit lines



Passed

Differential transmission coefficients

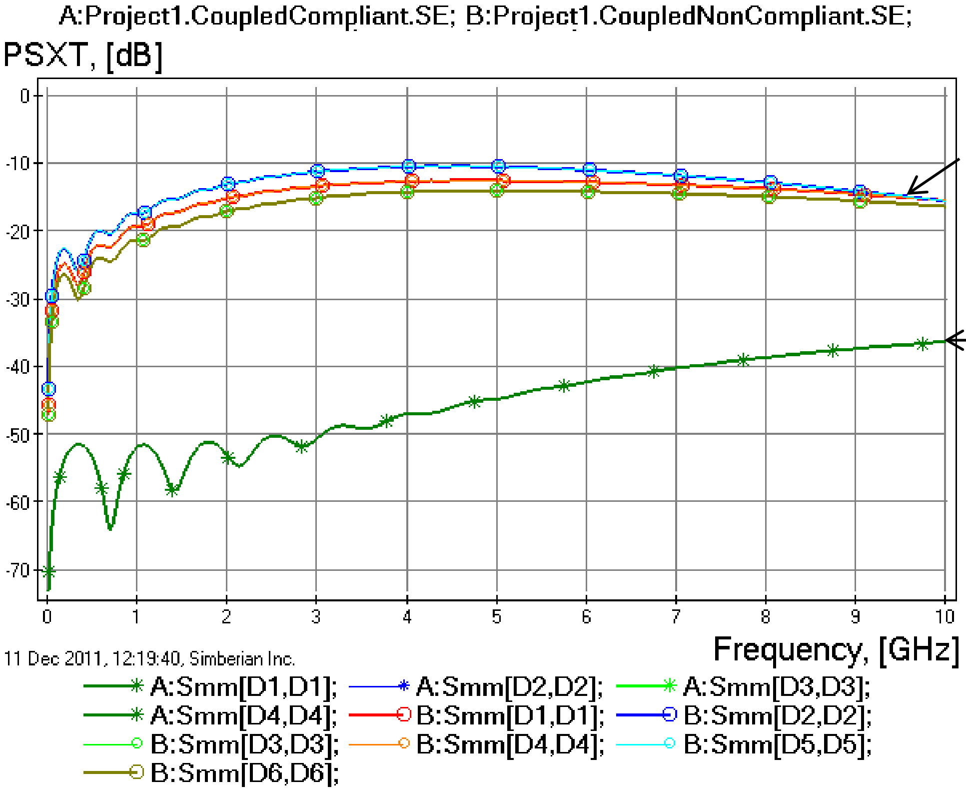
Example #113 at http://kb.simberian.com/SimbeorExamples.php

Power Sum Crosstalk (PSXT)

- Metric to set up crosstalk limit and compute ICR

$$PSXT_i = 10 \cdot \log\left(\sum_{j \in \Omega_{XT}} \left| S_{i,j} \right|^2 \right)$$

Power-sum cross-talks:
PSXT – total differential or single-ended crosstalk
PSNEXT – near end PSXT
PSFEXT – far-end PSXT
CDPSXT – common to differential PSXT



Large cross-talk (channel did not pass ICR)

Relatively small cross-talk (channel passed ICR specs)

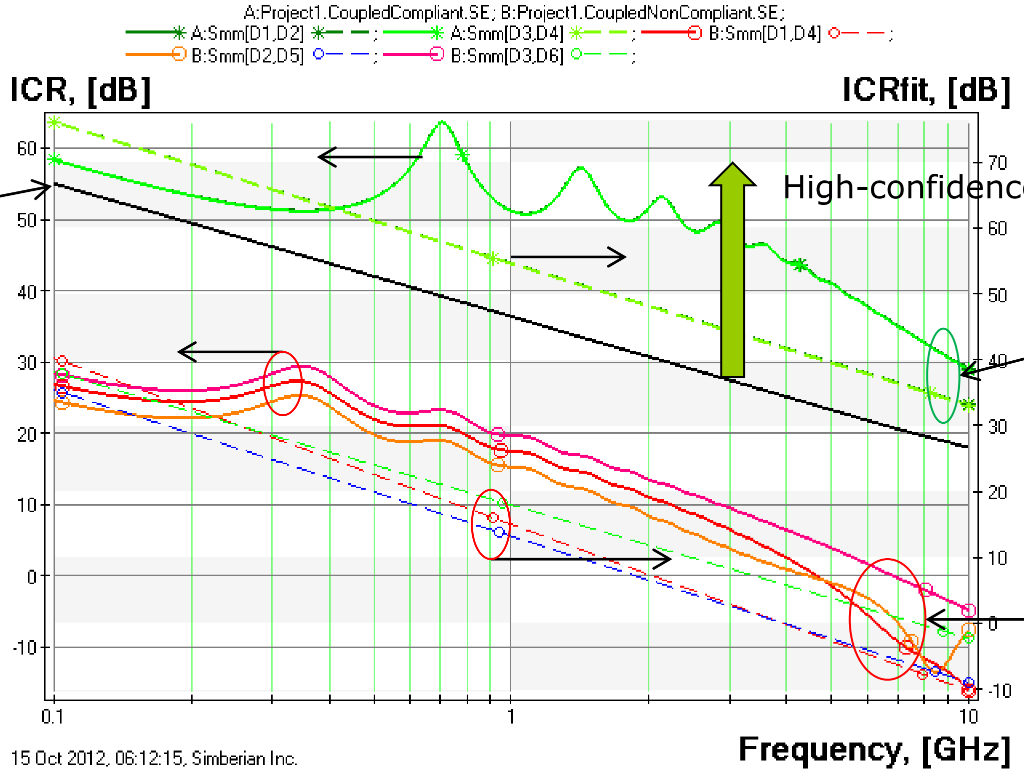
Example #113 at http://kb.simberian.com/SimbeorExamples.php

Insertion Loss to Crosstalk Ratio (ICR)

- Compliance with IEEE 802.03ap – Annex 69B

$$ICR_{i,j} = IL_{i,j} - PSXT_i$$

Only differential transmission coefficients



Compliance limit line

Passed

Failed

Example #113 at http://kb.simberian.com/SimbeorExamples.php

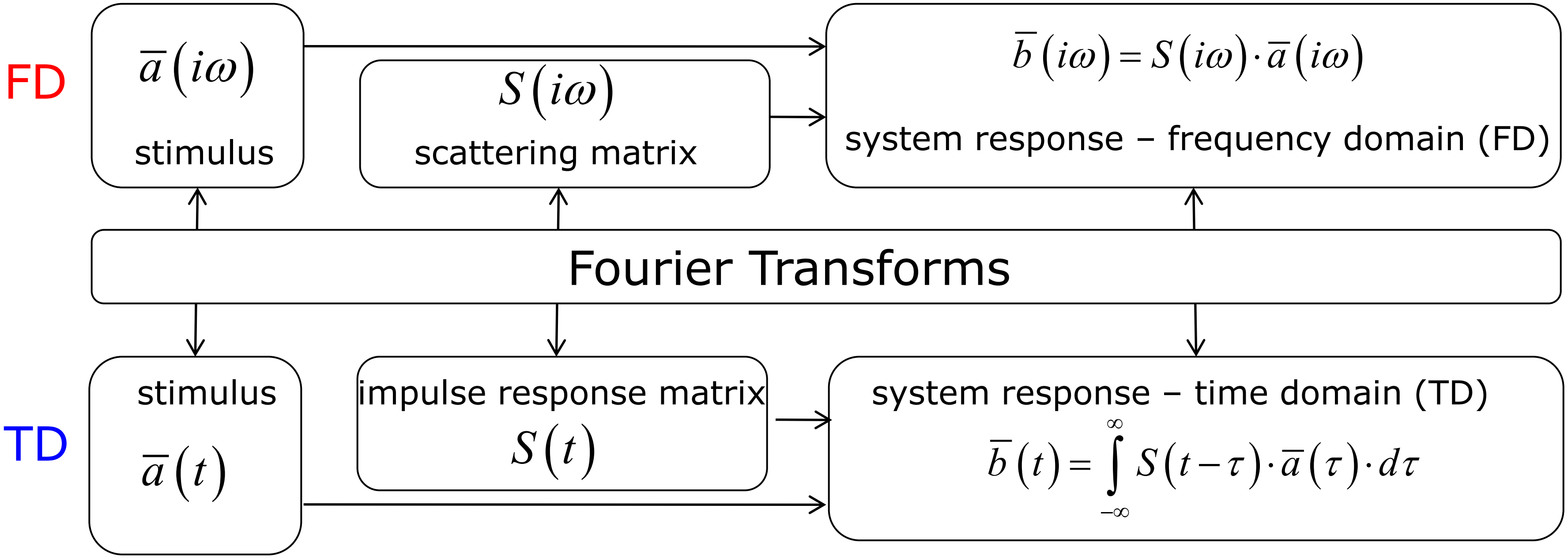
Multiport theory for interconnect analysis

Time-domain analysis with S-parameter models

System response computation requires frequency-continuous S-parameters from DC to infinity

Frequency domain is preferable for analysis of interconnects

$$S(i\omega) = \int_{-\infty}^{\infty} S(t) \cdot e^{-i\omega t} \cdot dt, \quad S(i\omega) \in C^{N \times N}$$



FD

$\bar{a}(i\omega)$

stimulus

$S(i\omega)$

scattering matrix

$$\bar{b}(i\omega) = S(i\omega) \cdot \bar{a}(i\omega)$$

system response – frequency domain (FD)

Fourier Transforms

TD

stimulus

$\bar{a}(t)$

impulse response matrix

$S(t)$

system response – time domain (TD)

$$\bar{b}(t) = \int_{-\infty}^{\infty} S(t-\tau) \cdot \bar{a}(\tau) \cdot d\tau$$

$$S(t) = \frac{1}{2\pi} \int_{-\infty}^{\infty} S(i\omega) \cdot e^{i\omega t} \cdot d\omega, \quad S(t) \in R^{N \times N}$$

Possible approximations for discrete models

- Inversed Discrete Fourier Transform (IDFT) and convolution (uncontrollable error)
  - Slow and may require interpolation and extrapolation of tabulated S-parameters
  - See more on typical problems with IDFT in
    P. Pupalaikis, “The Relationship Between Discrete-Frequency S-Parameters and Continuous-Frequency Responses”, DesignCon, Santa Clara CA, 2012
- Approximate discrete S-parameters with frequency-continuous rational functions (controllable error)
  - Accuracy control over defined frequency band (RMS error)
  - Causal functions (with passivity enforcement) defined from DC to infinity with analytical impulse response
  - Fast recursive convolution algorithm to compute TD response
  - Results consistent in time and frequency domains
- Not all models are suitable for either approach

Rational approximation of S-parameters as the frequency-continuous model

$$\bar{b} = S\cdot\bar{a},\quad S_{i,j} = \left.\frac{b_i}{a_j}\right|_{a_k=0\ k\neq j} \Rightarrow S_{i,j}(i\omega) = \left[d_{ij} + \sum_{n=1}^{N_{ij}}\left(\frac{r_{ij,n}}{i\omega - p_{ij,n}} + \frac{r^*_{ij,n}}{i\omega - p^*_{ij,n}}\right)\right]\cdot e^{-s\cdot T_{ij}}$$

$$s = i\omega,\ d_{ij} - values\ at\ \infty,\ N_{ij} - number\ of\ poles,$$

$$r_{ij,n} - residues,\ p_{ij,n} - poles\,(real\ or\ complex), T_{ij} - optional\ delay$$

Continuous functions of frequency defined from DC to infinity

- Impulse response is analytical, real and delay-causal:

$$S_{i,j}(t) = 0,\ t < T_{ij}$$

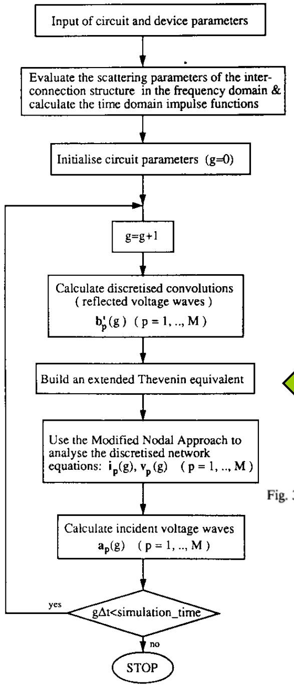
$$S_{i,j}(t) = d_{ij}\delta\left(t - T_{ij}\right) + \sum_{n=1}^{N_{ij}}\left[r_{ij,n}\cdot\exp\left(p_{ij,n}\cdot\left(t - T_{ij}\right)\right) + r^*_{ij,n}\cdot\exp\left(p^*_{ij,n}\cdot\left(t - T_{ij}\right)\right)\right],\ t \geq T_{ij}$$

- Stable $\mathrm{Re}\left(p_{ij,n}\right) < 0$
- Passive if $eigenvals\left[S(\omega)\cdot S^*(\omega)\right] \leq 1\ \forall\omega,\ from\ 0\ to\ \infty$
- Reciprocal if $S_{i,j}(\omega) = S_{j,i}(\omega)$

May require enforcement

Circuit analysis with S-parameter models in time domain

- T. Dhaene, L. Martens, D. De Zutter, IEEE Trans. On Circuit and Systems, v. 39, N 11, p. 928-937, 1992.
  IFFT+Convolution+MNA



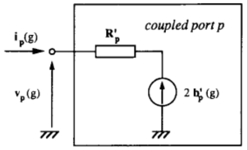


Fig. 3. Extended (time-domain) Thevenin equivalent of coupled port p at time $g\Delta t$.

Fig. 5. Flow chart of simulation algorithm.

- V. Dmitriev-Zdorov's algorithm (DesignCon 2006):
  Rational macro-model+Recursive convolution+MNA

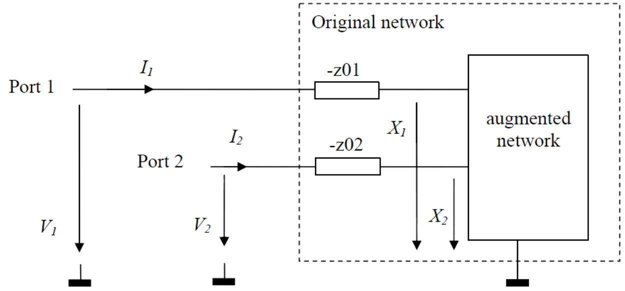


Fig. 1c. Original and augmented network

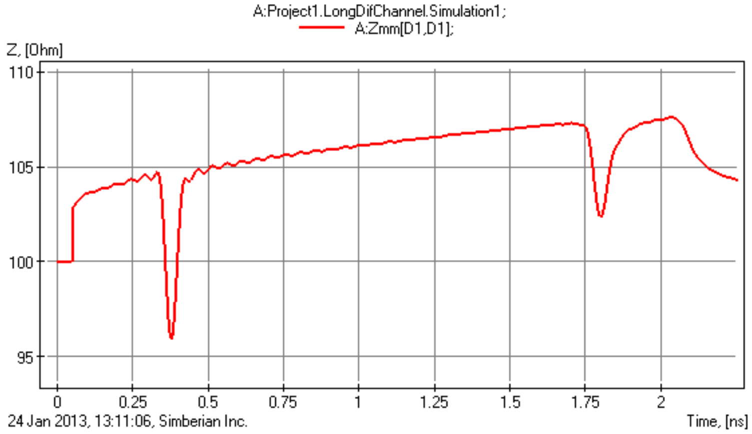
Model stamp:

$$\begin{bmatrix} I \\ 0 \end{bmatrix} = \begin{bmatrix} -Z_0^{-1} & Z_0^{-1} \\ Z_0^{-1} & -Z_0^{-1}+Y_{AN} \end{bmatrix} \begin{bmatrix} V \\ X \end{bmatrix}$$

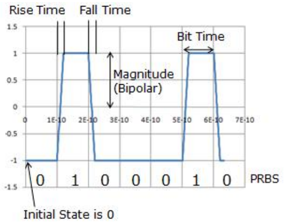
$$Y_{AN} = \frac{1}{2}Z_0^{-1} + H(s), \text{ where } H(s) = -\frac{1}{2}Z_0^{-1/2} S Z_0^{-1/2}.$$

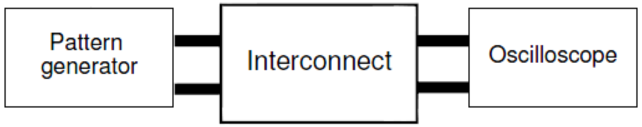
Interconnect simulation results in time domain

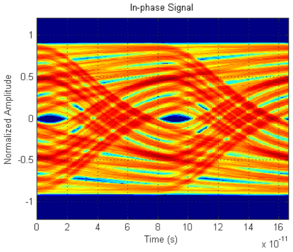
- Step response, TDR/TDT
- Impulse or pulse responses
- ….



- Eye diagrams
  - Bath tub diagram, Bit Error Rate, …
  - Jitter, Inter Symbol Interference (ISI) noise





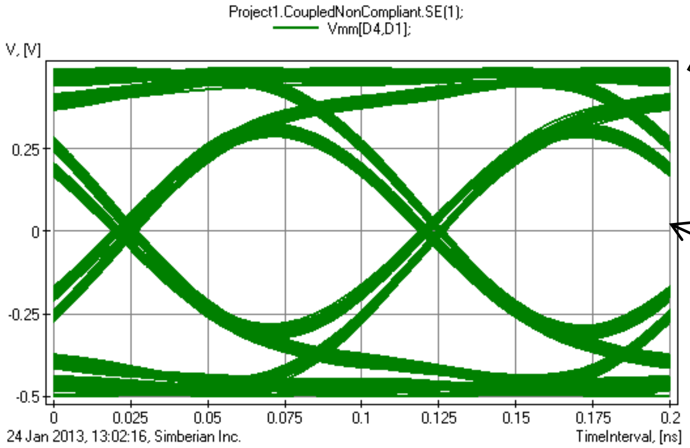


May be not usable due to high losses

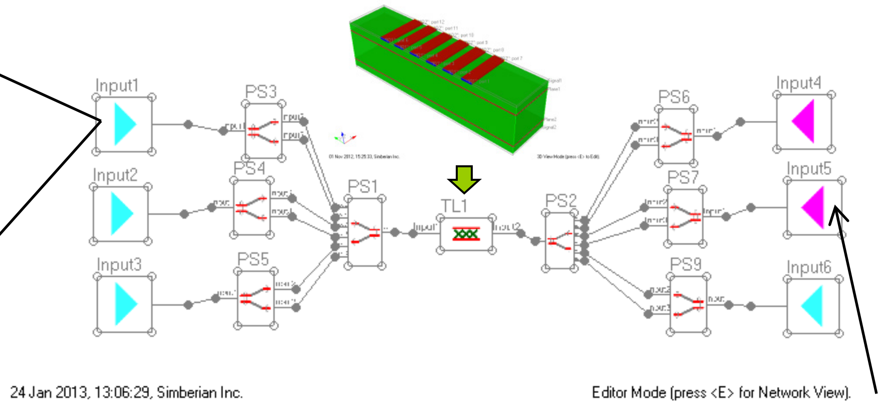
Analysis with Tx/Rx models may be required (not covered in this tutorial)

Example of TD interconnect analysis

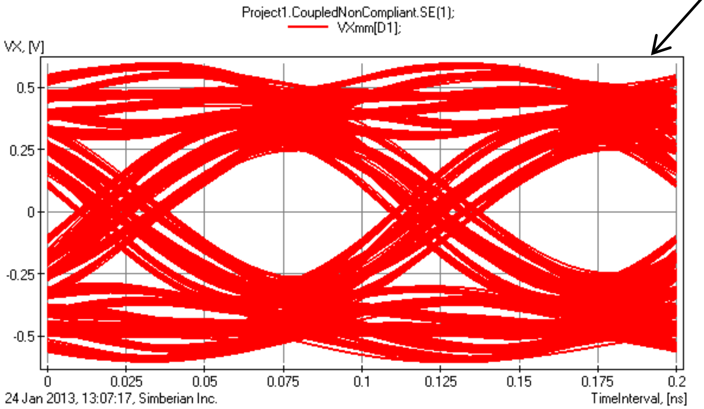
- X-talk between 10 G channels



Eye diagram without x-talk (no signal in coupled channels)



aggressor



Eye diagram with asynchronous 10G signal in coupled channel (FEXT)

Multiport theory for interconnect analysis

Quality of S-parameter models

Common S-parameter model defects

- Model bandwidth deficiency
  - Limited capabilities of solvers and measurement equipment
  - Need DC point or allow extrapolation
  - High frequencies must be defined by the signal spectrum
- Model discreteness
  - Touchstone models are matrix elements at a set of frequencies
  - Interpolation and extrapolation may be needed both for time and frequency domain analyses
- Model distortions due to
  - Measurement or simulation artifacts
  - Passivity violations and local “enforcements”
  - Causality violations and “enforcements”
- Human mistakes of model developers and users
- How to estimate quality of the models?

Good models of interconnects

- Must have sufficient bandwidth matching signal spectrum
- Must be appropriately sampled to resolve all resonances

- Must be passive (do not generate energy)

$$P_{in} = \overline{a}^* \cdot \left[U - S^* S \right] \cdot \overline{a} \ \geq 0 \quad \Rightarrow \quad eigenvals\left[S^* \cdot S \right] \leq 1$$

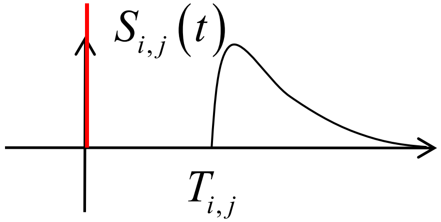
from DC to infinity!

- Must be reciprocal (linear reciprocal materials used in PCBs)

$$S_{i,j} = S_{j,i} \quad or \quad S = S^t$$

- Must be causal (have causal step or impulse response or satisfy KK relations)

$$S_{i,j}(t) = 0, \ \ t < T_{ij}$$



$$S(i\omega) = \frac{1}{i\pi} PV \int_{-\infty}^{\infty} \frac{S(i\omega')}{\omega - \omega'} \cdot d\omega'$$

Model bandwidth and sampling

- If no DC point, the lowest frequency in the sweep should be
  - Below the transition to skin-effect (1-50 MHz for PCB applications)
  - Below the first possible resonance in the system (important for cables, L is physical length)

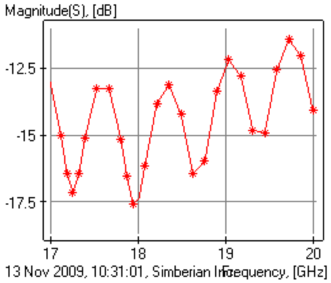
$$L < \frac{\lambda}{4} = \frac{c}{4 f_l \cdot \sqrt{\varepsilon_{eff}}} \Rightarrow f_l < \frac{c}{4L \cdot \sqrt{\varepsilon_{eff}}}$$

- The highest frequency in the sweep must be defined by the required resolution in time-domain or by spectrum of the signal (by rise time or data rate)

$$f_h > \frac{1}{2t_r} \qquad f_h > K \cdot f_{s1}$$

- The sampling is very important for DFT and convolution-based algorithms, but not so for algorithms based on fitting
  - There must be 4-5 frequency point per each resonance
  - The electrical length of a system should not change more than quarter of wave-length between two consecutive points

$$df < \frac{c}{4L \cdot \sqrt{\varepsilon_{eff}}}$$



Model quality metrics (0-100%)

First introduced at IBIS forum at DesignCon 2010

- Passivity Quality Measure:

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

$$PW_n = 0 \ if \ \ PM_n < 1.00001; \ otherwise \ PW_n = \frac{PM_n - 1.00001}{0.1}$$

$$PM_n = \sqrt{\max\left[eigenvals\left(S^*\left(f_n\right) \cdot S\left(f_n\right)\right)\right]}$$

should be >99%

- Reciprocity Quality Measure:

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

$$RW_n = 0 \ if \ \ RM_n < 10^{-6}; \ otherwise \ RW_n = \frac{RM_n - 10^{-6}}{0.1}$$

$$RM_n = \frac{1}{N_s}\sum_{i,j}\left|S_{i,j}\left(f_n\right) - S_{j,i}\left(f_n\right)\right|$$

should be >99%

- Causality Quality Measure: Minimal ratio of clockwise rotation measure to total rotation measure in % (should be >80% for numerical models)

Preliminary quality estimation metrics

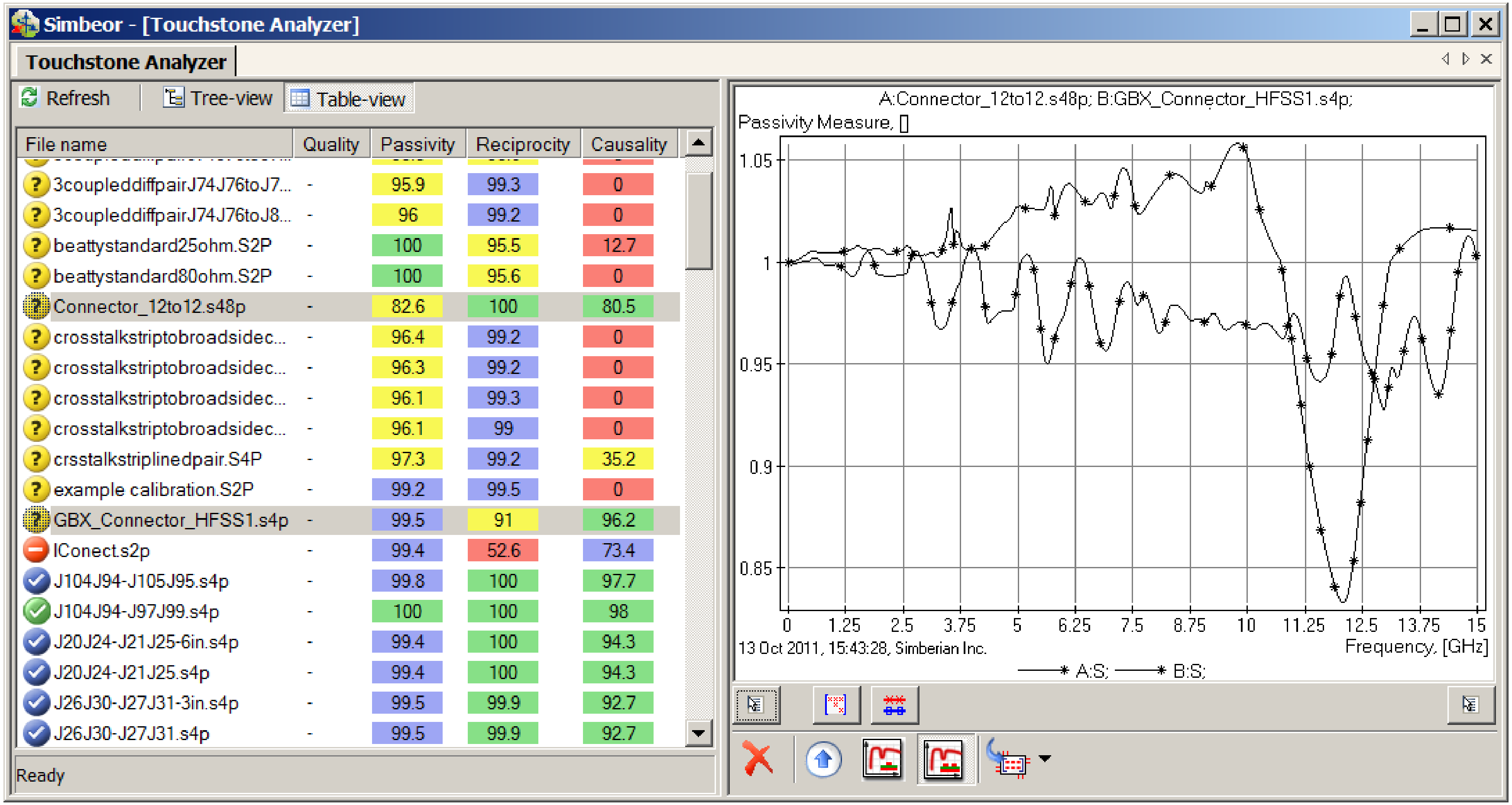
- Preliminary Touchstone model quality can be estimated with Passivity, Reciprocity and Causality quality metrics (PQM, RQM, CQM)

Metric/Model Icon	✅ - good	☑ - acceptable	? - inconclusive	⛔ - bad
Passivity	[100, 99.9]	(99.9, 99]	(99, 80]	(80, 0]
Reciprocity	[100, 99.9]	(99.9, 99]	(99, 80]	(80, 0]
Causality	[100, 80]	(80, 50]	(50, 0]	-----

Color code	Passivity (PQM)	Reciprocity (RQM)	Causality (CQM)
Green – good	[99.9, 100]	[99.9, 100]	[80, 100]
Blue – acceptable	[99, 99.9)	[99, 99.9)	[50, 80)
Yellow – inconclusive	[80, 99)	[80, 99)	[20, 50)
Red - bad	[0, 80)	[0, 80)	[0, 20)

Example of preliminary quality estimation in Simbeor Touchstone Analyzer™

Small passivity & reciprocity violations in most of the models
Low causality in some measured data due to noise at high frequencies



File name	Quality	Passivity	Reciprocity	Causality
? 3coupleddiffpairJ74J76toJ7...	-	95.9	99.3	0
? 3coupleddiffpairJ74J76toJ8...	-	96	99.2	0
? beattystandard25ohm.S2P	-	100	95.5	12.7
? beattystandard80ohm.S2P	-	100	95.6	0
? Connector_12to12.s48p	-	82.6	100	80.5
? crosstalkstriptobroadsidec...	-	96.4	99.2	0
? crosstalkstriptobroadsidec...	-	96.3	99.2	0
? crosstalkstriptobroadsidec...	-	96.1	99.3	0
? crosstalkstriptobroadsidec...	-	96.1	99	0
? crsstalkstriplinedpair.S4P	-	97.3	99.2	35.2
? example calibration.S2P	-	99.2	99.5	0
? GBX_Connector_HFSS1.s4p	-	99.5	91	96.2
⊖ IConect.s2p	-	99.4	52.6	73.4
✓ J104J94-J105J95.s4p	-	99.8	100	97.7
✓ J104J94-J97J99.s4p	-	100	100	98
✓ J20J24-J21J25-6in.s4p	-	99.4	100	94.3
✓ J20J24-J21J25.s4p	-	99.4	100	94.3
✓ J26J30-J27J31-3in.s4p	-	99.5	99.9	92.7
✓ J26J30-J27J31.s4p	-	99.5	99.9	92.7

Final quality estimation with rational approximation

- Accuracy of discrete S-parameters approximation with frequency-continuous macro-model, passive from DC to infinity

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

original tabulated data

$$S_{i,j}(i\omega)=\left[d_{ij}+\sum_{n=1}^{N_{ij}}\left(\frac{r_{ij,n}}{i\omega-p_{ij,n}}+\frac{r^*_{ij,n}}{i\omega-p^*_{ij,n}}\right)\right]\cdot e^{-s\cdot T_{ij}}$$

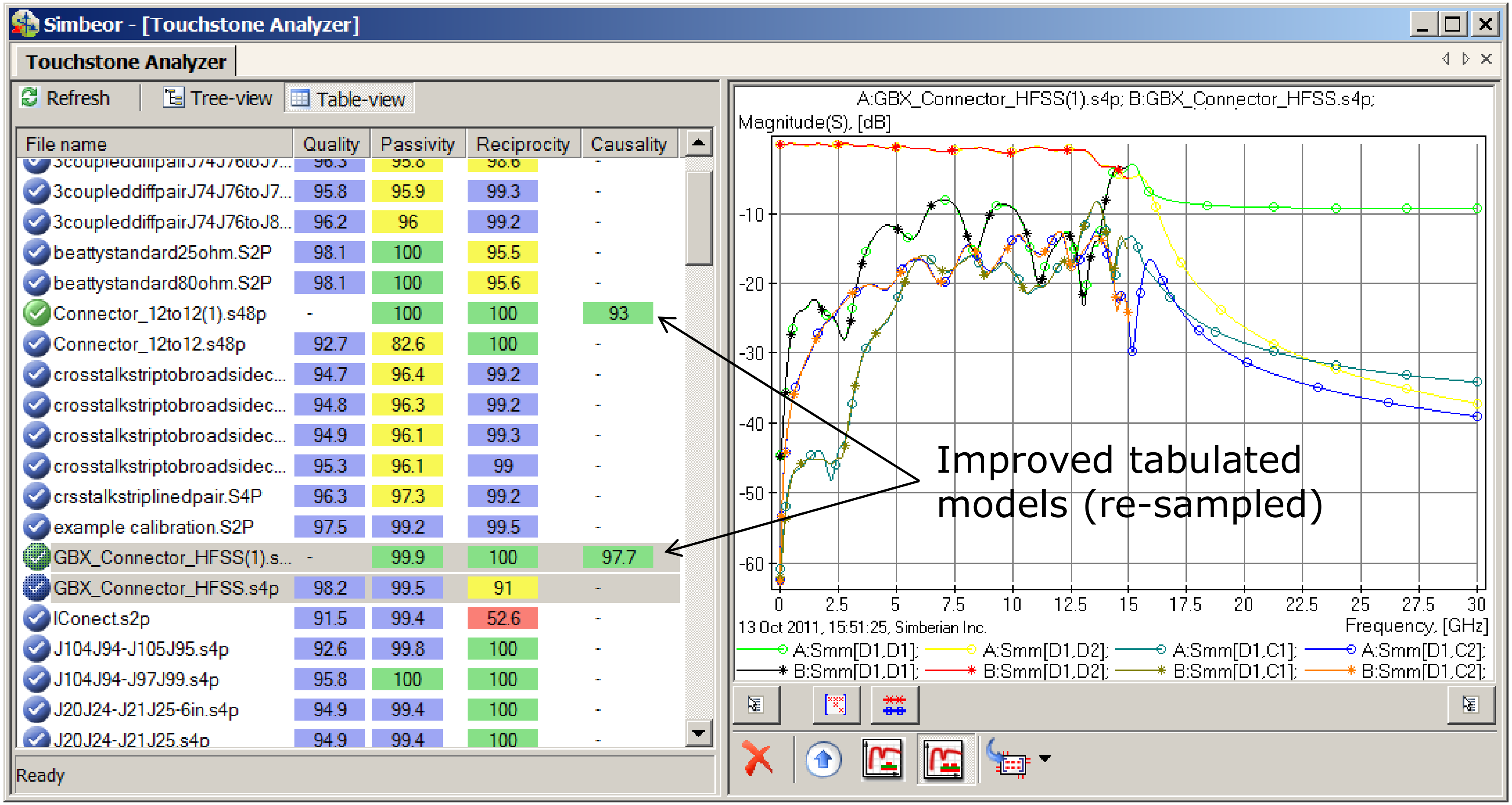
- Can be used to estimate quality of the original data

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

Model Icon/Quality	Quality Metric	RMSE
✓ - good	[99, 100]	[0, 0.01]
✓- acceptable	[90, 99)	(0.01, 0.1]
? - inconclusive	[50, 90)	(0.1, 0.5]
⊖ - bad	[0, 50)	> 0.5
? - uncertain	[0,100], not passive or not reciprocal	

Example of final quality estimation in Simbeor Touchstone Analyzer®

All rational macro-models are passive, reciprocal, causal and have acceptable accuracy (acceptable quality of original models)



File name	Quality	Passivity	Reciprocity	Causality
3coupleddiffpairJ74J76toJ7...	95.8	95.9	99.3	-
3coupleddiffpairJ74J76toJ8...	96.2	96	99.2	-
beattystandard25ohm.S2P	98.1	100	95.5	-
beattystandard80ohm.S2P	98.1	100	95.6	-
Connector_12to12(1).s48p	-	100	100	93
Connector_12to12.s48p	92.7	82.6	100	-
crosstalkstriptobroadsidec...	94.7	96.4	99.2	-
crosstalkstriptobroadsidec...	94.8	96.3	99.2	-
crosstalkstriptobroadsidec...	94.9	96.1	99.3	-
crosstalkstriptobroadsidec...	95.3	96.1	99	-
crsstalkstriplinedpair.S4P	96.3	97.3	99.2	-
example calibration.S2P	97.5	99.2	99.5	-
GBX_Connector_HFSS(1).s...	-	99.9	100	97.7
GBX_Connector_HFSS.s4p	98.2	99.5	91	-
IConect.s2p	91.5	99.4	52.6	-
J104J94-J105J95.s4p	92.6	99.8	100	-
J104J94-J97J99.s4p	95.8	100	100	-
J20J24-J21J25-6in.s4p	94.9	99.4	100	-
J20J24-J21J25.s4p	94.9	99.4	100	-

Improved tabulated models (re-sampled)

Basics of signal propagation in interconnects

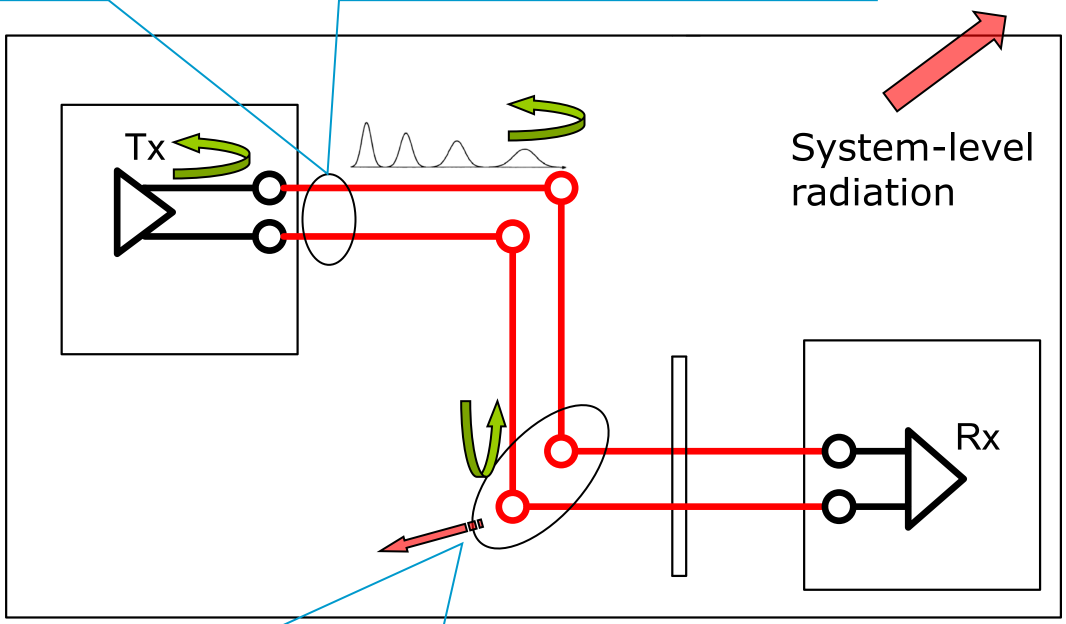
Signal degradation factors

Modeling transmission lines

Modeling via-holes and other discontinuities

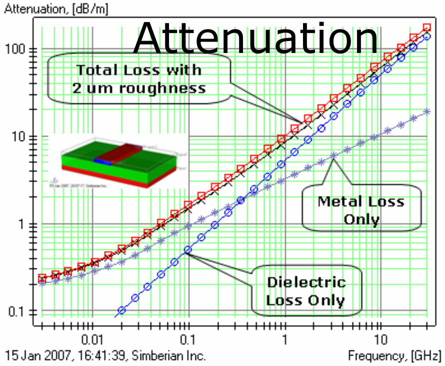
Major signal degradation factors

Transmission lines:
Attenuation and dispersion due to physical conductor and dielectric properties
High-frequency dispersion, coupling

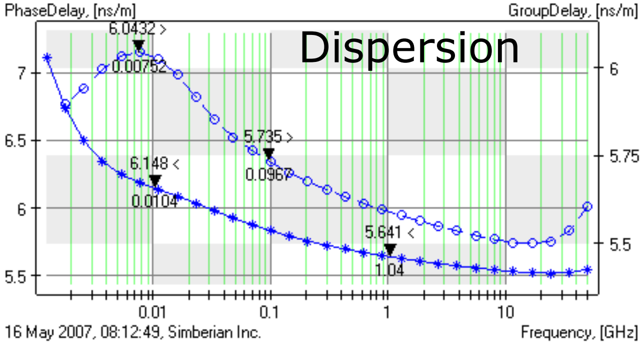


System-level radiation

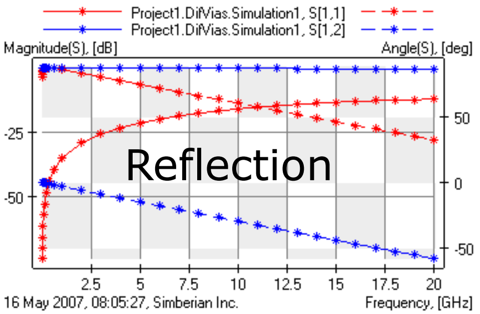
Via-hole transitions and discontinuities:
Reflection, radiation and coupling to parallel planes



15 Jan 2007, 16:41:39, Simberian Inc.



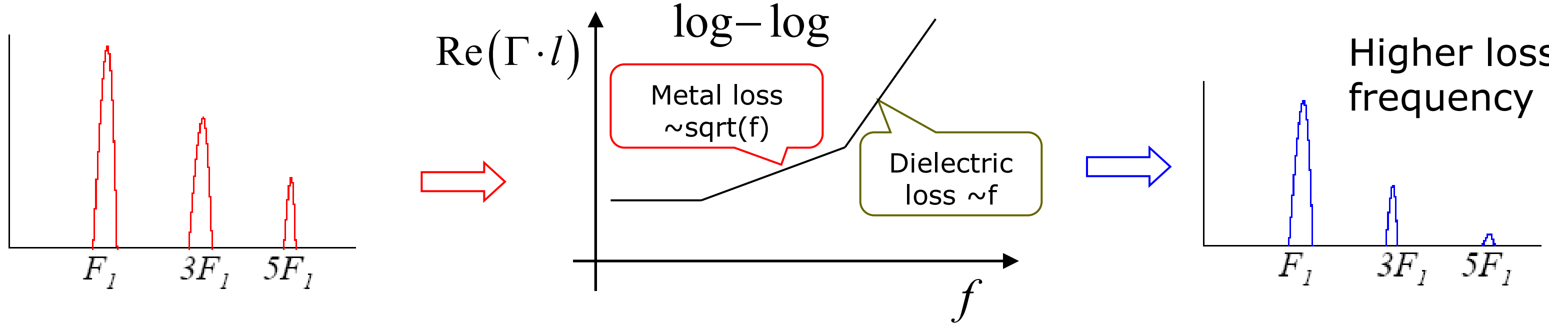
16 May 2007, 08:12:49, Simberian Inc.



16 May 2007, 08:05:27, Simberian Inc.

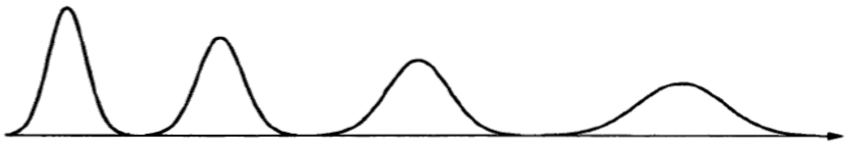
Effects of degradation factors on signal in transmission line segment (simplified view)

Attenuation effect in frequency domain



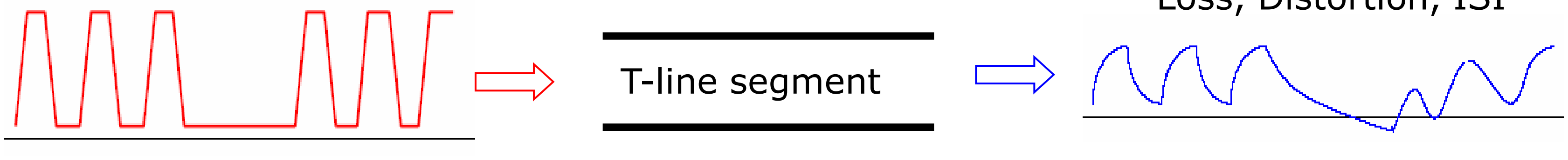
Higher loss in high-frequency harmonics

Group delay dispersion effect in time domain



Pulse becomes wider without energy loss

Combined effect in t-line segment



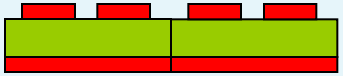
Loss, Distortion, ISI

Building models for transmission lines

2D Static and Magneto-Quasi-Static Solvers

$$\vec{E} = -\nabla\varphi - j\omega\vec{A}$$

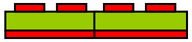
$$\vec{H} = \frac{1}{\mu}\nabla\times\vec{A}$$



2D Full-Wave Solvers

$$\nabla\times\vec{E} = -i\omega\mu\vec{H}$$

$$\nabla\times\vec{H} = i\omega\varepsilon\vec{E} + \sigma\vec{E} + \vec{J}$$



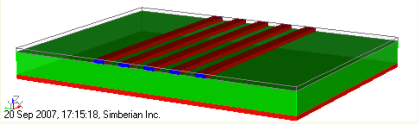
$$\vec{E} = \vec{E}_t\cdot\exp\left(-\Gamma\cdot l\right)$$

$$\vec{H} = \vec{H}_t\cdot\exp\left(-\Gamma\cdot l\right)$$

3D Full-Wave Solver

$$\nabla\times\vec{E} = -i\omega\mu\vec{H}$$

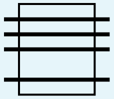
$$\nabla\times\vec{H} = i\omega\varepsilon\vec{E} + \sigma\vec{E} + \vec{J}$$

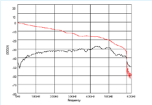


Generalized Telegrapher's equations (W-element or S-parameters)

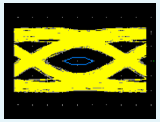
$$\frac{\partial V}{\partial x} = -\left(R(\omega) + i\omega L(\omega)\right)\cdot I$$

$$\frac{\partial I}{\partial x} = -\left(G(\omega) + i\omega C(\omega)\right)\cdot V$$





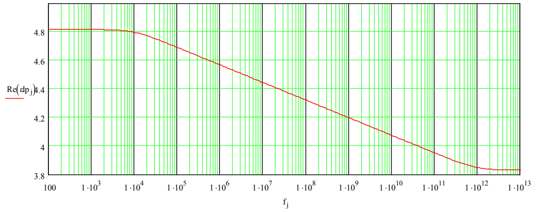
Decompositional Analysis



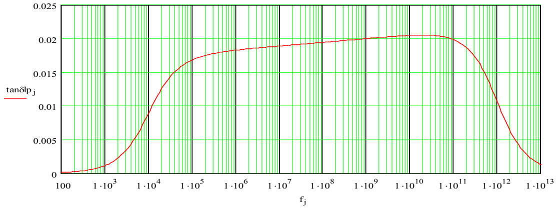
Dielectric attenuation and dispersion effects

- Dispersion of complex dielectric constant
  - Polarization changes with frequency
  - High frequency harmonics propagate faster
  - Almost constant loss tangent in broad frequency range – loss ~ frequency

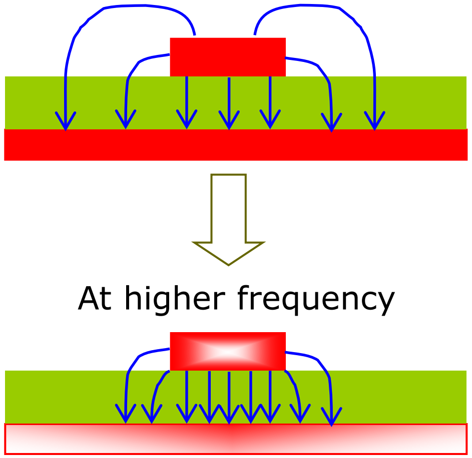
Dk vs. Frequency



Loss Tangent vs. Frequency



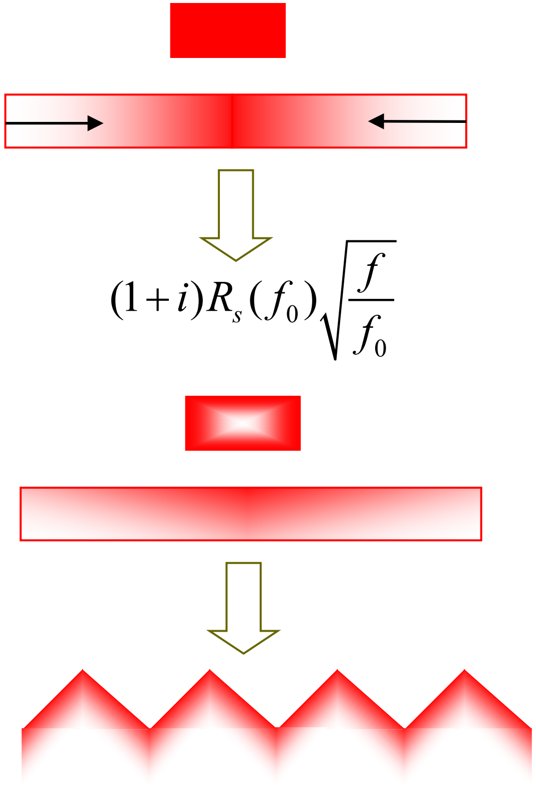
- High-frequency dispersion due to non-homogeneous dielectrics
  - TEM mode becomes non-TEM at high frequencies
  - Fields concentrate in dielectric with high Dk or lower LT
  - High-frequency harmonics propagate slower
  - Interacts with the conductor-related losses



At higher frequency

Metal attenuation and dispersion effects

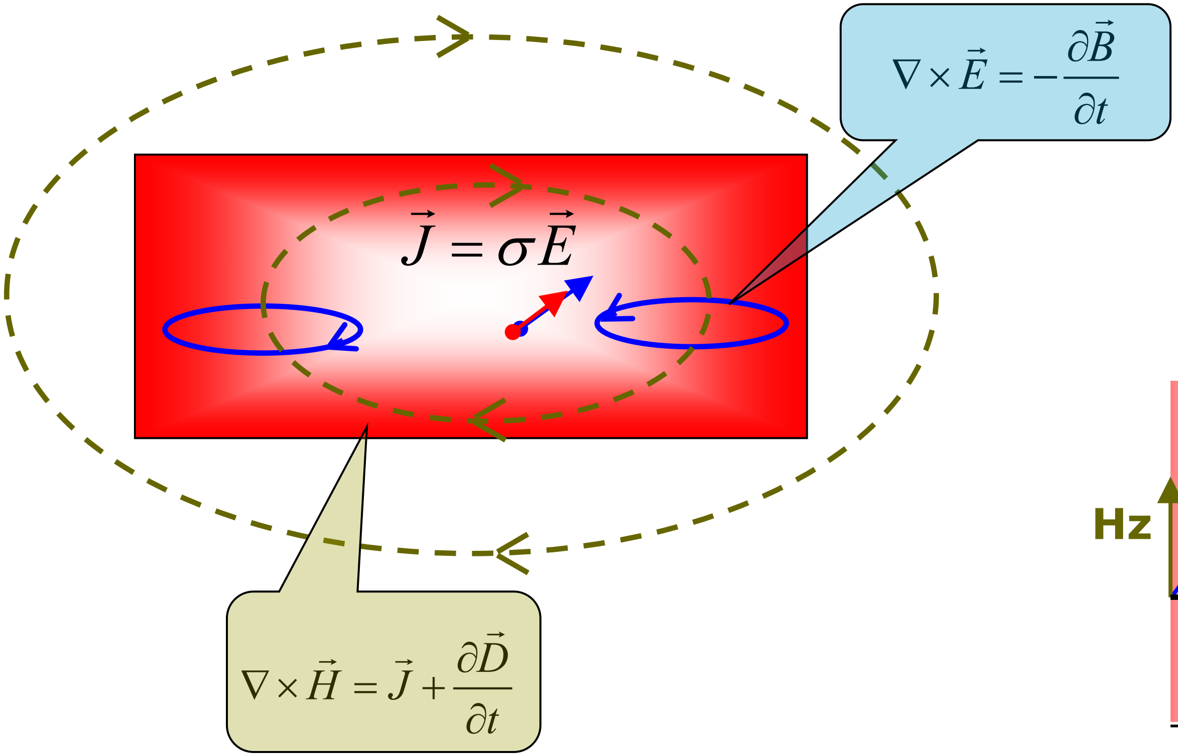
- Current crowding below strips
  - Around 10 KHz
  - Increases R and decreases L at very low frequencies, effectively at DC
- Skin-effect
  - Transition frequencies from 1 MHz to 100 GHz depending on technology
  - Wheeler's formula works for well-developed skin-effect - loss ~ sqrt(frequency)
- Skin-effect on rough surface
  - May be comparable with skin depth starting from 10 MHz
  - Increases both R and L (and possibly C)



$$(1+i)R_s(f_0)\sqrt{\frac{f}{f_0}}$$

Skin-effect: Maxwell’s equations+Ohm’s law

Current cancelation:

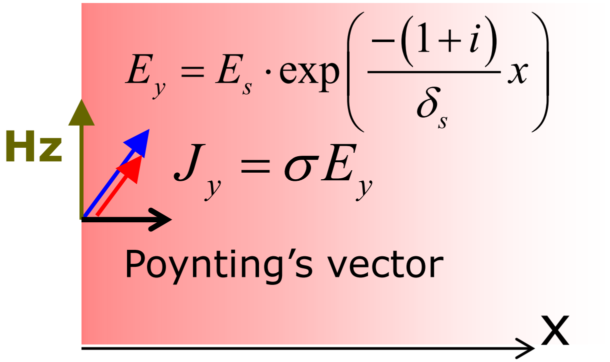


$$\vec{J} = \sigma \vec{E}$$

$$\nabla \times \vec{E} = -\frac{\partial \vec{B}}{\partial t}$$

$$\nabla \times \vec{H} = \vec{J} + \frac{\partial \vec{D}}{\partial t}$$

Plane-wave view:



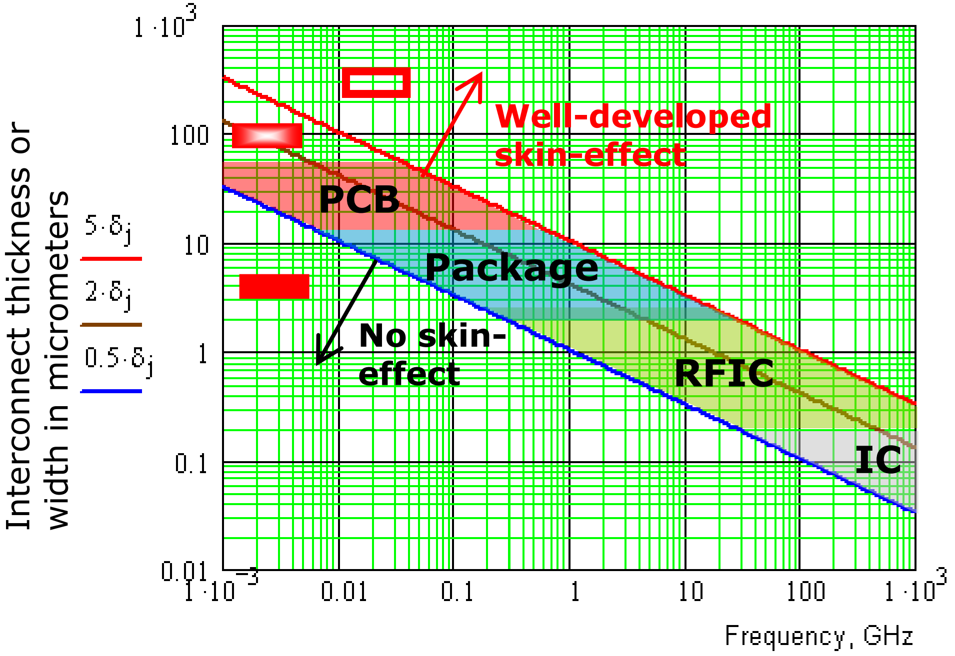
$$E_y = E_s \cdot \exp\left(\frac{-(1+i)}{\delta_s} x\right)$$

$$J_y = \sigma E_y$$

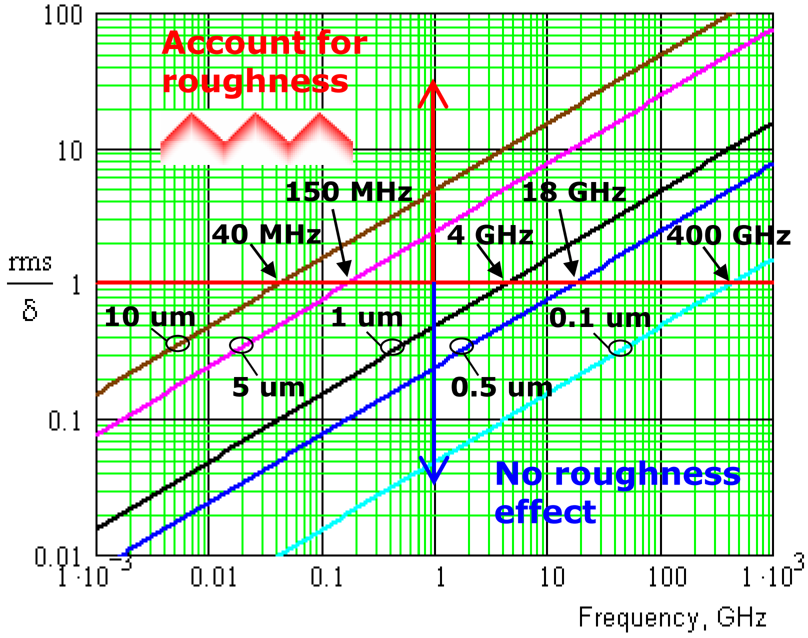
$$\delta_s = \frac{1}{\sqrt{\pi f \mu \sigma}}$$ Skin depth

Conductor skin-effect and roughness

Transition from 0.5 skin depth to 2 and 5 skin depths for copper interconnects on PCB, Package, RFIC and IC



Ratio of skin depth to r.m.s. surface roughness in micrometers vs. frequency in GHz

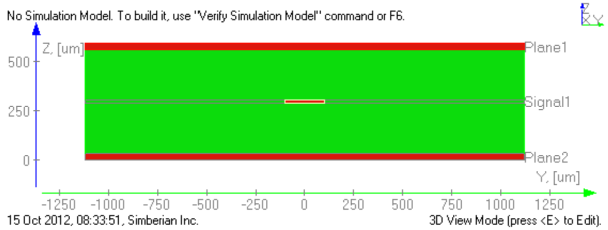


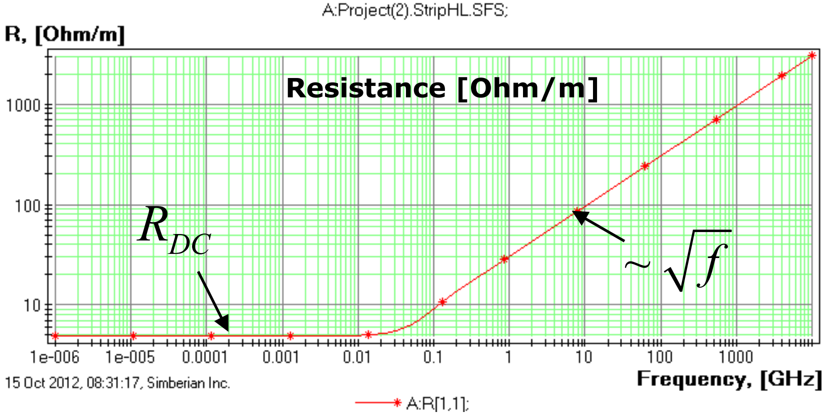
Roughness increases losses if rms value is comparable with the skin depth

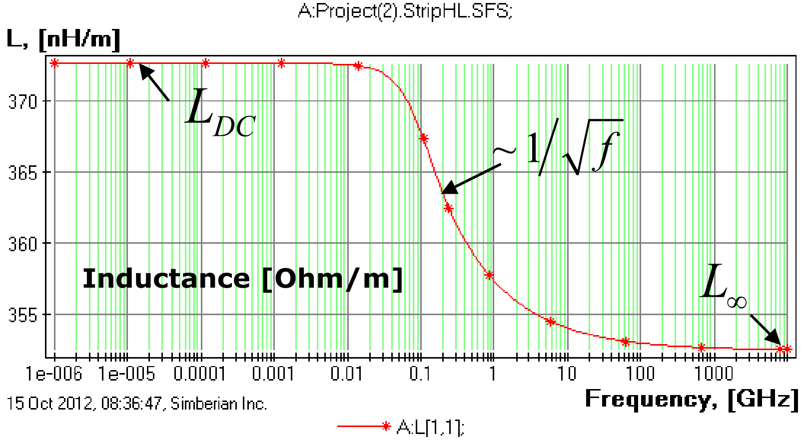
Example of causal R, L, G, C for a simple strip-line case (N=1)

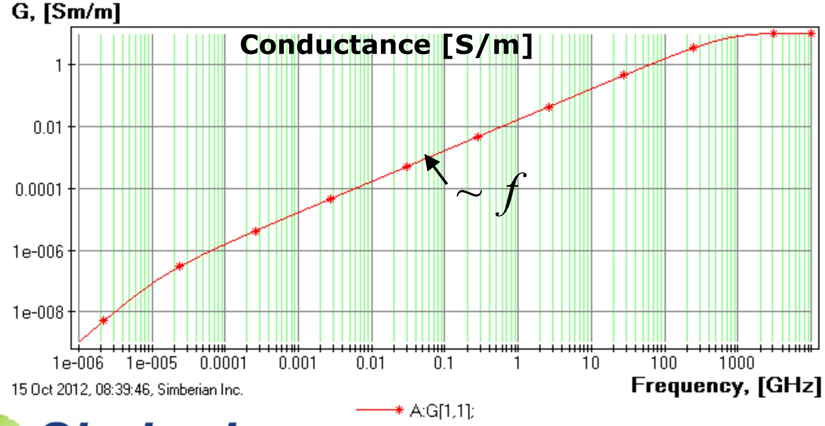
8-mil strip, 20-mil plane to plane distance, DK=4.2, LT=0.02 at 1 GHz, wideband Debye model.

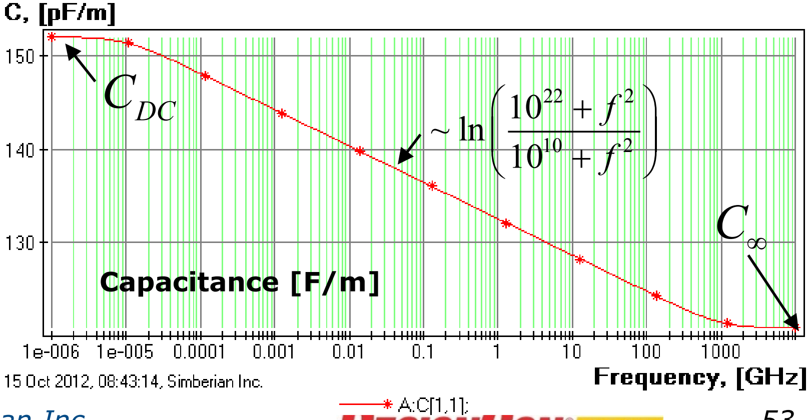
Strip and planes are made of copper, analysis in Simbeor SFS.











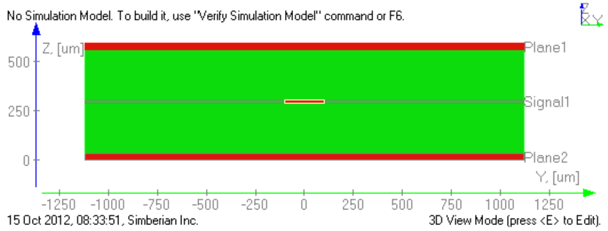
Example: Broadband characteristic impedance and propagation constant for a simple strip-line

$Z_0(\omega) = \sqrt{Z(\omega)/Y(\omega)}$ **Complex characteristic impedance [Ohm]**

$\Gamma(\omega) = \sqrt{Z(\omega)\cdot Y(\omega)} = \alpha + i\beta$

Attenuation Constant [Np/m]

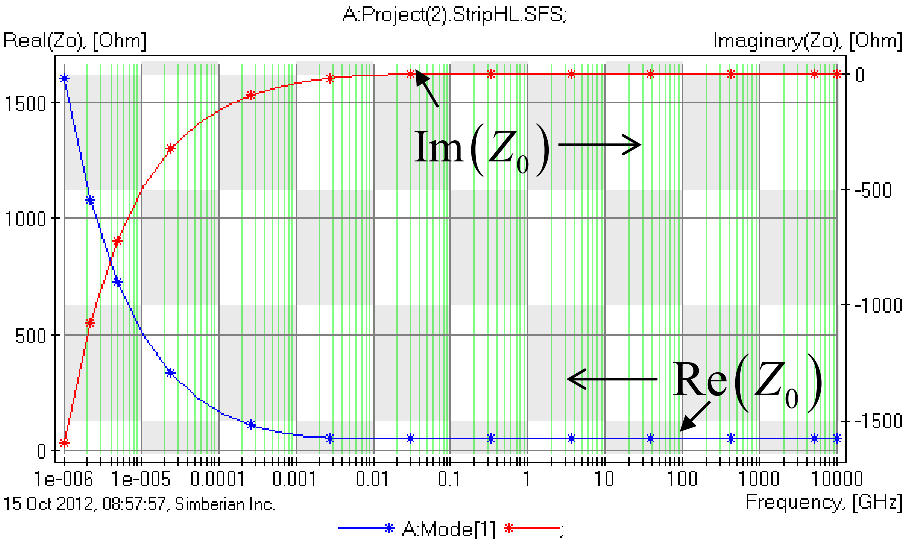
Phase Constant [rad/m]



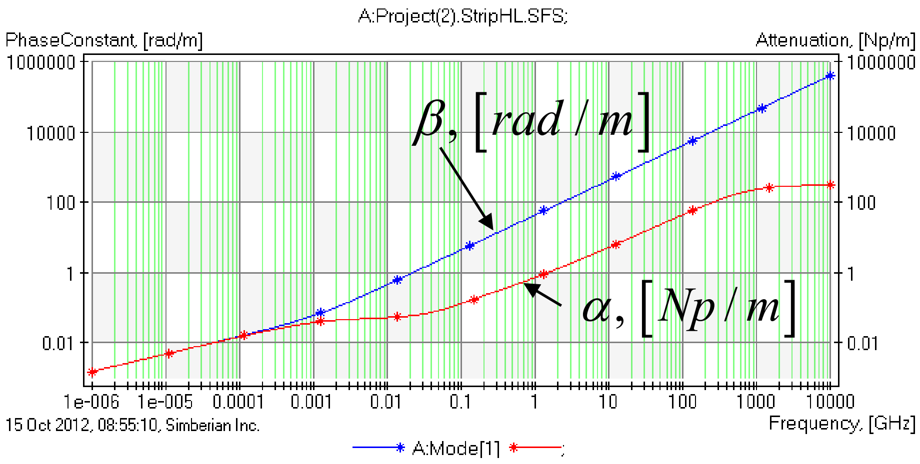
No Simulation Model. To build it, use "Verify Simulation Model" command or F6.

8-mil strip, 20-mil plane to plane distance. DK=4.2, LT=0.02 at 1 GHz, wideband Debye model. Strip and planes are made of copper, no high-frequency dispersion.

Characteristic impedance, [Ohm]

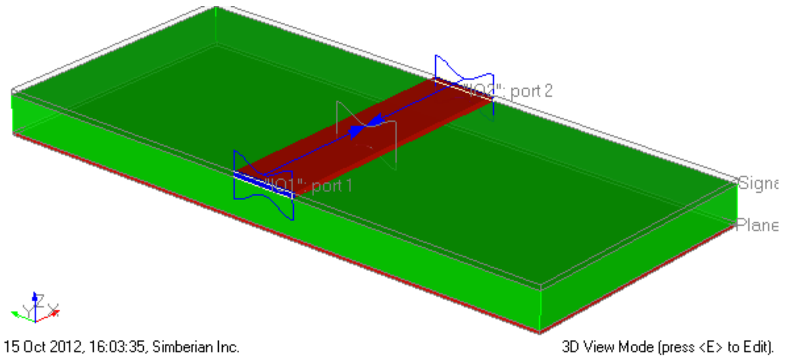


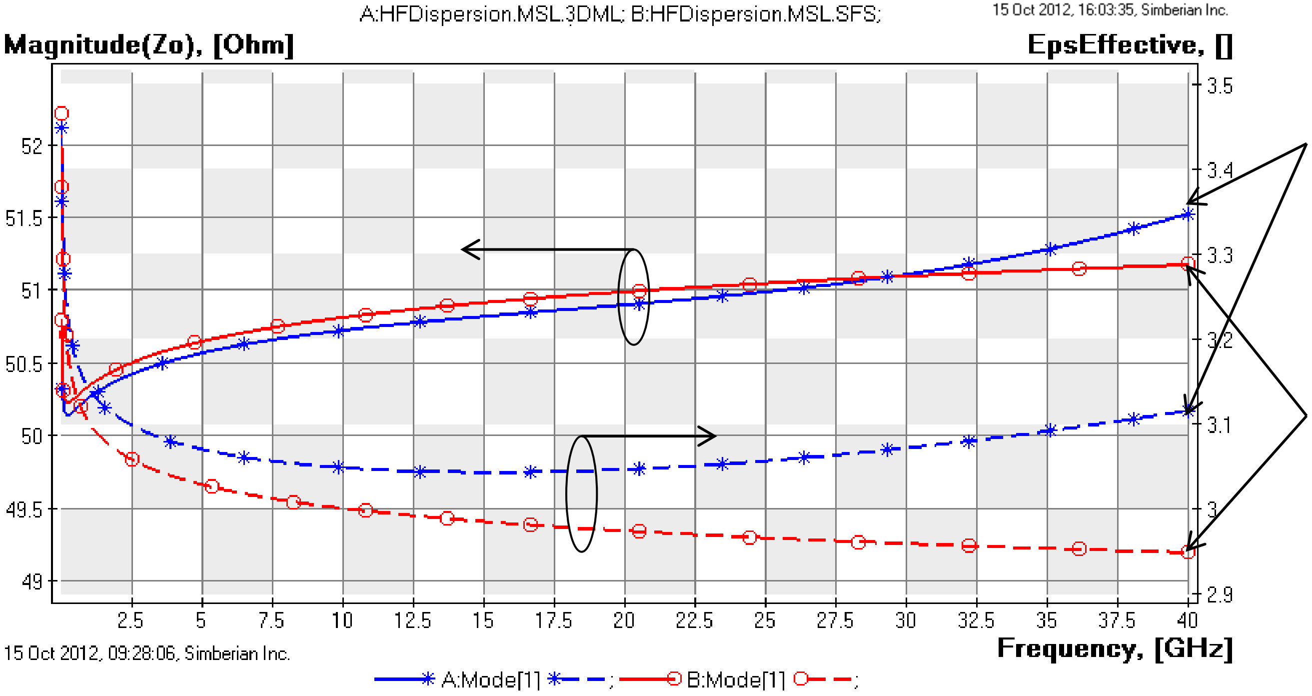
Propagation Constant



Example of high-frequency dispersion

- 14 mil microstrip line on 8 mil dielectric (Dk=4.2, LT=0.02 at 1 GHz, wideband Debye model), ½ Oz copper





Full-wave analysis (Simbeor 3DML)

Quasi-static analysis (Simbeor SFS)

Transmission line modeling tasks

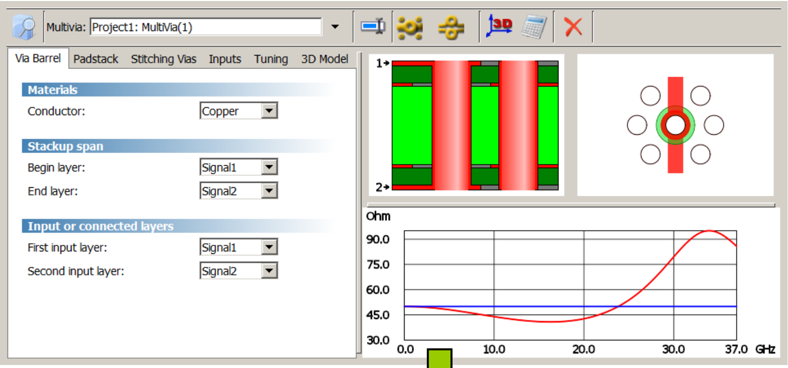
- Pre-layout tasks
  - Synthesize cross-sections with the target impedance
  - Estimate maximal possible line length (loss budget)
  - Estimate cross-talk and create design rules
  - Evaluate impact of manufacturing tolerances and weave effect
- Post-layout tasks
  - Identify transmission line segments
  - Identify coupled segments
  - Build models and simulate with the other elements of a channel

Modeling transmission lines (summary)

- Broadband material models is the most important element for transmission line models for data rates 10 Gbps and higher
- Such models must be identified - frequency-continuous models are not available from manufacturers
- Advanced quasi-static or full-wave solver can be used for strip lines
- Full-wave solver should be used for microstrip or CB-CPW lines (dispersion)
- Field solver for SI applications must have
  - Appropriate set of frequency-continuous dielectric models (wideband and multi-pole Debye for instance)
  - Conductor models valid over 4-5 frequency decades in general (to account for transition to skin-effect, skin-effect, skin-effect on rough surface)

Modeling discontinuities: Via-holes, breakouts, launches,…

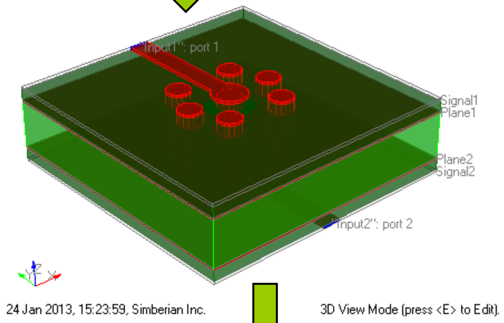
- Pre-layout tasks
  - Synthesize geometry for transitions into different layers with **minimal reflection and localization over the target frequency range**
  - Evaluate transitions impact on compliance metrics
  - Evaluate impact of manufacturing tolerances
- Post-layout tasks
  - Identify geometries of discontinuities
  - Build models and simulate with the other elements of a channel

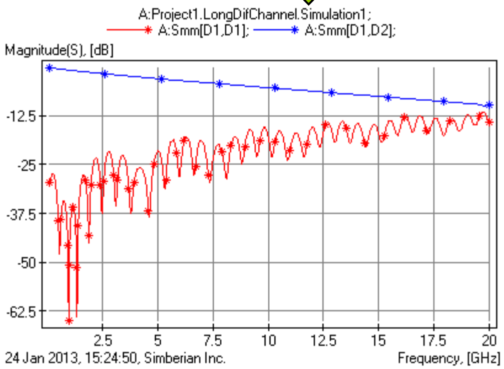

Multivia: Project1: MultiVia(1)

Materials
Conductor: Copper

Stackup span
Begin layer: Signal1
End layer: Signal2

Input or connected layers
First input layer: Signal1
Second input layer: Signal2





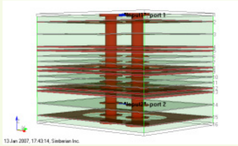
Models for discontinuities (via-holes, breakouts, launches,…)

3D Static and Magneto-Quasi-Static Solvers

$$\vec{E} = -\nabla\varphi - j\omega\vec{A}$$
$$\vec{H} = \frac{1}{\mu}\nabla\times\vec{A}$$

3D Full-Wave Solver

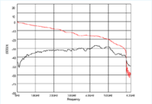
$$\nabla\times\vec{E} = -i\omega\mu\vec{H}$$
$$\nabla\times\vec{H} = i\omega\varepsilon\vec{E} + \sigma\vec{E} + \vec{J}$$



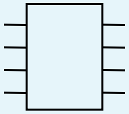
Transmission Plane Solvers

$$\frac{\partial J_{sx}}{\partial x} + \frac{\partial J_{sy}}{\partial y} = -Y_{\square}(\omega)\cdot V + J_z$$
$$\frac{\partial V}{\partial x} = -Z_{\square}(\omega)\cdot J_{sx}$$
$$\frac{\partial V}{\partial y} = -Z_{\square}(\omega)\cdot J_{sy}$$

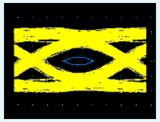
Simple lumped or distributed LC models



Multiport S-parameters



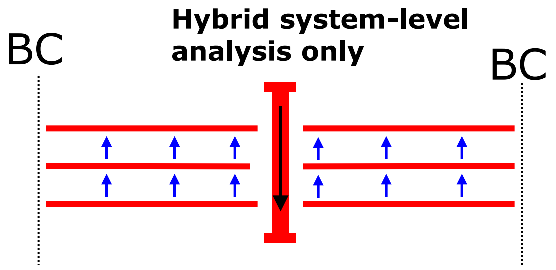
Hybrid 2D transmission plane + circuit solvers



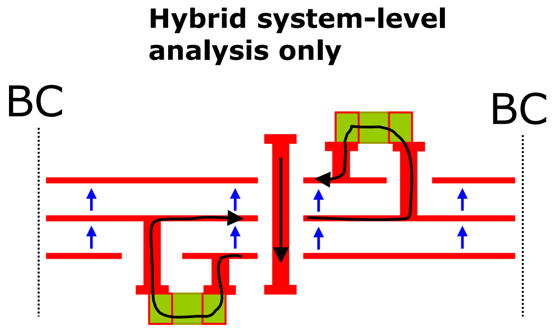
Decompositional Analysis

Localization of single vias going through multiple parallel planes

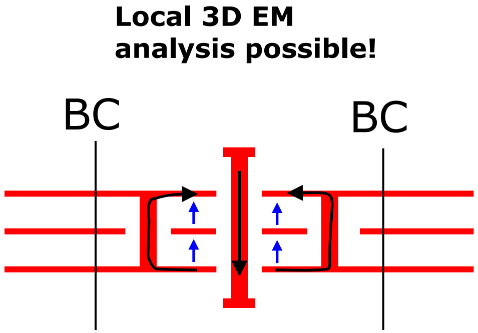
- Planes are not terminated and the return current is the “displacement” current between the planes
  - The problem is non-localizable – requires analysis of the whole board



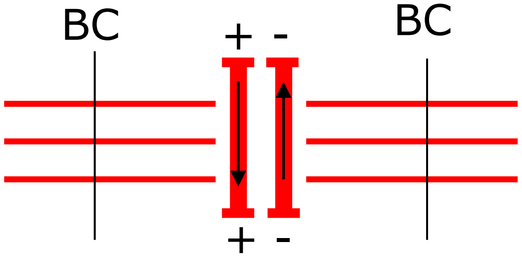
- Planes are terminated with the decoupling capacitors and the return current is a combination of the “displacement” currents through capacitors and planes
  - Decaps have low impedance only in a narrow band – thus the problem again is non-localizable for broadband EM analysis



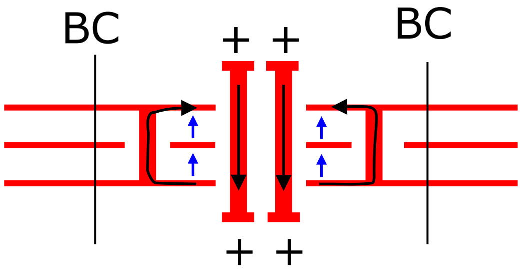
- Stitching vias are used to connect the reference planes for the connected layers and the return current is mostly conductive
  - Problem can be localized (conditionally localizable) and solved with any boundary conditions



Localization of differential vias going through multiple parallel planes

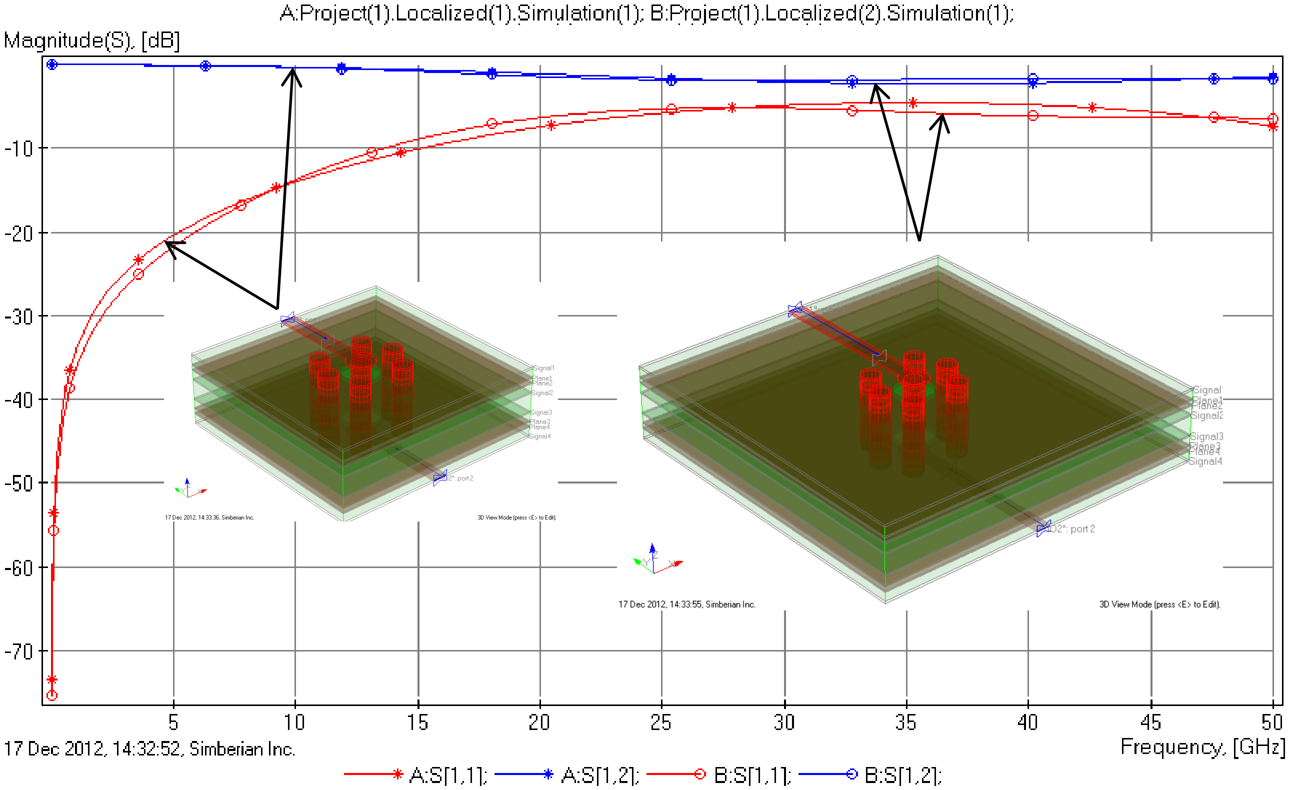


- Differential mode has two opposite currents on the via barrels
- The vias can be isolated from the rest of the board for the electromagnetic analysis with any boundary conditions (PEC, PMC, PML, ABC)
  - Distance from the vias to the simulation area boundaries should be selected to reduce the effect of sidewalls
  - In that case, the differential mode S-parameters are practically independent of the boundary conditions
- Common mode behavior is similar to the single-ended via case – see previous slide



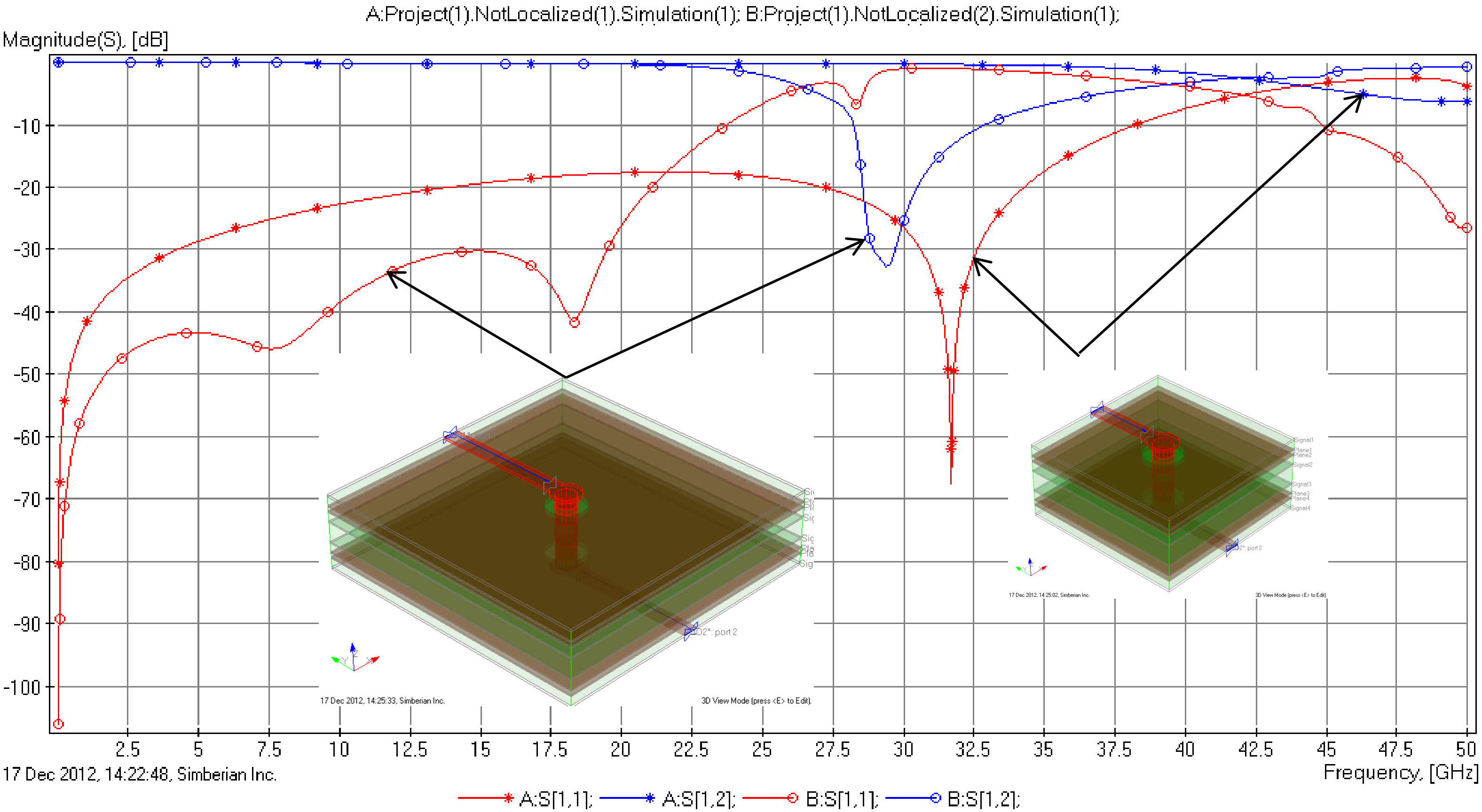
How estimate the localization?

- Change simulation area or simulate with different boundary conditions and observe changes
- Example of conditionally localized structure



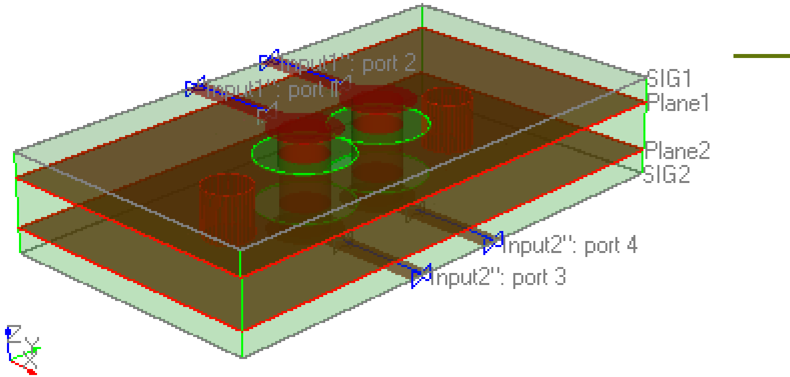
Example of non-localizable via

- Change of simulation area size causes huge differences in reflection and insertion loss – unpredictable “pathological” structure

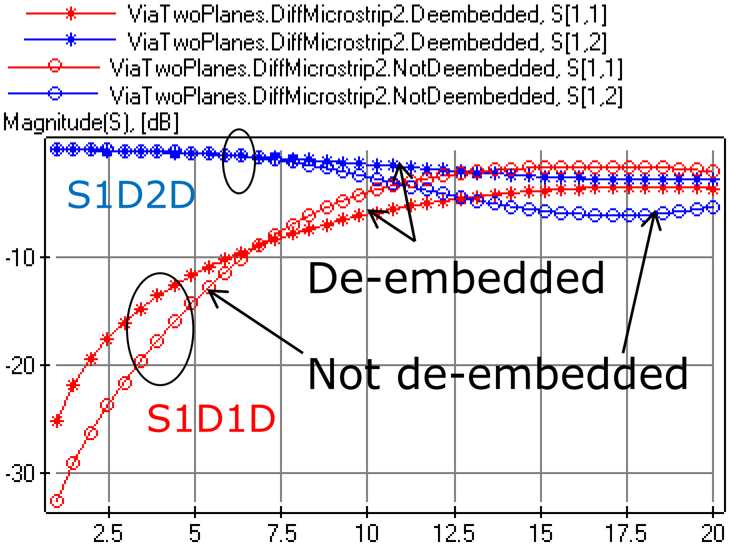


Effect of de-embedding on multiport parameters

Non-reflective excitation ports (lumped or wave-ports) increase the model quality



24 Jan 2008, 15:14:50, Simberian Inc.

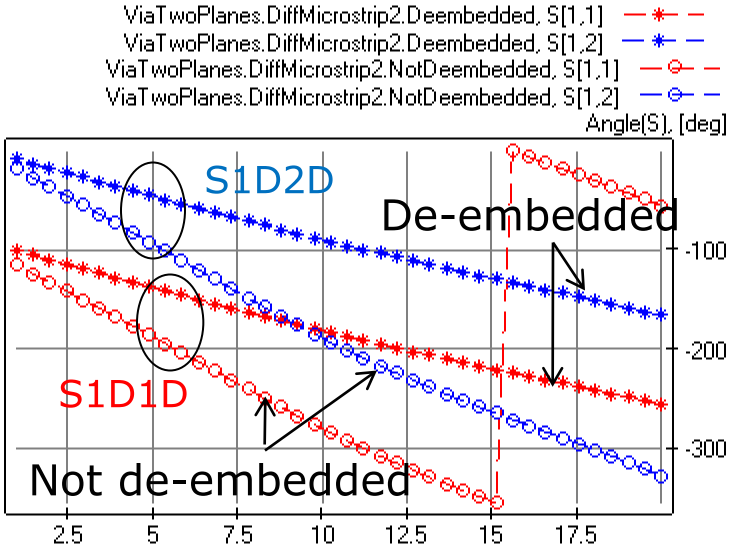


De-embedded

Not de-embedded

24 Jan 2008, 17:55:57, Simberian Inc.

Shift of reference planes makes model electrically smaller and reusable



De-embedded

Not de-embedded

24 Jan 2008, 17:58:26, Simberian Inc.

Simberian Electromagnetic Solutions

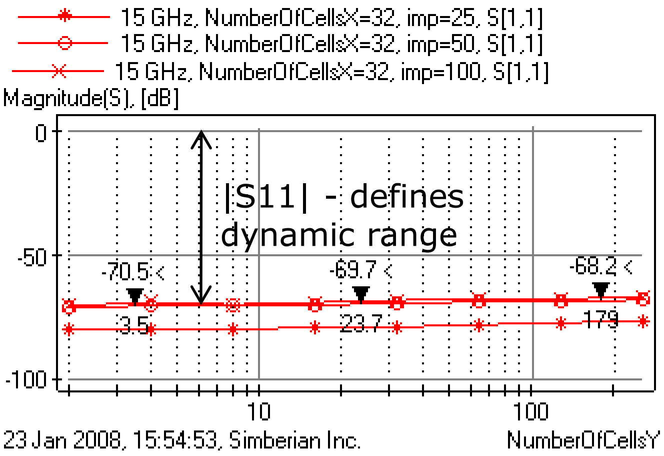
Estimation of de-embedding quality

- Analysis of a transmission line segment can reveal de-embedding defects
- Analysis of 25, 50 and 100-Ohm benchmark strip line segments can be used for this purpose

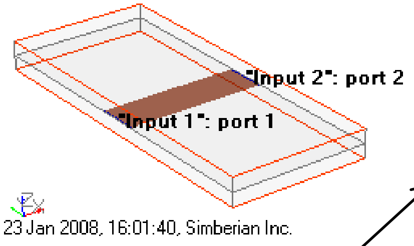
Benchmarks from J.C. Rautio, IEEE on MTT, v.42, N11, 1994, p. 2046-2050.

S-parameters normalized to 25, 50 and 100 Ohm, segment length is 90 deg. at 15 GHz

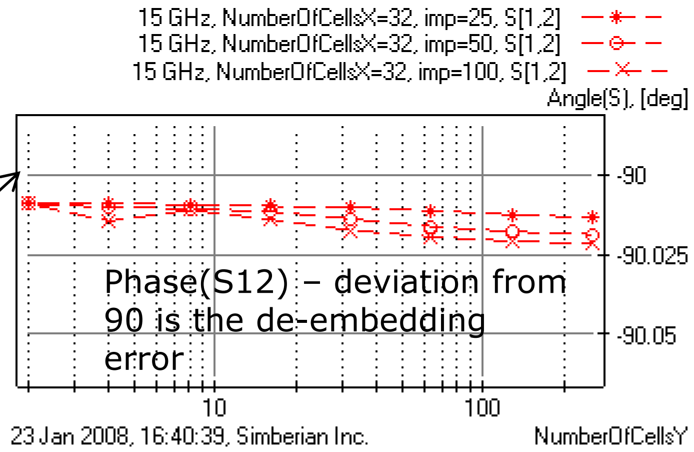
No losses, no dispersion – |S21| must be unit |S11| must be zero



|S11| - defines dynamic range



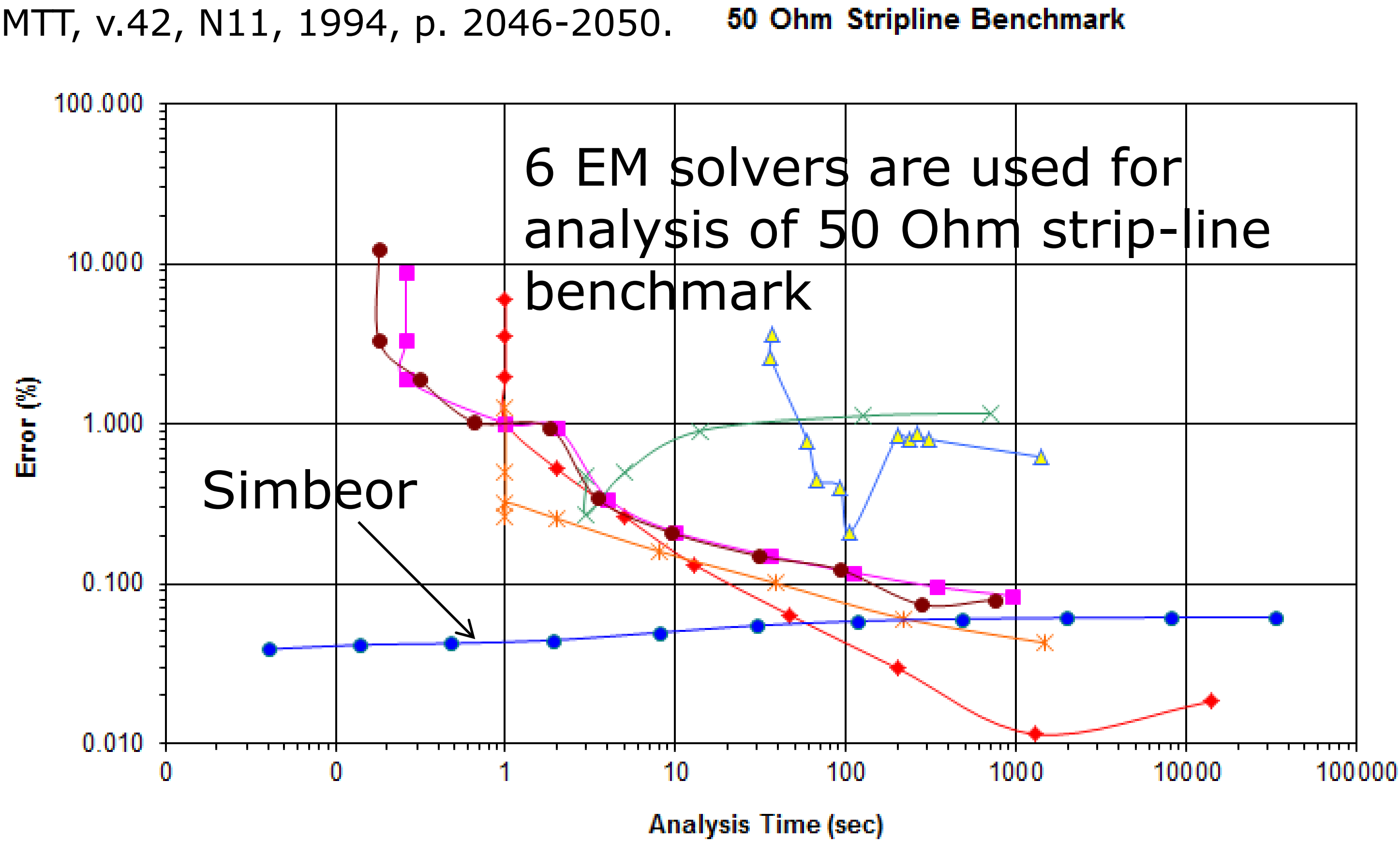
Exact value



Phase(S12) – deviation from 90 is the de-embedding error

De-embedding quality estimation example

Benchmark test from J.C. Rautio, IEEE on MTT, v.42, N11, 1994, p. 2046-2050.



50 Ohm Stripline Benchmark

6 EM solvers are used for analysis of 50 Ohm strip-line benchmark

Simbeor

Modeling discontinuities (summary)

- Localization is the most important element (predictability)
  - Planar discontinuities can be always simulated in isolation
  - Coupling between t-lines is a type of localization violation and must be avoided or accounted for
  - Vias, breakouts, connector launches have to be localized for analysis in isolation from the rest of the board
- Discontinuities must be appropriately de-embedded to avoid artificial (numerical) reflections
- Broadband material models are not so important as in the case of transmission lines
- Dielectric anisotropy may be important in analysis of vias

Broadband material models and material model identification

Broadband material models for PCB/packaging

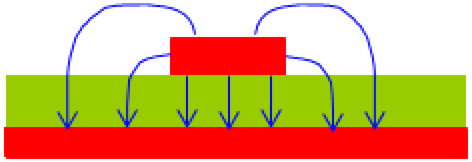
Material identification techniques

Identification with GMS-parameters

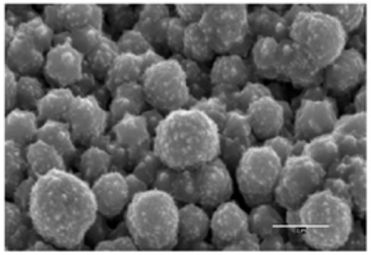
Practical examples

Why do we need broadband material models?

- Dielectrics are the media where signals propagate along the conductors of interconnects
  - Dielectric constant (DK) and loss tangent (Df or LT) may change substantially over the frequency band of multi-gigabit signal spectrum



- Interconnect conductors guide the signals but also absorb energy of the waves at the surface
  - Insertion loss at high frequencies can grow up to 50% due to surface roughness
  - Roughness can also increase group delay (increase cap.)



- *Broadband dielectric and conductor roughness models are needed for accurate electromagnetic analysis of multi-gigabit interconnects*
  - 10-20 Gb/s - from DC up to 20 GHz
  - 20-50 Gb/s - from DC up to 50 GHz

Why are obtaining material models so difficult?

- Manufacturers of dielectrics and PCBs provide measurements for dielectric parameters typically **without frequency or at 1-3 points in the best cases**
  - Simplified TDR-based methods and advanced microwave resonator-based methods do not produce broadband models
  - Only frequency-continuous models can describe dispersive behavior of PCB/packaging dielectrics over very wide bandwidth
- Conductor surface roughness is usually characterized with one number – RMS peak-to-valley (Rq) – not sufficient!
  - Practical all roughness models have multiple unknown parameters
- *Multi-gigabit interconnect design and compliance analysis must start with the identification of the dielectric and conductor properties over the frequency band of interest*

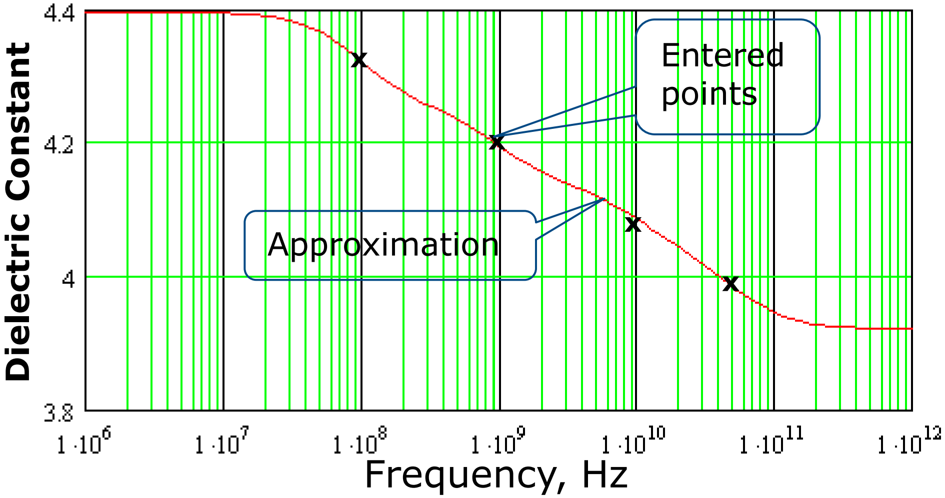
Dielectric models for PCB & packaging

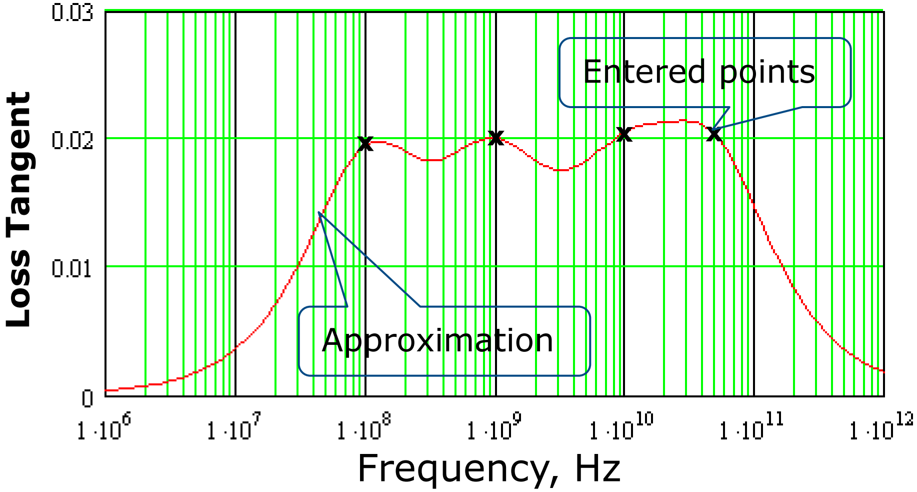
- Non-causal (frequency-independent Dk & LT)
- **Multi-pole Debye (real poles)**
- **Wideband Debye (Djordjevic-Sarkar)**
- Multi-pole with complex poles (Debye-Lorentz)
- Dielectric mixtures (Wiener, Hashin-Shtrickman, Maxwell-Garnet, Bruggeman)
- Anisotropic dielectrics (separate definition of Z, X, and Y components of permittivity tensor)

Multi-pole Debye model

$$\varepsilon(f) = \varepsilon(\infty) + \sum_{n=1}^{N} \frac{\Delta\varepsilon_n}{1 + i\dfrac{f}{fr_n}}$$

- Discrete-spectrum model
- Requires specification of value at infinity and poles/residues or DK and LT at multiple frequency points
- Can be used for any dielectric without resonances
- At least 4 poles (usually 10) are required for composite dielectrics for multi-gigabit signals





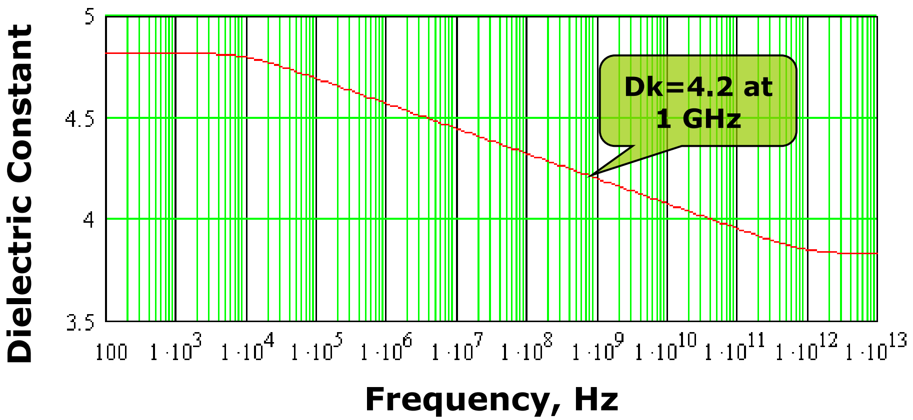
Wideband Debye model

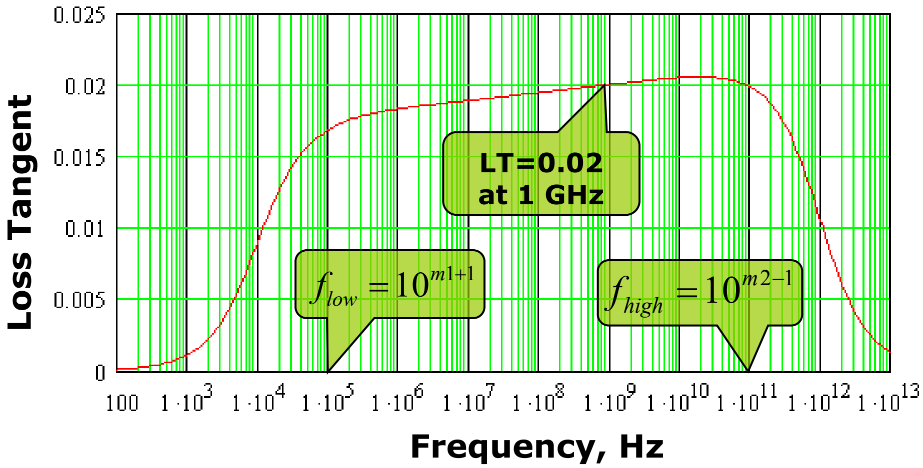
$$\varepsilon_{wd}(f) = \varepsilon_r(\infty) + \varepsilon_{rd} \cdot F_d(f)$$

$$F_d(f) = \frac{1}{(m_2 - m_1) \cdot \ln(10)} \cdot \ln\left[\frac{10^{m2} + if}{10^{m1} + if}\right]$$

- Continuous-spectrum model
- Requires specification of DK and LT at one frequency point
- Good match for high-loss FR-4 dielectrics (LT>0.01)
- May be not so good match for low-loss, high-frequency composites (LT<0.01)

Djordjevic, R.M. Biljic, V.D. Likar-Smiljanic, T.K.Sarkar, IEEE Trans. on EMC, vol. 43, N4, 2001, p. 662-667.





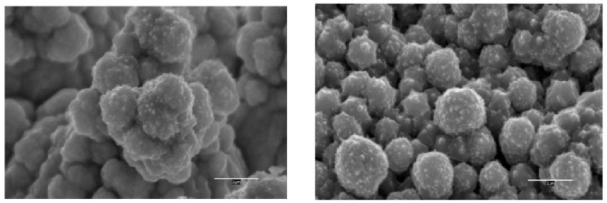
Simberian Electromagnetic Solutions

Roughness models

References and details are in DesignCon 2012 paper – see references at the end

- Attenuation correction coefficients
  - Hammerstad model (Hammerstad, Bekkadal, Jensen)
  - “Snowball” model (Hurray,…)
  - Hemispherical model (Hall, Pytel,…)
  - Stochastic models (Sanderson, Tsang,…)
  - Periodic structures (Lukic,…)
- Conductor loss separation by extrapolation
  - Koledintseva, Koul,…
- Equivalent boundary conditions
  - Holloway, Kuester
  - Koledintseva, Koul,…
- Direct electromagnetic analysis

P. G. Huray, at al., DesignCon 2010



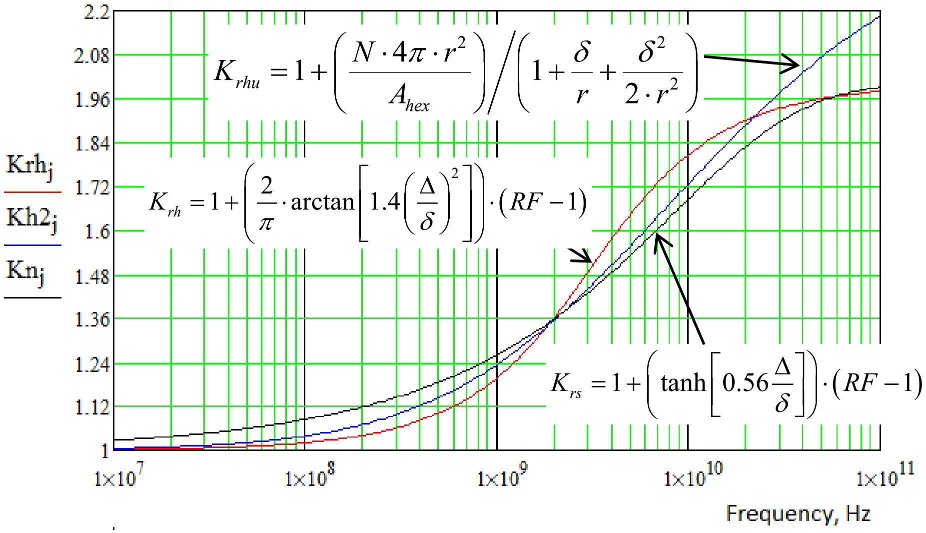
High Profile texture

Low Profile texture

Roughness correction coefficients

- Modified Hammerstad (red), Simbeor (black) and Huray's snowball (blue) models (shown for RTF/TWS foil as example)

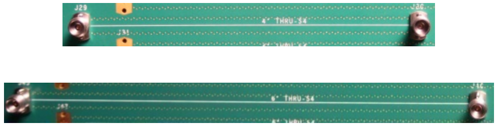
References and details are in DesignCon 2012 paper



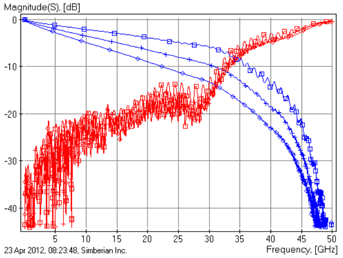
- If applied to conductor surface impedance operator – the model is causal!
- Where to get the model parameters?
  - SR and RF for Simbeor and MHCC
  - Number of balls, ball size and tile area for Huray's model

Material identification techniques

- For test structures …
  - **Transmission line segments**
  - Patch or parallel-plate resonators
  - Resonators coupled or connected to a transmission line

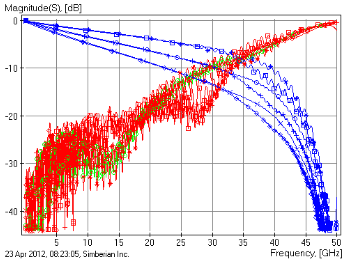


- … measurements …
  - **S-parameters** measured with VNA
  - TDR/TDT measurements
  - Combination of both



23 Apr 2012, 08:23:48, Simberian Inc.

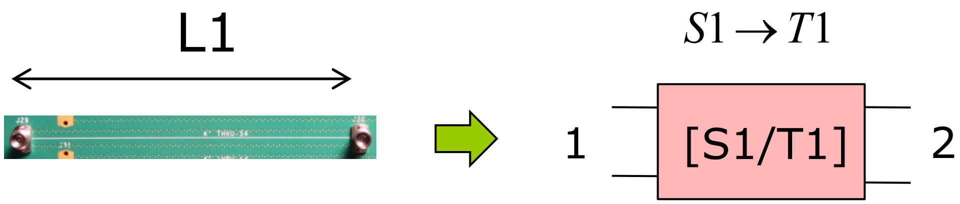
- … are correlated with a numerical model
  - Analytical or closed-form
  - Static or quasi-static field solvers
  - **3D full-wave solvers**

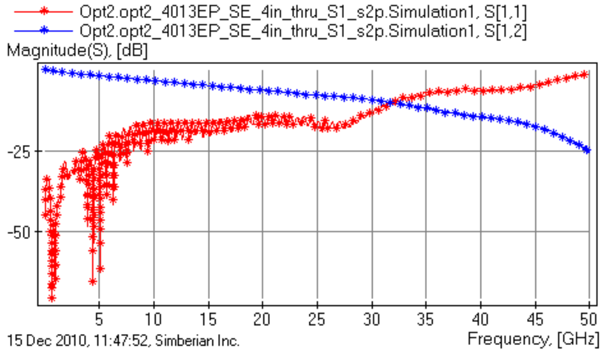


23 Apr 2012, 08:23:05, Simberian Inc.

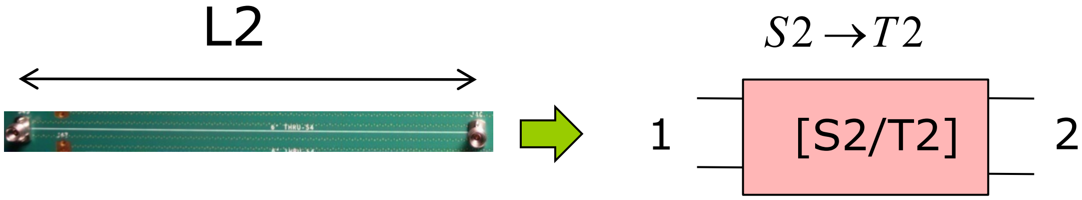
Measure S-parameters of two test fixtures with line segments (no SOLT calibration is required)

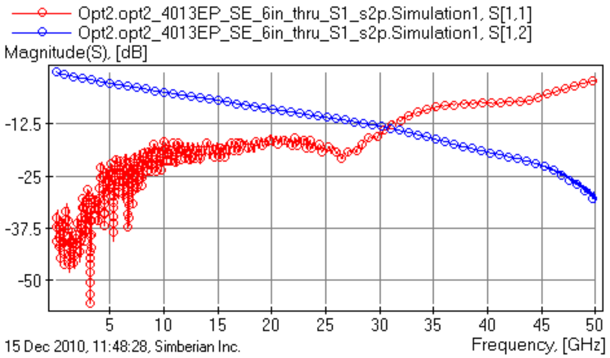
- S1 and T1 for line with length L1





- S2 and T2 for line with length L2



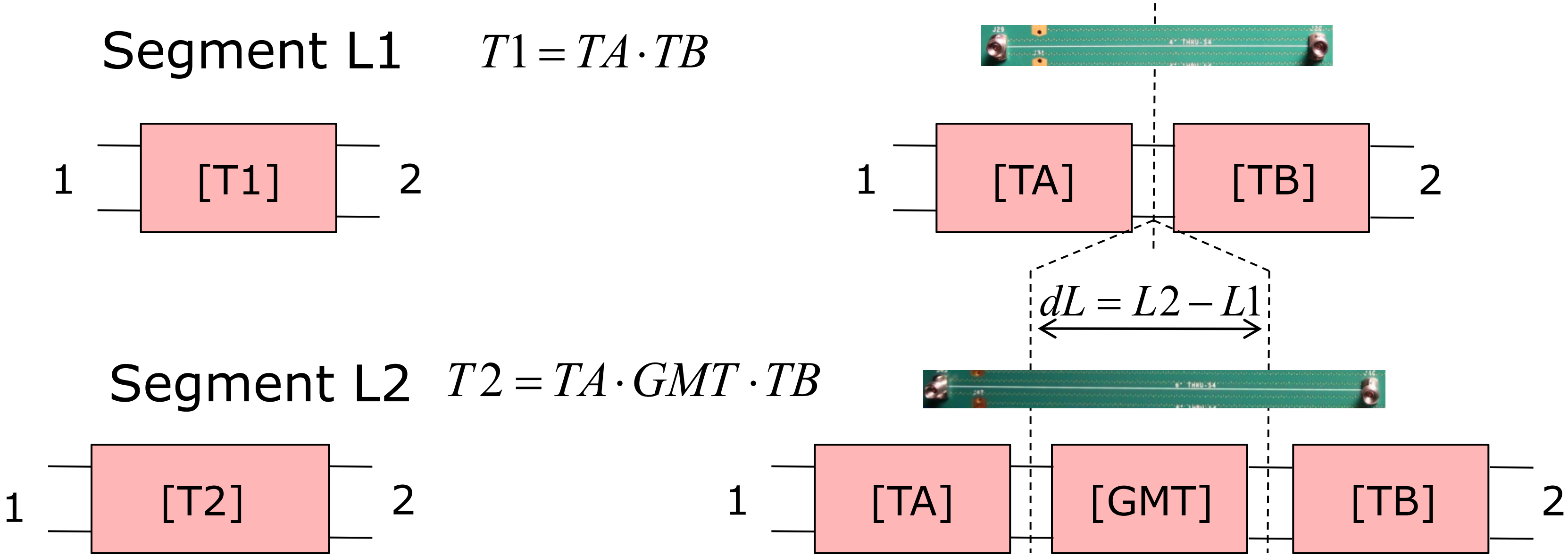


T1 and T2 matrices are scattering T-parameters (computed directly from S-parameters)

Extract Generalized Modal T-parameters (GMT) and then GMS-Parameters (1-conductor case)

Segment L1 $T1 = TA \cdot TB$

Segment L2 $T2 = TA \cdot GMT \cdot TB$



GMT is non-reflective modal T-matrix (normalized to the unknown characteristic impedances of the modes)

$$T2 \cdot T1^{-1} = TA \cdot GMT \cdot TA^{-1}$$

$$\Downarrow$$

$$GMT = eigenvals\left(T2 \cdot T1^{-1}\right)$$

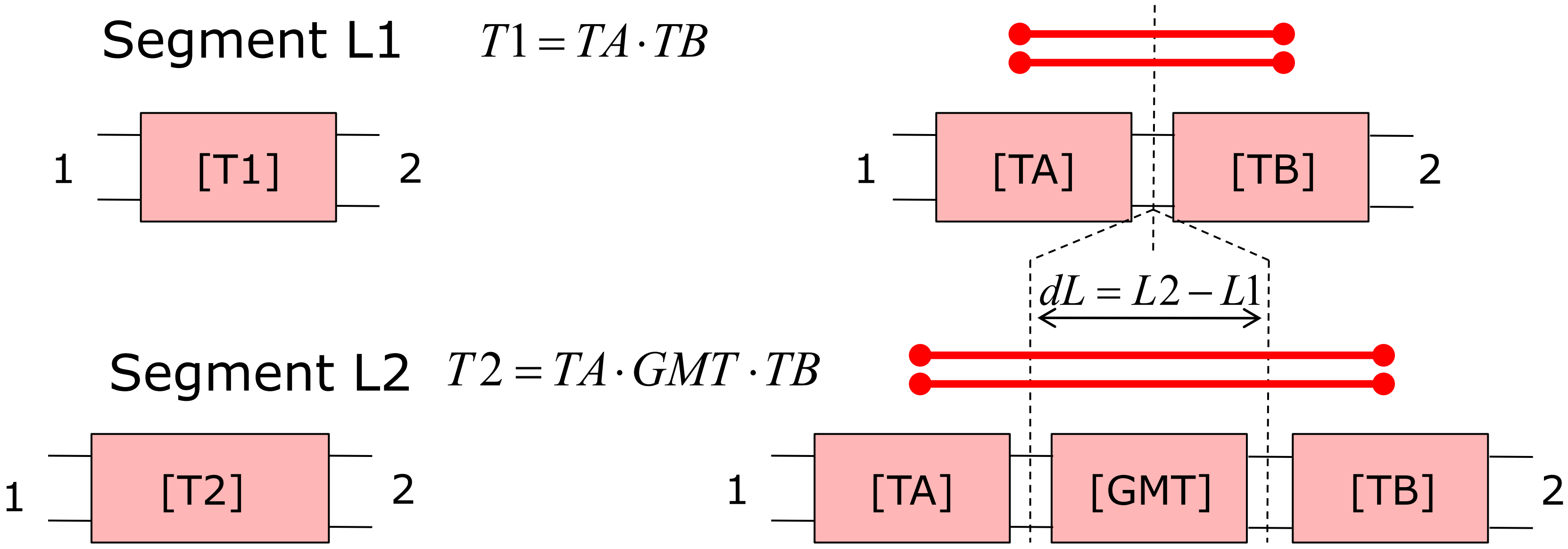
Easy to compute!

For 1-conductor line we get:

$$GMT = \begin{bmatrix} T_{11} & 0 \\ 0 & T_{11}^{-1} \end{bmatrix} \Rightarrow GMSm = \begin{bmatrix} 0 & T_{11} \\ T_{11} & 0 \end{bmatrix}$$

Just 1 complex function!

Extract Generalized Modal T-parameters (GMT) and then GMS-Parameters (2-conductor case)



$$T1 = TA \cdot TB$$

$$dL = L2 - L1$$

$$T2 = TA \cdot GMT \cdot TB$$

GMT is non-reflective modal T-matrix (normalized to the unknown characteristic impedances of the modes)

$$T2 \cdot T1^{-1} = TA \cdot GMT \cdot TA^{-1}$$

$$\Downarrow$$

$$GMT = eigenvals\left(T2 \cdot T1^{-1}\right)$$

For 2-conductor line we get:

$$GMT = \begin{bmatrix} T_{11} & 0 & 0 & 0 \\ 0 & T_{22} & 0 & 0 \\ 0 & 0 & T_{11}^{-1} & 0 \\ 0 & 0 & 0 & T_{22}^{-1} \end{bmatrix} \Rightarrow GMSm = \begin{bmatrix} 0 & 0 & T_{11} & 0 \\ 0 & 0 & 0 & T_{22} \\ T_{11} & 0 & 0 & 0 \\ 0 & T_{22} & 0 & 0 \end{bmatrix}$$

Just 2 complex functions!

Identifying dielectrics by fitting GMS-parameters (1-conductor case)

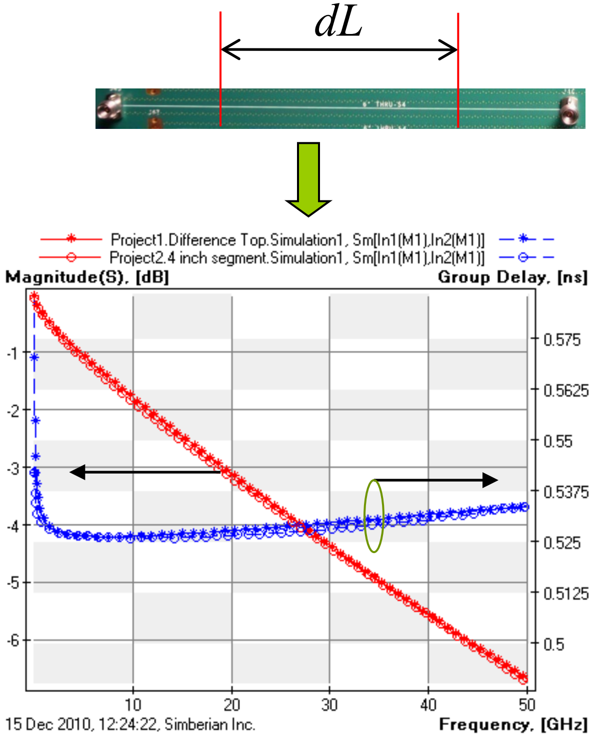
- Solve Maxwell’s equations for 1-conductor line:

$$GMSc = \begin{bmatrix} 0 & \exp\left(-\Gamma \cdot dL\right) \\ \exp\left(-\Gamma \cdot dL\right) & 0 \end{bmatrix}$$

- Fit measured data:

Only 1 complex function!

$$GMSm = \begin{bmatrix} 0 & T_{11} \\ T_{11} & 0 \end{bmatrix}$$

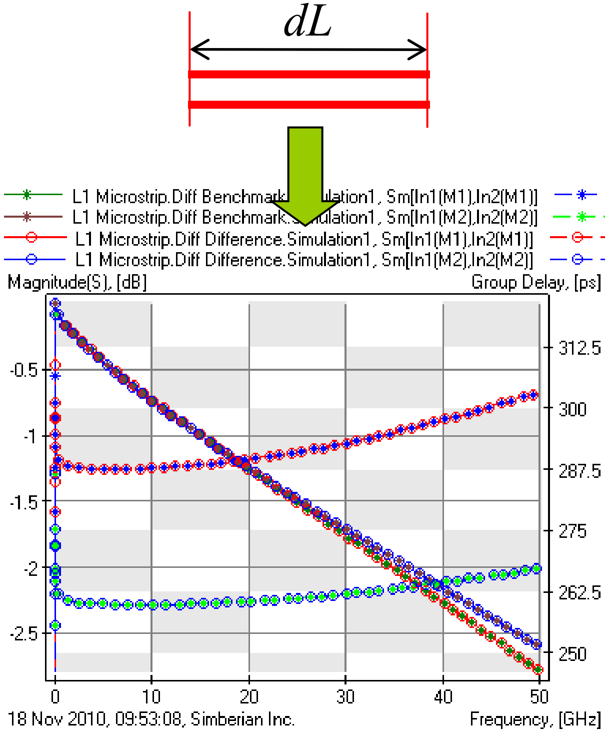


- Measured GMS-parameters of the segment can be directly fitted with the calculated GMS-parameters for material parameters identification
- Phase or group delay can be used to identify DK and insertion loss to identify LT or conductor roughness!

Identifying dielectrics by fitting GMS-parameters (2-conductor case)

- Solve Maxwell's equations for 2-conductor line:

$$GMSc = \begin{bmatrix} 0 & 0 & \exp\left(-\Gamma_1 \cdot dL\right) & 0 \\ 0 & 0 & 0 & \exp\left(-\Gamma_2 \cdot dL\right) \\ \exp\left(-\Gamma_1 \cdot dL\right) & 0 & 0 & 0 \\ 0 & \exp\left(-\Gamma_2 \cdot dL\right) & 0 & 0 \end{bmatrix}$$



- Fit measured data:

Only 2 complex functions!

$$GMSm = \begin{bmatrix} 0 & 0 & T_{11} & 0 \\ 0 & 0 & 0 & T_{22} \\ T_{11} & 0 & 0 & 0 \\ 0 & T_{22} & 0 & 0 \end{bmatrix}$$

- Measured GMS-parameters of the segment can be directly fitted with the calculated GMS-parameters for material parameters identification
- Two functions can be used to identify 2 dielectrics!

Material parameters identification with GMS-parameters

1. Measure S-parameters of two test fixtures with different length of line segments S1 and S2
2. Extract Generalized Modal S-parameters of the line difference
3. Select material model and guess values of the model parameters
4. Compute GMS-parameters of the line difference segment by solving Maxwell's equation for t-line cross-section
5. Adjust material parameters until computed GMS parameters fit measured GMS-parameters with the computed

Procedure is implemented in Simbeor software
Simberian's patent pending #**13/009,541**

The GMS-parameters technique is the simplest possible

- Needs ECAL-calibrated measurements for 2 t-lines with any geometry of cross-section and transitions
  - No extraction of propagation constants (Gamma) from measured data (difficult, error-prone)
  - No de-embedding of connectors and launches (difficult, error-prone)
- Needs the simplest numerical model
  - Requires computation of only propagation constants
  - No 3D electromagnetic models of the transitions
- Minimal number of smooth complex functions to match
  - One parameter for single and two parameters for differential
  - All reflection and modal transformation parameters are exactly zeros

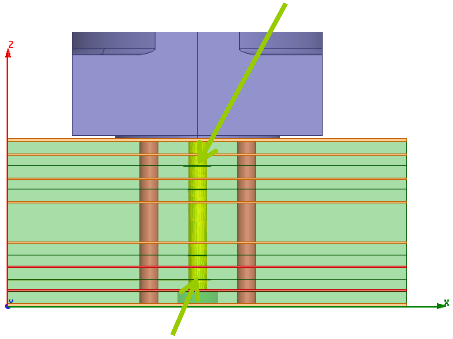
Example: Nelco N4000-13EP

- Example for the original board made with Nelco 4000-13EP investigated in: *D. Dunham, J. Lee, S. McMorrow, Y. Shlepnev, 2.4mm Design/Optimization with 50 GHz Material Characterization, DesignCon2011*

Test structures are pre-qualified for the identification up to 50 GHz in the paper

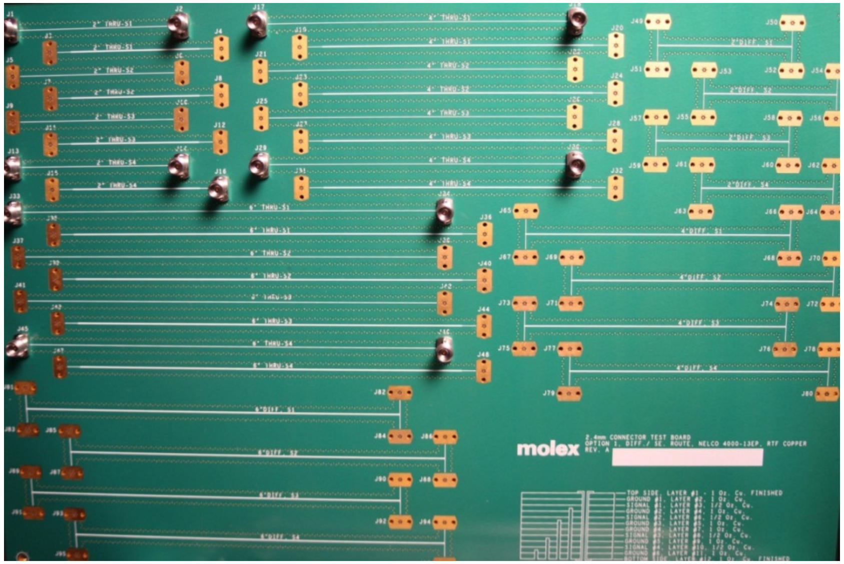
6 test fixtures with 2, 4 and 6 inch strip line segments in Layer 1 (S1) and Layer 4 (S4)

Signal Layer 1



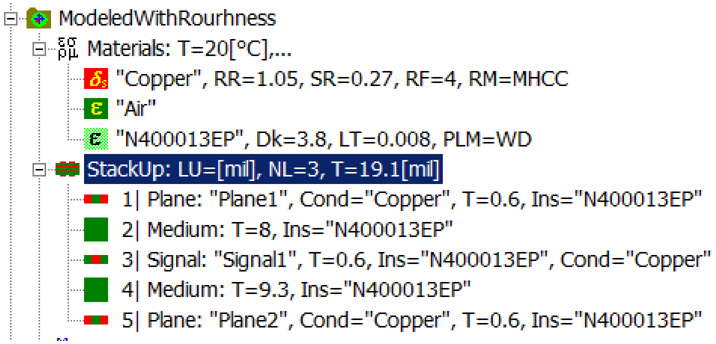
Signal Layer 4

Scott McMorrow from Teraspeed Consulting Group designed launches for 2.4mm Molex connectors, board made by Molex and measurements done by David Dunham, Molex



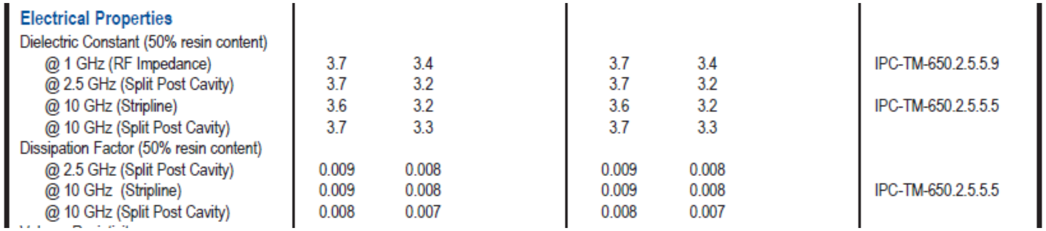
Test board and cross-section

- Strip line segments in Nelco N4000-13EP
- 2 inch, 4 inch and 6 inch segments with launches and Molex 2.6 mm connectors to identify material parameters



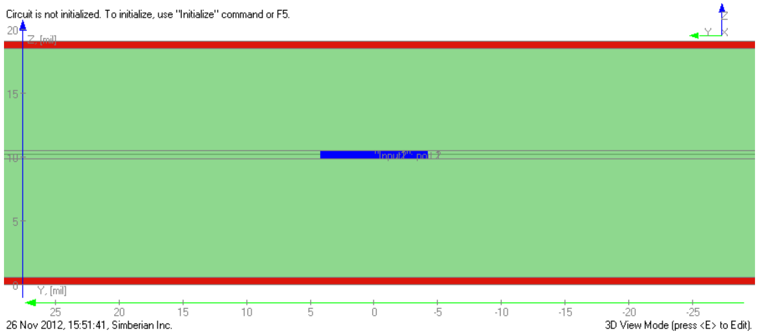
ModeledWithRourhness
- Materials: T=20[°C],....
  - "Copper", RR=1.05, SR=0.27, RF=4, RM=MHCC
  - "Air"
  - "N400013EP", Dk=3.8, LT=0.008, PLM=WD
- StackUp: LU=[mil], NL=3, T=19.1[mil]
  - 1| Plane: "Plane1", Cond="Copper", T=0.6, Ins="N400013EP"
  - 2| Medium: T=8, Ins="N400013EP"
  - 3| Signal: "Signal1", T=0.6, Ins="N400013EP", Cond="Copper"
  - 4| Medium: T=9.3, Ins="N400013EP"
  - 5| Plane: "Plane2", Cond="Copper", T=0.6, Ins="N400013EP"

From datasheet Dk is 3.6-3.7 and LT 0.008-0.009



Electrical Properties					
Dielectric Constant (50% resin content)					
@ 1 GHz (RF Impedance)	3.7	3.4	3.7	3.4	IPC-TM-650.2.5.5.9
@ 2.5 GHz (Split Post Cavity)	3.7	3.2	3.7	3.2	
@ 10 GHz (Stripline)	3.6	3.2	3.6	3.2	IPC-TM-650.2.5.5.5
@ 10 GHz (Split Post Cavity)	3.7	3.3	3.7	3.3	
Dissipation Factor (50% resin content)					
@ 2.5 GHz (Split Post Cavity)	0.009	0.008	0.009	0.008	
@ 10 GHz (Stripline)	0.009	0.008	0.009	0.008	IPC-TM-650.2.5.5.5
@ 10 GHz (Split Post Cavity)	0.008	0.007	0.008	0.007	

Strip width 8.5 mil (both S1 and S4)



Circuit is not initialized. To initialize, use "Initialize" command or F5.

26 Nov 2012, 15:51:41, Simberian Inc.

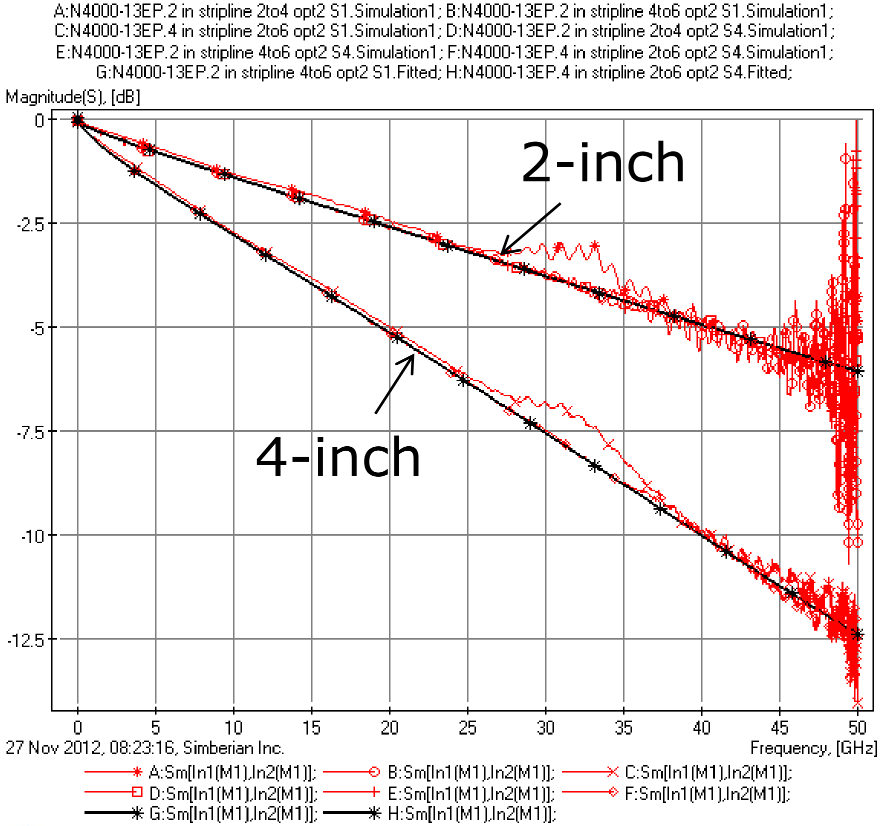
3D View Mode (press <E> to Edit).

Different methods produce slightly different parameters
Which one to use?
What model to use?

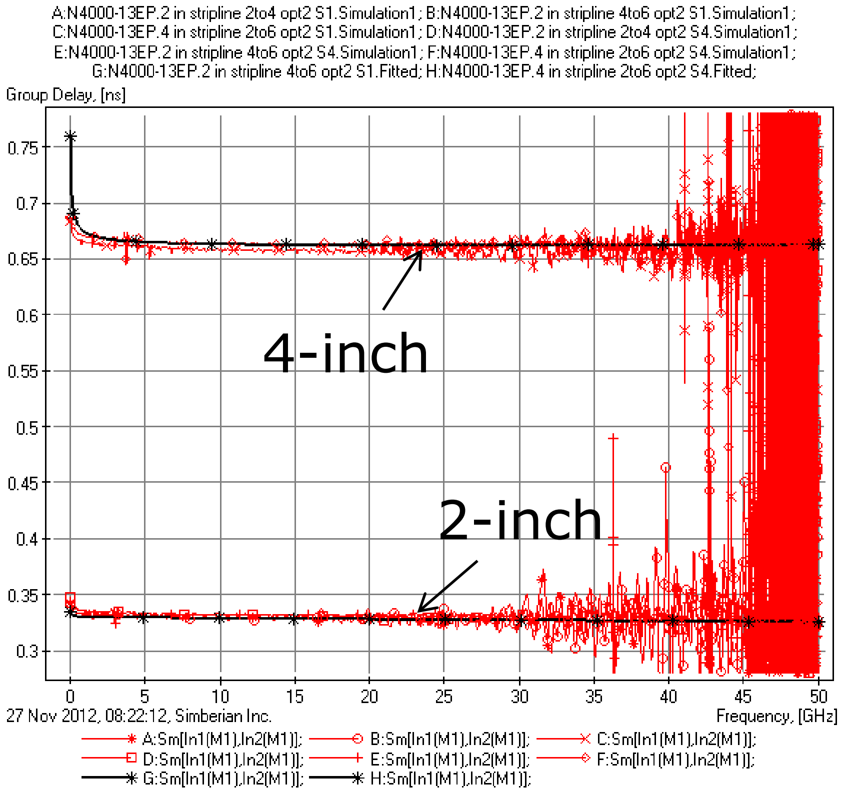
N4000-13EP board measured and post-processed GMS-parameters

- Red lines – GMS from all 6 combinations of lines, black lines – data post-processed for the identification

GMS Insertion Loss



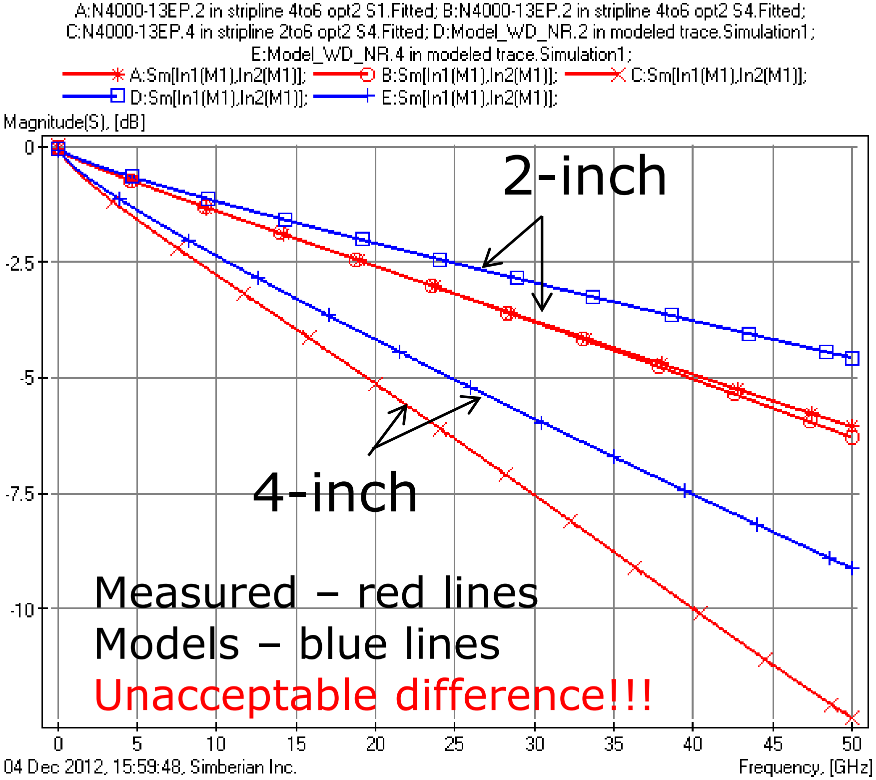
GMS Group Delay



Use material parameters from specs

- Dk=3.8, LT=0.008 @ 10 GHz, WD model, no roughness

GMS Insertion Loss

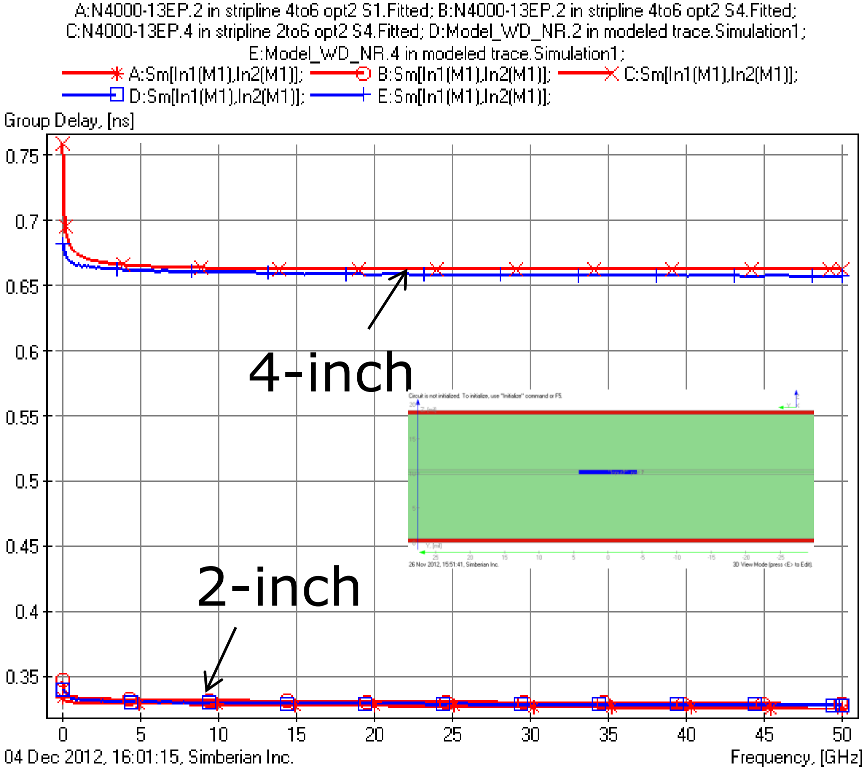


2-inch

4-inch

Measured – red lines
Models – blue lines
Unacceptable difference!!!

GMS Group Delay



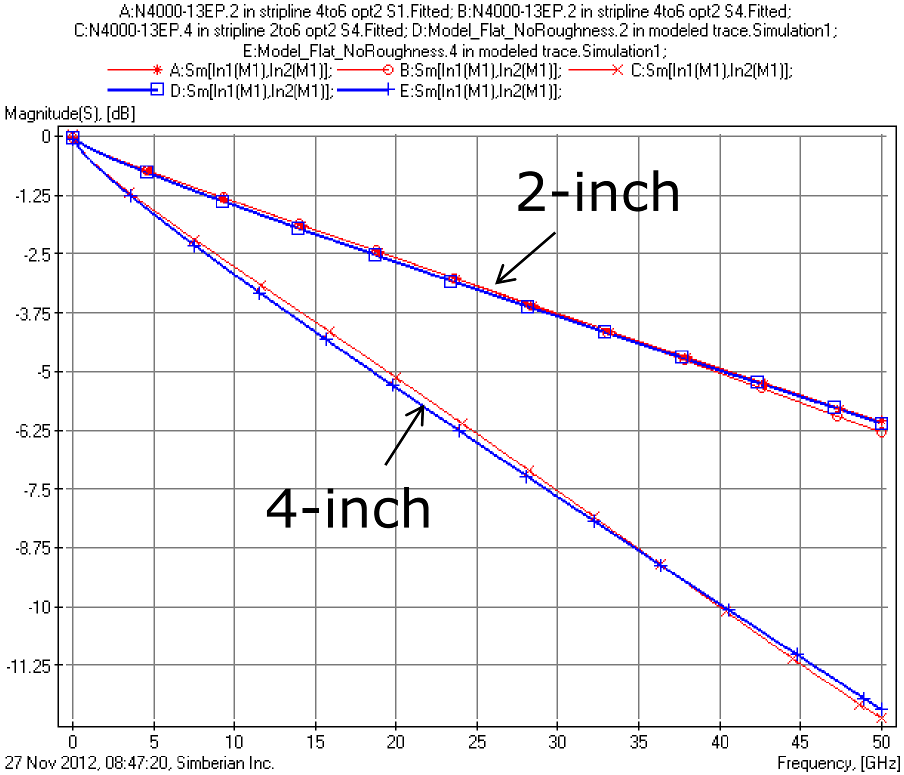
4-inch

2-inch

Flat non-causal dielectric model
No roughness

- Dk=3.8, LT=0.0112 – acceptable fit (blue lines) to measured GMS-parameters (red lines)

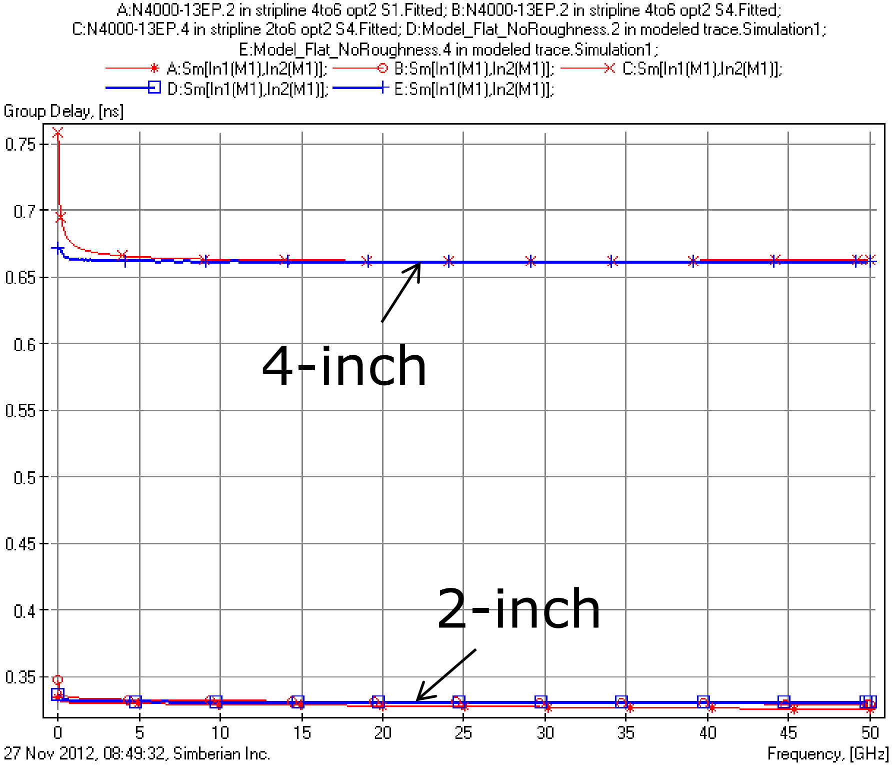
GMS Insertion Loss



2-inch

4-inch

GMS Group Delay



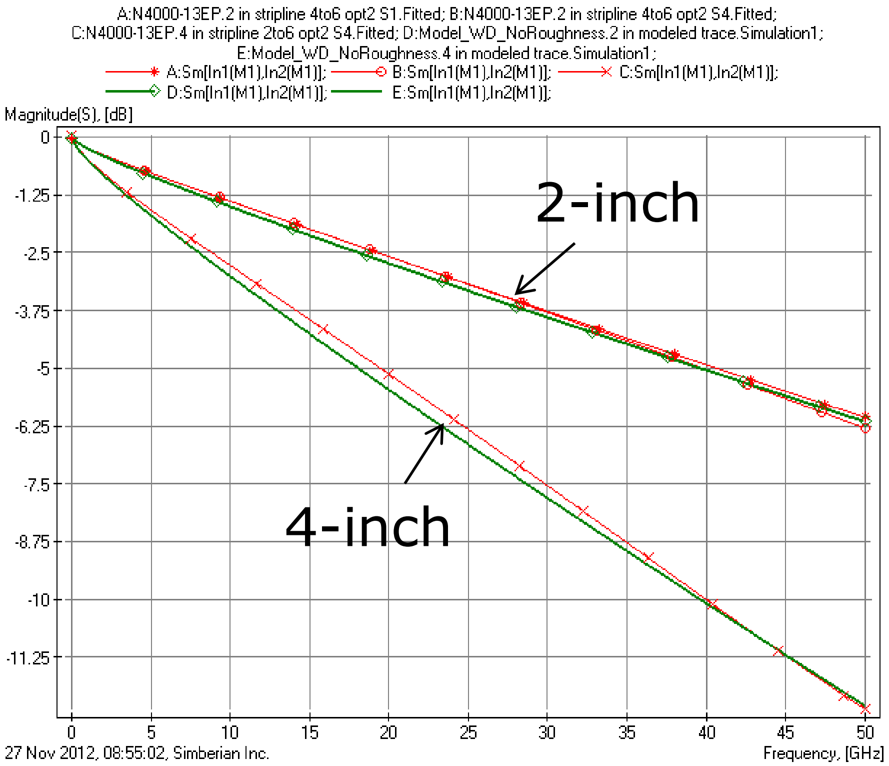
4-inch

2-inch

Wideband Debye dielectric model
No roughness

- Dk=3.84, LT=0.0115 @ 10 GHz, no adjustment for low frq. – acceptable fit (green lines) to measured GMS-parameters (red lines)

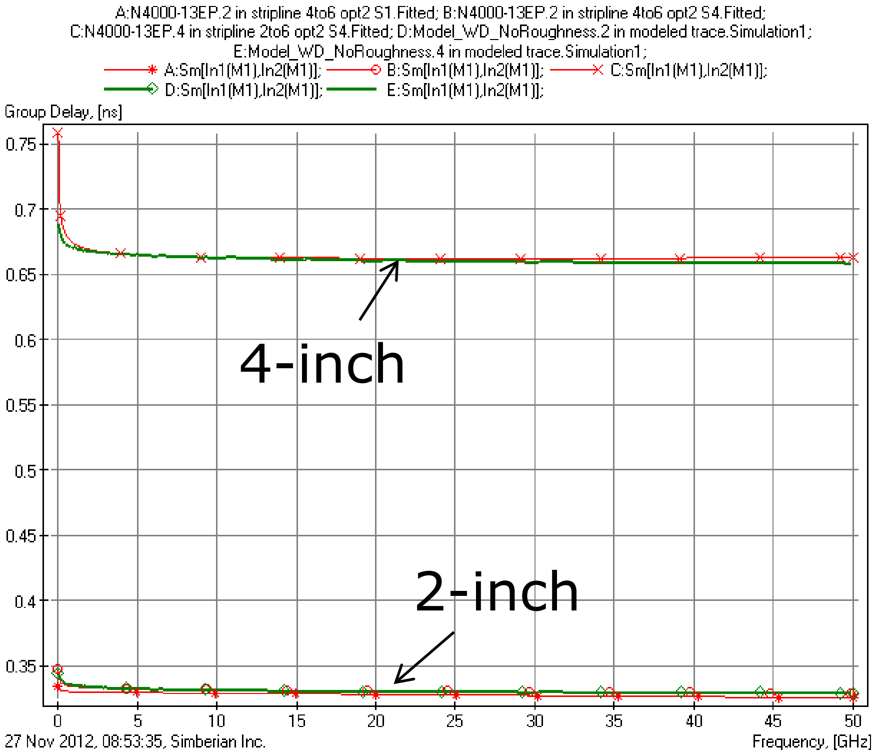
GMS Insertion Loss



2-inch

4-inch

GMS Group Delay



4-inch

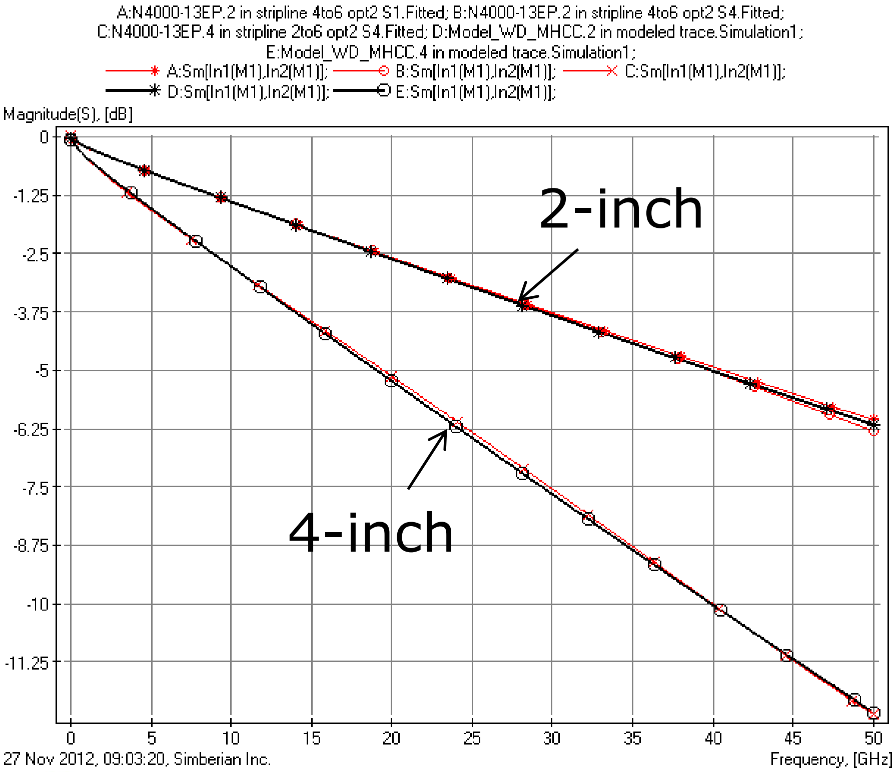
2-inch

Comparable with flat non-causal due to low dispersion

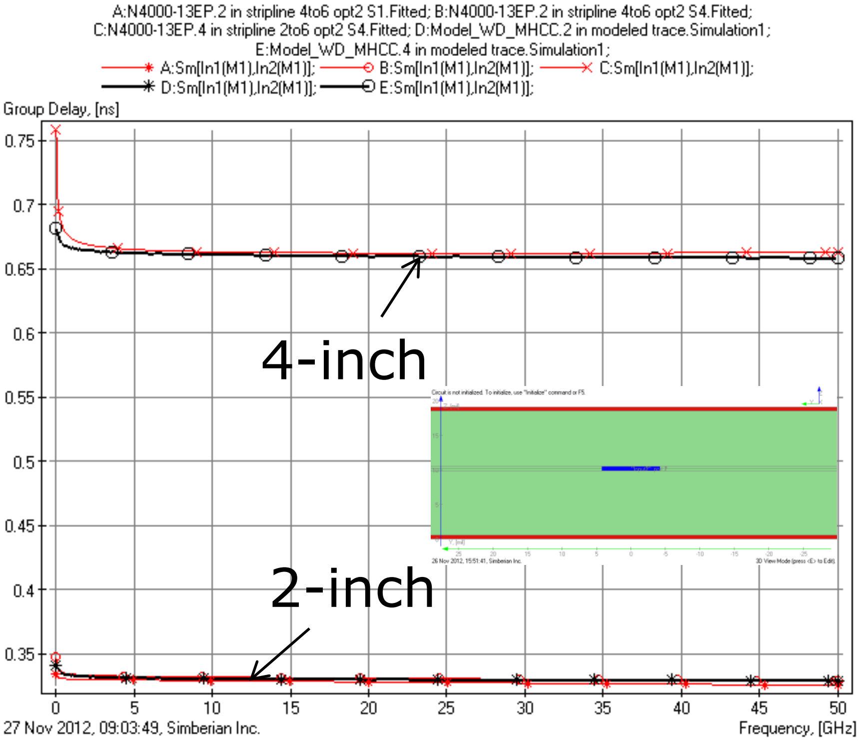
Wideband Debye dielectric model With roughness

- Dk=3.8, LT=0.008 @ 10 GHz – as in specs, modified Hammerstadt correction coefficient SR=0.27, RF=4 (relative resistivity 1.05) produces good fit (black lines) to measured GMS-parameters (red lines)

GMS Insertion Loss

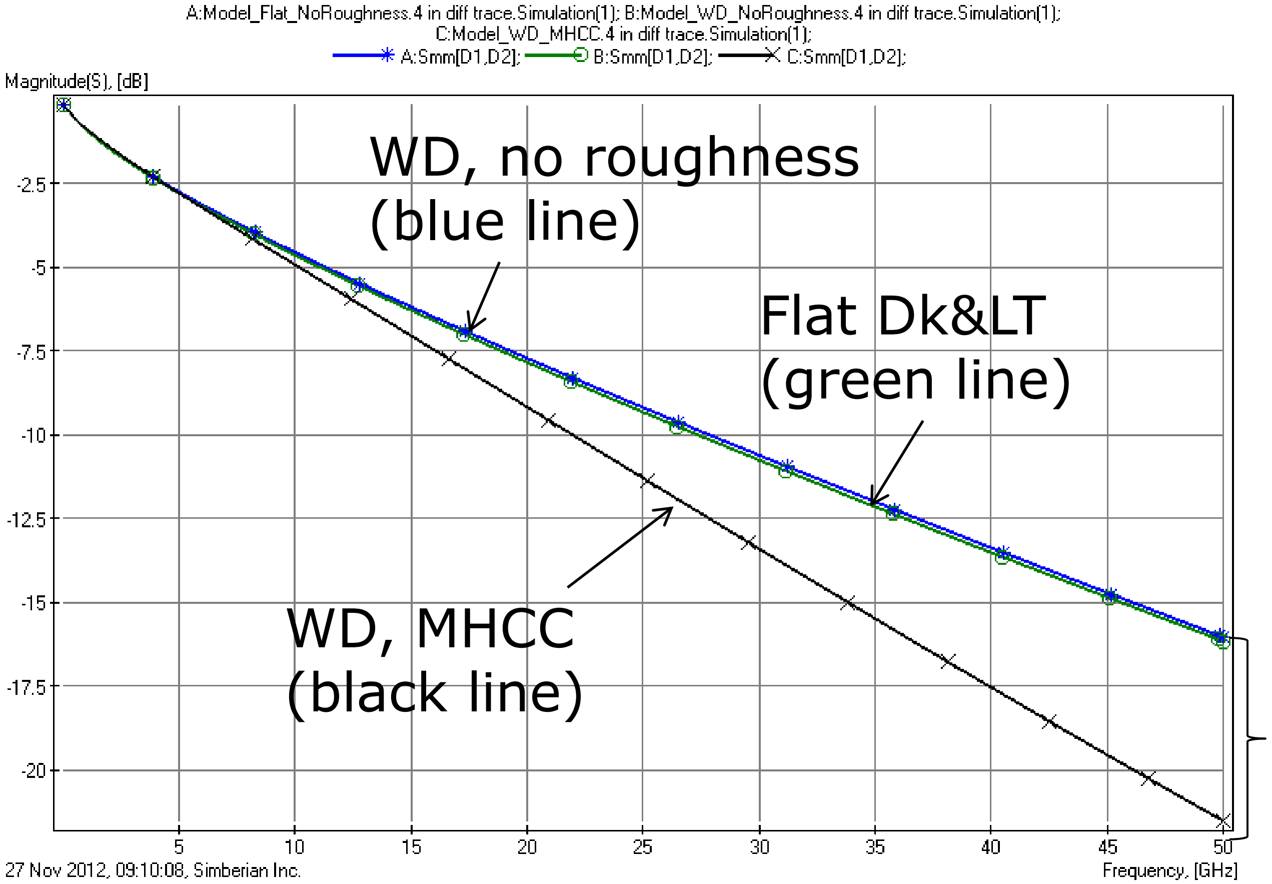


GMS Group Delay



Models for differential strips (4 mil wide, 4 mil distance)

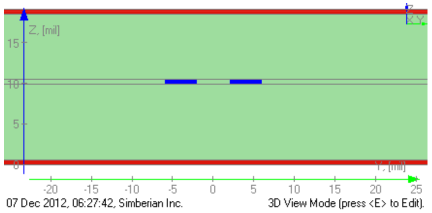
- Differential IL for flat and WD models without roughness are close, but model with the roughness predict much more loss!!!



WD, no roughness (blue line)

Flat Dk&LT (green line)

WD, MHCC (black line)



All models predict close phase and GD

Over 40% difference!!!

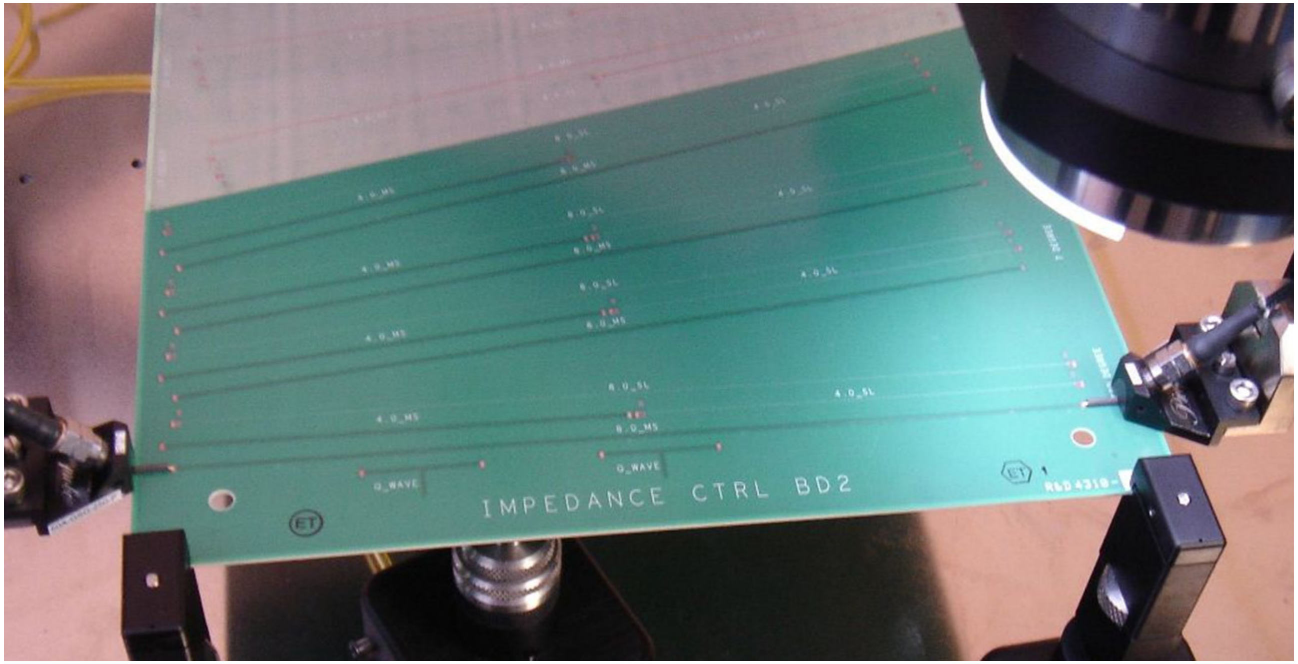
Summary on N4000-13EP example

- Identified dielectric constants (DK) are close to the specifications
  - Small differences due to anisotropy and non-identities of fixtures
- Very large difference in LT if copper assumed smooth
  - Small dielectric dispersion points at small increase of dielectric loss (consistent with specs, no presence of water)
  - The rest of the losses are due to the conductor roughness
- Separation of conductor roughness model from dielectric model is important in case if traces with different widths are used on the same board
- Without roughness model, dielectric models must be built for every cross section!

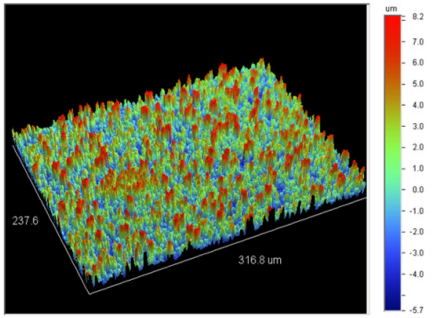
Isola’s test board (designed with Simbeor)

- 8 layer stackup with two microstrip layers (Top and Bottom) and 2 strip-line layers (L3, and L6)
- Microstrip Top - TWS copper foil, 1080 prepreg, no solder mask
- Strip L3 - TWS copper foil, laminate 1080 core and prepreg
- Strip L6 – LP3 copper foil, laminate 2116 core and prepreg
- Microstrip Bottom – LP3 copper foil, laminate 2116 prepreg

Test structures – 4 and 8 inch line segment with transitions to probe pads



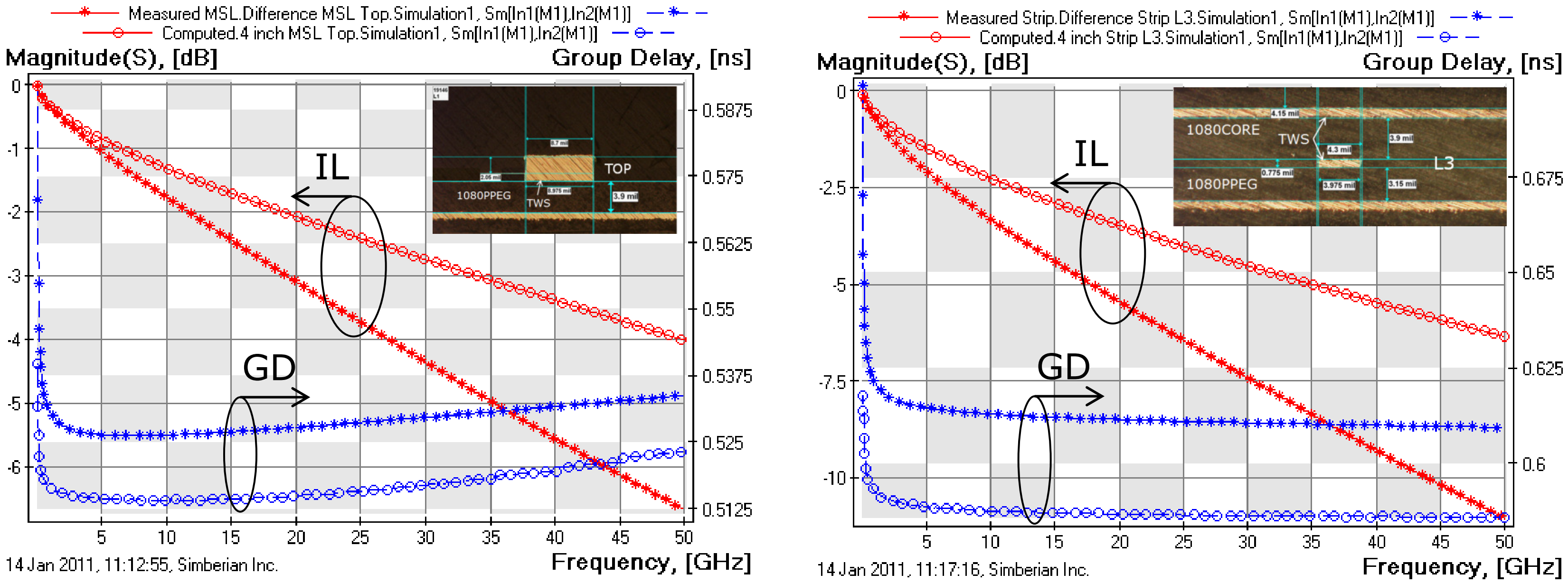
TWS surface (Rq=2.6 um):



Complete description is in: Y. Shlepnev, C. Nwachukwu, Practical methodology for analyzing the effect of conductor roughness on signal losses and dispersion in interconnects, DesignCon2012, Feb. 1st, 2012, Santa Clara, CA

TWS & IS680-1080 – No Roughness

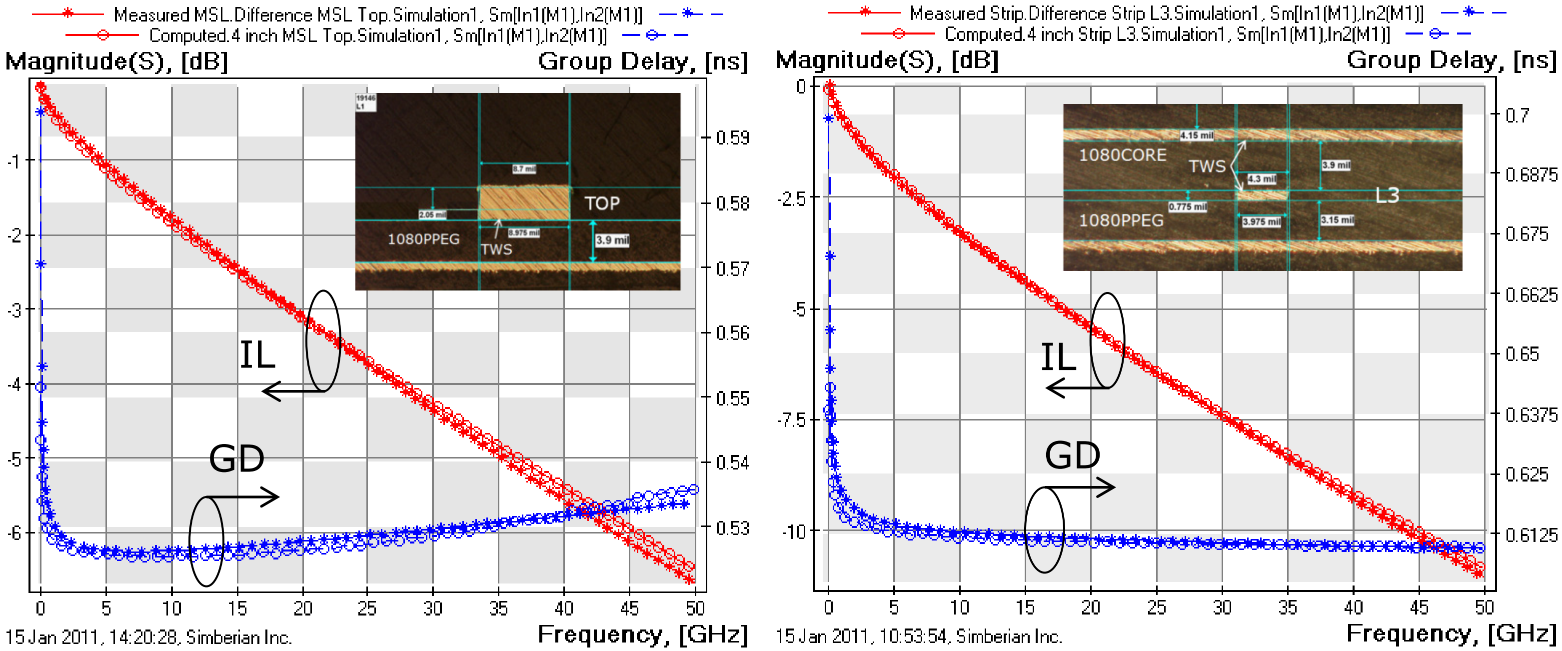
- Berezkin method: Dk=3.0+-0.05, LT=0.003+-0.0005 @ 2.5 GHz
- Huge difference in insertion loss (IL) and in Group Delay both in microstrip and strip-line configurations (GMS, 4-inch)



Stars – measured and fitted, Circles - modeled

TWS & IS680-1080 – Adjusted roughness parameters to fit the measurements (Simbeor)

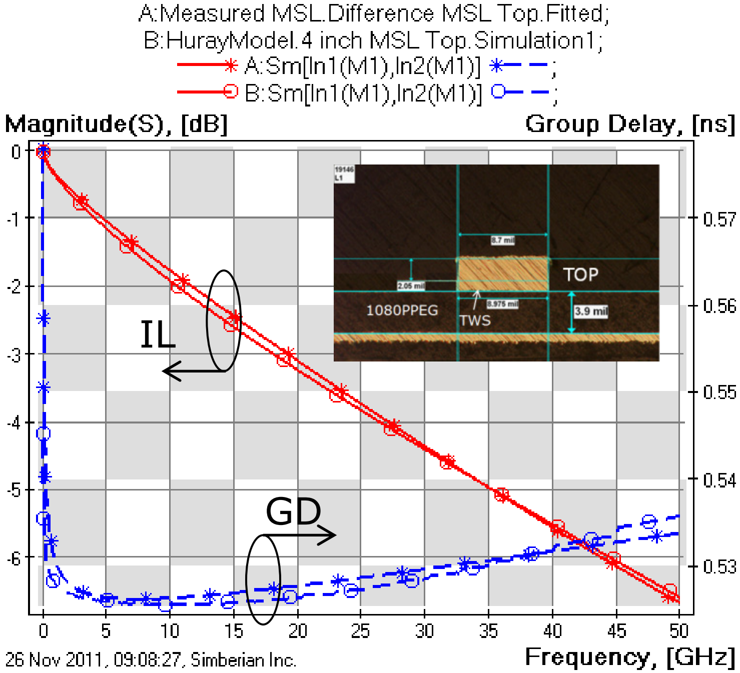
- **Dielectric constants are adjusted 3 -> 3.15 for 1080 prepreg, 3-> 3.35 for 1080 core**
- Roughness parameters: Rq=0.35 um, RF=2.8 for all surfaces
- Both insertion loss and group delay now match well!

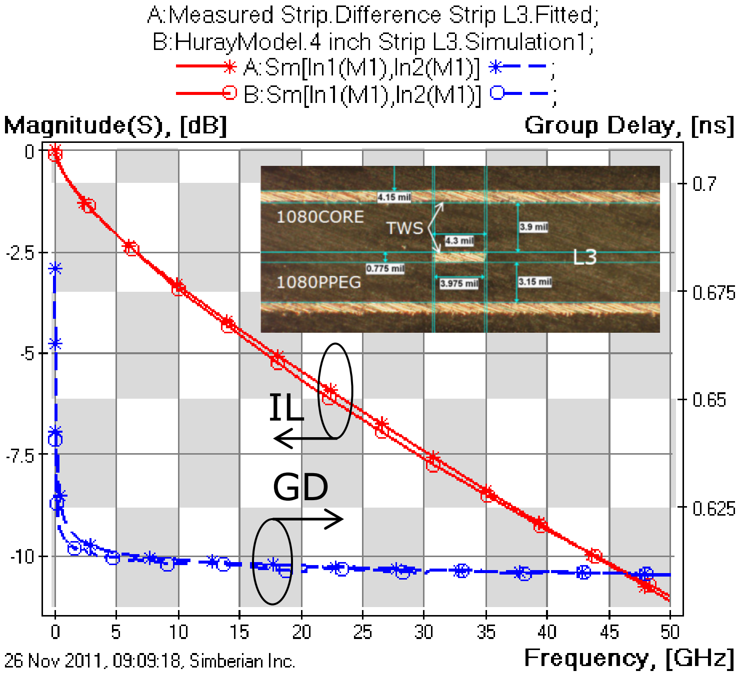


Stars – measured and fitted, Circles - modeled

TWS & IS680-1080 – Adjusted roughness parameters to fit the measurements (Huray's snowball model)

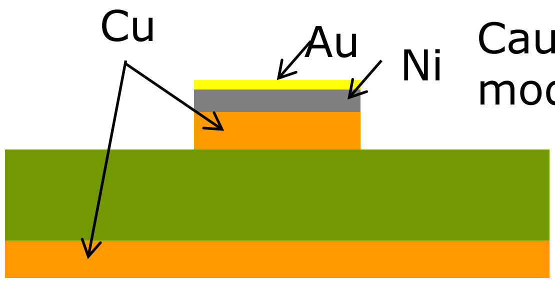
- Dielectric constants are adjusted 3 -> 3.15 for 1080 prepreg, 3-> 3.35 for 1080 core
- Roughness parameters: Ball radius 0.8 um, tile size 9.9 um, Nb=20, Rr=1.14
- Acceptable accuracy!



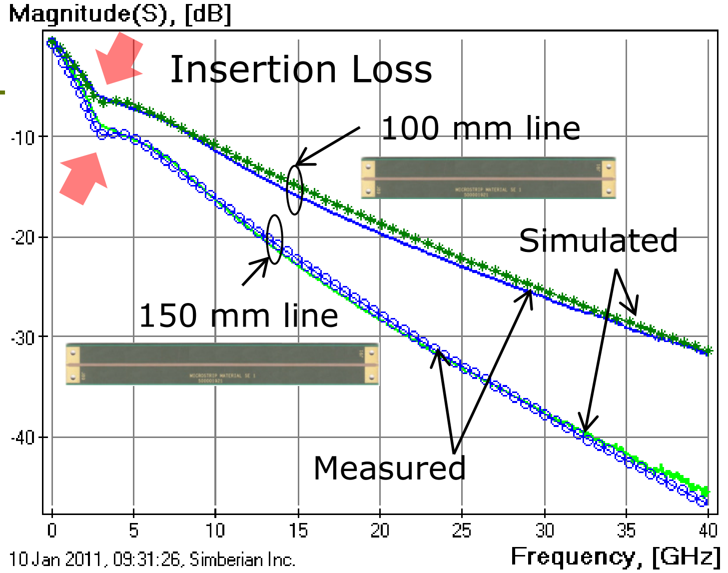


Stars – measured and fitted, Circles - modeled

Plated nickel model identification for ENIG finish (EMC 2011)



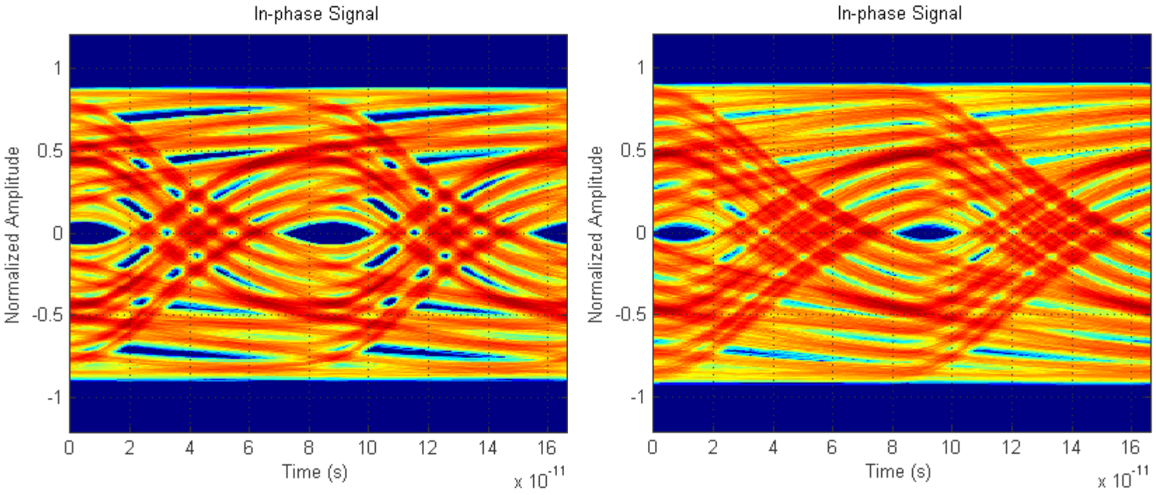
Causal Landau-Lifshits model for nickel layer

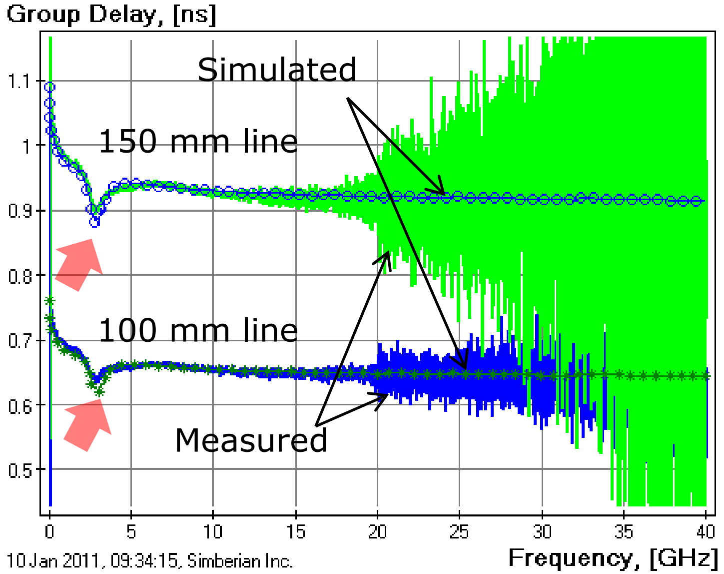


150 mm MSL link, 12 Gb/s

Simulated

Measured





Computations and measurements provided by Teraspeed Consulting Group

Material identification (summary)

- Any interconnect design project at 10 Gbps and above must start from the dielectric and roughness parameters identification
- Material parameters identification with GMS-parameters is the simplest and the most accurate for PCB
  - Verified in multiple projects and implemented in Simbeor software
- For successful identification S-parameters and test fixtures have to be pre-qualified
  - Pass the quality metrics in Simbeor Touchstone Analyzer
  - Have consistent impedance on TDR plots

Validation of analysis with measurements (benchmarking)

Building benchmark boards

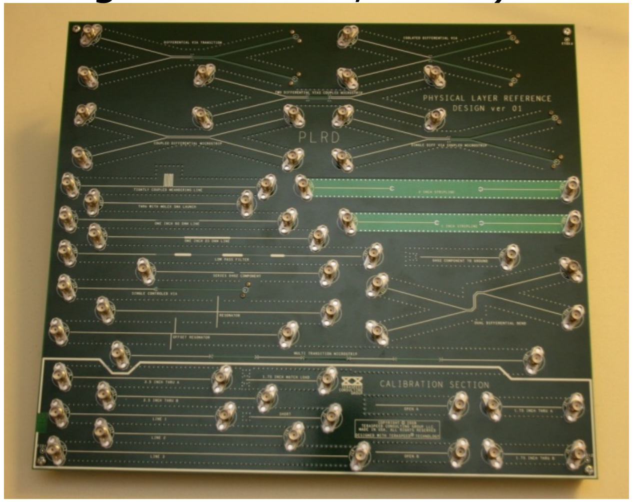
To understand capabilities and limitation of a solver

Benchmarking board

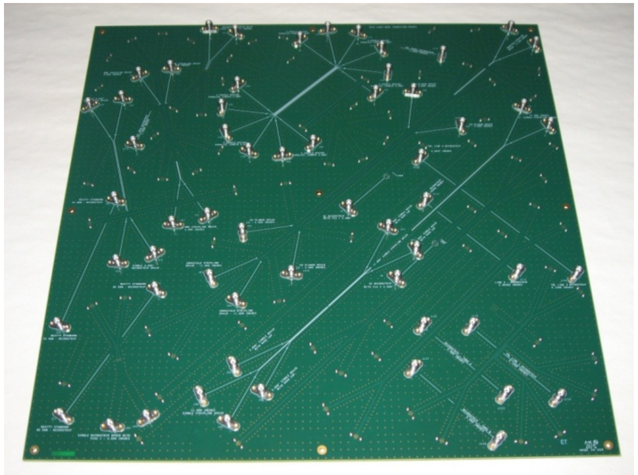
- Controlled board manufacturing is the key for success – fiber type, resin content, copper roughness must be strictly specified or fixed!!!
- A set of structures to identify one material model at a time
    - Solder mask, core and prepreg, resin and glass, roughness, plating,…
- A set of structures to identify accuracy for transmission line (with possible coupling) and typical discontinuities
    - Use identified material models for all structures on the board consistently
    - No tweaking - discrepancies should be investigated
- A set of structures for TRL-type de-embedding
    - Simple T-matrix de-embedding does not work on PCBs!
- Alternatively, build models for launches (jitter de-embedding approach)
    - Probe launches are the most accurate (require probe station)
- Use VNA/TDNA measurements and compare both magnitude and phase (or group delay) of all S-parameters

Example of benchmarking boards

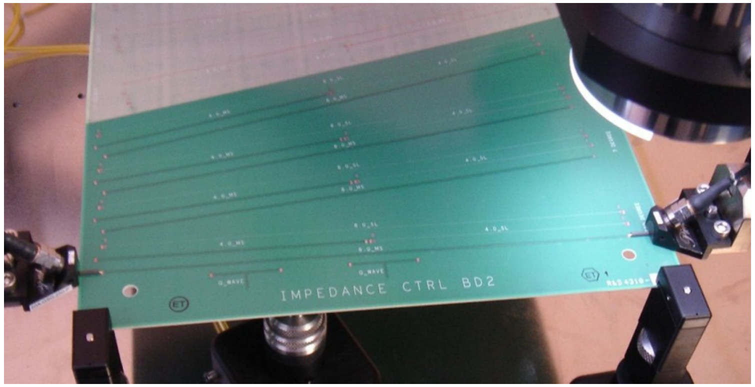
PLRD-1 (Teraspeed Consulting, DesignCon 2009, 2010)



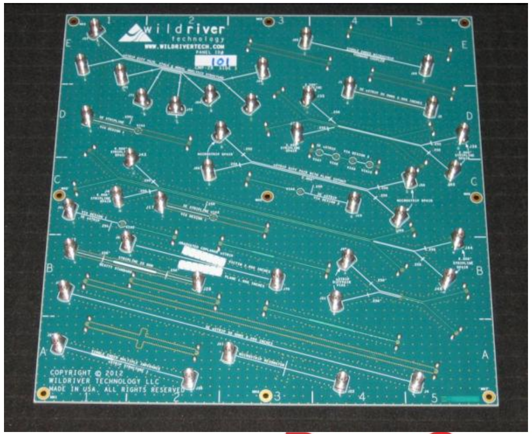
CMP-08 (Wild River Technology & Teraspeed Consulting, DesignCon 2011)



Isola, EMC 2011, DesignCon 2012



CMP-28, Wild River Technology, DesignCon 2012

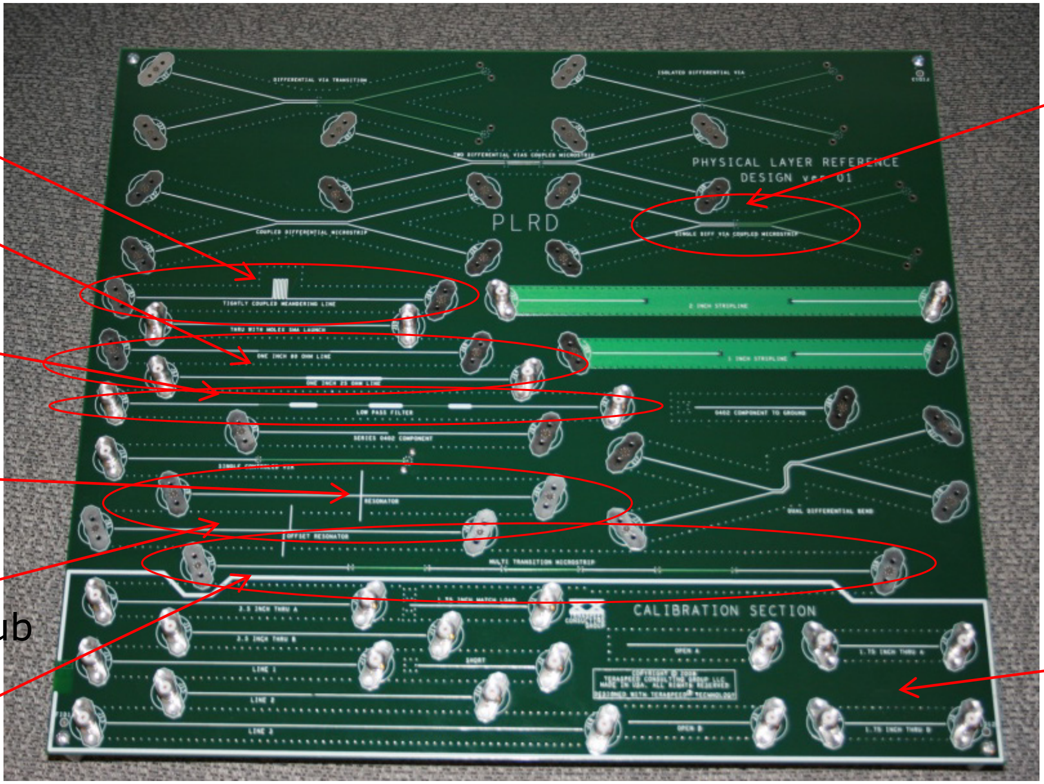


Hybrid simulation technology is used to illustrate benchmarking (Simbeor software)

- Method of Lines (MoL)
  - More accurate than finite element method (FEM) and finite integration technique (FIT) for problems with multiple dielectric and metal layers
  - Provides conductor interior solution for metal planes
- Trefftz Finite Elements (TFE)
  - Used to model strip conductor interior with rough surface
- Method of Simultaneous Diagonalization (MoSD)
  - Extracts modal and per unit length parameters of lossy multi-conductor lines and periodic structures
  - Allows precise non-reflective de-embedding
  - Provides 3D observable definition of the characteristic impedance

PLRD-1 benchmark board examples

- 4-layer stackup with two planes and 2 signal layers
- 30 test structures – all equipped with SMA connectors



Meander

Beatty standards

Low-pass filter

Two-stub resonator

Two offset stub resonators

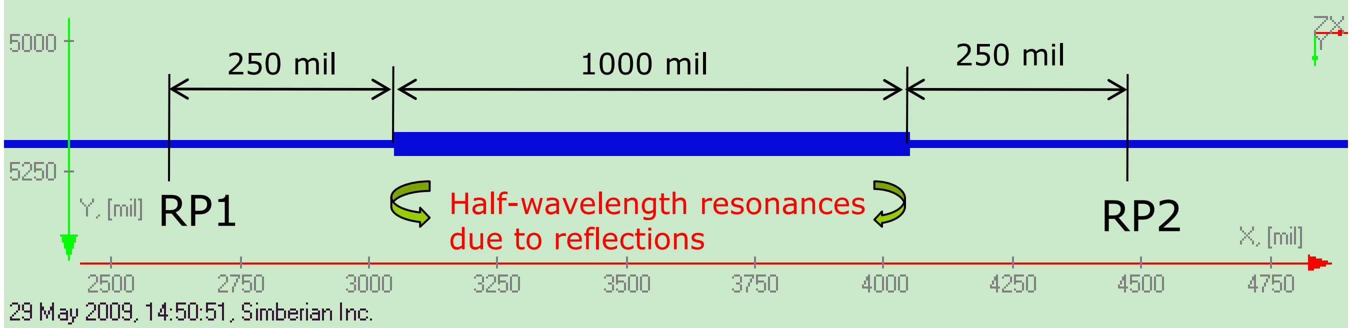
Single-ended channel with 6 vias

Differential vias

PLRD-1 board created and independently investigated by Teraspeed Consulting Group
www.teraspeed.com

Structures for TRL de-embedding and material identification

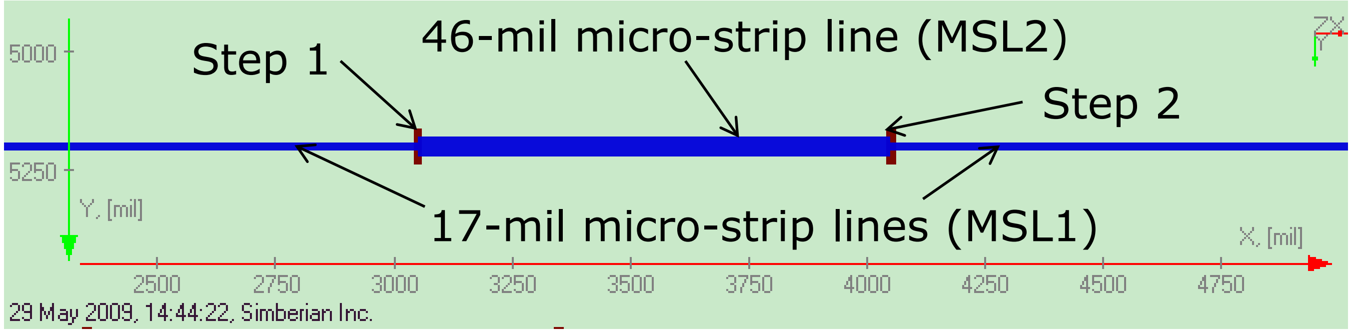
25-Ohm micro-strip Beatty standard



- 1-inch 46 mil wide micro-strip line segment connected in series into 17-mil wide micro-strip line
- DK=4.0, LT=0.018 @ 1 GHz, WD model – lower DK for wider line (anisotropy)
- Conductor roughness 0.5 um, RF=2
- De-embedded to reference planes to keep 250 mil micro-strip segments on both sides of the structure
- Can be analyzed as a whole or with decomposition into two step discontinuities and line segments
- De-compositional analysis is faster and more accurate

De-composition of 25-Ohm Beatty standard

Two rectangular discontinuity selectors created to de-compose the structure in 5 elements



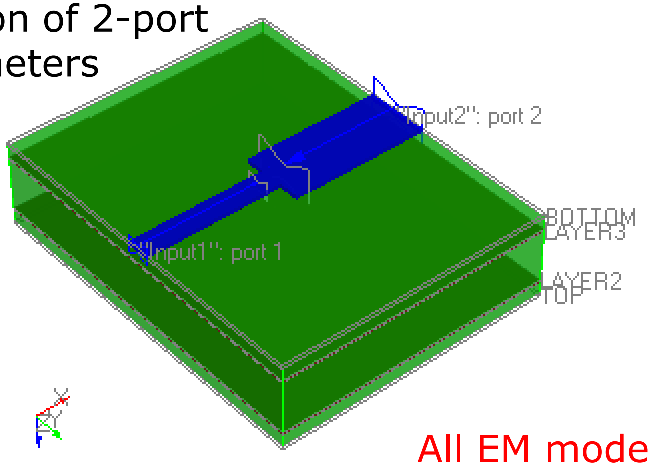
Simbeor de-compositional model (linear network)


Port 1 — MSL1 — Step 1 (S-parameters) — MSL2 — Step 2 (S-parameters) — MSL1 — Port 2

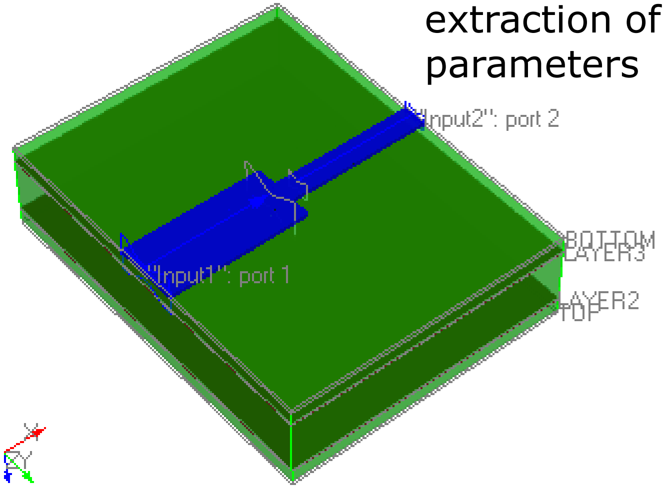
Auto-decomposition is used here as demonstrated in screen-cast #2009_03 at http://www.simberian.com/ScreenCasts.php

Circuit elements automatically created for the electromagnetic extraction

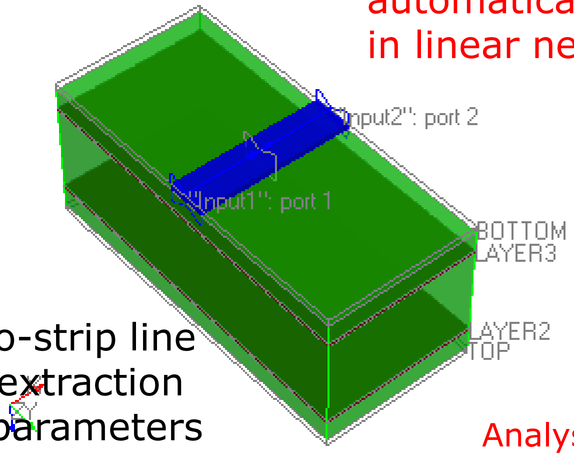
Step 1 – full-wave extraction of 2-port S-parameters



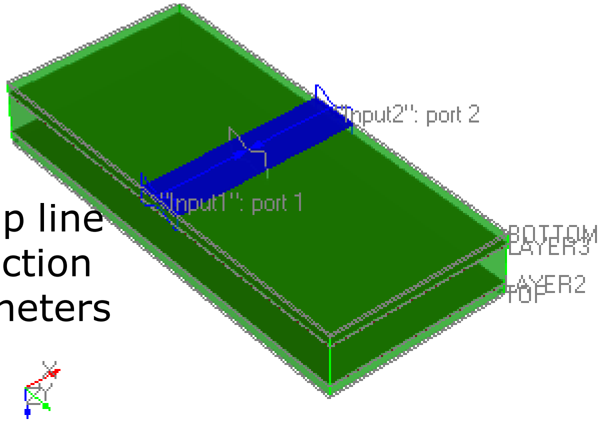
Step 2 – full-wave extraction of 2-port S-parameters



All EM models are automatically re-composed in linear network “Beatty25”



17-mil micro-strip line – full-wave extraction of RLGC(f) parameters

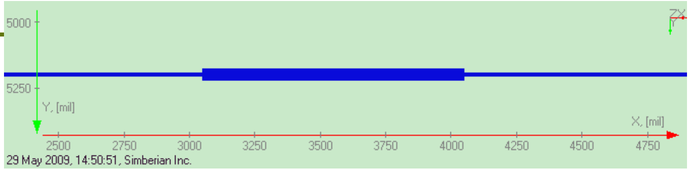


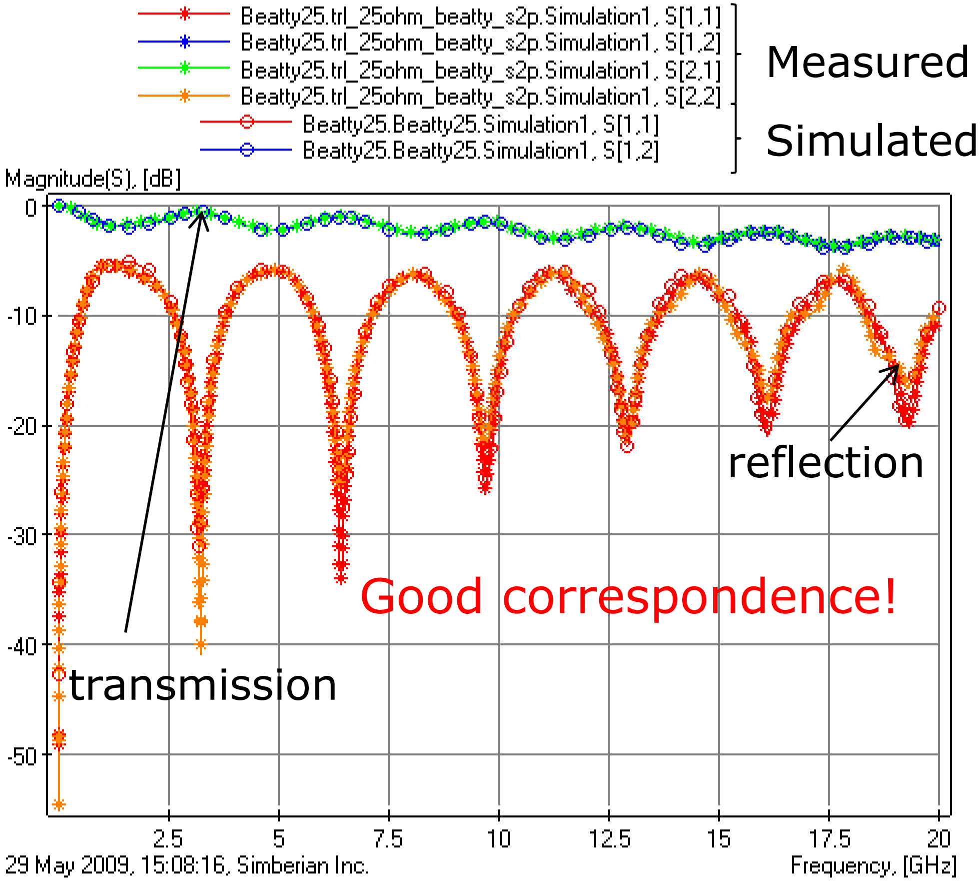
46-mil micro-strip line – full-wave extraction of RLGC(f) parameters

Analysis takes less than 1 min and all models are re-usable for possible fast “tuning” by adjustment of the t-line lengths

Comparison with measurement results de-embedded with TRL

- Magnitudes of S-parameters





Measured (stars)

Simulated (circles)

reflection

transmission

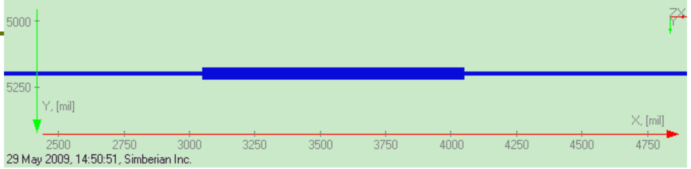
Good correspondence!

Measured Data Quality Metric:
Passivity QM=99.9999%
Reciprocity QM=99.21%
Symmetry QM=38.6%

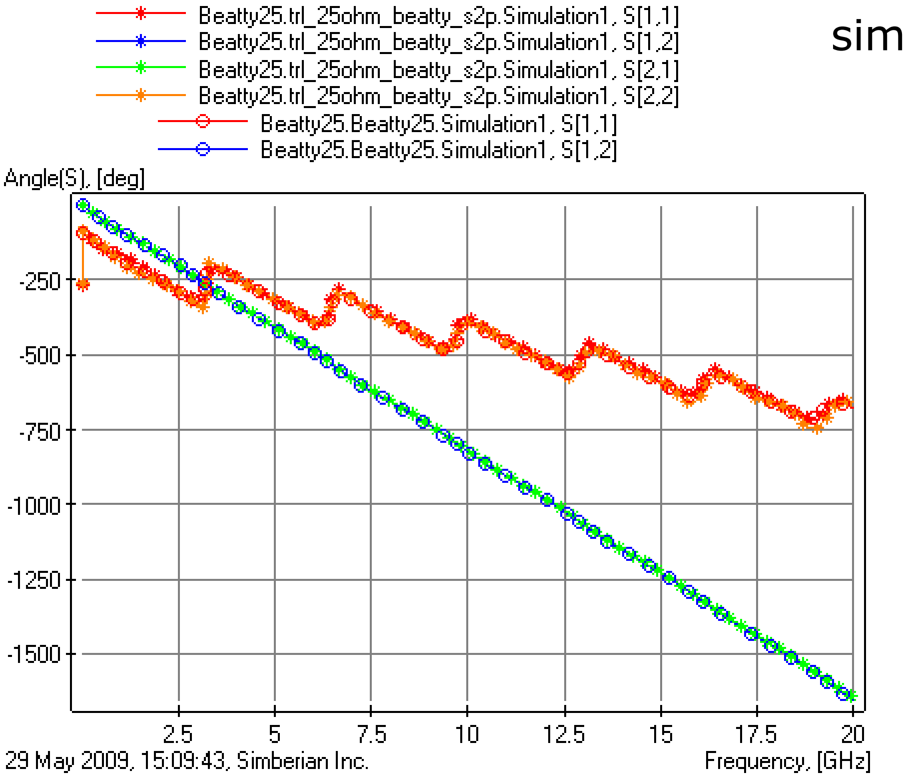
Visible difference in |S11| and |S22| - the actual structure has mirror geometric symmetry violations (manufacturing variations and the weave effect)

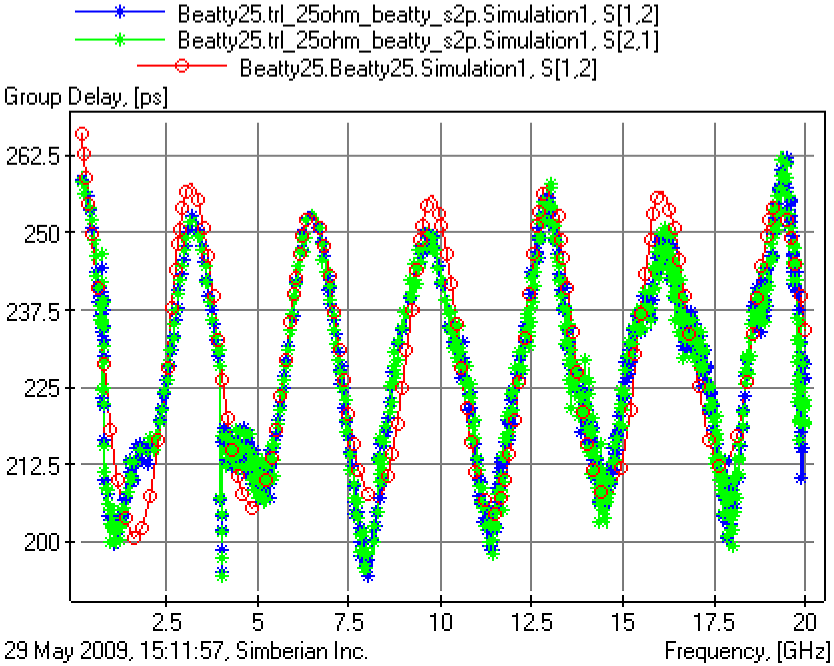
Comparison with measurement results de-embedded with TRL

- Phase and group delay



Measured – stars, simulated - circles

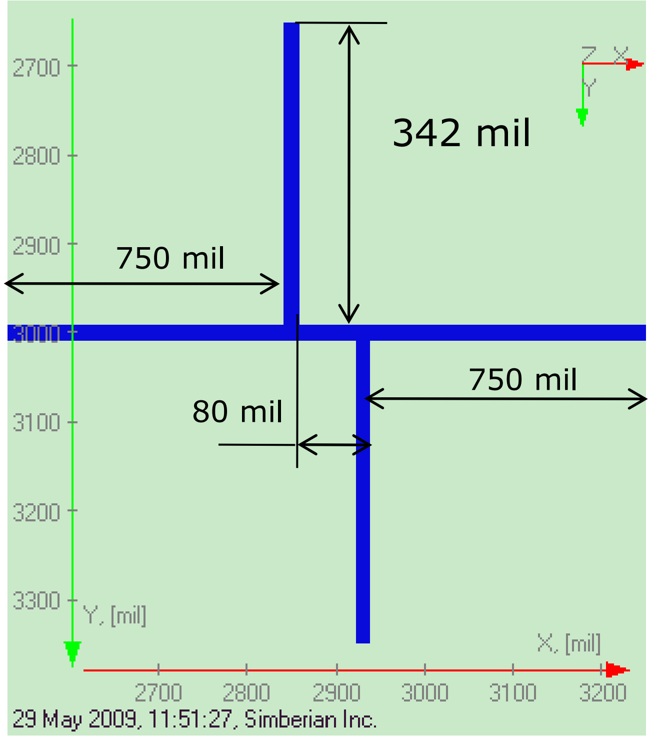




Good correspondence!

Micro-strip resonator with two offset stubs

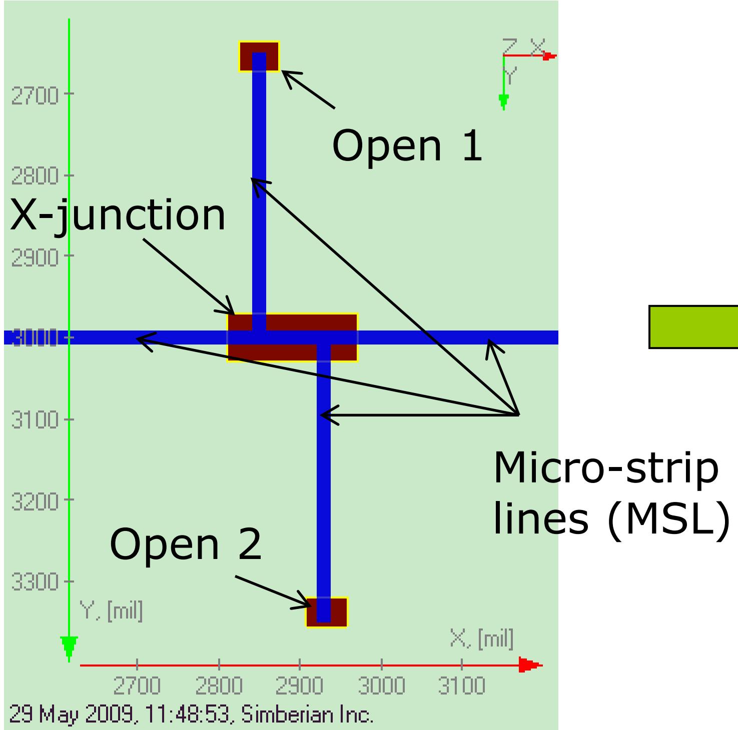
- Two 17-mil wide micro-strip stubs separated by 80 mil as shown
- DK=4.2, LT=0.018 @ 1 GHz, WD model
- Conductor roughness 0.5 um
- De-embedded to reference planes to keep 750 mil micro-strip segments on both sides of the structure
- Can be analyzed as a whole or with decomposition into three discontinuities and line segments
- De-compositional analysis is faster and more accurate



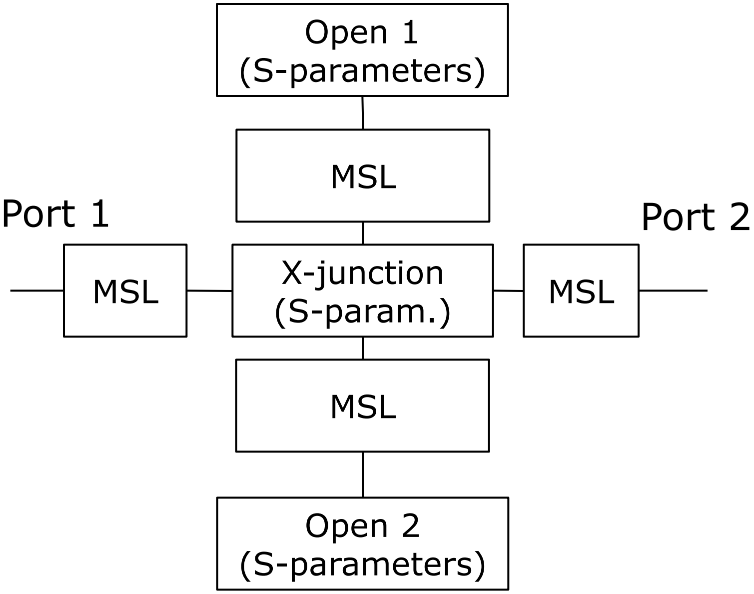
29 May 2009, 11:51:27, Simberian Inc.

De-composition of two-stub resonator

Three rectangular discontinuity selectors created to de-compose the structure in 7 elements



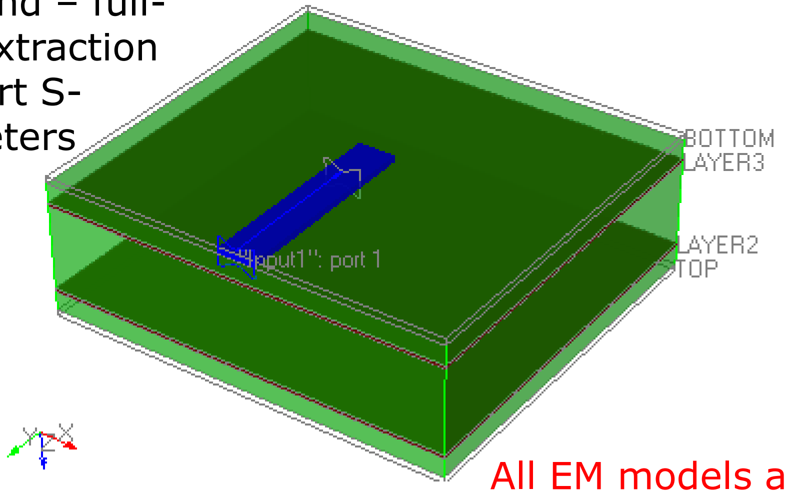
Simbeor de-compositional model (linear network)



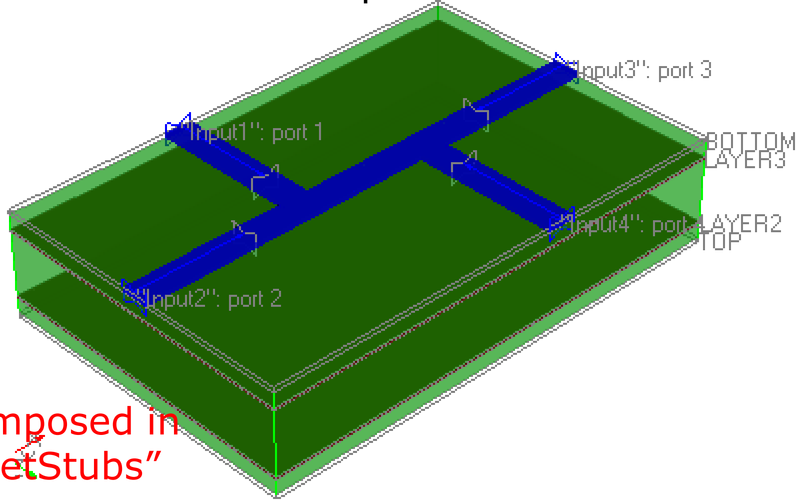
Auto-decomposition is used here as demonstrated in screen-cast #2009_03 at http://www.simberian.com/ScreenCasts.php

Circuit elements are automatically created for the electromagnetic extraction

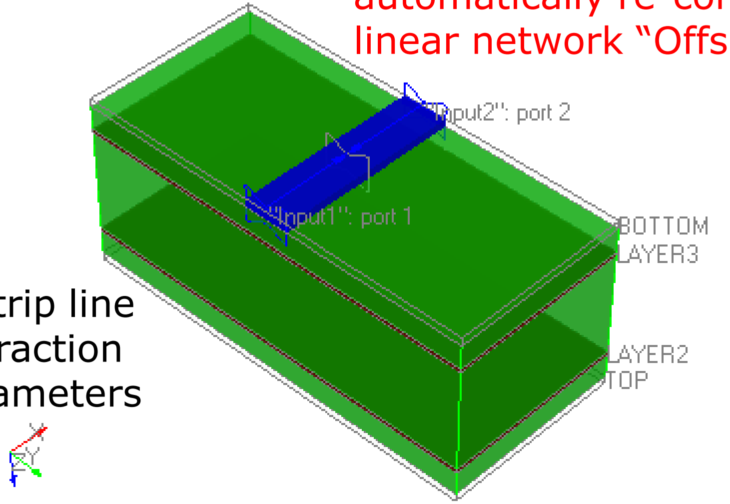
Open end – full-wave extraction of 1-port S-parameters



X-junction – full-wave extraction of 4-port S-parameters



All EM models are automatically re-composed in linear network "OffsetStubs"

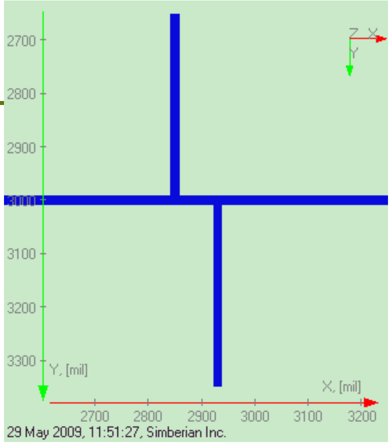


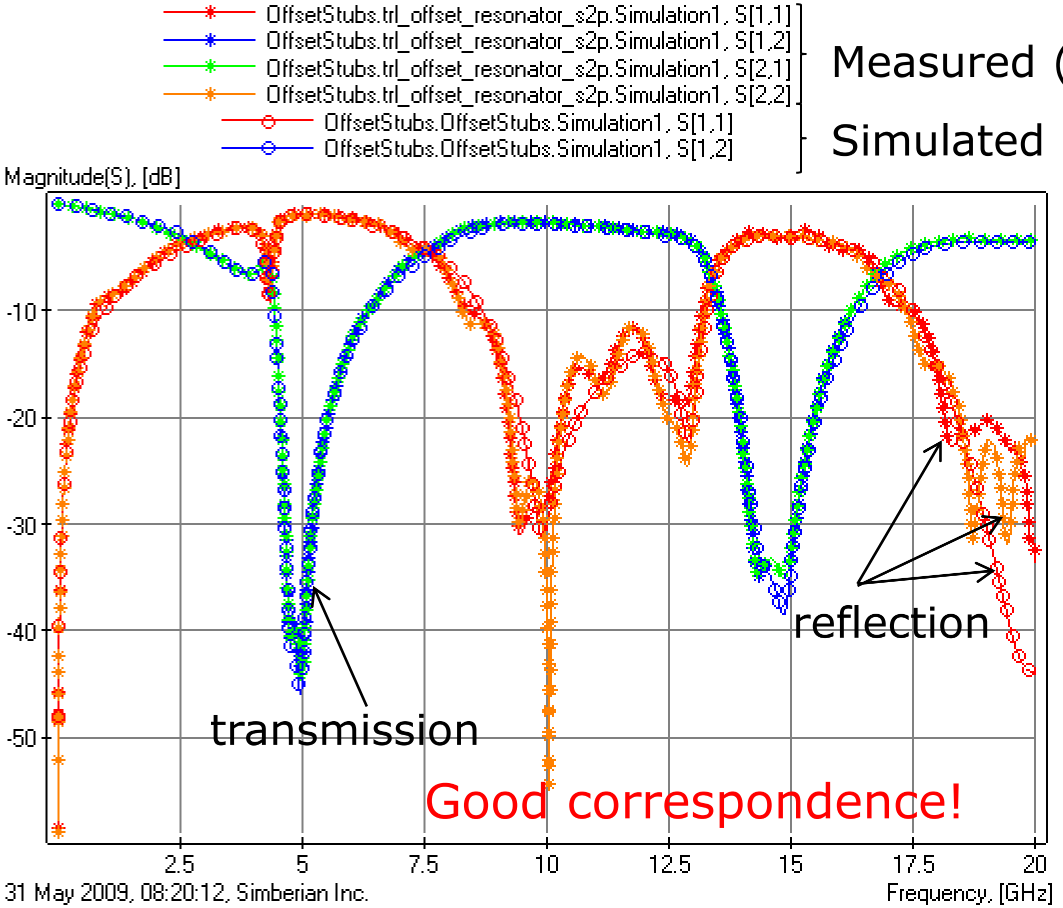
17-mil micro-strip line – full-wave extraction of RLGC(f) parameters

Analysis takes less than 1 min and all models are re-usable for possible fast "tuning" by adjustment of the t-line lengths

Comparison with measurement results de-embedded with TRL

- Magnitudes of S-parameters





Measured (stars)

Simulated (circles)

transmission

reflection

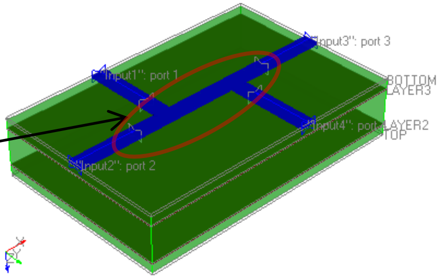
Good correspondence!

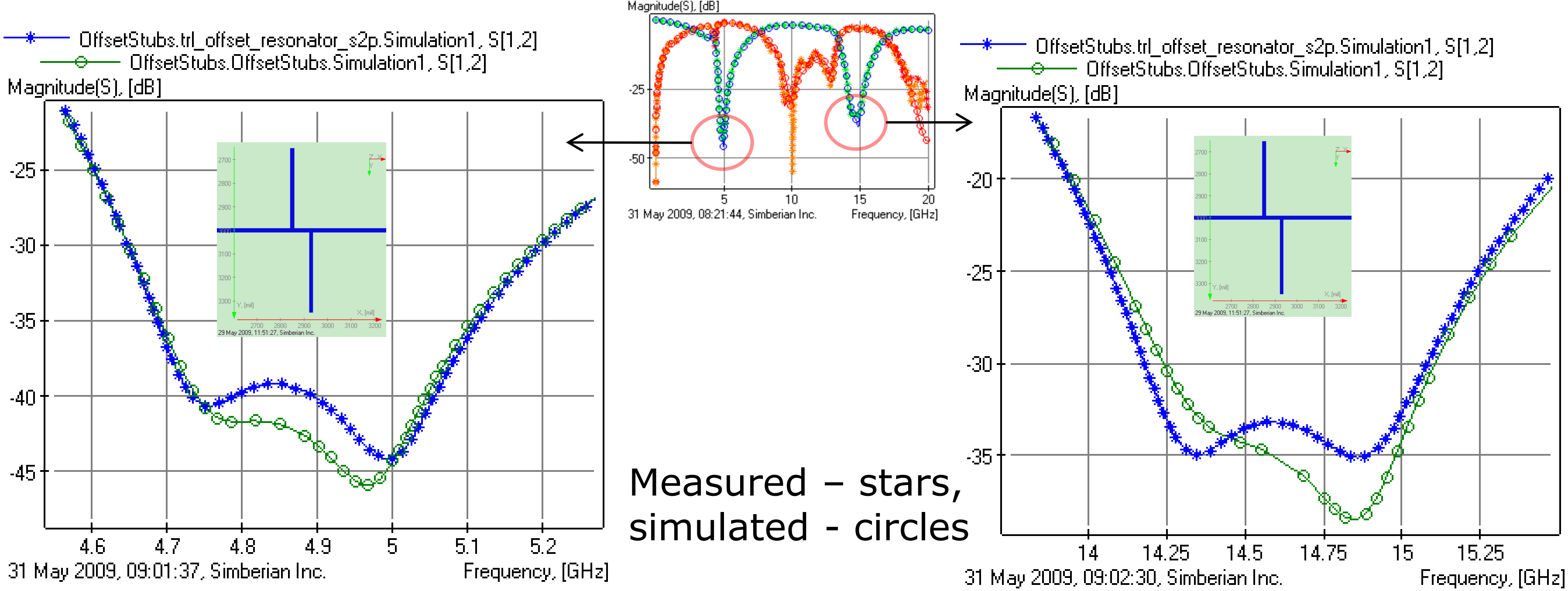
Measured Data Quality Metric:
Passivity QM=100%
Reciprocity QM=99.41%
Symmetry QM=0%

Visible difference in |S11| and |S22| - the actual structure has rotational geometric symmetry violations (manufacturing variations and the weave effect)

Double resonance effect

- The effect of interaction between the resonators first observed by M. Goldfarb and A. Platzker in “The effects of electromagnetic coupling on MMIC design”, Int. J. of Microwave and Millimeter-wave Computer-Aided Engineering, 1991, v.1, p. 38-47
- The de-compositional analysis proves that the effect is due to the interactions of two T-junctions through the high-order modes in micro-strip line connecting them



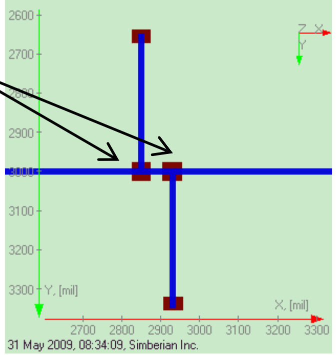


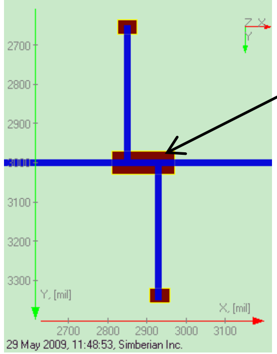
Measured – stars, simulated - circles

The effect cannot be observed without the EM analysis!

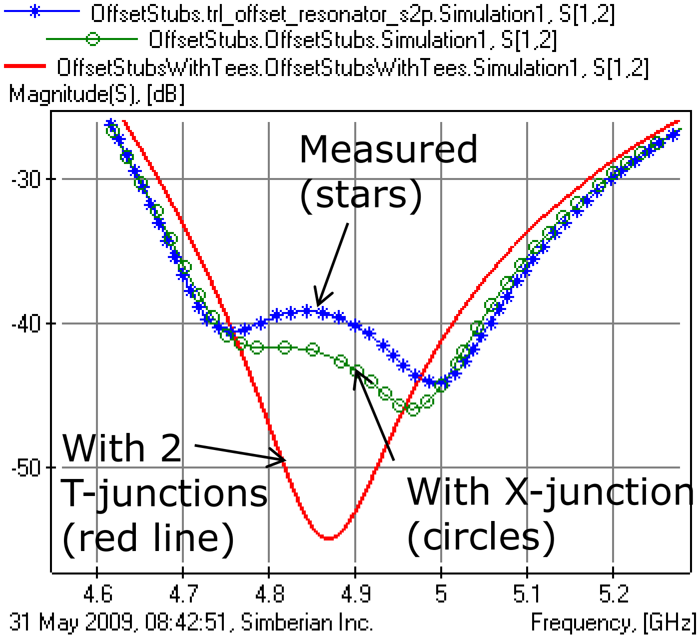
What if the interaction is ignored?

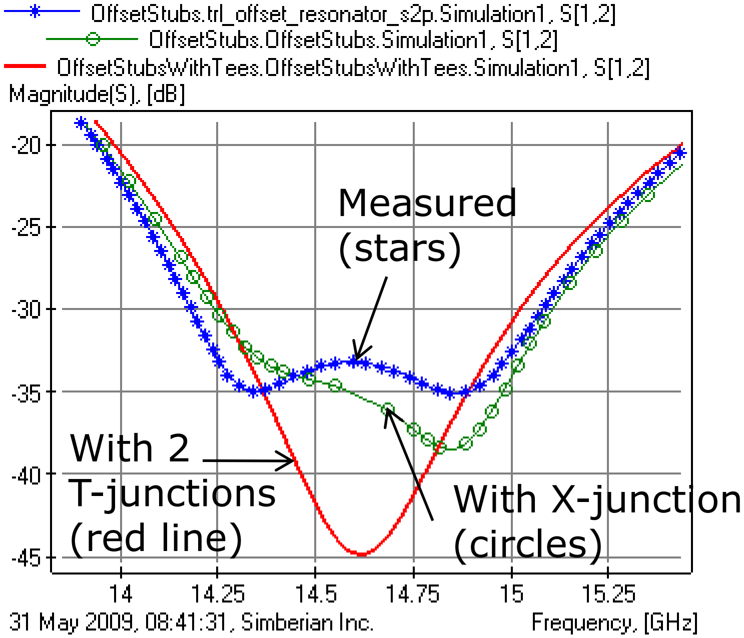
With two separate T-junction (no high-order mode interactions) – red lines on graphs





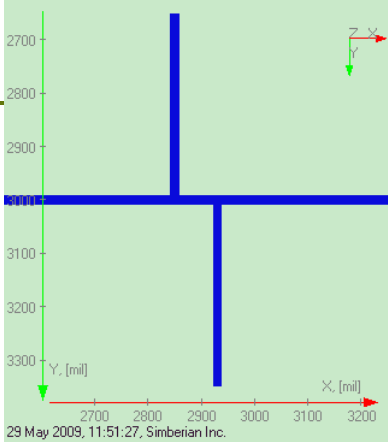
With single X-junction (high-order mode interactions) – green circles on graphs



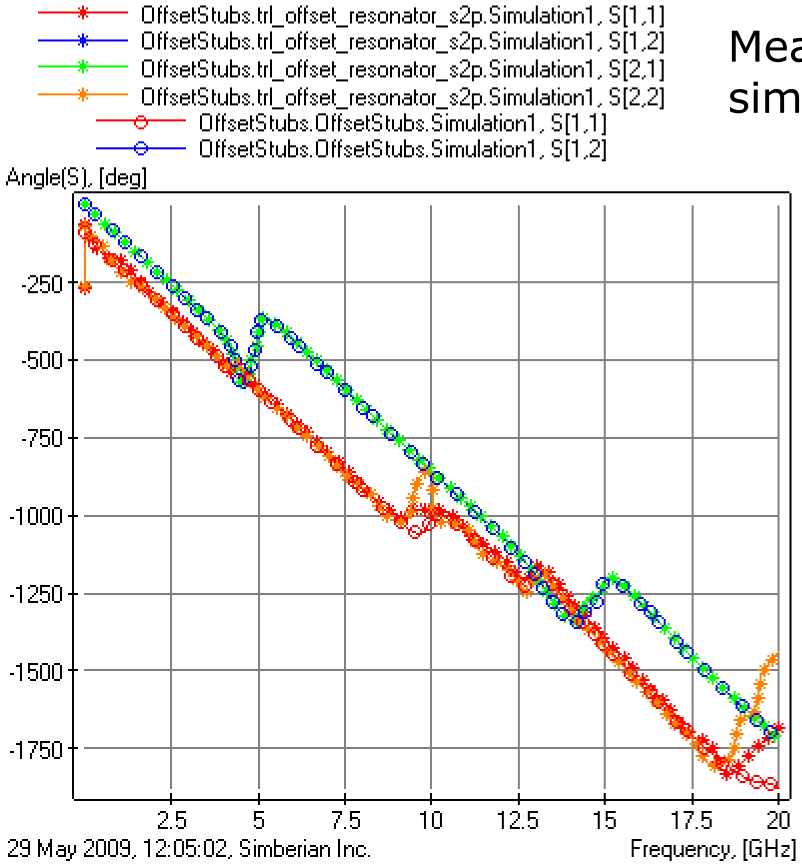


The effect cannot be observed without coupled discontinuities!

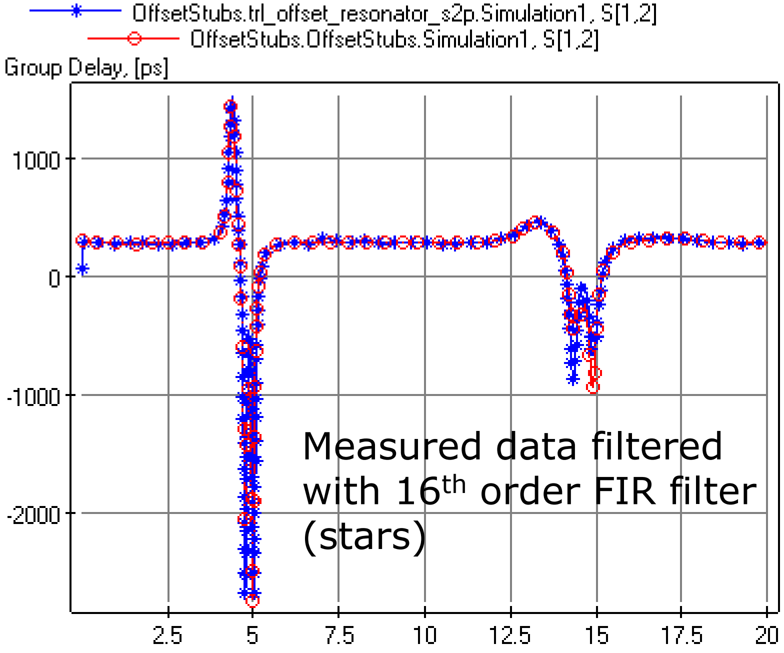
Comparison with measurement results de-embedded with TRL



- Phase and group delay



Measured – stars, simulated - circles

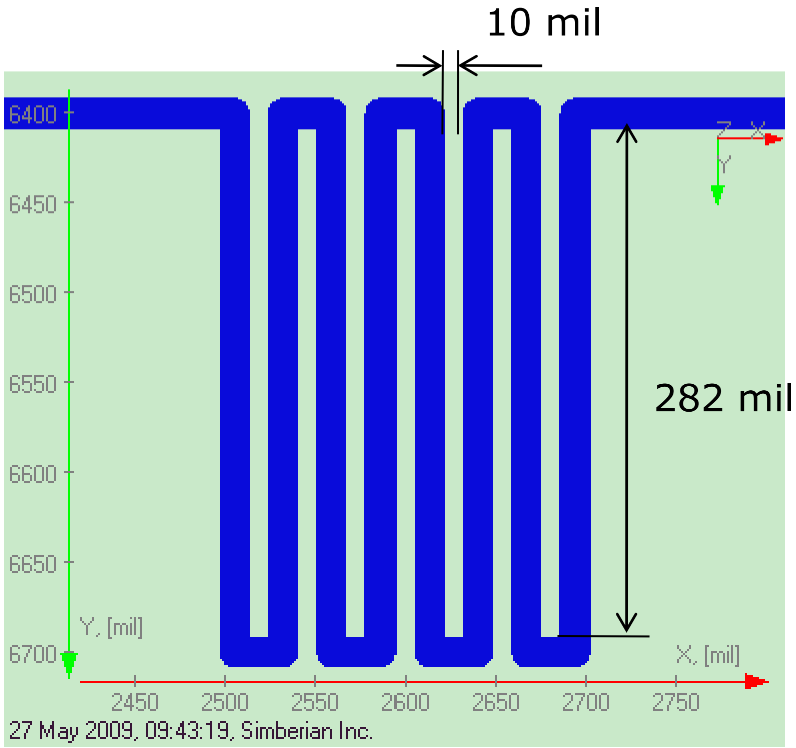


Measured data filtered with 16th order FIR filter (stars)

Good correspondence!

Meandering micro-strip line

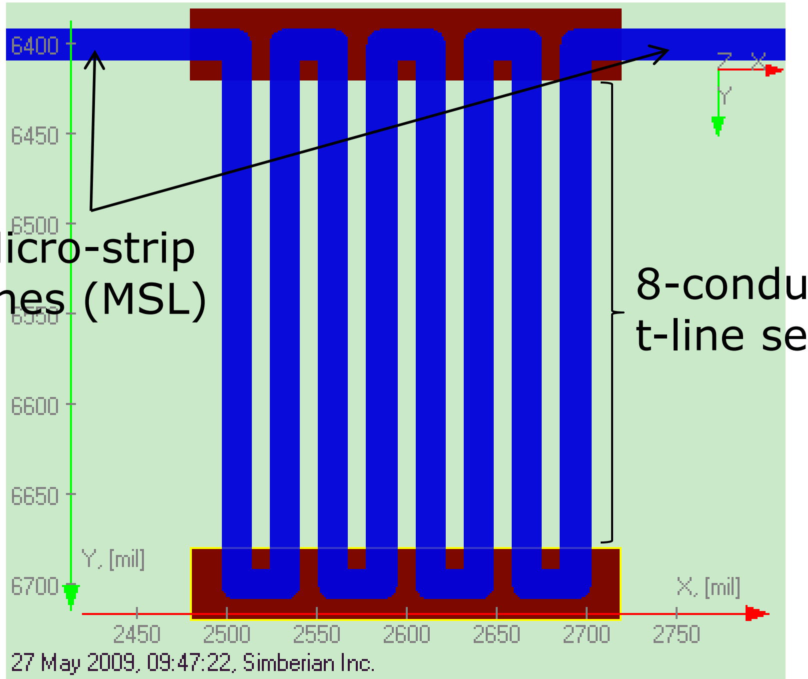
- Meandering 17-mil 2.6 inch long micro-strip line
- DK=4.2, LT=0.018 @ 1 GHz, WD model
- Conductor roughness 0.5 um
- De-embedded to reference planes to keep 390 mil micro-strip segments on both sides of the meander – total length of the line is 3380 mil
- Can be analyzed as a whole or with decomposition into two discontinuities and line segments
- De-compositional analysis is faster and more accurate



De-composition of the meander

Two rectangular discontinuity selectors created to de-compose the meander in 5 elements

Mender Top

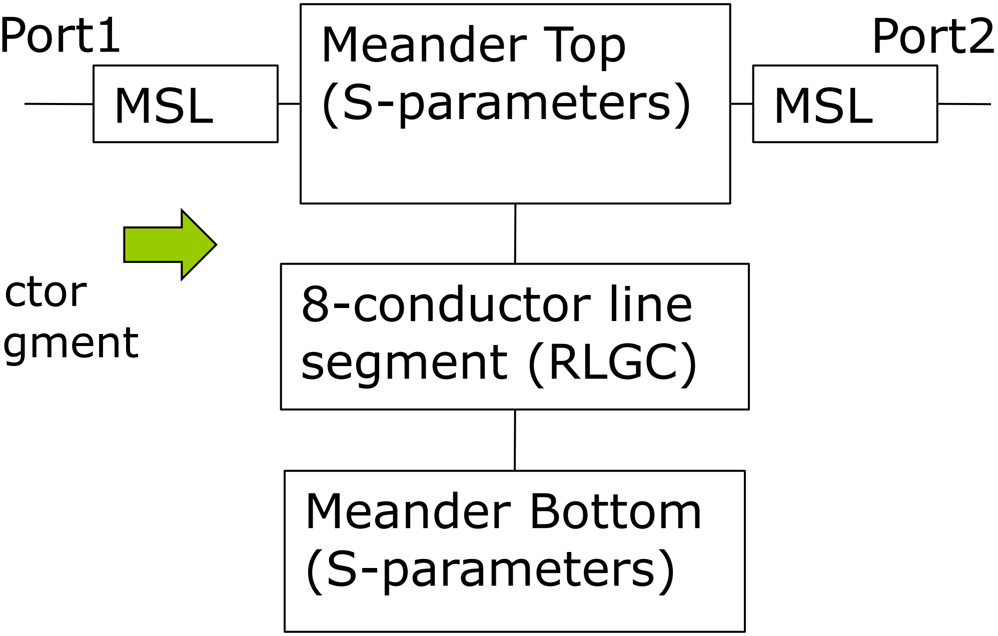


Micro-strip lines (MSL)

8-conductor t-line segment

Meander Bottom

Simbeor de-compositional model (linear network)



Port1

MSL

Meander Top (S-parameters)

Port2

MSL

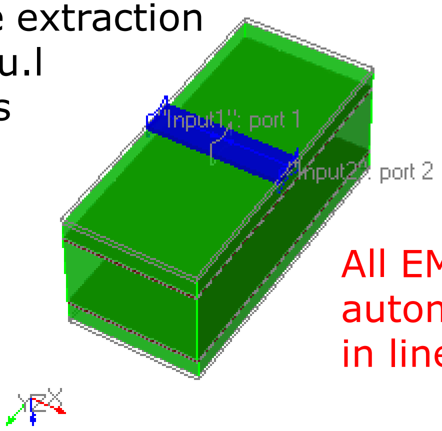
8-conductor line segment (RLGC)

Meander Bottom (S-parameters)

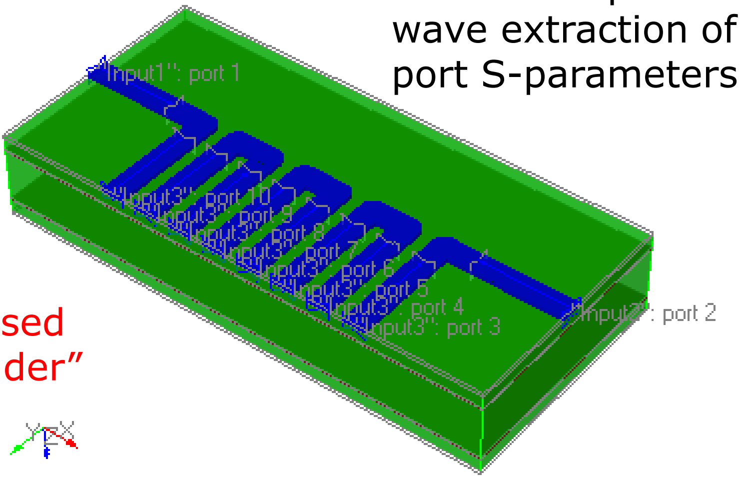
Auto-decomposition is used here as demonstrated in screen-cast #2009_02 at http://www.simberian.com/ScreenCasts.php

Circuit elements automatically created for the electromagnetic extraction

17-mil micro-strip line – full-wave extraction of RLGC p.u.l parameters

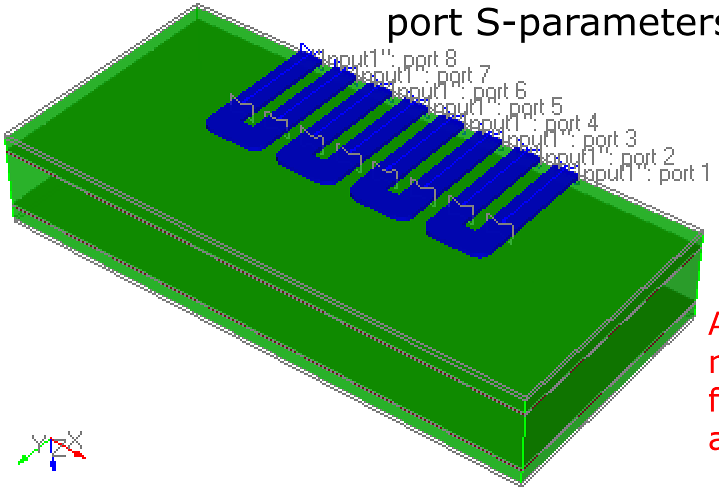


Meander Top – full-wave extraction of 10-port S-parameters

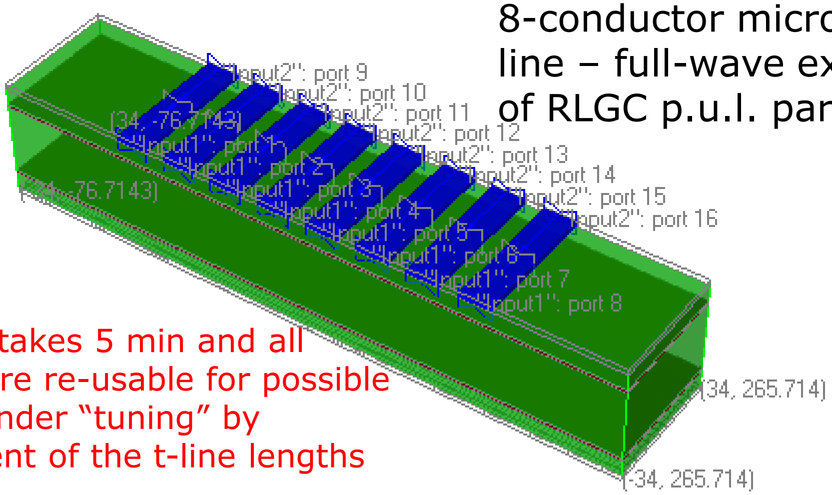


All EM models are automatically re-composed in linear network "Meander"

Meander Bottom – full-wave extraction of 10-port S-parameters

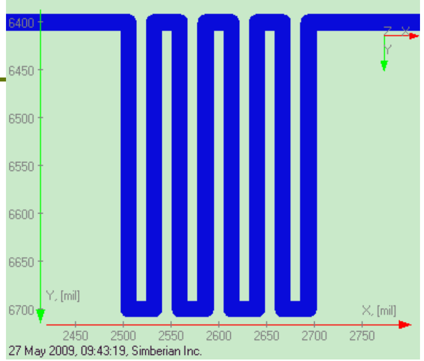


8-conductor micro-strip line – full-wave extraction of RLGC p.u.l. parameters

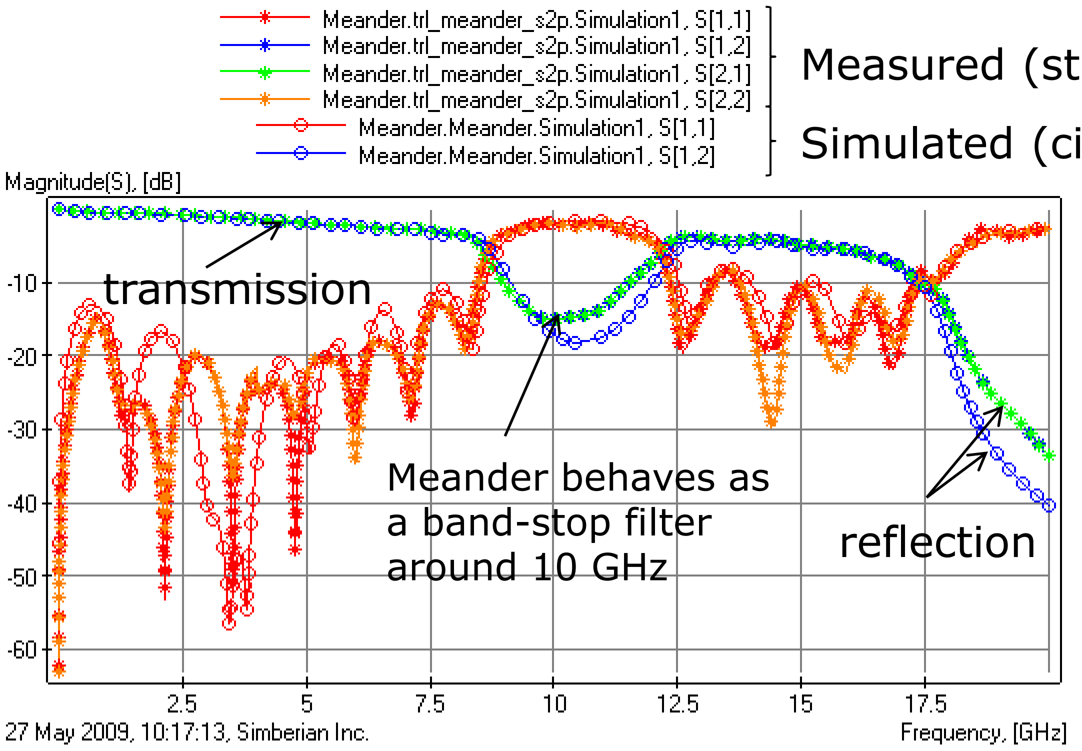


Analysis takes 5 min and all models are re-usable for possible fast meander "tuning" by adjustment of the t-line lengths

Comparison with measurement results de-embedded with TRL



- Magnitudes of S-parameters



Measured (stars)

Simulated (circles)

transmission

Meander behaves as a band-stop filter around 10 GHz

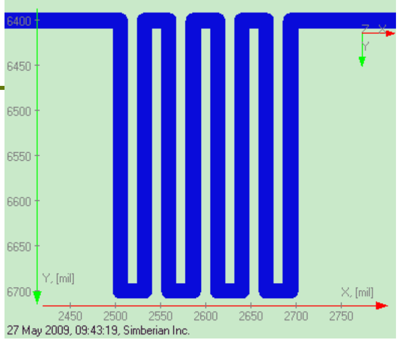
reflection

Measured Data Quality Metric:
Passivity QM=100%
Reciprocity QM=99.6%
Symmetry QM=49.3%

Some visible differences in |S11| and |S22| - the actual structure has mirror geometric symmetry violations (manufacturing variations and the weave effect)

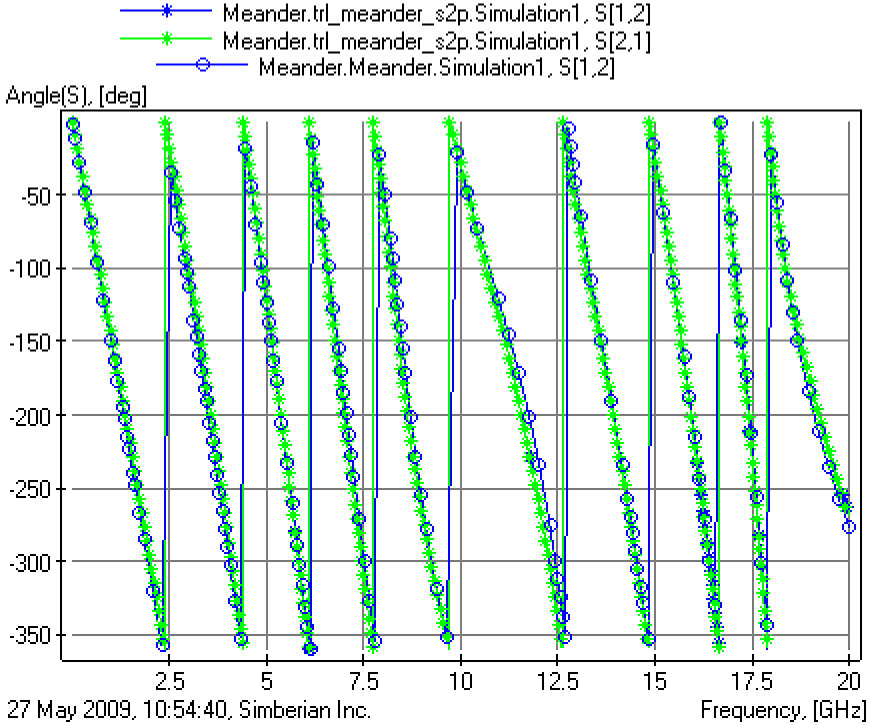
Good correspondence!

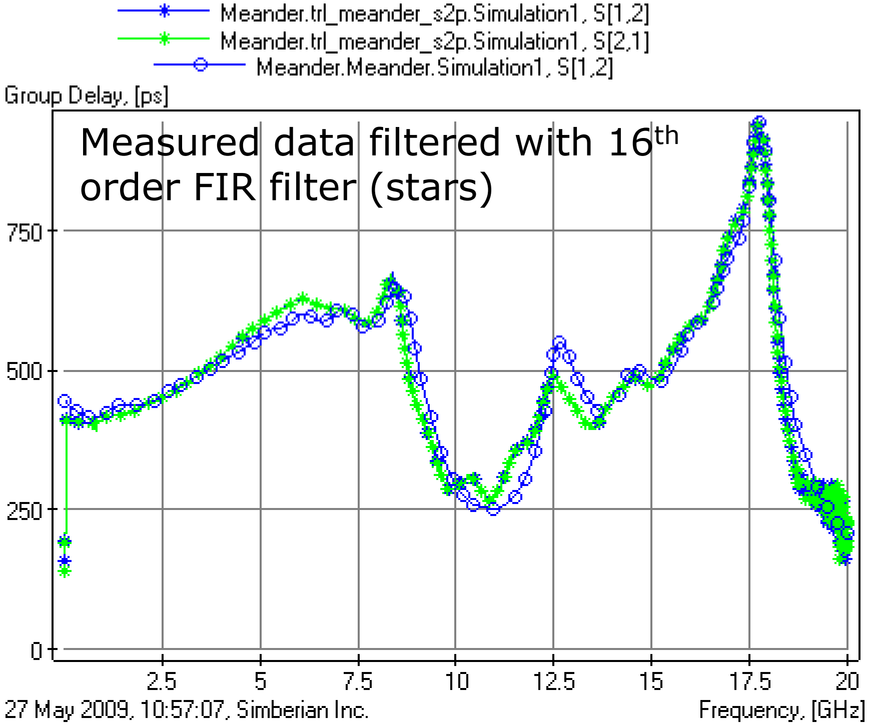
Comparison with measurement results de-embedded with TRL



- Transmission coefficient phase (angle) and group delay

Measured – stars, simulated - circles



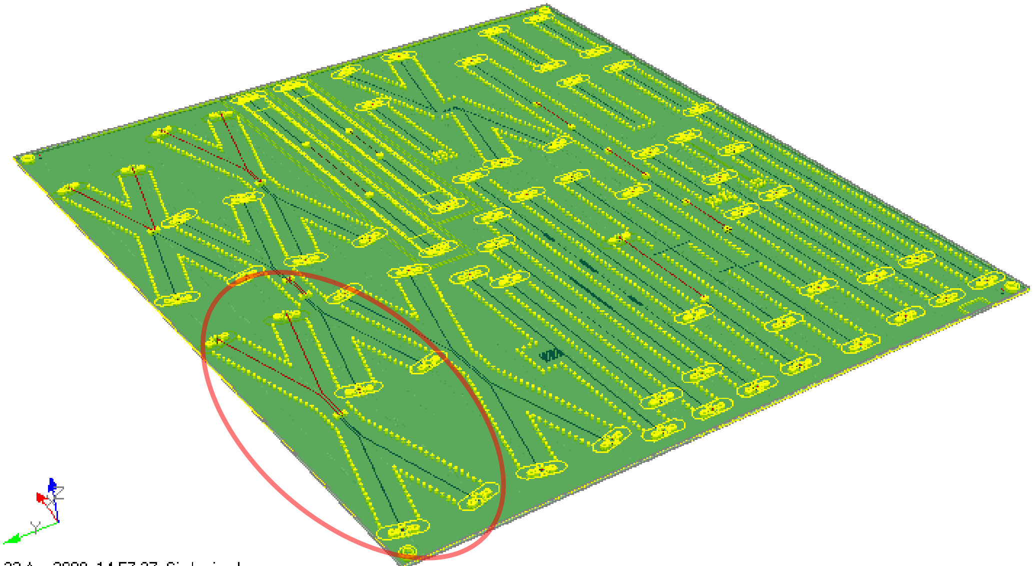


Measured data filtered with 16th order FIR filter (stars)

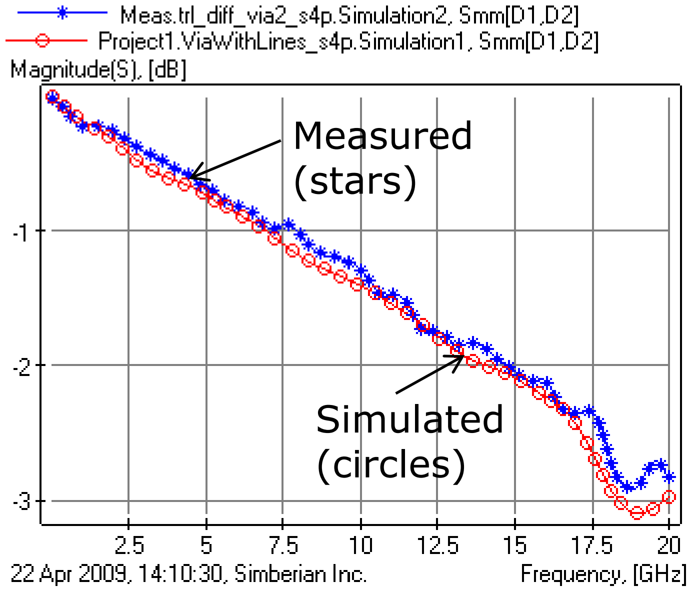
Good correspondence!

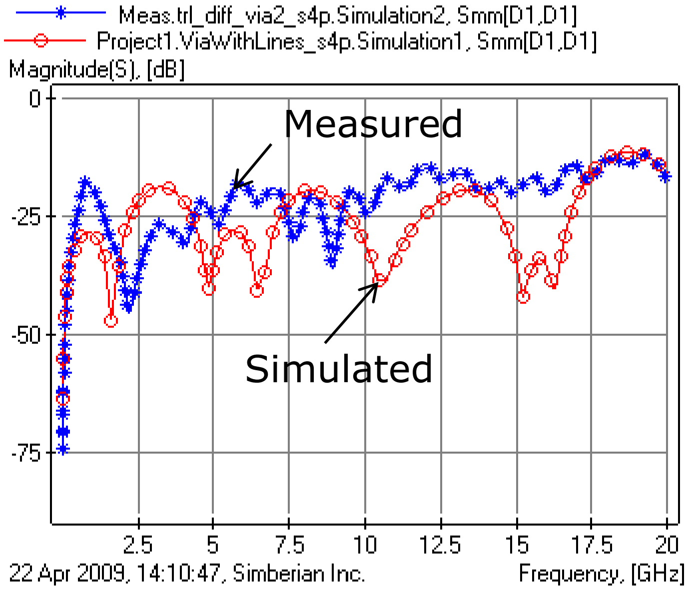
PLRD-1: S-parameters of differential vias

Good correspondence in magnitude of the transmission and reflection coefficients between Simbeor model and measurements



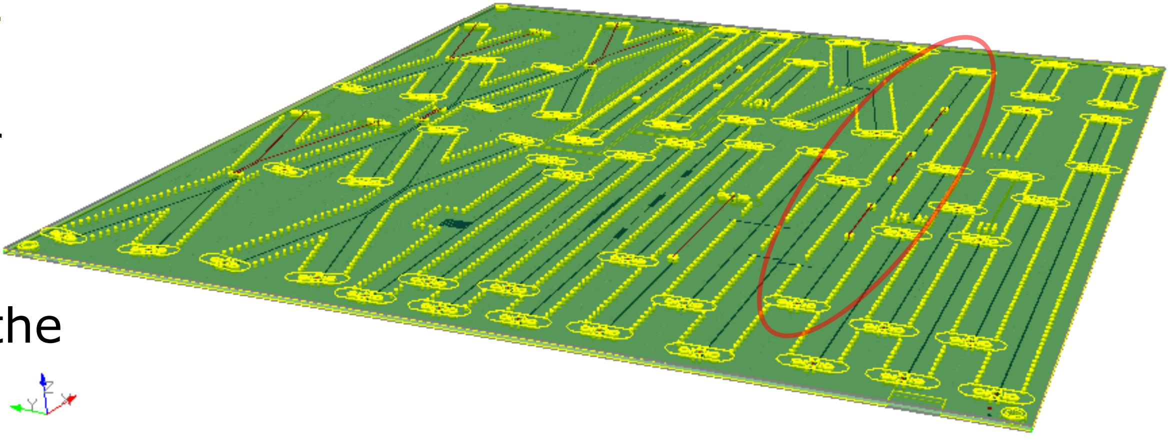
22 Apr 2009, 14:57:37, Simberian Inc.



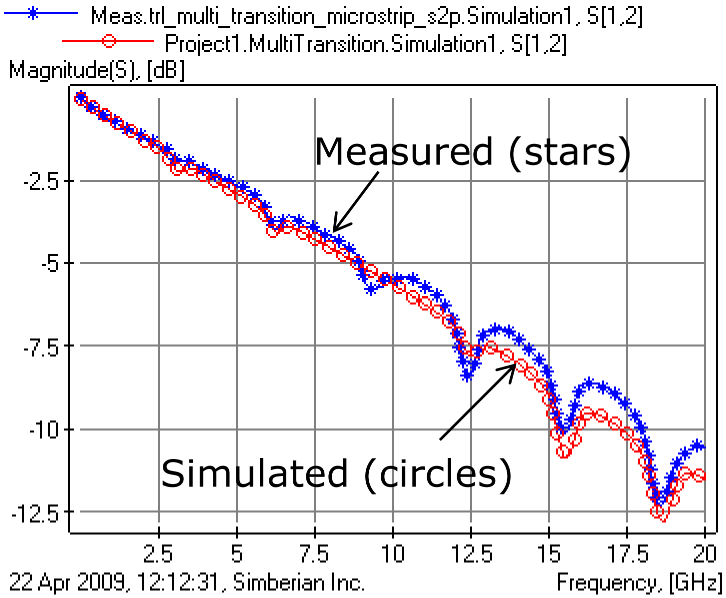


PLRD-1: S-parameters of micro-strip channel with 6 localized vias

Good correspondence in magnitude and phase of the transmission coefficient (though reflection was larger in the experiment)



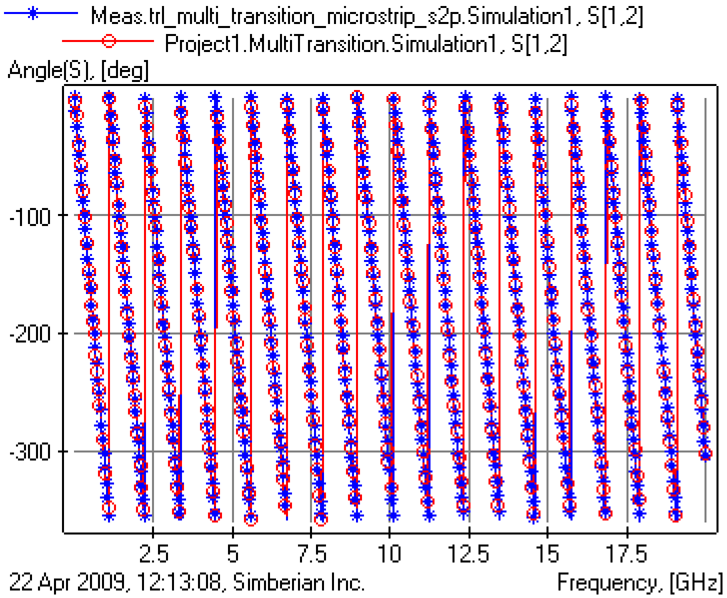
22 Apr 2009, 12:29:27, Simberian Inc.



Measured (stars)

Simulated (circles)

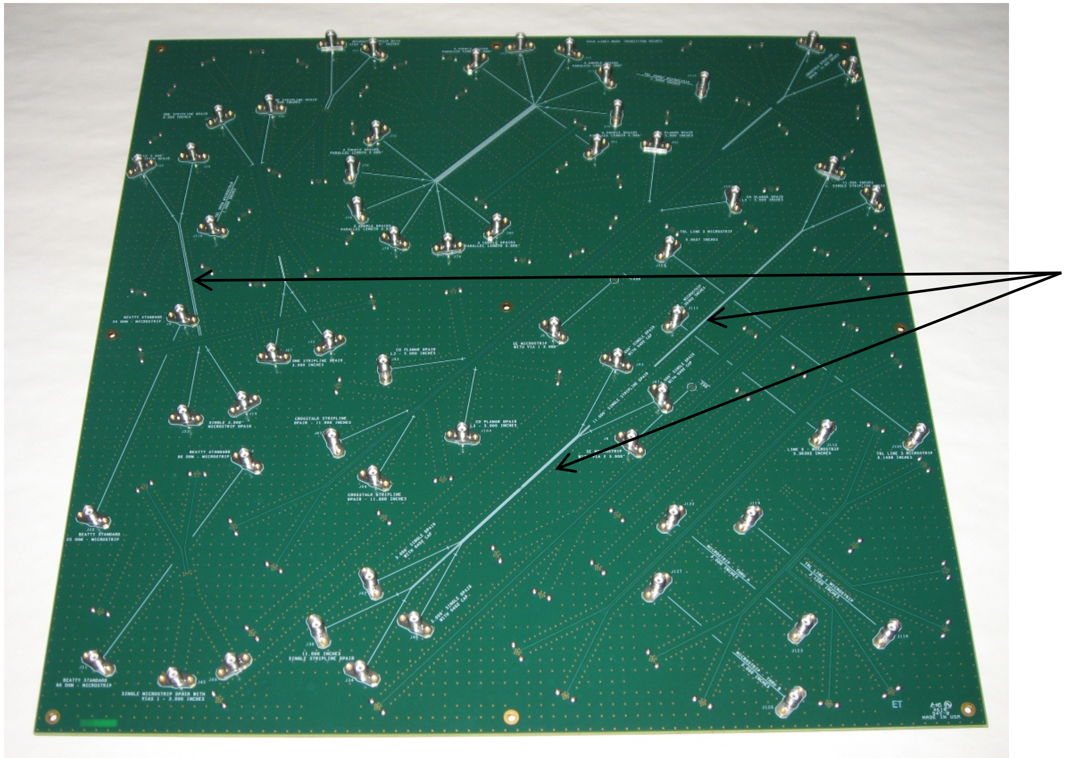
22 Apr 2009, 12:12:31, Simberian Inc.



22 Apr 2009, 12:13:08, Simberian Inc.

Channel Modeling Platform CMP-08

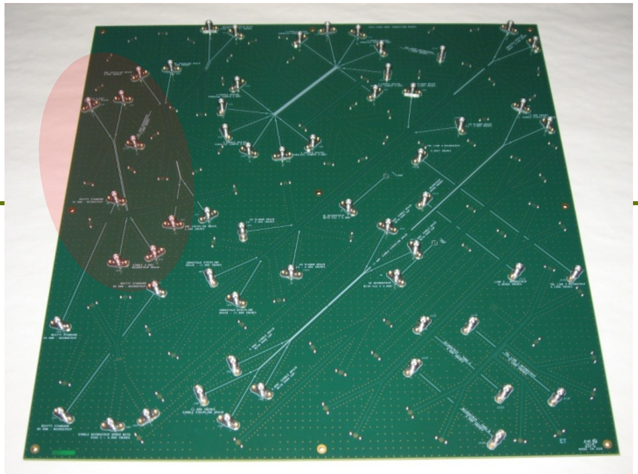
- **Validation board with coupled microstrip and strip structures** designed with Simbeor software by Wild River Technology
  - J. Bell, S. McMorrow, M. Miller, A. P. Neves, Y. Shlepnev, Unified Methodology of 3D-EM/Channel Simulation/Robust Jitter Decomposition, DesignCon2011 (also App Note #2011_02 at www.simberian.com)



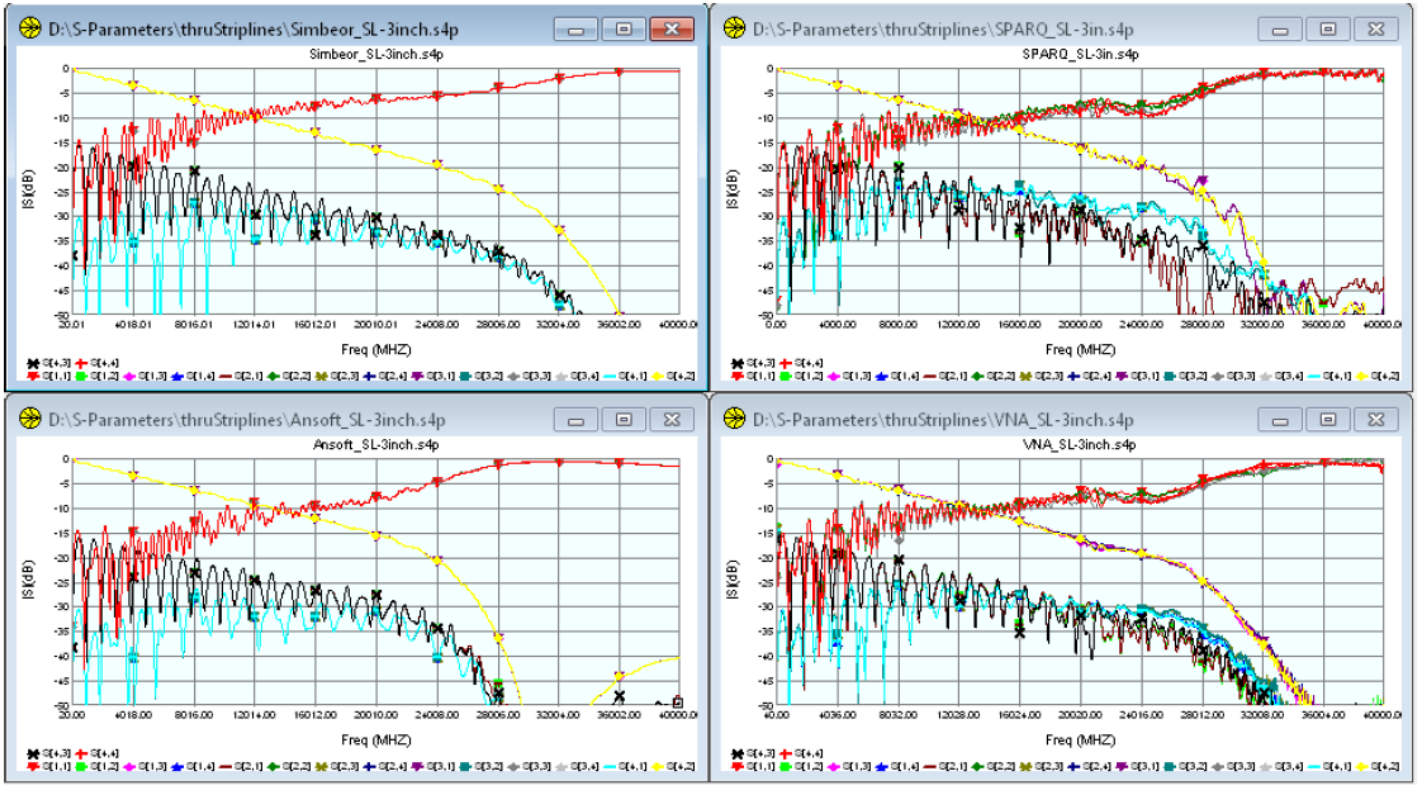
Analysis to measurement correlation investigation on 38 structures up to 30 GHz!

3″, 6″, and 11″ Differential THRU structures are used to benchmark simulations-measurements, and jitter tools

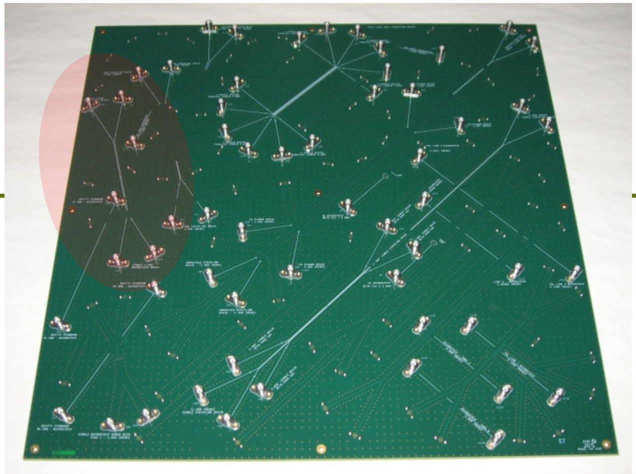
CMP-08 examples



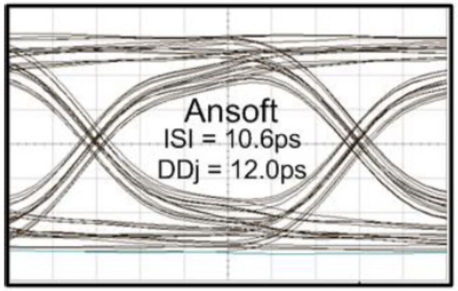
- Three-inch stripline differential traces
- Results of S-parameter comparisons from models and from VNA and TDNA for the 3 inch differential stripline

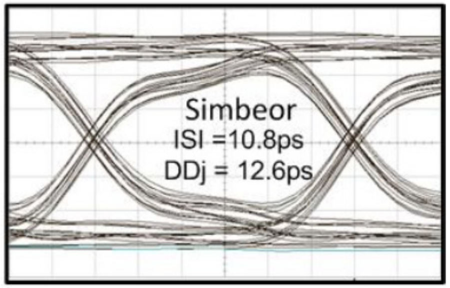


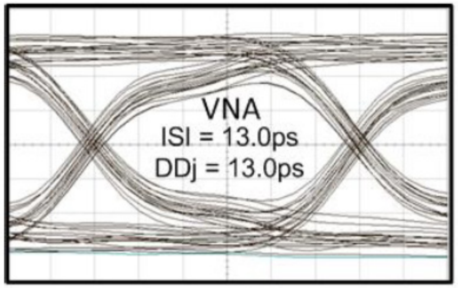
CMP-08 examples

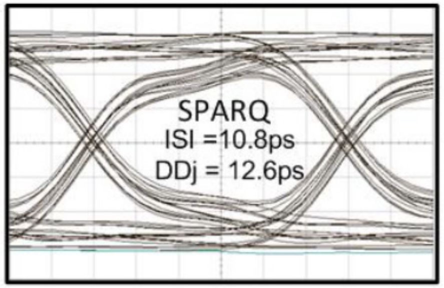


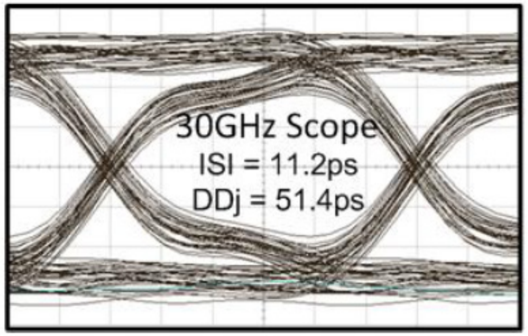
- Three-inch stripline differential traces
- Using recorded differential stimulus
- Two co-simulations with “modeled” S-parameters
- Two co-simulations with “measured” S-parameters
- One direct measurement
- Illustrating “good” agreement





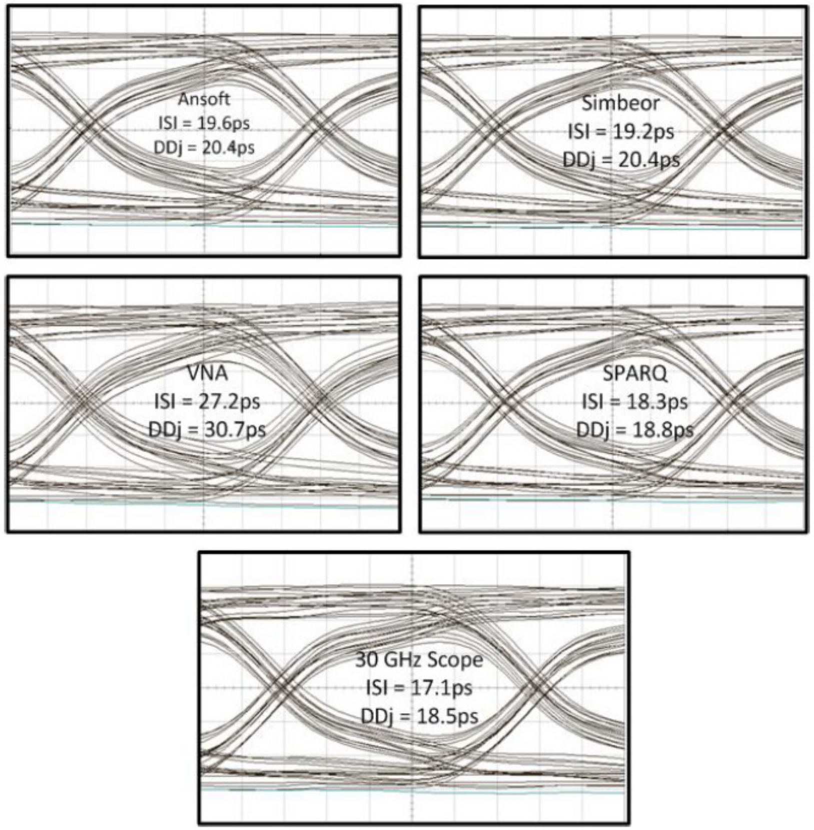




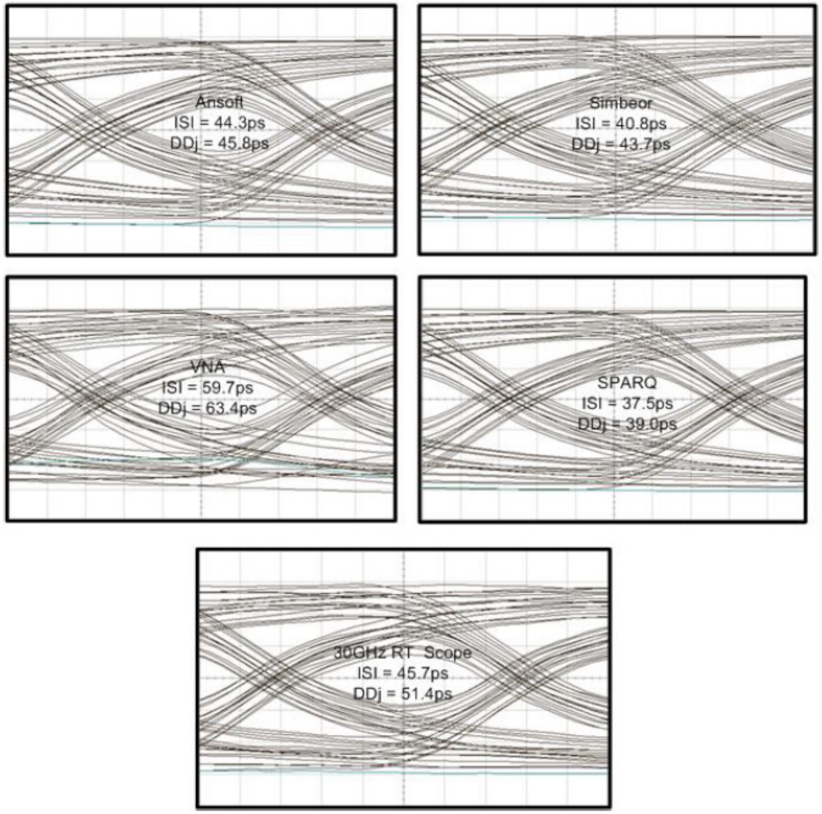


CMP-08 examples

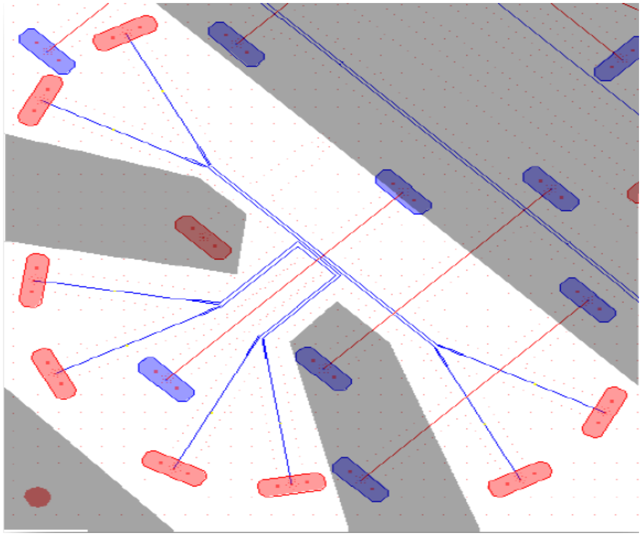
6 inch Differential Stripline Comparison of Noiseless Eye Diagrams, ISI and DDj jitter figures.



11 inch Differential Stripline Comparison of Noiseless Eye Diagrams, ISI and DDj jitter figures.

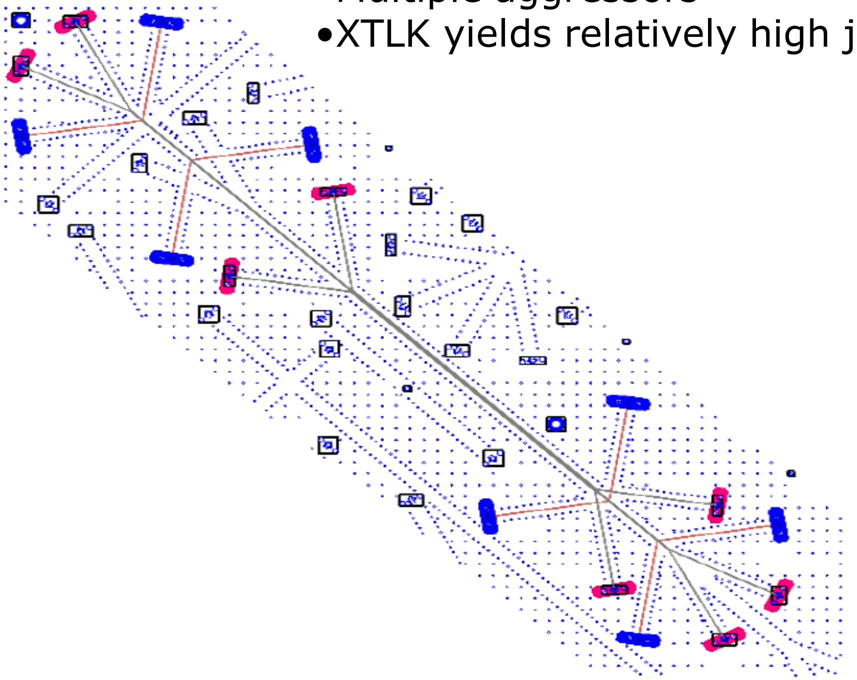


CMP-08 examples of structures for x-talk analysis to measurement correlation



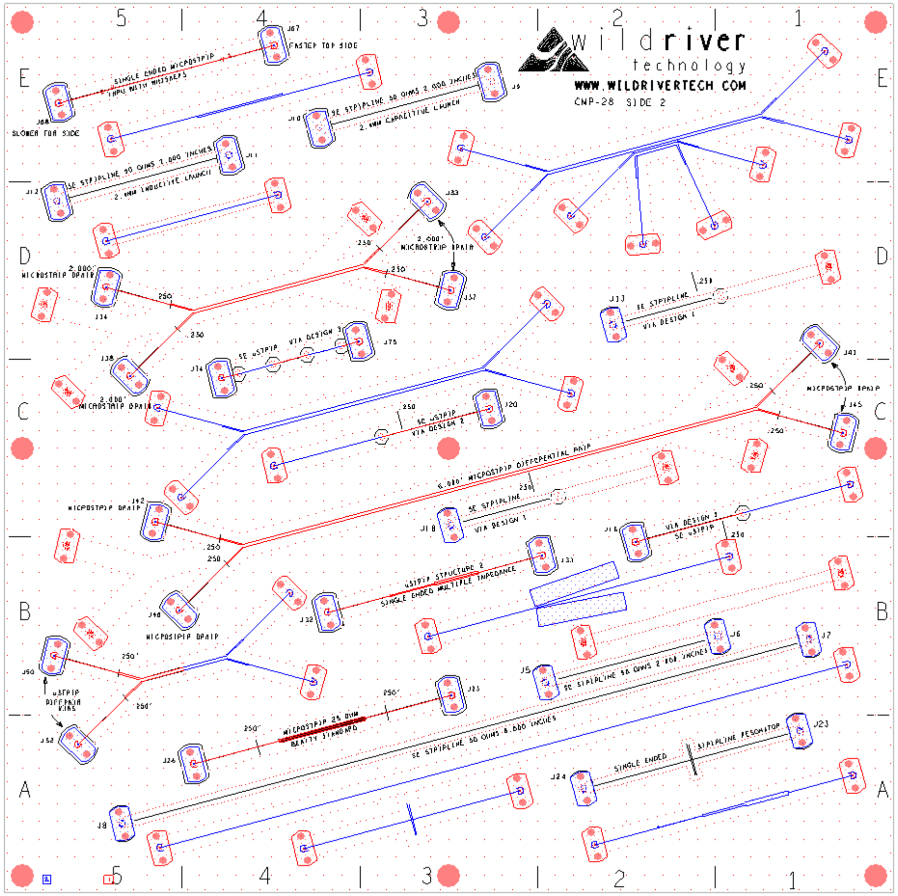
Neves Pathological
- Microstrip Traces
- Mode Conversion
- Impedance Variation
- XTLK yields relatively low jitter

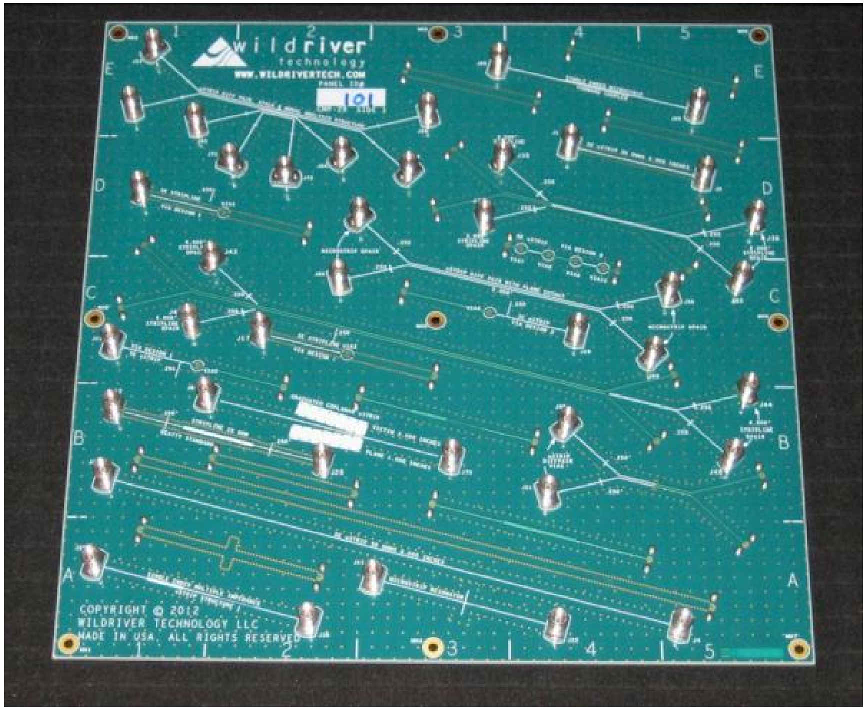
McMorrow Broadside Coupler
- Mimics Backplane
- Stripline Traces
- Multiple aggressors
- XTLK yields relatively high jitter



See more in J. Bell, S. McMorrow, M. Miller, A. P. Neves, Y. Shlepnev, Unified Methodology of 3D-EM/Channel Simulation/Robust Jitter Decomposition, DesignCon2011

CMP-28 benchmark board





Validation board with microstrip and strip structures designed with Simbeor software and available from Wild River Technology – http://wildrivertech.com/

Conclusion

- Decompositional electromagnetic analysis is the fastest and the most accurate technique for signal integrity analysis
  - Predictable interconnects must be designed as localized wave-guiding channels
  - Via-holes, breakouts and connector launches must be localized to make models independent of the board geometry
- Always start project with material parameters identification
  - Accuracy of transmission line models depends mostly on the dielectric and conductor surface roughness models and they may be not available
- Ensure S-parameter model quality (created and from vendors)
- Always validate your analysis with measurements
  - Use VNA or TDNA and compare both magnitudes and angles

Contact and resources

- Yuriy Shlepnev, Simberian Inc., Booth #626

  shlepnev@simberian.com

  Tel: 206-409-2368
- **Webinars on decompositional analysis, S-parameters quality and material identification http://www.simberian.com/Webinars.php**
- Simberian web site and contacts www.simberian.com
- Demo-videos http://www.simberian.com/ScreenCasts.php
- App notes http://www.simberian.com/AppNotes.php
- Technical papers http://kb.simberian.com/Publications.php
- Presentations http://kb.simberian.com/Presentations.php
- Download Simbeor® from www.simberian.com and try it on your problems for 15 days

References on multiport theory

- H. J. Carlin, A. B. Giordano, Network theory: An introduction to reciprocal and nonreciprocal circuits, Prentice Hall, 1964.
- N. Balabanian, T.A. Bickart, S. Seshu, Electrical network theory, John Wiley & Sons, 1968.
- J. Choma, Electrical networks: Theory and analysis, John Wiley & Sons, 1985.
- S. Ramo, J. R. Whinnery, T. Van Duzer, Fields and waves in communication electronics, John Wiley & Sons, 1994.
- D.E. Bockelman, W.R. Eisenstadt, Combined differential and common-mode scattering parameters: Theory and simulation, IEEE Trans. on MTT, vol. 43, 1995, N7, p. 1530-1539.
- A. Ferrero, M. Priola, Generalized mixed-mode S-parameters, IEEE Trans. on MTT, vol. 54, 2006, N1, p. 458-463.
- W.K. Gwarek, M. Celuch-Marcysiak, Wide-band S-parameter extraction from FD-TD simulations for propagating and evanescent modes in inhomogeneous guides, IEEE Trans. on MTT, vol. 51, 2003, N 8, p. 1920-1928.
- P. Triverio S. Grivet-Talocia, M.S. Nakhla, F.G. Canavero, R. Achar, Stability, Causality, and Passivity in Electrical Interconnect Models, IEEE Trans. on Advanced Packaging, vol. 30. 2007, N4, p. 795-808.

References on macro-modeling

- B. Gustavsen, A. Semlien, Rational approximation of frequency domain responses by vector fitting, IEEE Trans. on Power Delivery, v. 14, 1999, N3, p. 1052-1061.
- G. Antonini, SPICE equivalent circuits of frequency-domain responses, IEEE Trans. on EMC, vol. 45, 2003, N3, p. 502-512.
- S. Grivet-Talocia, A. Uboli, A comparative study of passivity enforcement schemes for linear lumped macromodels, IEEE Trans. on Adv. Packaging, v. 31, 2008, N4, p. 673-683.
- J. De Geest, S. Sercu, C. Clewell, J. Nadolny, Making S-parameters suitable for SPICE modeling, - DesignCon2004
- N. Stevens, T. Dhaene, Generation of rational model based SPICE circuits for transient simulations, - SPI2008

References on transmission line theory

- F. Olyslager, D. DeZutter, A. T. de Hoop, New reciprocal circuit model for lossy waveguide structures based on the orthogonality of the eigenmodes, IEEE Trans. on MTT, v. 42, No. 12, 1994, p. 2261-2269.
- C. R. Paul, Decoupling the multiconductor transmission line equations, IEEE Trans. on MTT, v. 44, No. 8, 1996, p. 1429-1440.
- D. F. Williams, L. A. Hayden, R. B. Marks, A complete multimode equivalent-circuit theory for electrical design, Journal of Research of the National Institute of Standards and Technology, v. 102, No. 4, 1997, p. 405-423.
- D. F. Williams, J. E. Rogers, C. L. Holloway, Multiconductor transmission line characterization: Representations, approximations, and accuracy, IEEE Trans. on MTT, v. 47, No. 4, 1999, p. 403-409.
- G. G. Gentili, M. Salazar-Palma, The definition and computation of modal characteristic impedance in quasi-TEM coupled transmission lines. IEEE Trans. on MTT, v. 43, No. 2, p. 338-343.

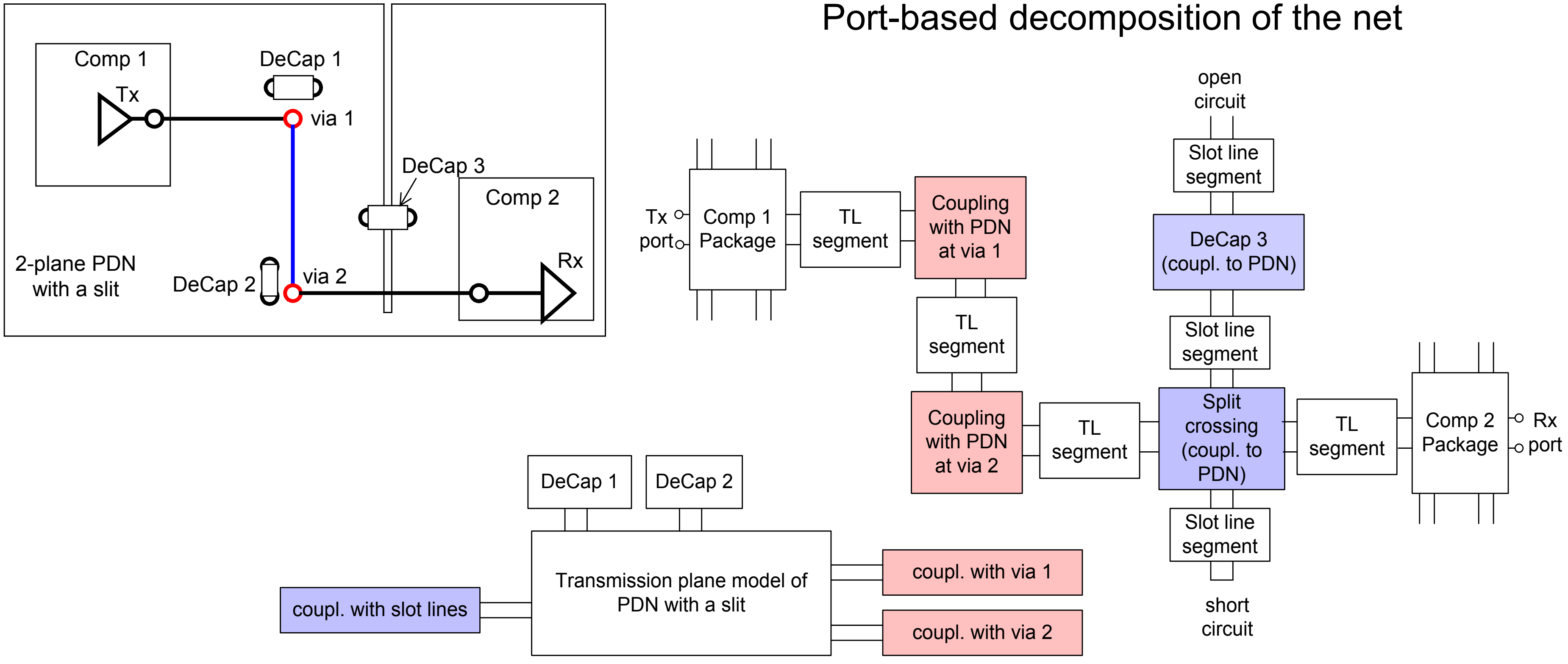
References on materials identification
(available at http://www.simberian.com/AppNotes.php)

- 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 (App Note #2010_01)
- Sensitivity of PCB Material Identification with GMS-Parameters to Variations in Test Fixtures, Simberian App Note #2010_03
- Material Identification With GMS-Parameters of Coupled Lines, Simberian App Note #2010_04
- J. Bell, S. McMorrow, M. Miller, A. P. Neves, Y. Shlepnev, Unified Methodology of 3D-EM/Channel Simulation/Robust Jitter Decomposition, DesignCon2011, (App Note #2011_02)
- D. Dunham, J. Lee, S. McMorrow, Y. Shlepnev, 2.4mm Design/Optimization with 50 GHz Material Characterization, DesignCon2011 (App Note #2011_03)
- Y. Shlepnev, S. McMorrow, Nickel characterization for interconnect analysis. - Proc. of the 2011 IEEE International Symposium on Electromagnetic Compatibility, Long Beach, CA, USA, August, 2011, p. 524-529.
- 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
- Y. Shlepnev, C. Nwachukwu, Practical methodology for analyzing the effect of conductor roughness on signal losses and dispersion in interconnects, DesignCon2013, Feb. 1st, 2012, Santa Clara, CA.

Backup slides

Example of decompositional analysis of non-localizable link

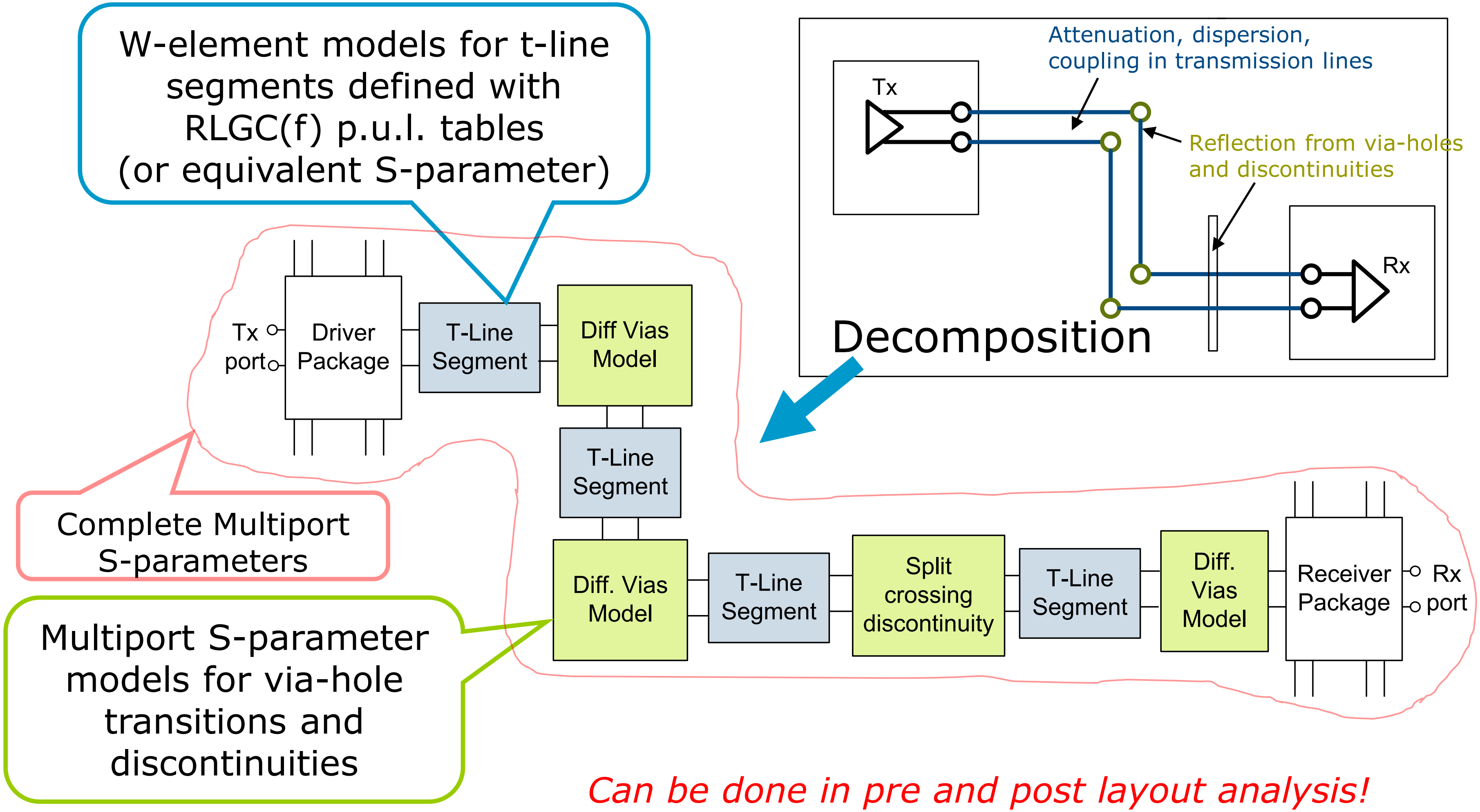
A net on 4-layer board with two parallel planes (S-G-P-S) to illustrate the port-based decomposition process



Port-based decomposition of the net

Not possible in pre-layout analysis!

Example of decompositional analysis of localizable link



W-element models for t-line segments defined with RLGC(f) p.u.l. tables (or equivalent S-parameter)

Attenuation, dispersion, coupling in transmission lines

Reflection from via-holes and discontinuities

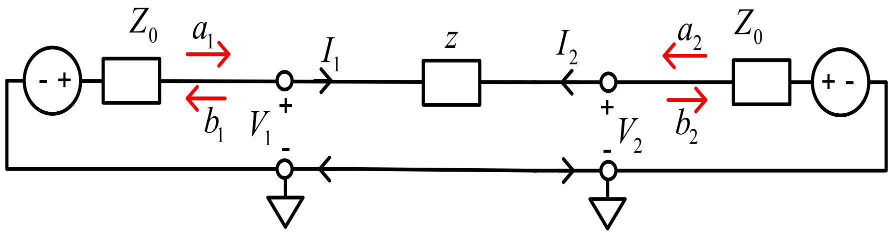
Decomposition

Complete Multiport S-parameters

Multiport S-parameter models for via-hole transitions and discontinuities

Can be done in pre and post layout analysis!

Example: Series impedance, two-port



z is a complex impedance

$$S \in C^{2\times 2}$$

We just use known Y and transform it to S

$$Y = \frac{1}{z}\begin{bmatrix} 1 & -1 \\ -1 & 1 \end{bmatrix} \Rightarrow Y_N = Z_0 \cdot Y = \frac{Z_0}{z}\begin{bmatrix} 1 & -1 \\ -1 & 1 \end{bmatrix}$$

$$S = \left(U - Y_N\right)\cdot\left(U + Y_N\right)^{-1} = \frac{1}{z + 2\cdot Z_0}\begin{bmatrix} z & 2\cdot Z_0 \\ 2\cdot Z_0 & z \end{bmatrix}$$

Short-circuit:

$$z = 0 \Rightarrow S_{1,1} = \begin{bmatrix} 0 & 1 \\ 1 & 0 \end{bmatrix}$$

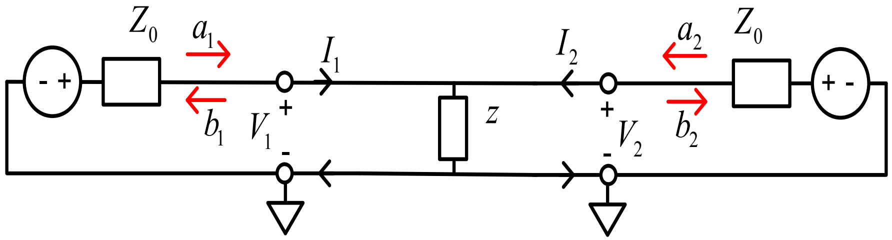
Open-circuit:

$$z = \infty \Rightarrow S_{1,1} = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}$$

Passivity: $\left|eigenvals\left[S\right]\right| = \left\{\left|\frac{z - 2\cdot Z_0}{z + 2\cdot Z_0}\right|, 1\right\} \le 1 \Rightarrow \mathrm{Re}\left(z\right) \ge 0$ For real normalization impedance

S-matrix is always symmetric (reciprocal system) and non-singular for any z

Example: Parallel impedance, two-port



z is a complex impedance

$$S \in C^{2\times 2}$$

We just use known Z and transform it to S

$$Z = z\begin{bmatrix} 1 & 1 \\ 1 & 1 \end{bmatrix} \Rightarrow Z_N = \frac{1}{Z_0}\cdot Z = \frac{z}{Z_0}\begin{bmatrix} 1 & 1 \\ 1 & 1 \end{bmatrix}$$

$$S = (Z_N - U)\cdot(U + Z_N)^{-1} = \frac{1}{y + 2\cdot Y_0}\begin{bmatrix} -y & 2\cdot Y_0 \\ 2\cdot Y_0 & -y \end{bmatrix}$$

$$y = \frac{1}{z}, \quad Y_0 = \frac{1}{Z_0}$$

Short-circuit:

$$z = 0 \Rightarrow S_{1,1} = \begin{bmatrix} -1 & 0 \\ 0 & -1 \end{bmatrix}$$

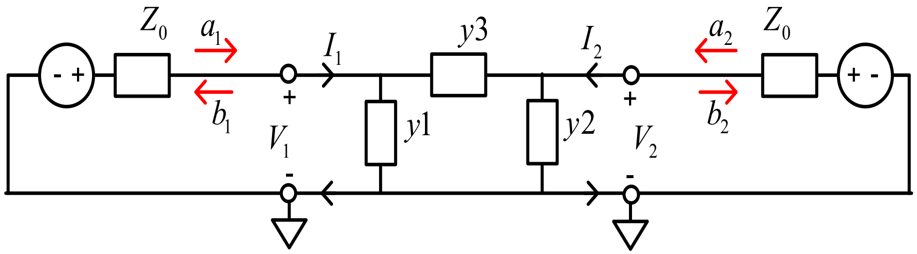
Open-circuit:

$$z = \infty \Rightarrow S_{1,1} = \begin{bmatrix} 0 & 1 \\ 1 & 0 \end{bmatrix}$$

Passivity: $\left|eigenvals[S]\right| = \left\{\left|\frac{y - 2\cdot Y_0}{y + 2\cdot Y_0}\right|, 1\right\} \le 1 \Rightarrow \mathrm{Re}(z) \ge 0$ For real normalization impedance

S-matrix is always symmetric (reciprocal system) and non-singular for any z

Example: PI-circuit, two-port



y1, y2, y3 are complex admittances

$$S \in C^{2\times 2}$$

We just use known Z and transform it to S

$$Y = \begin{bmatrix} y1+y3 & -y3 \\ -y3 & y2+y3 \end{bmatrix} \quad Y_N = Z_0 \cdot \begin{bmatrix} y1+y3 & -y3 \\ -y3 & y2+y3 \end{bmatrix}$$

$$S = (U - Y_N)\cdot(U + Y_N)^{-1} = \frac{1}{A}\begin{bmatrix} Y_0^2 - (y1-y2)\cdot Y_0 - B & 2\cdot y3\cdot Y_0 \\ 2\cdot y3\cdot Y_0 & Y_0^2 + (y1-y2)\cdot Y_0 \end{bmatrix} \qquad Y_0 = \frac{1}{Z_0}$$

$$A = Y_0^2 + (y1 + y2 + 2\cdot y3)\cdot Y_0 + B \quad B = y1\cdot y2 + y2\cdot y3 + y1\cdot y3$$

S is always symmetric (reciprocal system) and non-singular

Passivity: $\left|eigenvals[S]\right| \le 1$ Always satisfied for nets composed of R,L,C

Waves in multiconductor t-lines

$$Z_{0n}(\omega) = \sqrt{z_{n,n}(\omega) / y_{n,n}(\omega)}$$

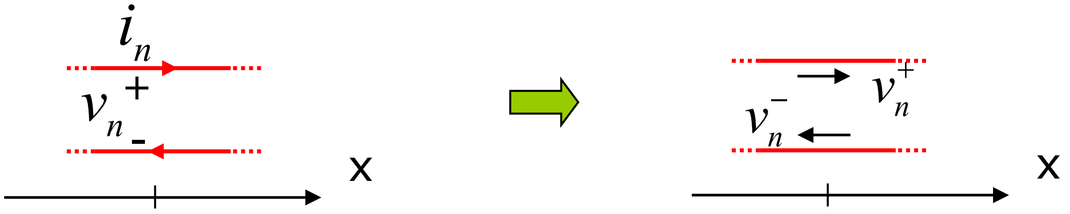
$$\Gamma_n(\omega) = \sqrt{z_{n,n}(\omega) \cdot y_{n,n}(\omega)}$$

Modal complex characteristic impedance and propagation constant

---

Current and voltage of mode number n (n=1,…,N)

Voltage waves for mode number n (n=1,…,N)



$$v_n(x) = v_n^+ \cdot \exp(-\Gamma_n \cdot x) + v_n^- \cdot \exp(\Gamma_n \cdot x)$$

$$i_n(x) = \frac{1}{Z_{0n}} \left[v_n^+ \cdot \exp(-\Gamma_n \cdot x) - v_n^- \cdot \exp(\Gamma_n \cdot x) \right]$$

$$P_n^+ = \frac{|v_n^+|^2}{Z_{0n}}$$

$$P_n^- = \frac{|v_n^-|^2}{Z_{0n}}$$

Passivity:

$$\mathrm{Re}(Z_{0n}(\omega)) \geq 0$$

$$\alpha_n = \mathrm{Re}(\Gamma_n(\omega)) \geq 0$$

$$\bar{V} = M_V \cdot \bar{v}$$

$$\bar{I} = M_I \cdot \bar{i}$$

Voltage and current in multiconductor line can be expressed as a superposition of modal currents and voltages

One and two-conductor lines

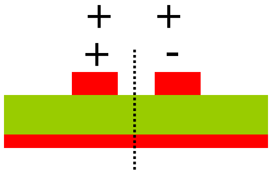
One-conductor case

$$M_V = M_I = 1$$

$$Z_0(\omega) = \sqrt{Z(\omega)/Y(\omega)}$$

$$\Gamma(\omega) = \sqrt{Z(\omega) \cdot Y(\omega)}$$

Symmetric two-conductor case – even and odd mode normalization



$$M_V = M_I = M_{eo} = \frac{1}{\sqrt{2}} \begin{bmatrix} 1 & 1 \\ -1 & 1 \end{bmatrix}$$

$$y_{eo} = M_{eo} \cdot Y(\omega) \cdot M_{eo}$$

$$z_{eo} = M_{eo} \cdot Z(\omega) \cdot M_{eo}$$

$$Z_{odd}(\omega) = \sqrt{z_{eo1,1}/y_{eo1,1}} \qquad \Gamma_{odd}(\omega) = \sqrt{z_{eo2,2} \cdot y_{eo2,2}}$$

$$Z_{even}(\omega) = \sqrt{z_{eo2,2}/y_{eo2,2}} \qquad \Gamma_{even}(\omega) = \sqrt{z_{eo2,2} \cdot y_{eo2,2}}$$

Common and differential mode normalization

$$M_V = M_{Vmm} = \begin{bmatrix} 1 & 0.5 \\ -1 & 0.5 \end{bmatrix}, \quad M_I = M_{Imm} = \begin{bmatrix} 0.5 & 1 \\ -0.5 & 1 \end{bmatrix}$$

$$y_{mm} = M_{Imm}^{-1} \cdot Y(\omega) \cdot M_{Vmm}$$

$$z_{mm} = M_{Vmm}^{-1} \cdot Z(\omega) \cdot M_{Imm}$$

$$Z_{differential}(\omega) = \sqrt{z_{mm1,1}/y_{mm1,1}} \qquad Z_{differential} = 2 \cdot Z_{odd} \qquad \Gamma_{differential} = \Gamma_{odd}$$

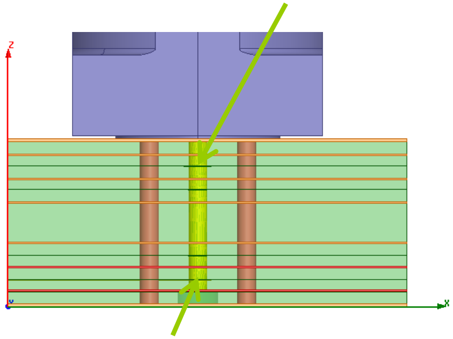
$$Z_{common}(\omega) = \sqrt{z_{mm2,2}/y_{mm2,2}} \qquad Z_{common} = 0.5 \cdot Z_{even} \qquad \Gamma_{common} = \Gamma_{even}$$

Material identification board from Molex/Teraspeed Consulting Group

- Board made with Nelco 4000-EP have been investigated and featured in:
  D. Dunham, J. Lee, S. McMorrow, Y. Shlepnev, 2.4mm Design/Optimization with 50 GHz Material Characterization, DesignCon2011
- Similar board was made with Panasonic Megtron 6 dielectric, VLP copper

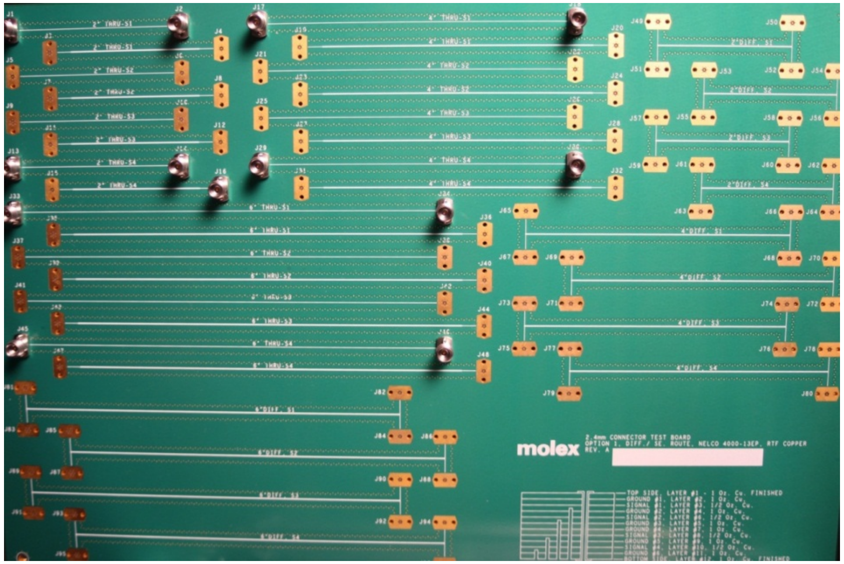
6 test fixtures with 2, 4 and 6 inch strip line segments in Layer 1 (S1) and Layer 4 (S4)

Signal Layer 1



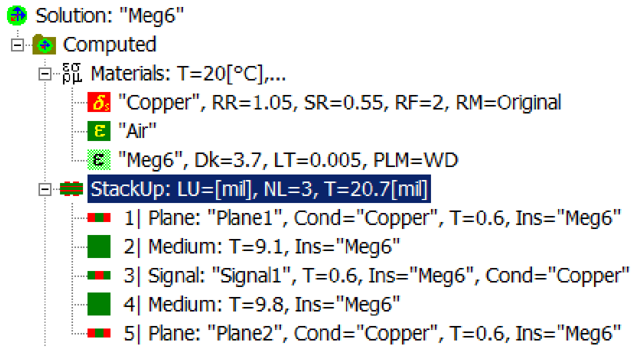
Signal Layer 4

Scott McMorrow from Teraspeed Consulting Group designed launches for 2.4mm Molex connectors, board made by Molex and measurements done by David Dunham, Molex



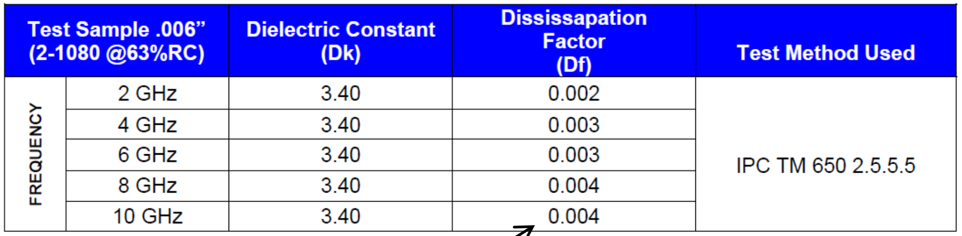
Test board and cross-section

- Strip line segments in Panasonic's Megtron 6
- 2 inch, 4 inch and 6 inch segments on board to identify parameters



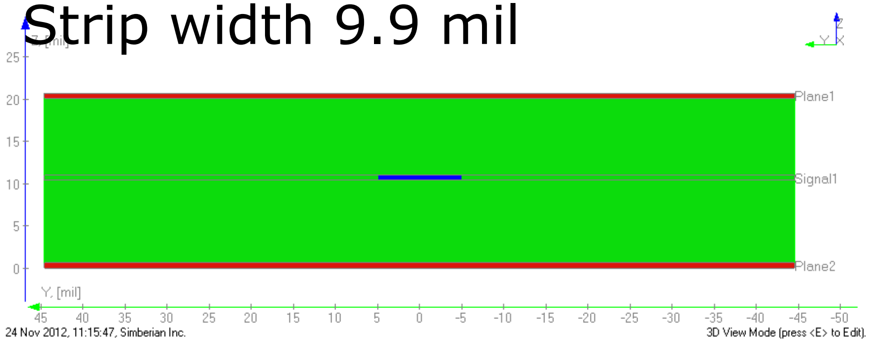
Solution: "Meg6"
Computed
Materials: T=20[°C],...
"Copper", RR=1.05, SR=0.55, RF=2, RM=Original
"Air"
"Meg6", Dk=3.7, LT=0.005, PLM=WD
StackUp: LU=[mil], NL=3, T=20.7[mil]
1| Plane: "Plane1", Cond="Copper", T=0.6, Ins="Meg6"
2| Medium: T=9.1, Ins="Meg6"
3| Signal: "Signal1", T=0.6, Ins="Meg6", Cond="Copper"
4| Medium: T=9.8, Ins="Meg6"
5| Plane: "Plane2", Cond="Copper", T=0.6, Ins="Meg6"

From datasheet (Dk is 3.6-3.7 for 2116 glass style, LT=0.002):



	Test Sample .006" (2-1080 @63%RC)	Dielectric Constant (Dk)	Dississapation Factor (Df)	Test Method Used
FREQUENCY	2 GHz	3.40	0.002	IPC TM 650 2.5.5.5
	4 GHz	3.40	0.003	
	6 GHz	3.40	0.003	
	8 GHz	3.40	0.004	
	10 GHz	3.40	0.004	

Strip width 9.9 mil



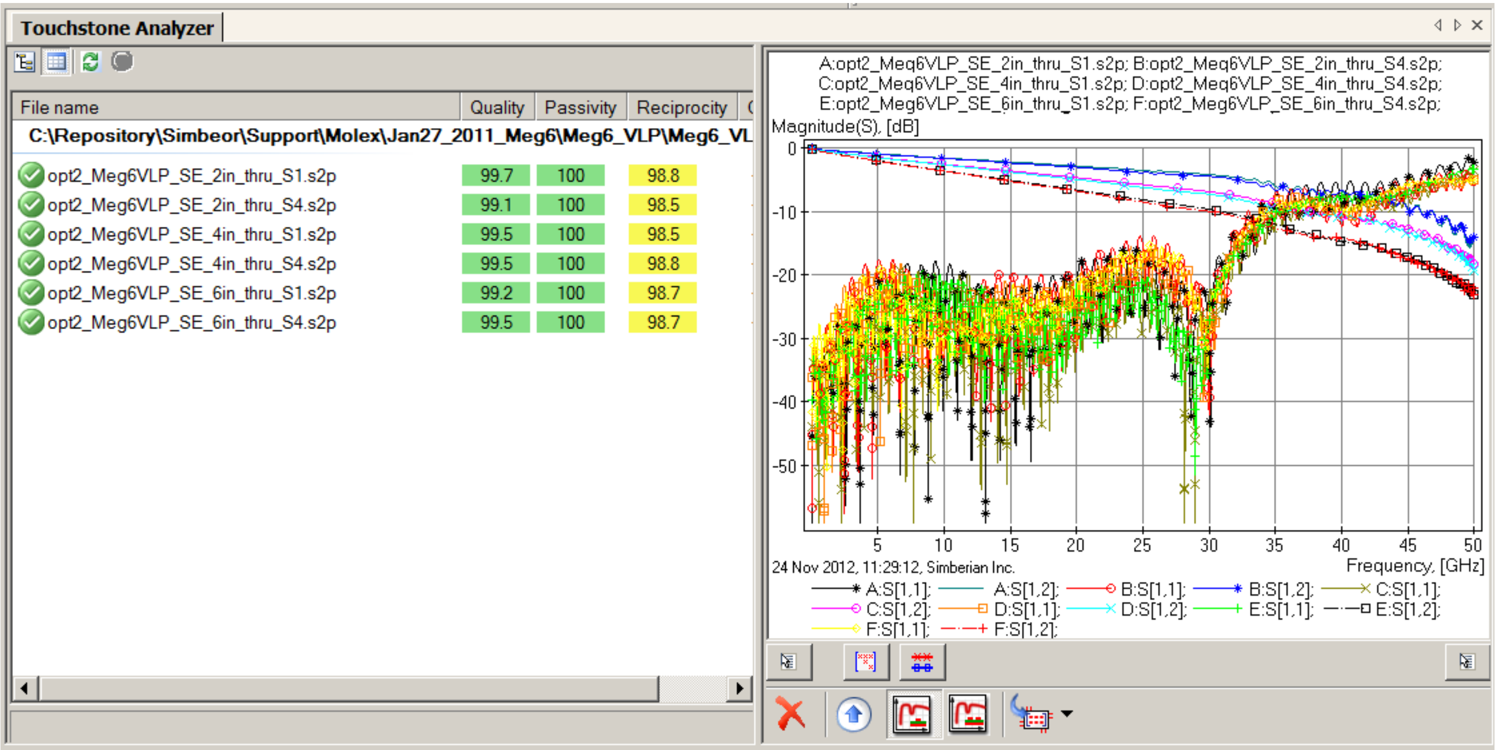
24 Nov 2012, 11:15:47, Simberian Inc.

3D View Mode (press <E> to Edit).

Constant Dk and growing LT – NON CAUSAL!

Measurements pre-qualification

- Good quality of frequency-domain models for all six test structures



Touchstone Analyzer

C:\Repository\Simbeor\Support\Molex\Jan27_2011_Meg6\Meg6_VLP\Meg6_VL

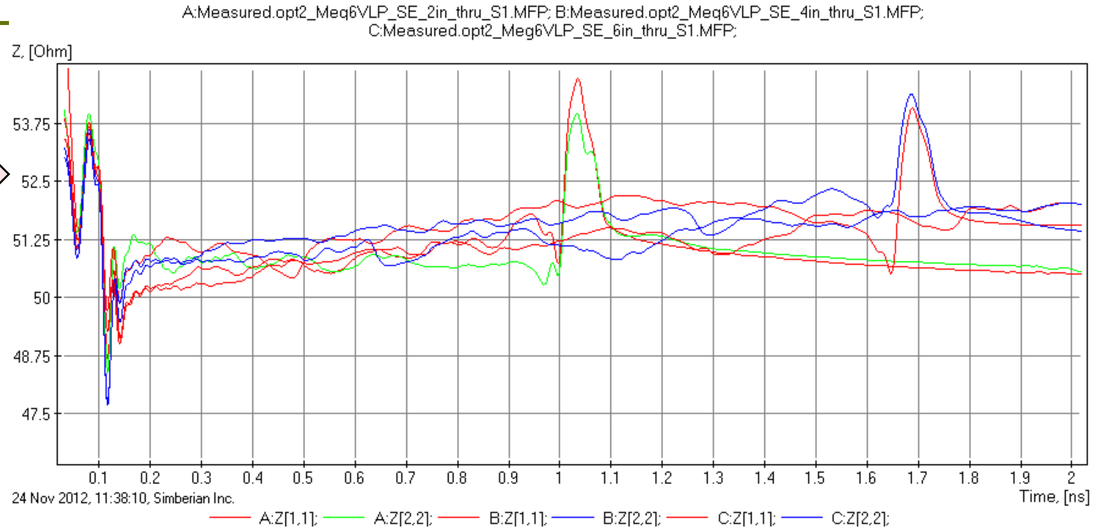
File name	Quality	Passivity	Reciprocity
opt2_Meg6VLP_SE_2in_thru_S1.s2p	99.7	100	98.8
opt2_Meg6VLP_SE_2in_thru_S4.s2p	99.1	100	98.5
opt2_Meg6VLP_SE_4in_thru_S1.s2p	99.5	100	98.5
opt2_Meg6VLP_SE_4in_thru_S4.s2p	99.5	100	98.8
opt2_Meg6VLP_SE_6in_thru_S1.s2p	99.2	100	98.7
opt2_Meg6VLP_SE_6in_thru_S4.s2p	99.5	100	98.7

A:opt2_Meg6VLP_SE_2in_thru_S1.s2p; B:opt2_Meg6VLP_SE_2in_thru_S4.s2p;
C:opt2_Meg6VLP_SE_4in_thru_S1.s2p; D:opt2_Meg6VLP_SE_4in_thru_S4.s2p;
E:opt2_Meg6VLP_SE_6in_thru_S1.s2p; F:opt2_Meg6VLP_SE_6in_thru_S4.s2p;

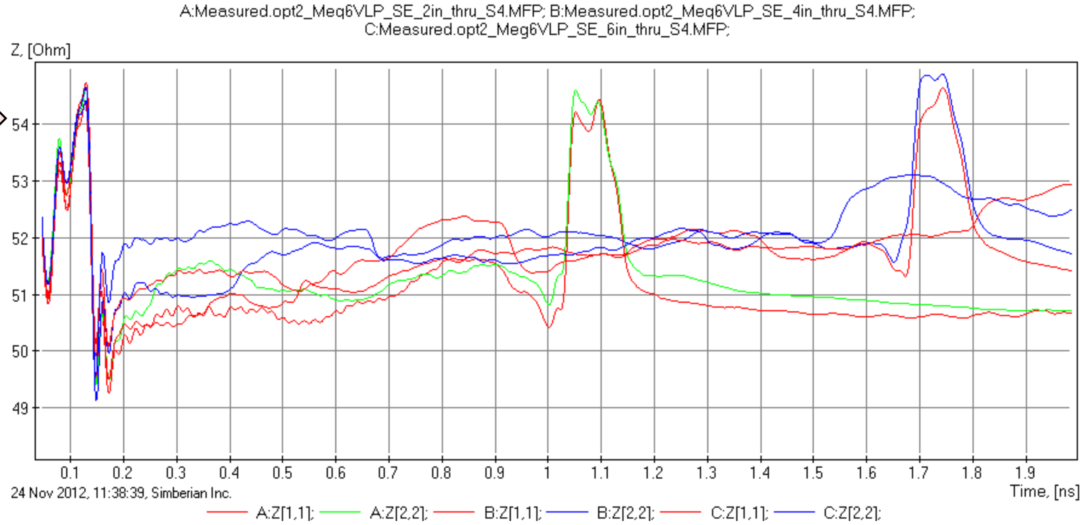
24 Nov 2012, 11:29:12, Simberian Inc.

TDR pre-qualification

Strip lines in layer S1 – about 2 Ohm difference at launches



Strip lines in layer S4 – about 2 Ohm difference along the lines

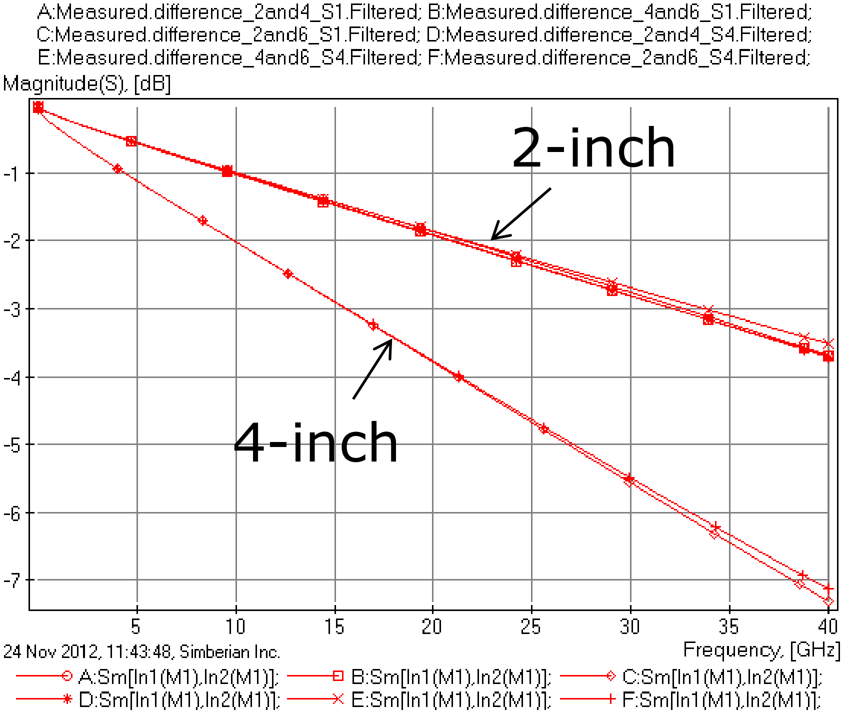


Data acceptable for material identification up to 40 GHz

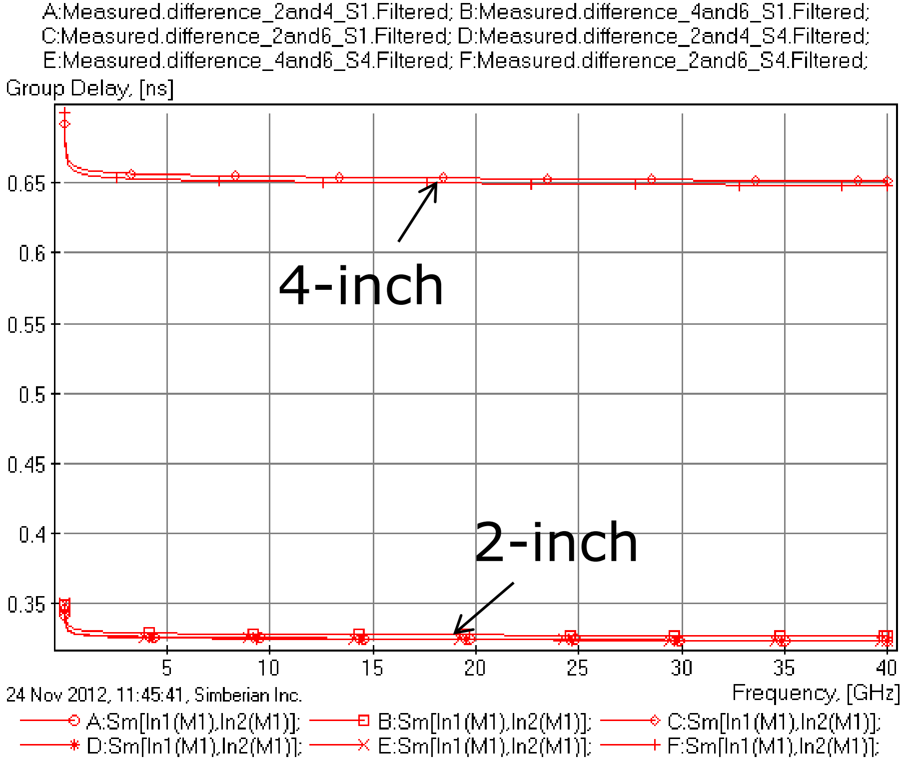
GMS-parameters to fit

- 2 inch and 4 inch differences for all possible combinations

GMS Insertion Loss



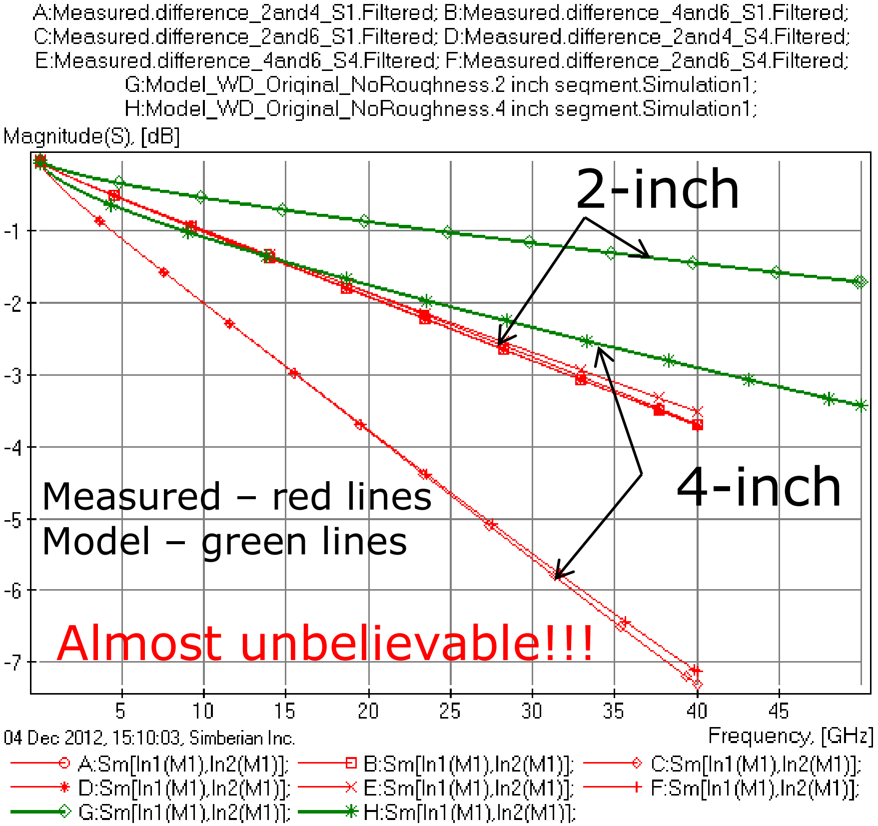
GMS Group Delay



Use material parameters from specs

- Dk=3.7, LT=0.002, @ 2 GHz, WD model

GMS Insertion Loss



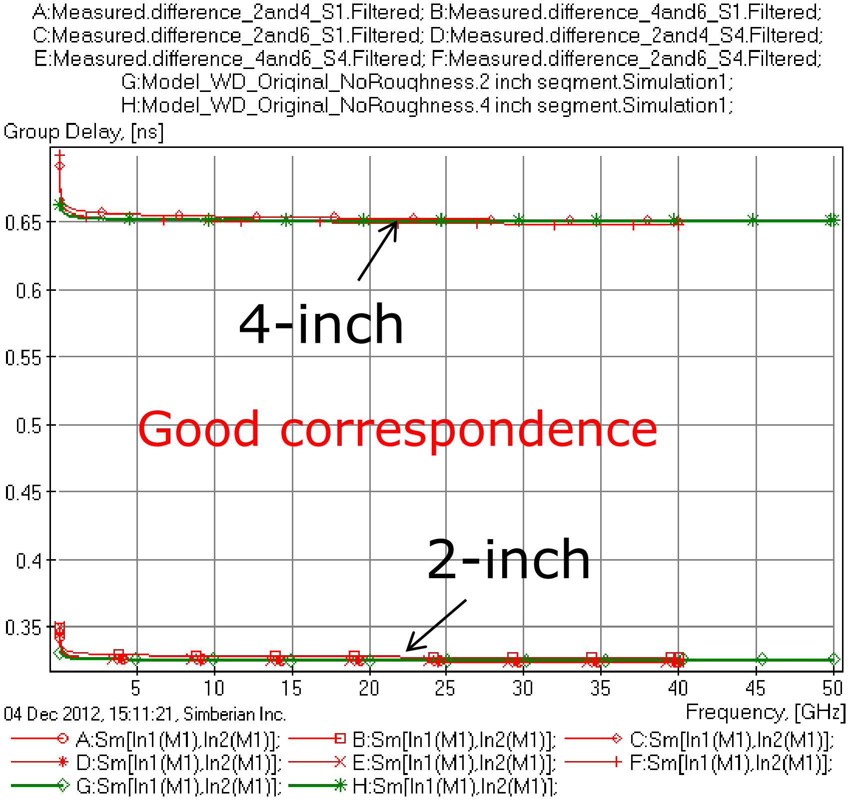
2-inch

4-inch

Measured – red lines
Model – green lines

Almost unbelievable!!!

GMS Group Delay



4-inch

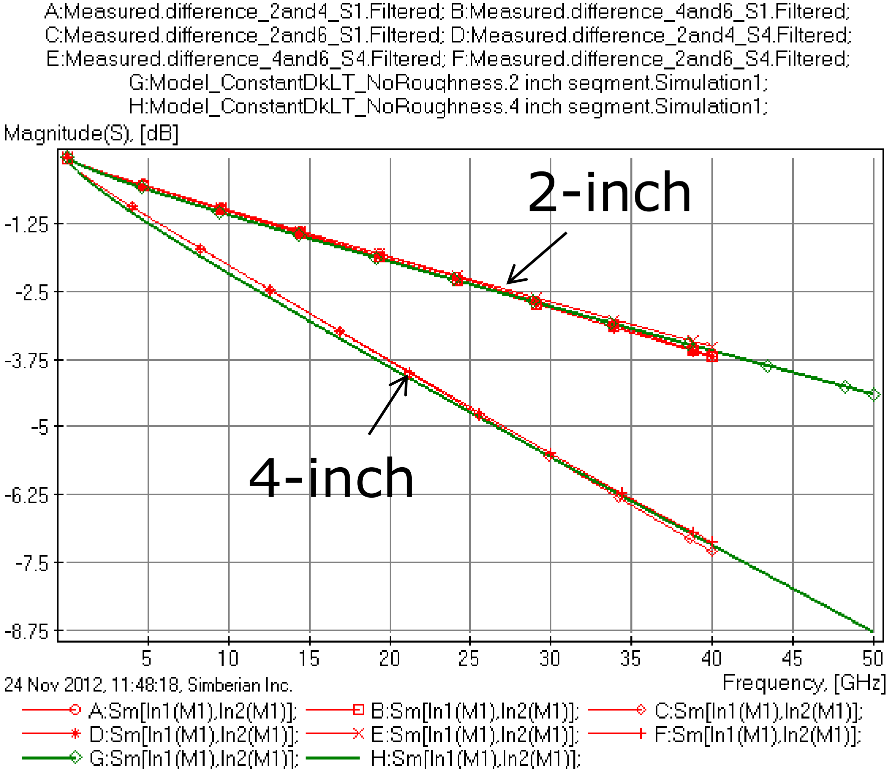
Good correspondence

2-inch

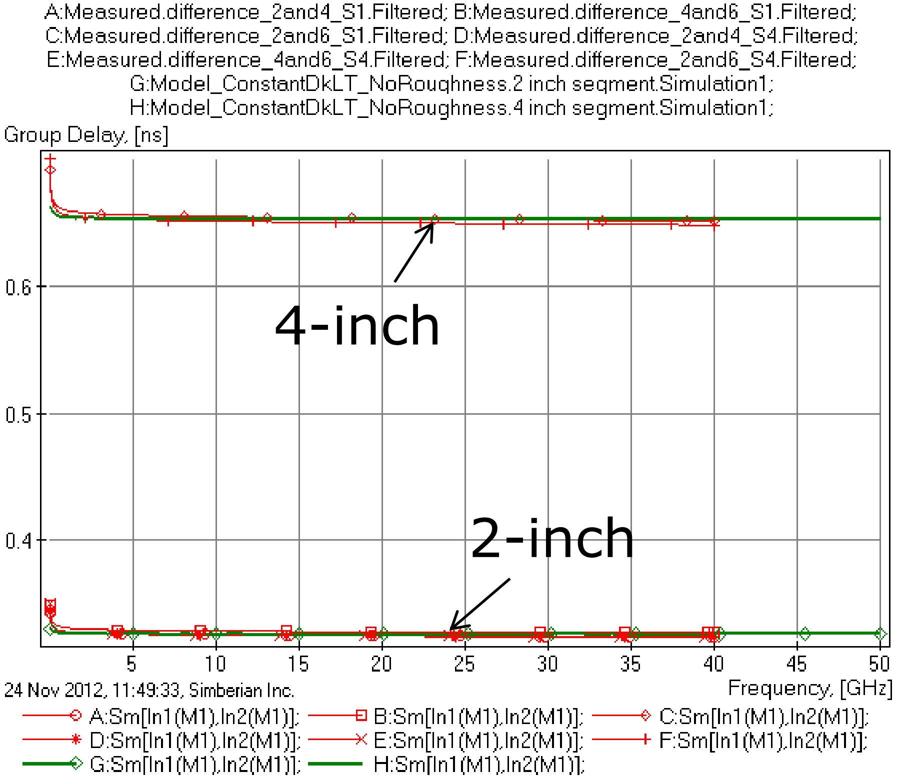
Flat non-causal dielectric model, no roughness

- Dk=3.7, LT=0.0082 – acceptable fit (green line)

GMS Insertion Loss

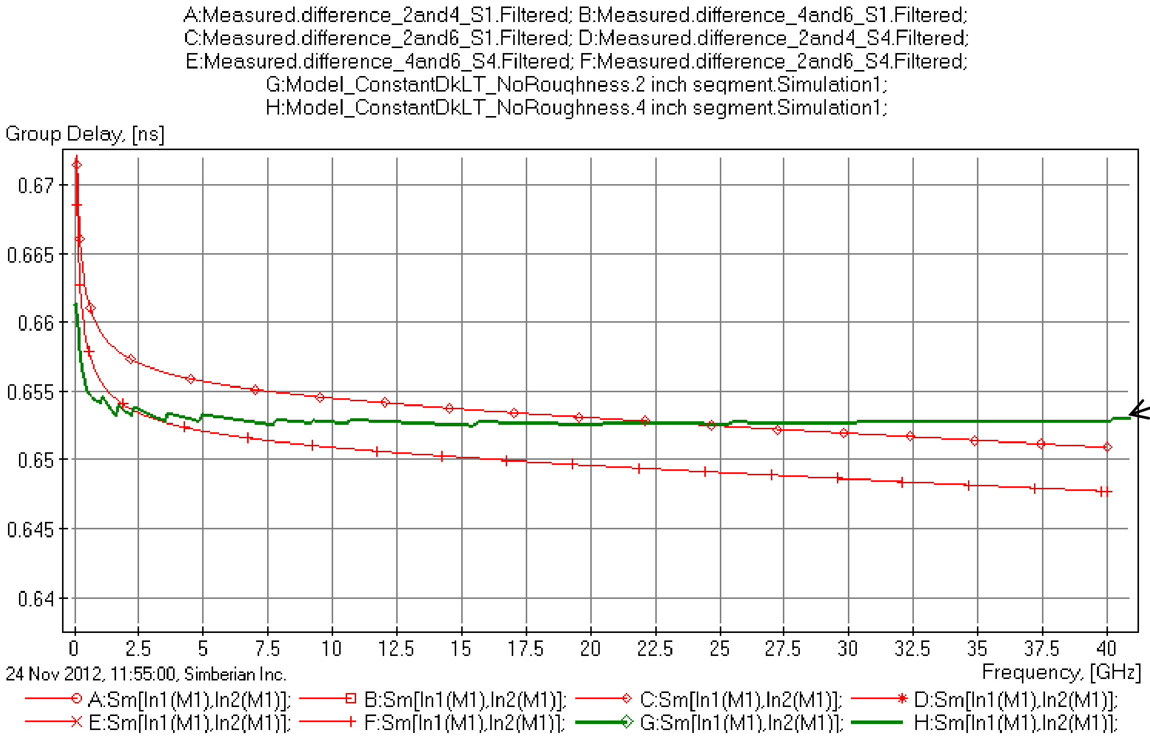


GMS Group Delay



Flat model defects

- Difficult to build rational macro-model and possible defects in impulse response (due to non-causality)
- No dispersion due to dielectric properties

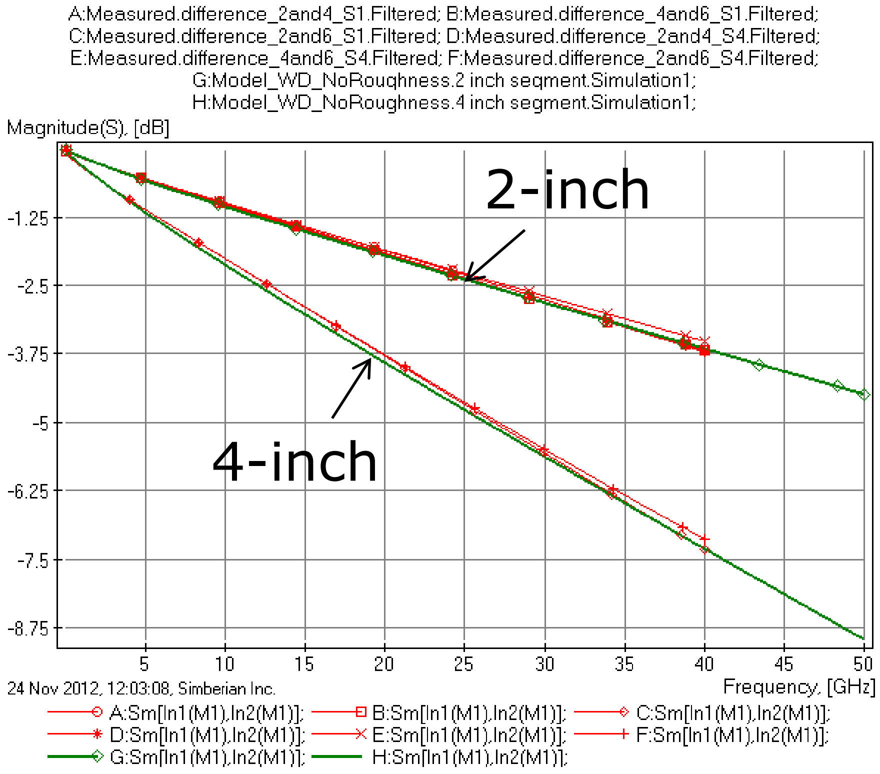


No dispersion (green line) as in measured GD (red lines) Though, the difference is small for practical concerns

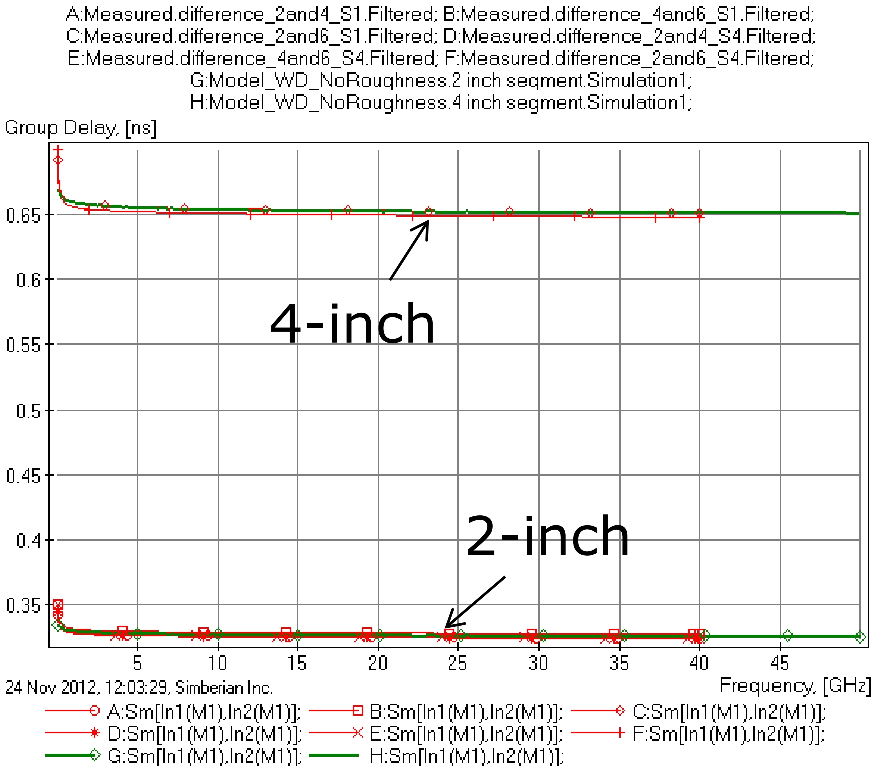
Causal Wideband Debye model, no roughness

- WD mode, DK=3.7, LT=0.0082 at 50 GHz, WD Low frequency is set to 10 GHz – good fit (green lines)

GMS Insertion Loss

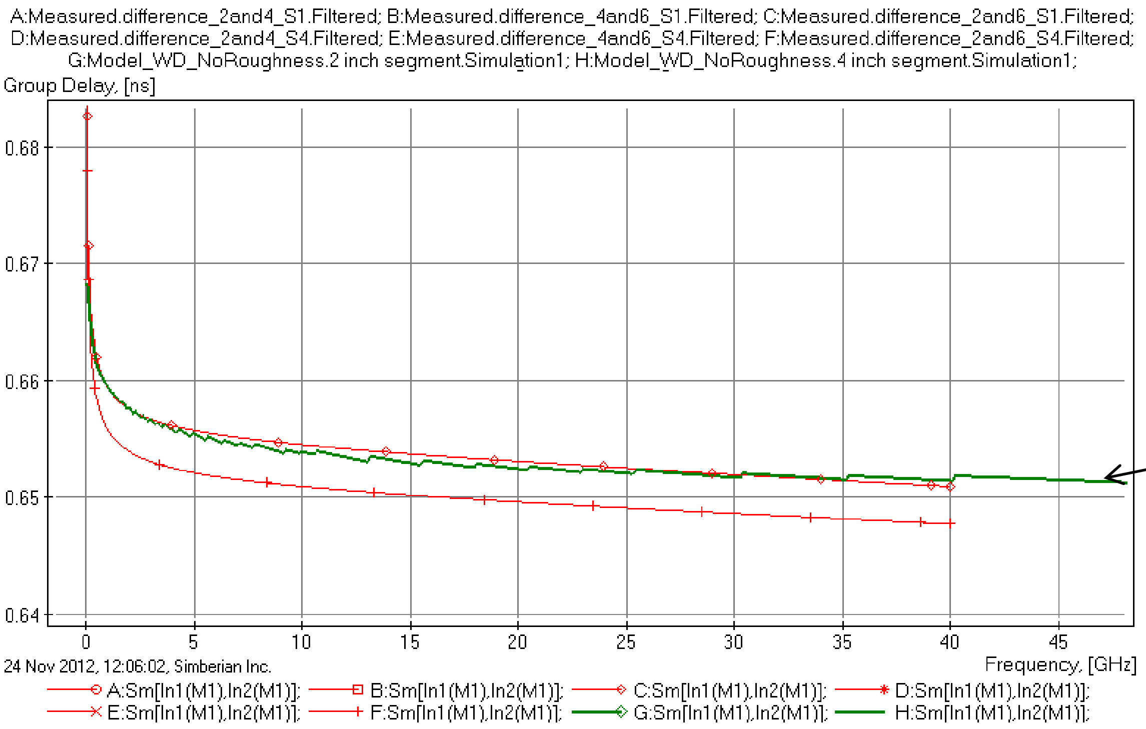


GMS Group Delay



Dispersive model

- No defects in rational approximation and impulse response
- Group delay decreases as in the measured data

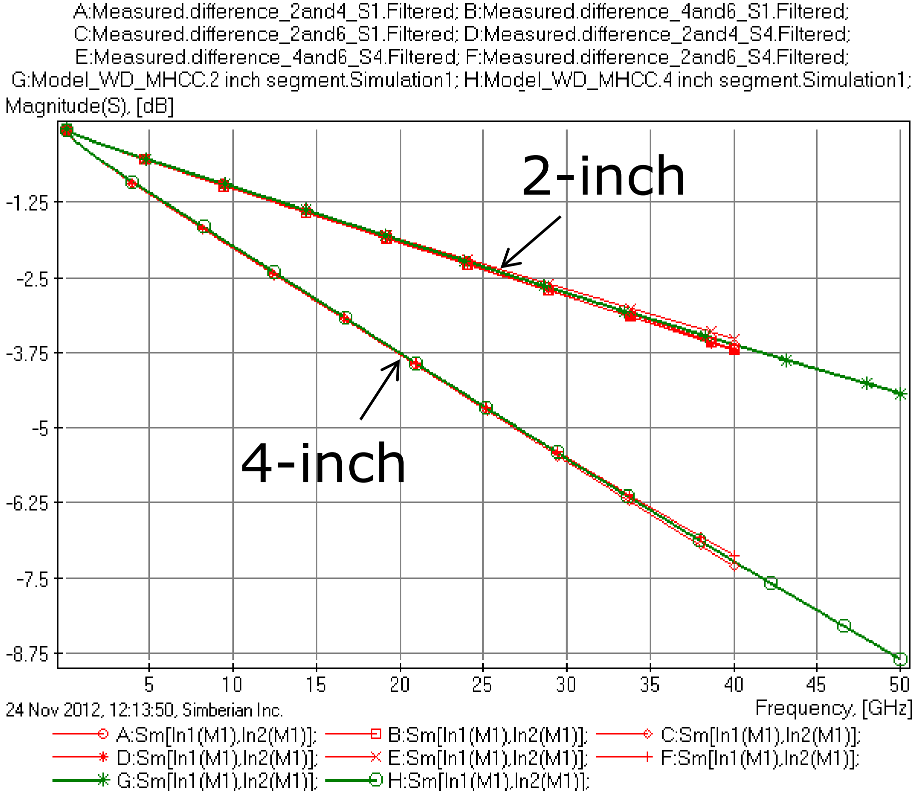


Dispersion (green line) as in measured GD (red lines)

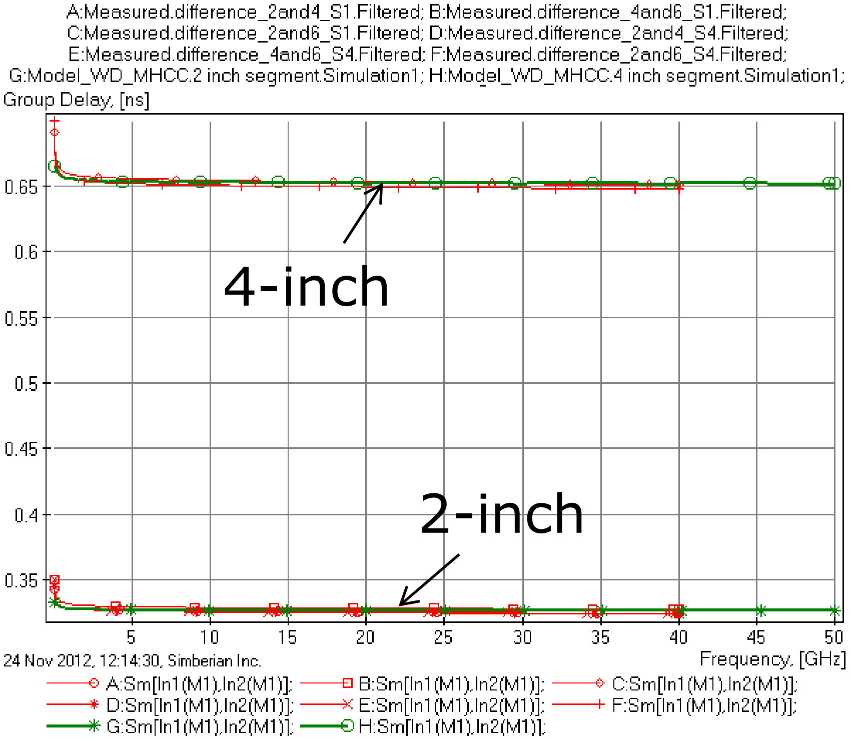
Dielectric and roughness model (MHCC)

- Dielectric: regular Wideband Debye, DK=3.7, LT=0.002 @ 2 GHz (as in specs)
- Roughness: Modified Hammerstadt Correction Coefficient, SR=0.3 um, RF=5 – excellent fit (green lines)

GMS Insertion Loss

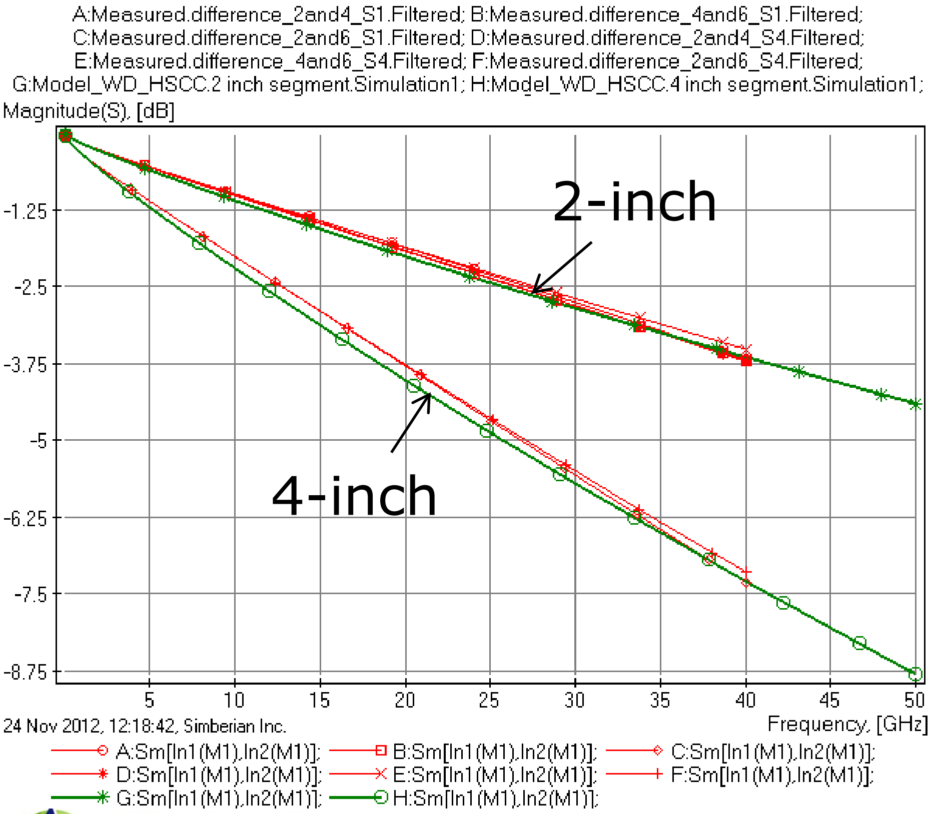


GMS Group Delay

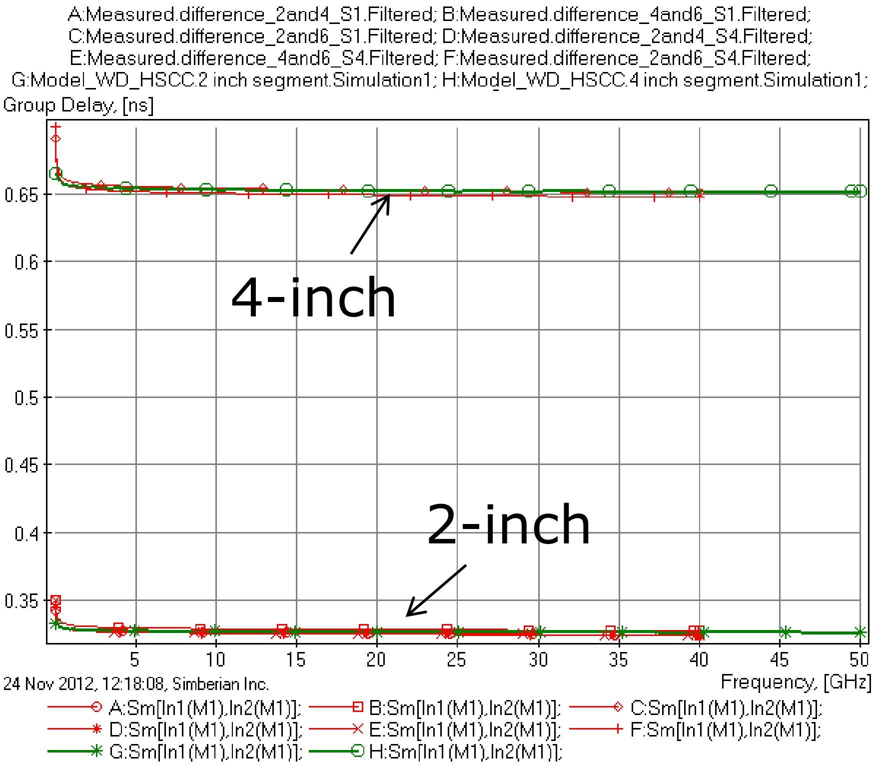


Dielectric and roughness model (HSCC)

- Dielectric: regular Wideband Debye, DK=3.7, LT=0.002 @ 2 GHz (as in specs)
- Roughness: Huray Snowball Correction Coefficient, BS=10 um, BD=0.7 um, Nb=330, good fit (green lines), multi-ball model needed for better fit



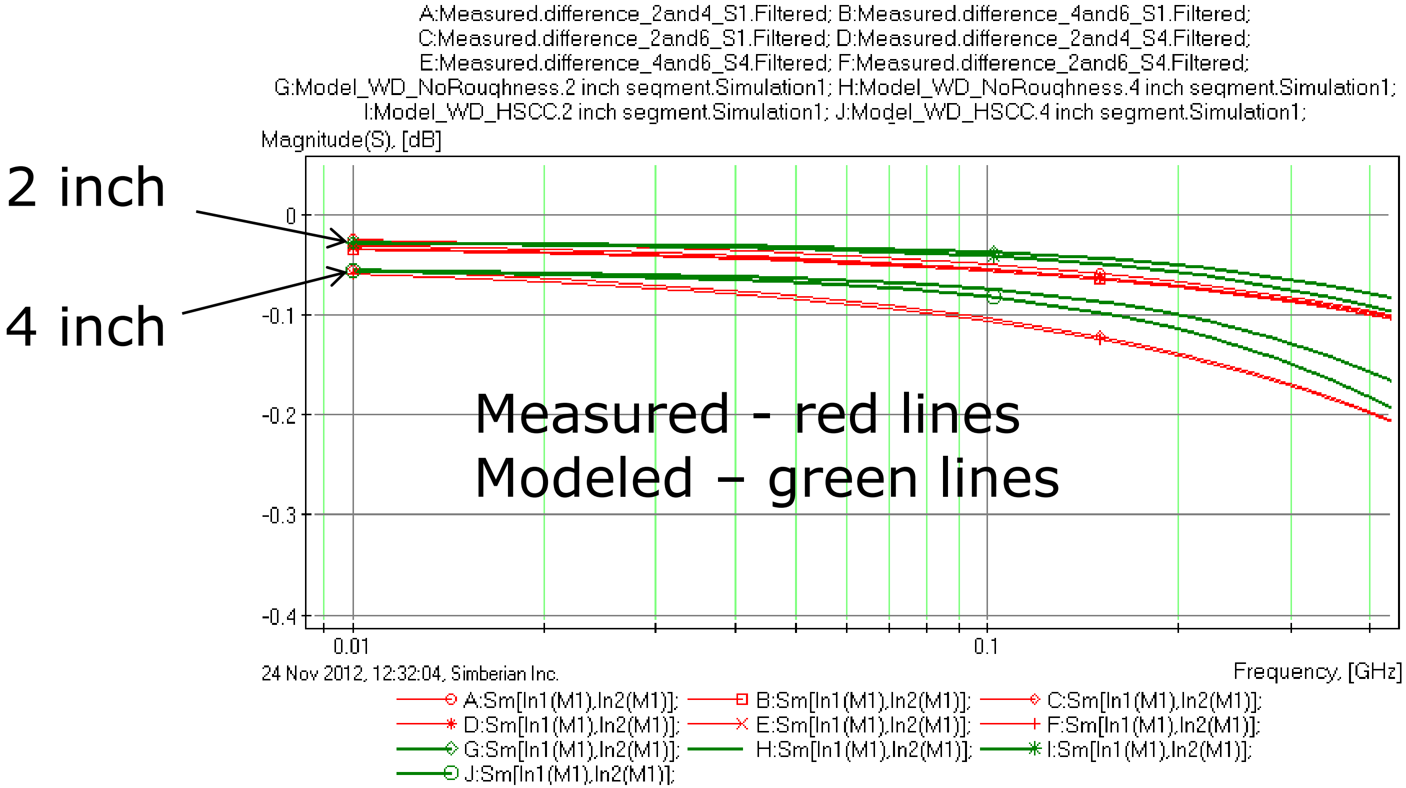
GMS Insertion Loss



GMS Group Delay

Resistivity at DC

- Copper resistivity was adjusted to 1.1 or annealed copper to match measured data at very low frequencies



2 inch

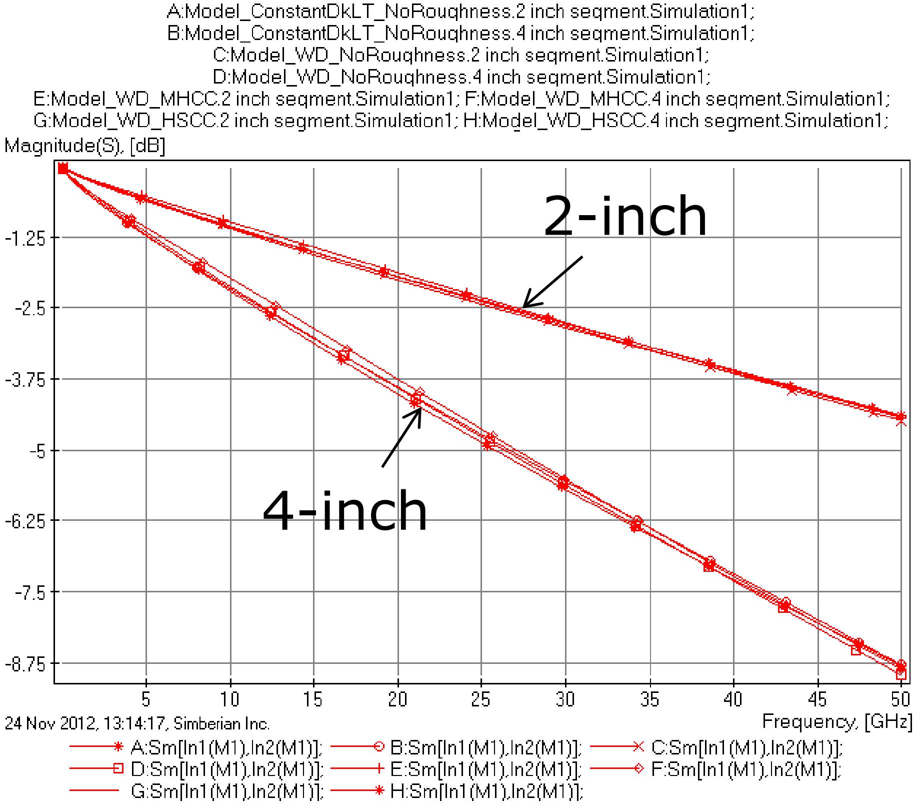
4 inch

Measured - red lines
Modeled – green lines

Model comparison up to 50 GHz

- All models produce close IL and GD

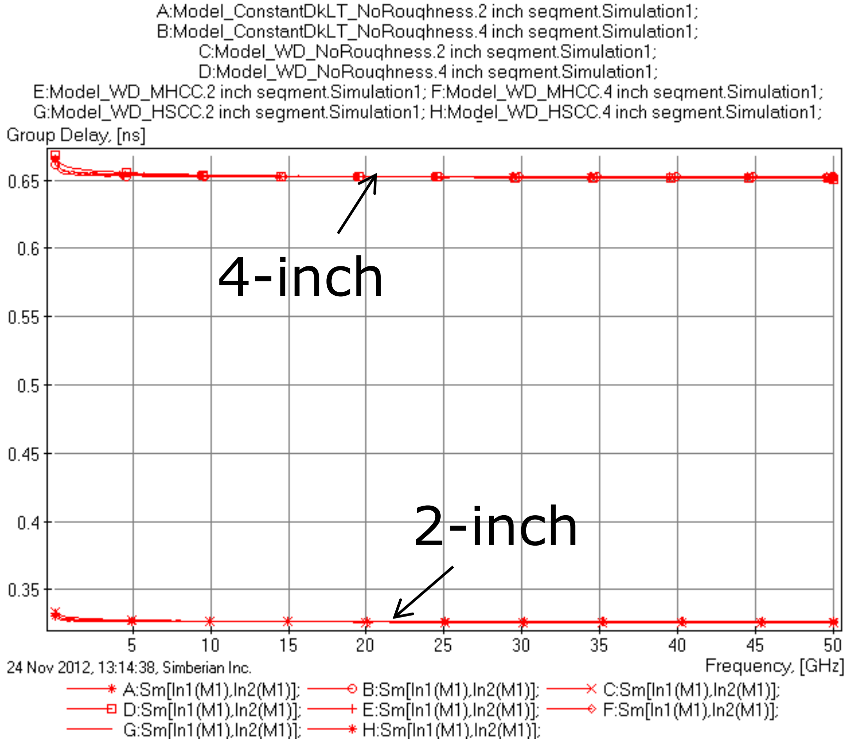
GMS Insertion Loss



2-inch

4-inch

GMS Group Delay

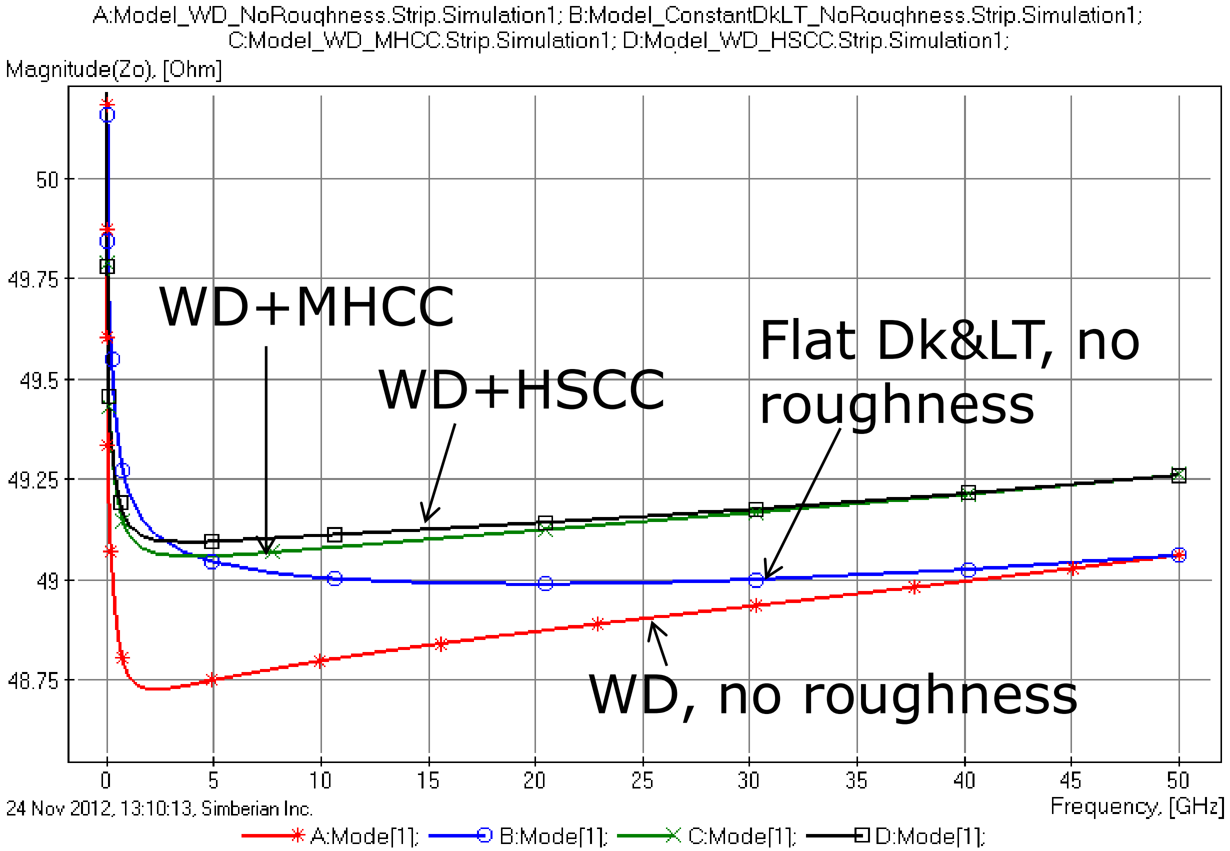


4-inch

2-inch

Characteristic impedance comparison

- GMS-parameters are close, but Zo are different!!!



WD+MHCC

WD+HSCC

Flat Dk&LT, no roughness

WD, no roughness

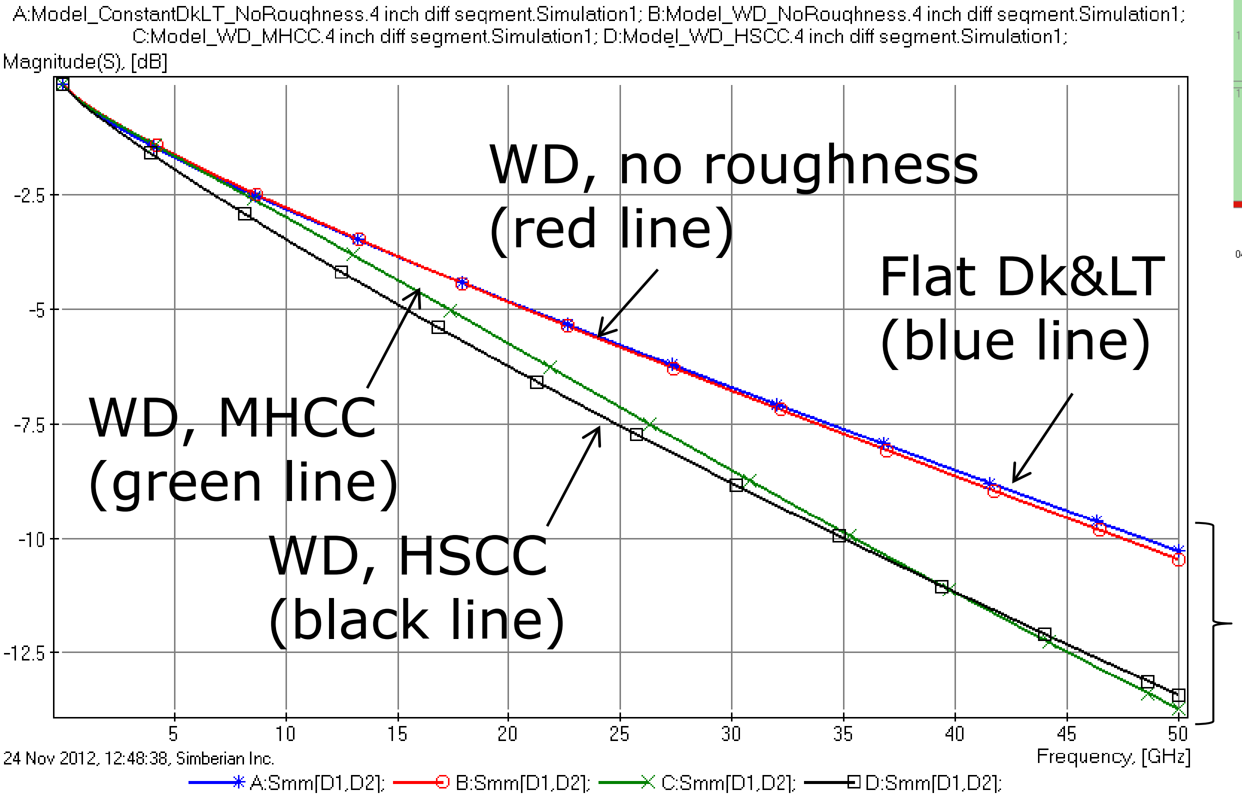
Though, less then 0.5 Ohm is within manufacturing tolerances

What model is right?

- All models are suitable for the practical analysis of 9.9 mil strip line in this dielectric
  - Non-causality in the flat model can be easily fixed with the rational approximation
  - Group delay dispersion concerns are not important for practical reasons (small differences)
- Even static field solver with flat dielectric model can produce acceptable accuracy for strip line!!!
- But, if cross-section changed models without roughness introduce larger errors

Differential 5 mil strips, 4.6 mil distance

- All models have very close results for 9.9 mil strip, but produce large difference for diff strips in the insertion loss!!!



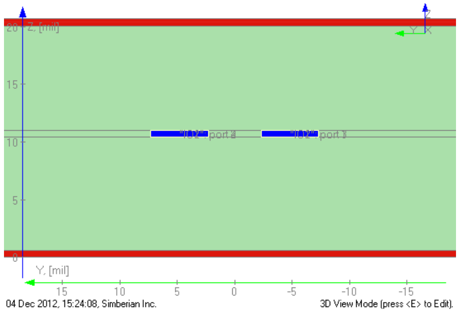
WD, no roughness (red line)

Flat Dk&LT (blue line)

WD, MHCC (green line)

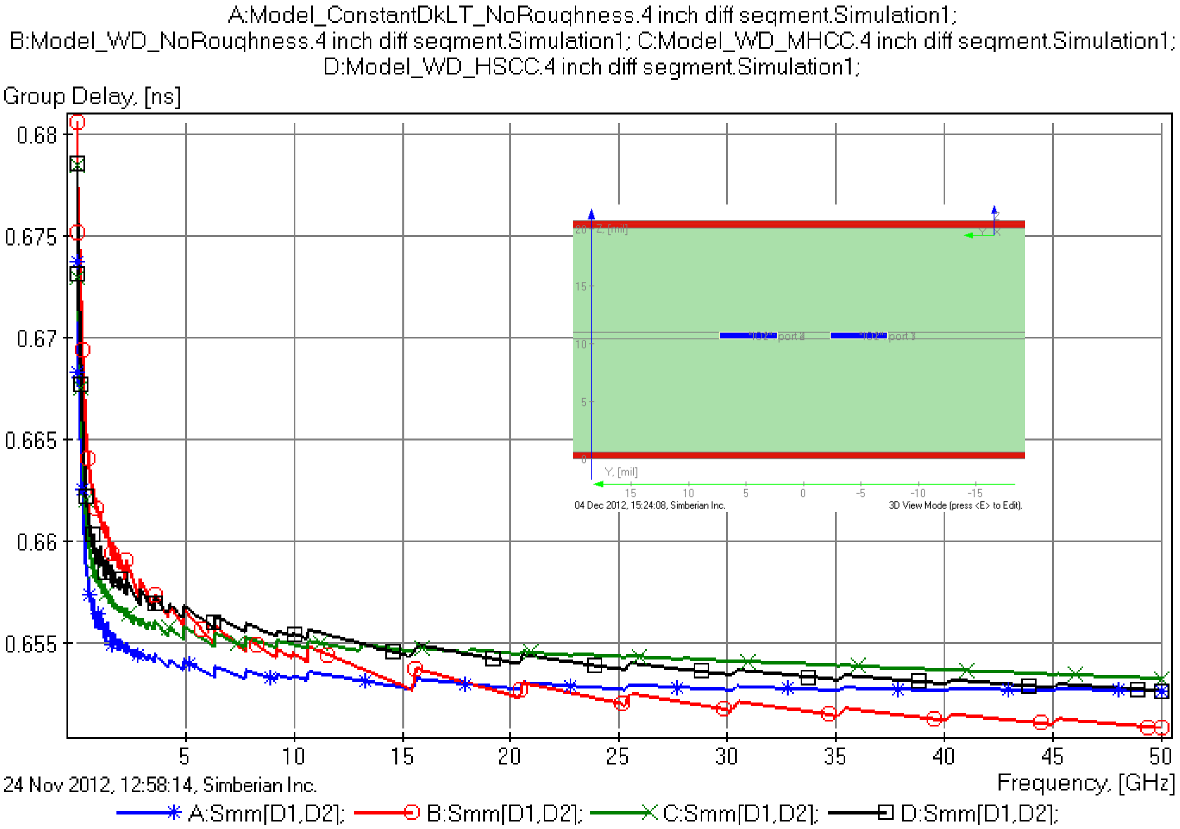
WD, HSCC (black line)

Over 30% difference!!!



Differential 5 mil strips, 4.6 mil distance

- Group delays for differential transmission through 4 inch line segment are within 5 ps



It is all about the losses!!!