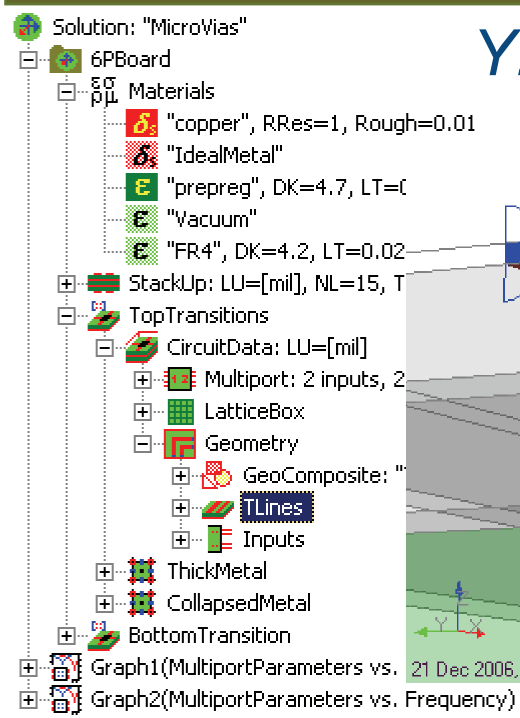
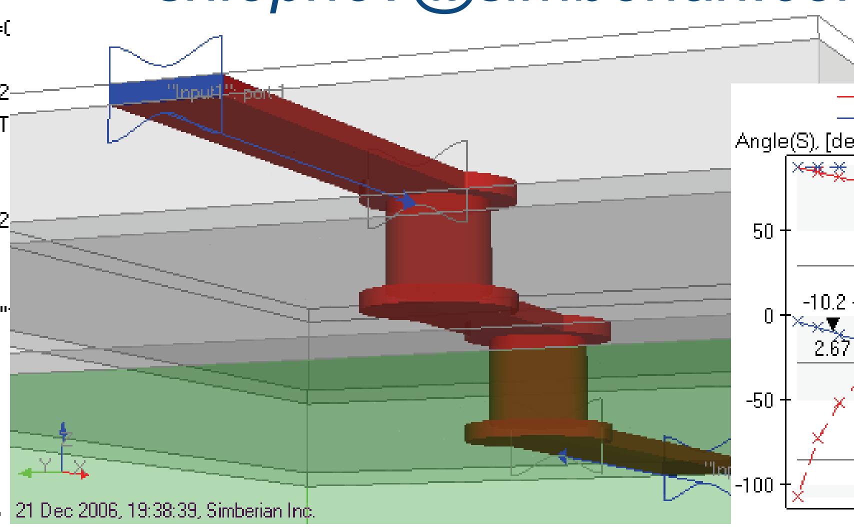
Simberian
Electromagnetic Solutions

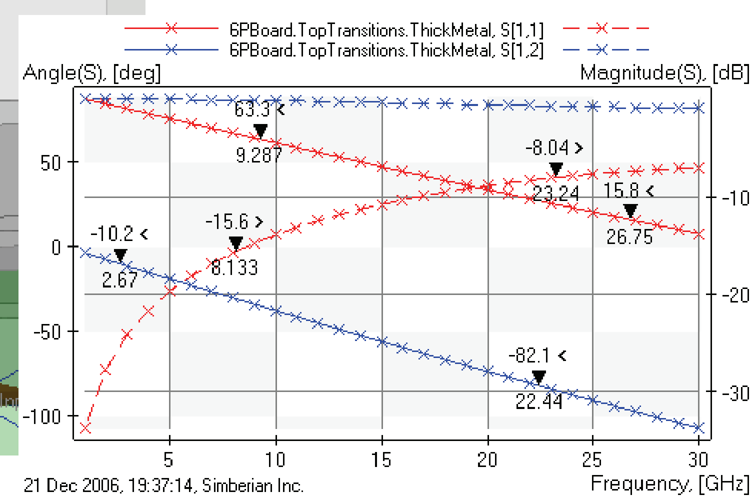
Broadband transmission line models for analysis of serial data channel interconnects

Y. O. Shlepnev, Simberian, Inc.

shlepnev@simberian.com







Agenda

- Introduction
- Broadband transmission line theory
  - Modal superposition
  - S-parameters of t-line segment
- Signal degradation factors
  - Conductor effects
  - Dielectric effects
- RLGC parameters extraction technologies
  - 2D static field solvers
  - 2D magneto-static field solvers
  - 3D full-wave solvers
- Examples of broadband parameters extraction
- Conclusion

Introduction

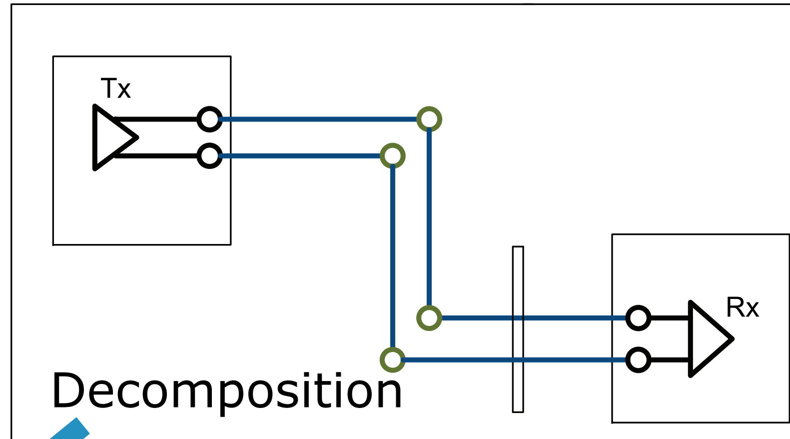
- Faster data rates drive the need for accurate electromagnetic models for multi-gigabit data channels
- Without the electromagnetic models, a channel design may require
  - Test boards, experimental verification, …
  - Multiple iterations to improve performance
- No models or simplified static models may result in the design failure, project delays, increased cost …

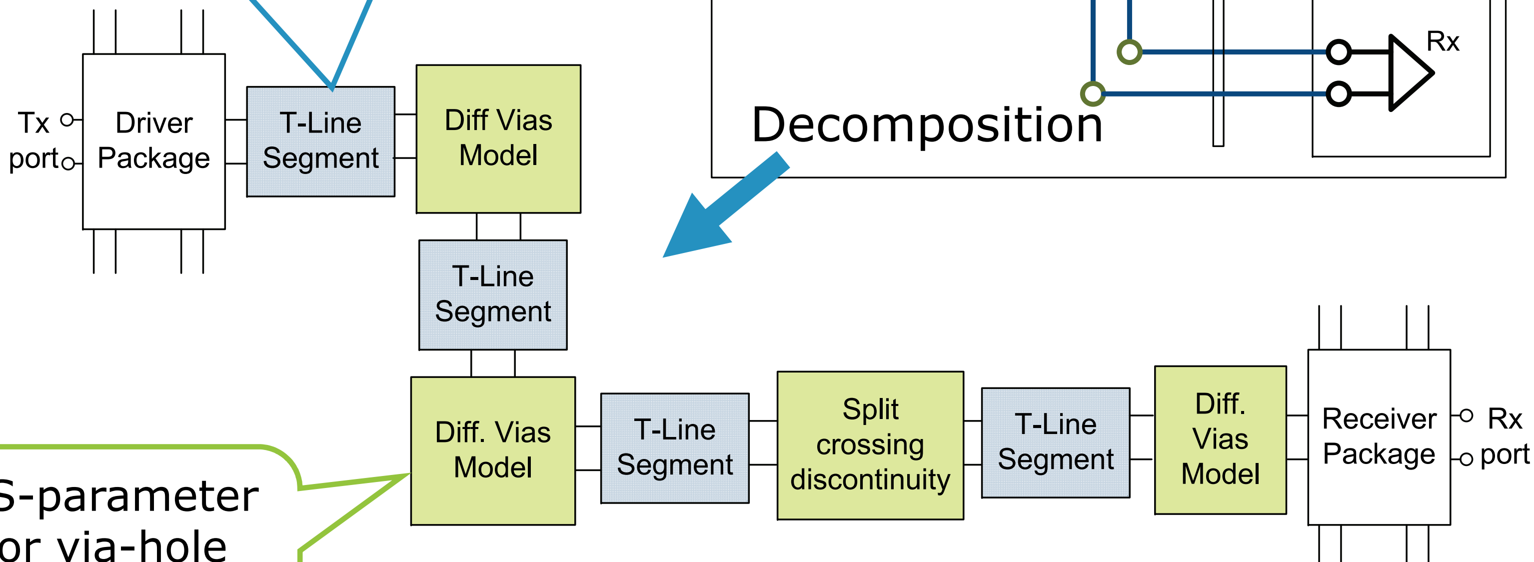
Trends in the Signal Integrity Analysis

- 80-s
  - Static field solvers for parameters extraction, simple frequency-independent loss models
  - Simple time-domain line segment simulation algorithms
    - Behavioral models, lumped RLGC models, finite differences,…
- Since 90-s
  - Static field solvers for parameters extraction, frequency-dependent analytical loss models
  - W-element for line segment analysis
- Trends in this decade
  - Transition to full-wave electromagnetic field solvers
  - Method of characteristic type algorithms or W-elements with tabulated RLGC per unit length parameters for analysis of line segment

De-compositional analysis of a channel

W-element models for t-line segments defined with RLGC per unit length parameters





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

Transmission line and discontinuity models are required for successful analysis of multi-gigabit channels!

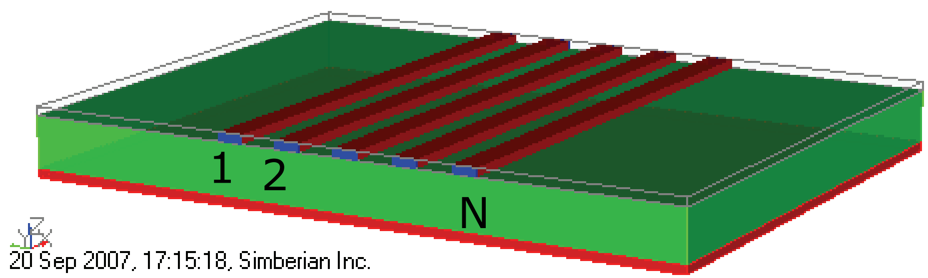
Agenda

- Introduction
- Broadband transmission line theory
  - Modal superposition
  - S-parameters of t-line segment
- Signal degradation factors
  - Conductor effects
  - Dielectric effects
- RLGC parameters extraction technologies
  - 2D static field solvers
  - 2D magneto-static field solvers
  - 3D full-wave solvers
- Examples of broadband parameters extraction
- Conclusion

Transmission line description with generalized Telegrapher's equations

$$\frac{\partial \bar{V}(x)}{\partial x} = -Z(\omega)\cdot \bar{I}(x)$$

$$\frac{\partial \bar{I}(x)}{\partial x} = -Y(\omega)\cdot \bar{V}(x)$$



Plus boundary conditions at the ends of the segment

$$Z(\omega) = R(\omega) + i\omega \cdot L(\omega)$$

$$Y(\omega) = G(\omega) + i\omega \cdot C(\omega)$$

I – complex vector of N currents

V – complex vector of N voltages

Z [Ohm/m] and *Y* [S/m] are complex NxN matrices of impedances and admittances per unit length

R [Ohm/m], L [Hn/m] – real NxN frequency-dependent matrices of resistance and inductance per unit length

G [S/m], C [F/m] – real NxN frequency-dependent matrices of conductance and capacitance per unit length

Transformation to modal space

$$Z(\omega) = R(\omega) + i\omega \cdot L(\omega)$$
$$Y(\omega) = G(\omega) + i\omega \cdot C(\omega)$$

Per unit length matrix parameters (NxN complex matrices)

$$y(\omega) = M_I^{-1} \cdot Y(\omega) \cdot M_V$$
$$z(\omega) = M_V^{-1} \cdot Z(\omega) \cdot M_I$$

Matrices of impedances and admittances per unit length are transformed into diagonal form with current M_I and voltage M_V transformation matrices (both are frequency-dependent)

$$\overline{V} = M_V \cdot \overline{v}$$
$$\overline{I} = M_I \cdot \overline{i}$$

Definition of terminal voltage and current vectors through modal voltage and current vectors and transformation matrices

$$W = M_I^t \cdot M_V = M_V^t \cdot M_I$$

Diagonal modal reciprocity matrix

$$P = M_V^t \cdot M_I^*$$

Complex power transferred by modes along the line

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

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

Modal complex characteristic impedance and propagation constant are defined by elements of the diagonal impedance and admittance matrices

Waves in multiconductor t-lines

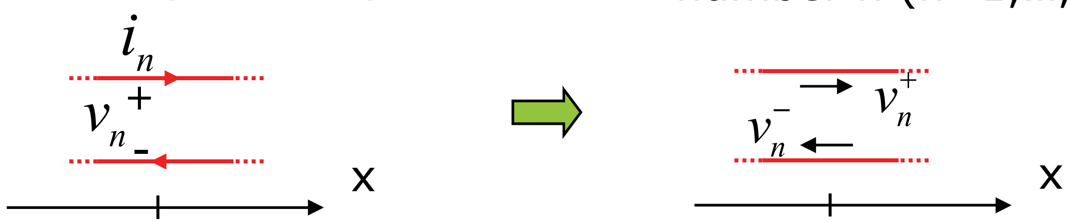
$$Z_{0n}(\omega) = \sqrt{z_{n,n}(\omega) / y_{n,n}(\omega)}$$

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

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)



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

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

$$\overline{V} = M_V \cdot \overline{v}$$
$$\overline{I} = M_I \cdot \overline{i}$$

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

One and two-conductor lines

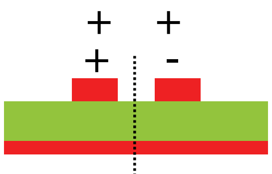
One-conductor case

$$M_V = M_I = 1$$

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

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

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

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

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

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

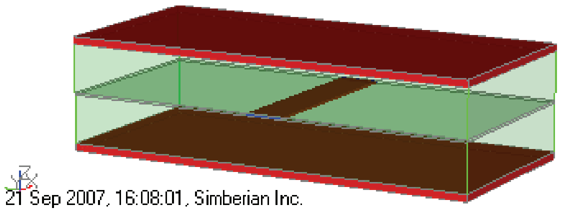
$$Z_{differential}(\omega) = \sqrt{z_{mm1,1}/y_{mm1,1}} \qquad Z_{differential} = 2\cdot Z_{odd} \qquad \Gamma_{differential} = \Gamma_{odd}$$

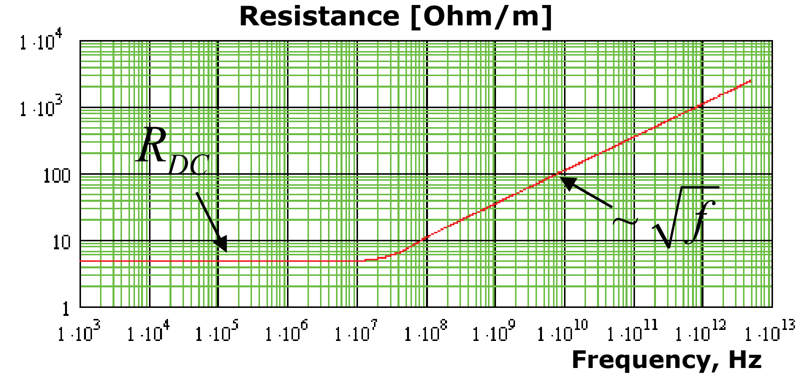
$$Z_{common}(\omega) = \sqrt{z_{mm2,2}/y_{mm2,2}} \qquad Z_{common} = 0.5\cdot Z_{even} \qquad \Gamma_{common} = \Gamma_{even}$$

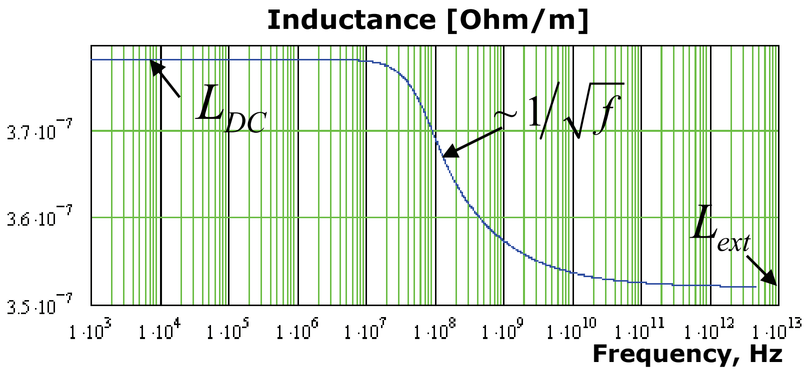
Example of causal R, L, G, C for a simple strip-line case (N=1)

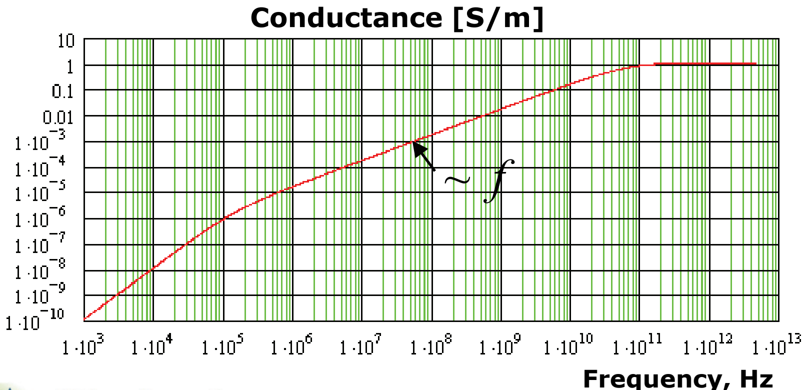
8-mil strip, 20-mil plane to plane distance, DK=4.2, LT=0.02 at 1 GHz, no dielectric conductivity.

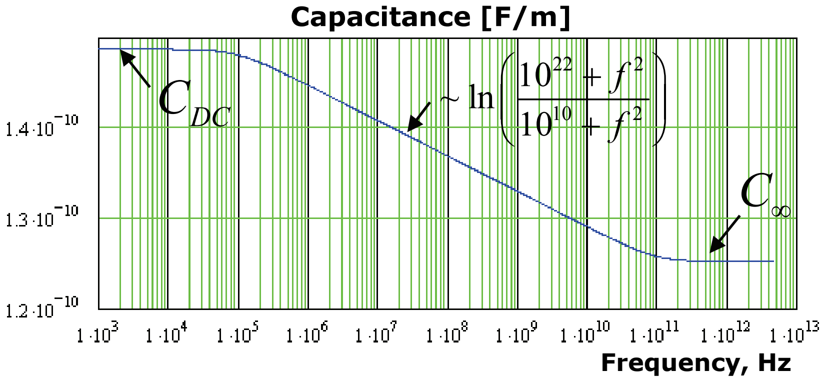
Strip is made of copper, planes are ideal, no roughness, no high-frequency dispersion.











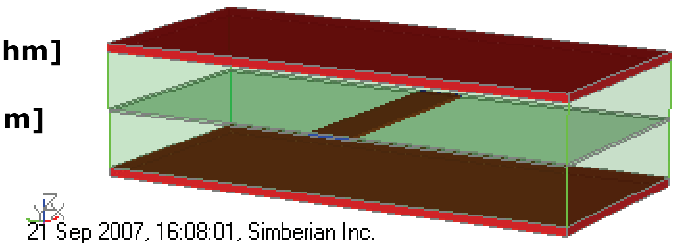
Broadband characteristic impedance and propagation constant for a simple strip-line

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

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

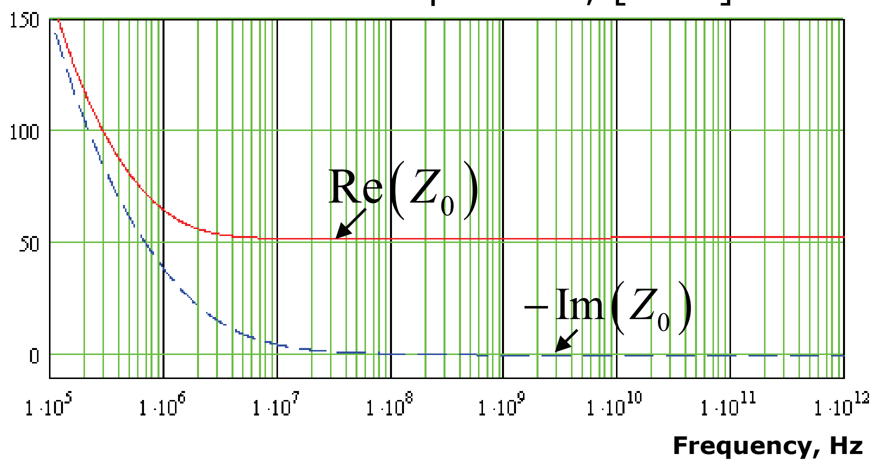
Attenuation Constant [Np/m] (α)

Phase Constant [rad/m] (β)

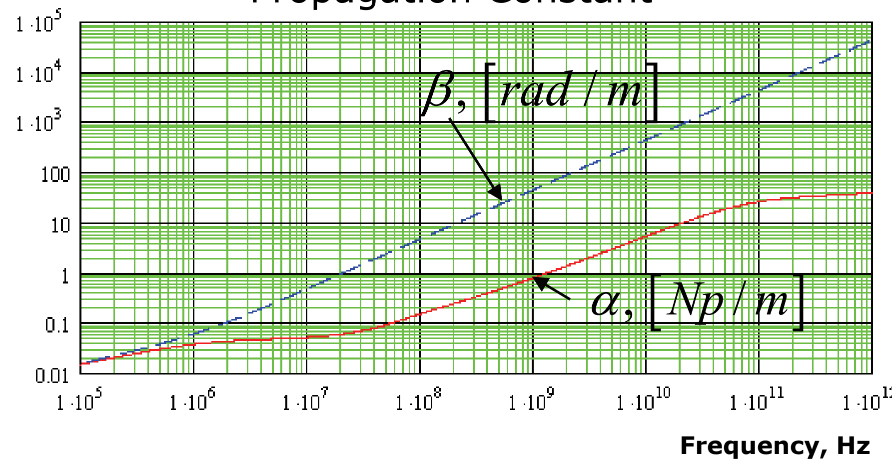


8-mil strip, 20-mil plane to plane distance. DK=4.2, LT=0.02 at 1 GHz, no dielectric conductivity. Strip is made of copper, planes are ideal, no roughness, no high-frequency dispersion.

Characteristic impedance, [Ohm]



Propagation Constant



Definitions of modal parameters

$$\Gamma_n(\omega) = \sqrt{z_{n,n}(\omega)\cdot y_{n,n}(\omega)} = \alpha_n + i\beta_n$$

$\alpha = \mathrm{Re}(\Gamma)$ attenuation constant [Np/m]

$\alpha_{dB} = \dfrac{20\cdot\alpha}{\ln(10)} \approx 8.686\cdot\alpha$ attenuation constant [dB/m]

$\beta = \mathrm{Im}(\Gamma)$ phase constant [rad/m]

$\Lambda = \dfrac{2\pi}{\beta}$ wavelength [m]

$\varepsilon_{eff} = \mathrm{Re}\left[-\left(\dfrac{c\cdot\Gamma}{\omega}\right)^2\right]$ effective dielectric constant

$p = \dfrac{c}{\upsilon_p} = \dfrac{c\cdot\beta}{\omega}$ slow-down factor, *c* is the speed of electromagnetic waves in vacuum

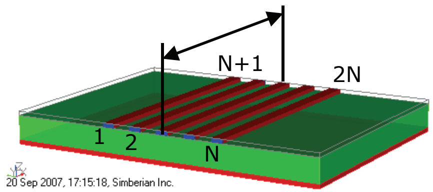
$\upsilon_p = \dfrac{\omega}{\beta}$ phase velocity [m/sec]

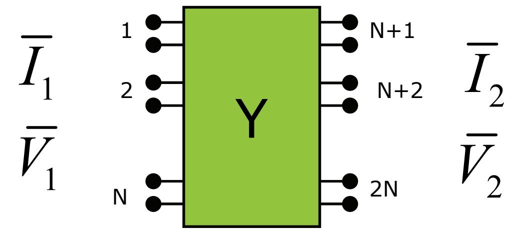
$\tau_p = \dfrac{\beta}{\omega}$ phase delay [sec/m]

$\upsilon_g = \dfrac{\partial\omega}{\partial\beta}$ group velocity [m/sec]

$\tau_g = \dfrac{\partial\beta}{\partial\omega}$ group delay [sec/m]

Admittance parameters of multiconductor line segment





$$\tilde{Y}(\omega,l)=\begin{bmatrix} diag\left(\dfrac{cth(\Gamma_n l)}{Z_{0n}}\right) & diag\left(-\dfrac{csh(\Gamma_n l)}{Z_{0n}}\right) \\ diag\left(-\dfrac{csh(\Gamma_n l)}{Z_{0n}}\right) & diag\left(\dfrac{cth(\Gamma_n l)}{Z_{0n}}\right) \end{bmatrix}$$

2N x 2N three-diagonal admittance matrix of the line segment in the modal space

$$Y(\omega,l)=\begin{bmatrix} M_I & 0 \\ 0 & M_I \end{bmatrix}\cdot\tilde{Y}(\omega,l)\cdot\begin{bmatrix} M_V^{-1} & 0 \\ 0 & M_V^{-1} \end{bmatrix}$$

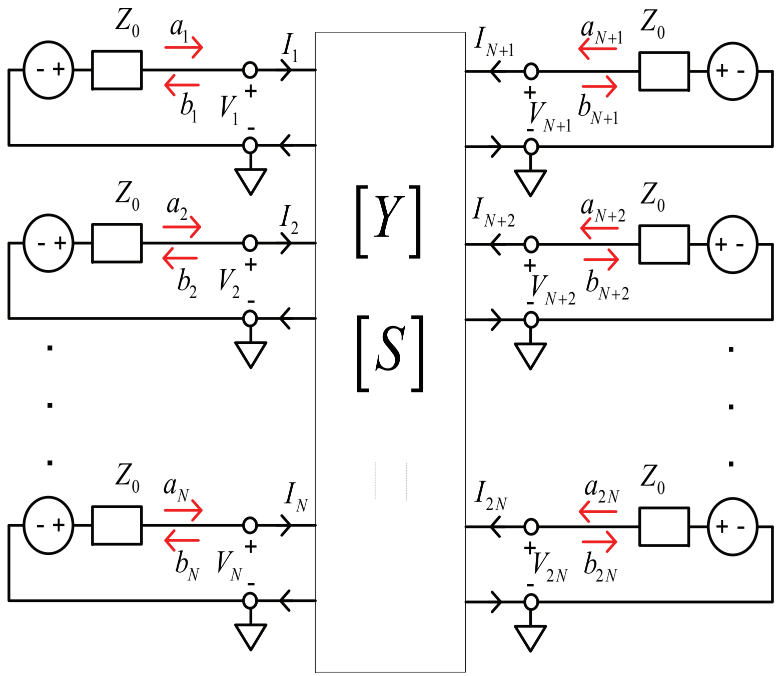
2N x 2N admittance matrix of the line segment in the terminal space

$$\begin{bmatrix} \overline{I}_1 \\ \overline{I}_2 \end{bmatrix}=Y(\omega,l)\cdot\begin{bmatrix} \overline{V}_1 \\ \overline{V}_2 \end{bmatrix}$$

Admittance matrix leads to a system of linear equations with voltages and currents at the external line terminals

Alternative equivalent formulation with admittance and propagation operators is used in W-element to facilitate integration in time domain

Scattering parameters of multiconductor line segment



$$\begin{bmatrix}\bar{I}_1\\ \bar{I}_2\end{bmatrix} = Y(\omega,l)\cdot\begin{bmatrix}\bar{V}_1\\ \bar{V}_2\end{bmatrix}$$ Admittance parameters

$$\bar{a}_{1,2} = \frac{1}{2\sqrt{Z_0}}\left(\bar{V}_{1,2} + Z_0\cdot\bar{I}_{1,2}\right)$$ Vectors of incident waves

$$\bar{b}_{1,2} = \frac{1}{2\sqrt{Z_0}}\left(\bar{V}_{1,2} - Z_0\cdot\bar{I}_{1,2}\right)$$ Vectors of reflected waves

$$\begin{bmatrix}\bar{b}_1\\ \bar{b}_2\end{bmatrix} = S(\omega,l)\cdot\begin{bmatrix}\bar{a}_1\\ \bar{a}_2\end{bmatrix}$$ Scattering parameters

$$Y_N = Z_0^{1/2}\cdot Y(\omega,l)\cdot Z_0^{1/2}$$

Normalization matrix is diagonal matrix usually with 50-Ohm values at the diagonal

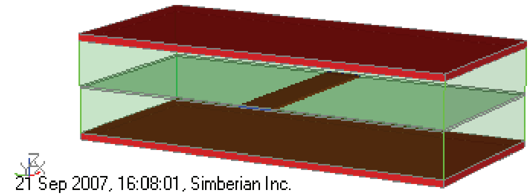
$$S(\omega,l) = (U - Y_N)\cdot(U + Y_N)^{-1}$$

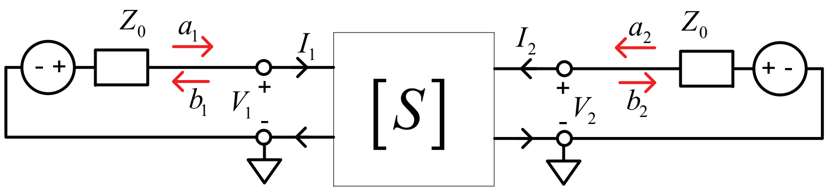
S-matrix of the line segment computed as the Cayley transform of the normalized Y-matrix

Simple strip-line segment example

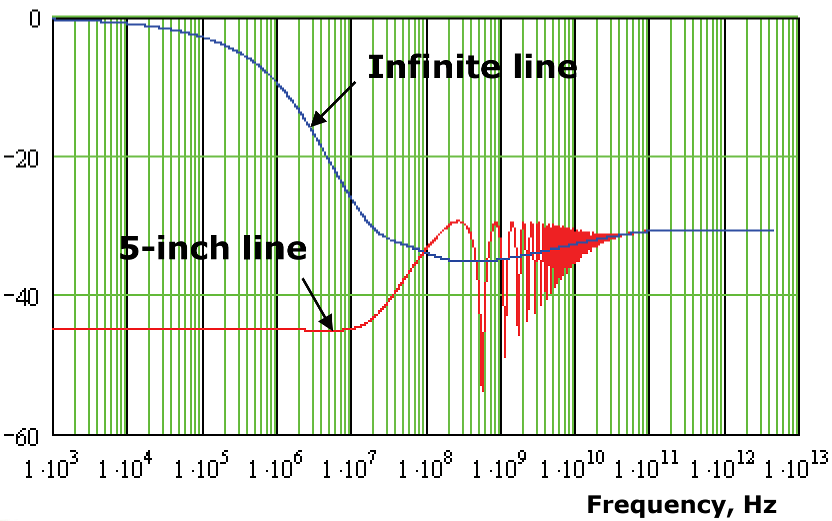
8-mil strip, 20-mil plane to plane distance, DK=4.2, LT=0.02 at 1 GHz, no dielectric conductivity.

Strip is made of copper, planes are ideal, no roughness, no high-frequency dispersion.

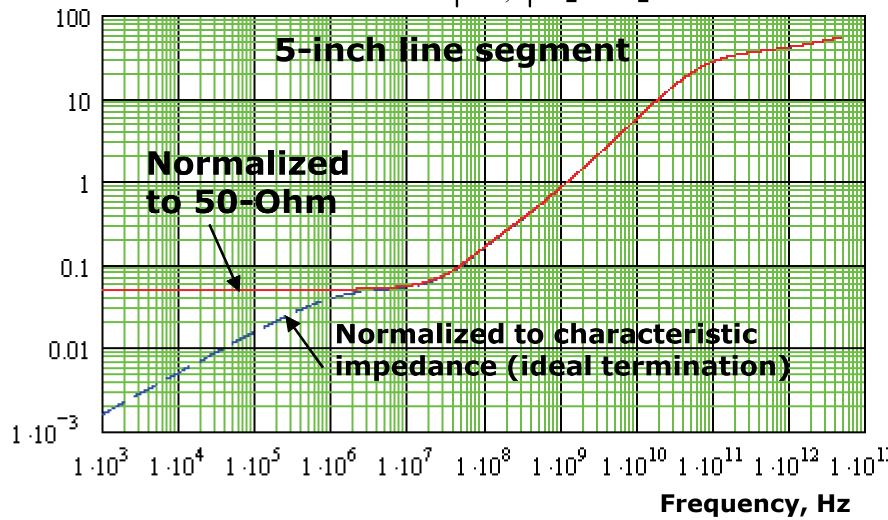




$$20\log\left|S_{1,1}\right|, [dB]$$



$$-20\log\left|S_{2,1}\right|, [dB]$$

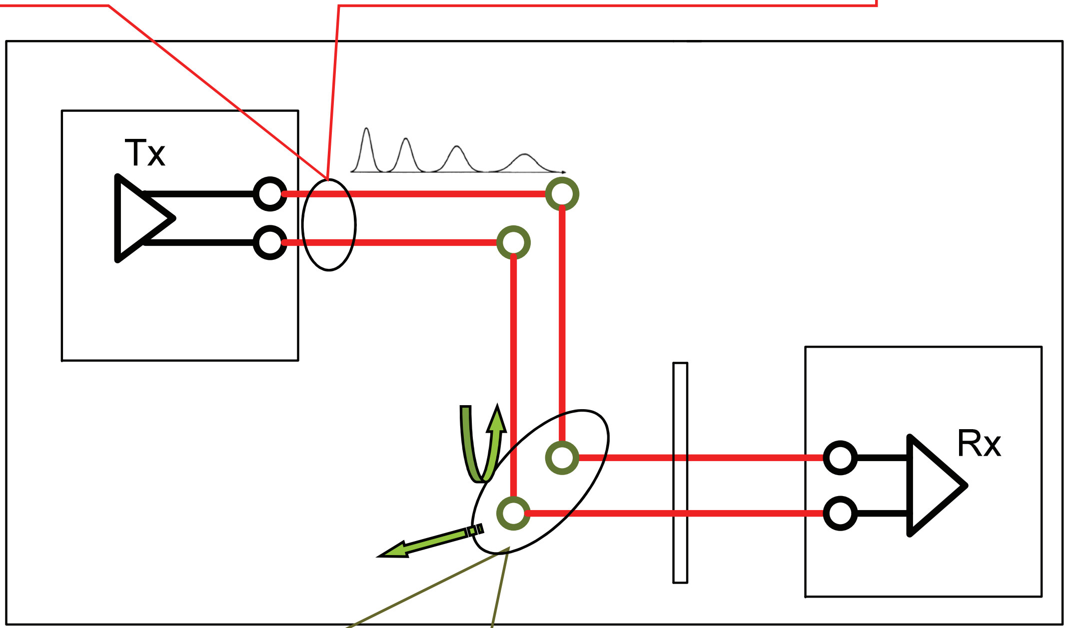


Agenda

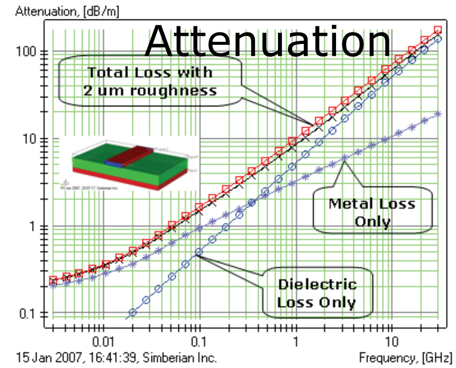
- Introduction
- Broadband transmission line theory
  - Modal superposition
  - S-parameters of t-line segment
- Signal degradation factors
  - Conductor effects
  - Dielectric effects
- RLGC parameters extraction technologies
  - 2D static field solvers
  - 2D magneto-static field solvers
  - 3D full-wave solvers
- Examples of broadband parameters extraction
- Conclusion

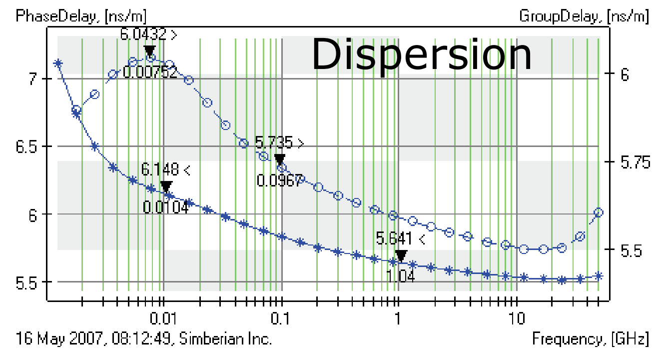
Signal degradation factors

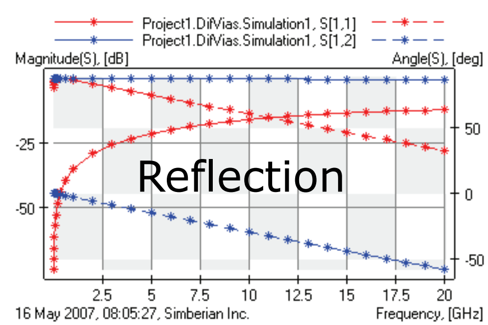
Transmission lines:
Attenuation and dispersion due to physical conductor and dielectric properties
High-frequency dispersion



Via-hole transitions and discontinuities:
Reflection, radiation and impedance mismatch

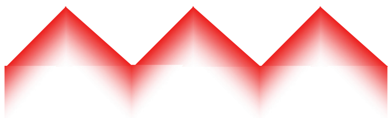






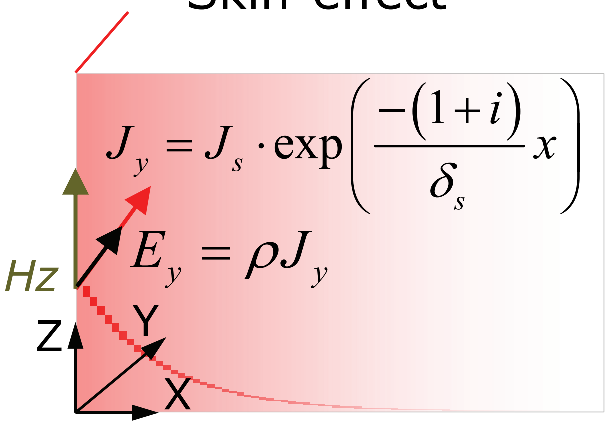
Conductor attenuation and dispersion effects

Roughness
~40 MHz



R(f) and L(f)
increase

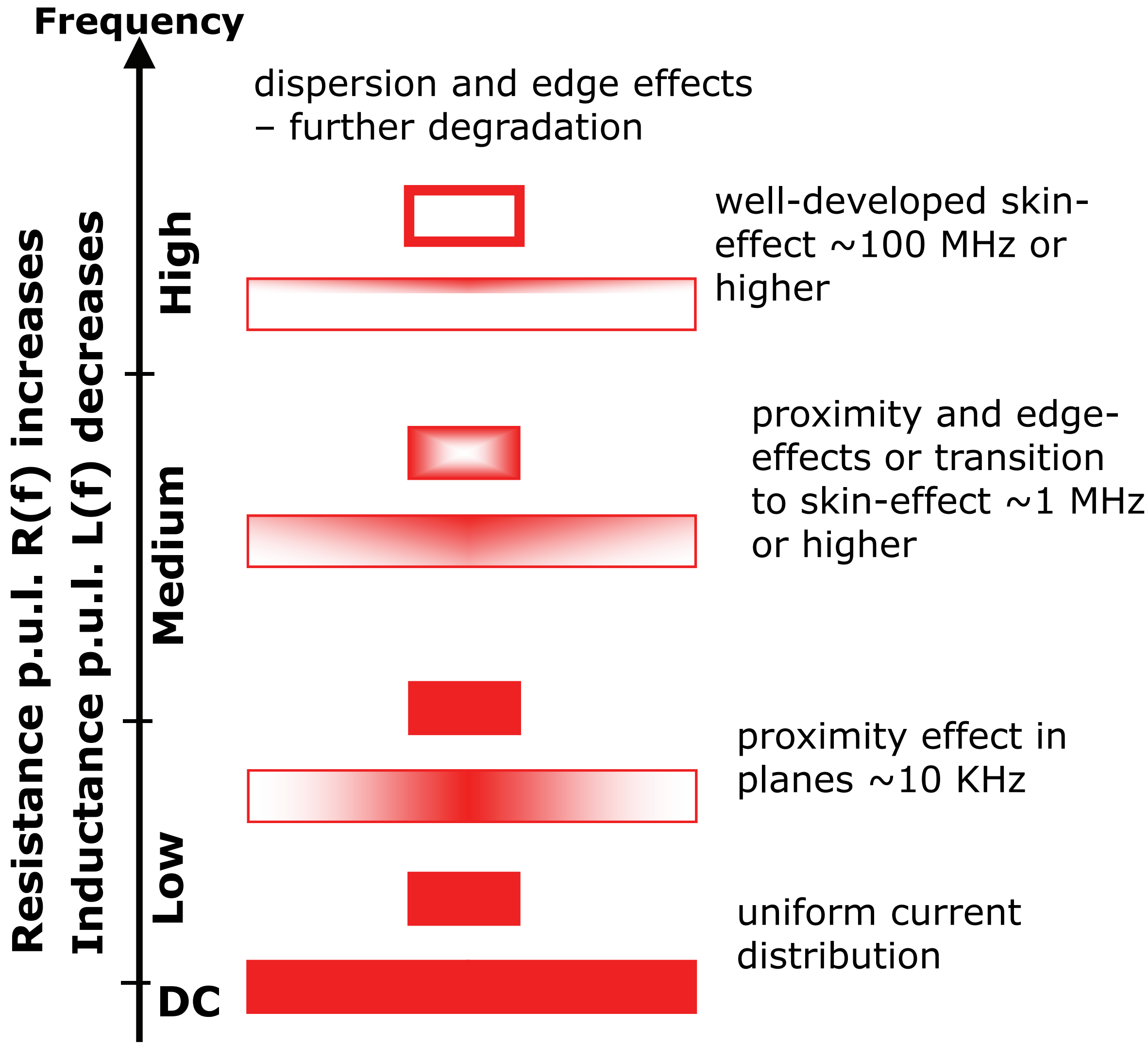
Skin-effect



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

$$E_y = \rho J_y$$

$$\delta_s = \sqrt{\frac{\rho}{\pi \mu f}} \; [m]$$

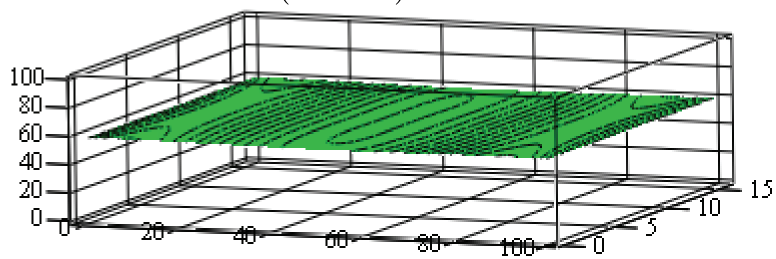


Current distribution in rectangular conductor

Current density distribution in 10 mil wide and 1 mil thick copper at different frequencies

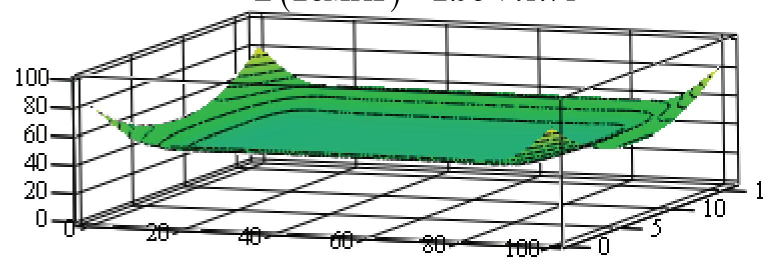
1.7 MHz, t/s=0.5, Jc/Je=0.999

$$Z(1.7MHz) = 2.67 + i0.11$$



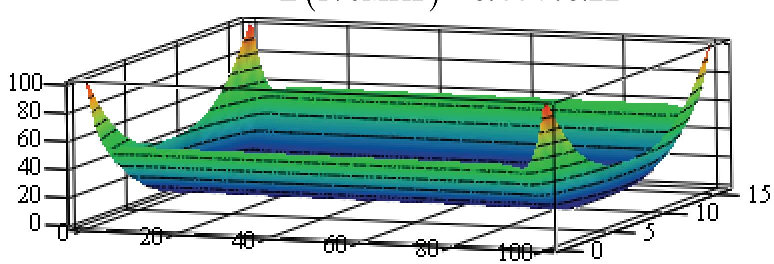
28 MHz, t/s=2.0, Jc/Je=0.76

$$Z(28MHz) = 2.95 + i1.71$$

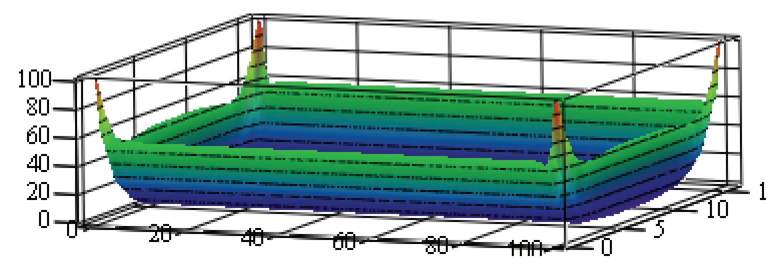


170 MHz, t/s=5, Jc/Je=0.16

$$Z(170MHz) = 6.44 + i6.22$$



1 GHz, t/s=12.2, Jc/Je=0.005

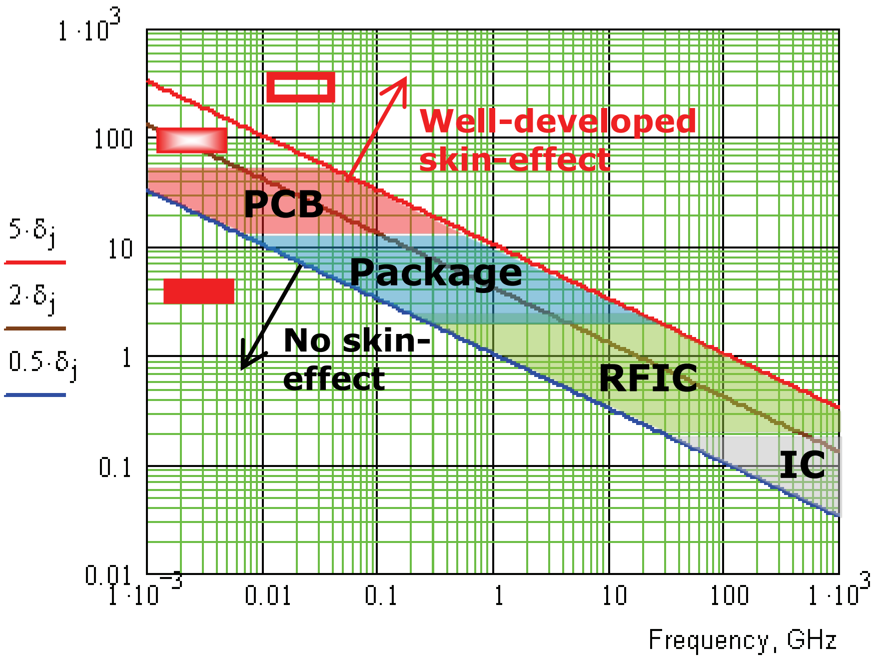


t/s is the strip thickness to the skin depth ratio

Jc/Je is the ratio of current density at the edge to the current in the middle.

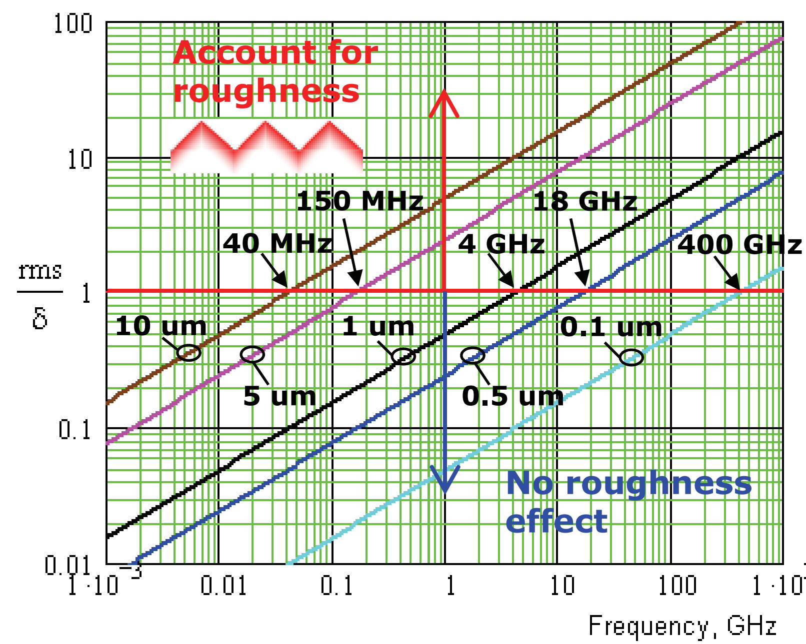
Transition to skin-effect and roughness

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



Interconnect or plane thickness in micrometers vs. Frequency in GHz

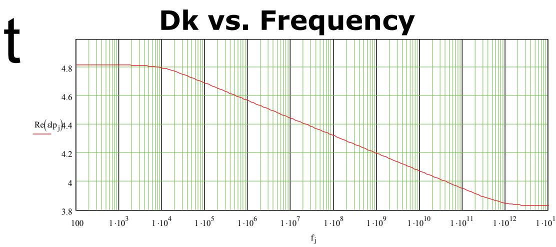
Ratio of skin depth to r.m.s. surface roughness in micrometers vs. frequency in GHz

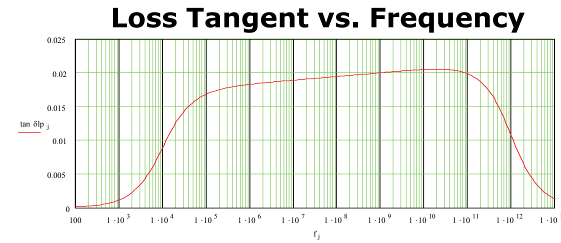


Roughness has to be accounted if rms value is comparable with the skin depth

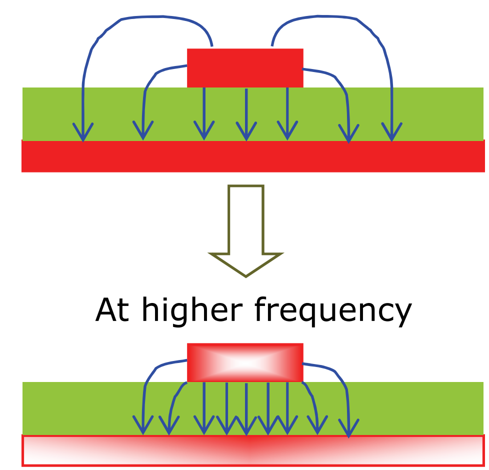
Dielectric attenuation and dispersion effects

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





- 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



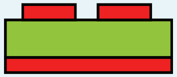
Agenda

- Introduction
- Broadband transmission line theory
  - Modal superposition
  - S-parameters of t-line segment
- Signal degradation factors
  - Conductor effects
  - Dielectric effects
- RLGC parameters extraction technologies
  - 2D static field solvers
  - 2D magneto-static field solvers
  - 3D full-wave solvers
- Examples of broadband parameters extraction
- Conclusion

Extraction of RLGC parameters

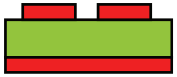
2D Static and Magneto-Quasi-Static Solvers

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



2D Full-Wave Solvers

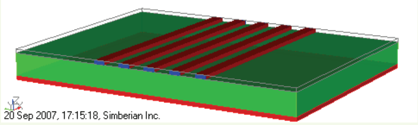
$$\nabla\times\vec{E} = -i\omega\mu\vec{H}$$
$$\nabla\times\vec{H} = i\omega\varepsilon\vec{E} + \sigma\vec{E} + \vec{J}$$



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

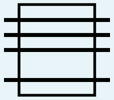
3D Full-Wave Solver

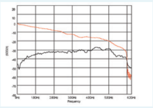
$$\nabla\times\vec{E} = -i\omega\mu\vec{H}$$
$$\nabla\times\vec{H} = i\omega\varepsilon\vec{E} + \sigma\vec{E} + \vec{J}$$



W-element model

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

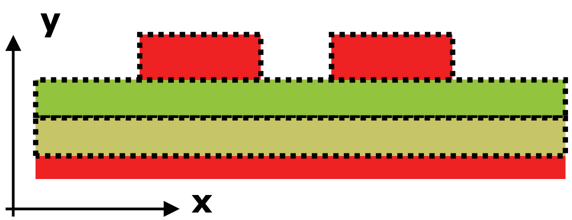




System-level simulator

2D static field solvers

Solve Laplace's equations for a transmission line cross-section to find capacitance and conductance p.u.l. matrices and distribution or charge on metal boundaries



$$\nabla^2 \varphi\left(x, y, \varepsilon' - i\varepsilon''\right) = 0$$
$$\varphi\big|_{S_i} = \varphi_i \ (i = 1, \ldots, N)$$

Plus additional boundary conditions at the boundaries between dielectrics

$$\Rightarrow$$

$$C\left(\varepsilon'\left(f_0\right)\right),\ G\left(\varepsilon''\left(f_0\right)\right) = G_d \cdot f_0$$
$$C\left(\varepsilon_0\right) \Rightarrow L_{ext} = \mu_0 \varepsilon_0 C\left(\varepsilon_0\right)^{-1}$$
$$q\left(x, y\right) \Rightarrow R_s\left(f_0\right)$$

Integral equation or boundary element methods with meshing of conductor and dielectric boundaries are usually used to solve the problem.

Conductor loss accounted for with diagonal *RDC* and with *R* computed at 1 GHz with the perturbation method, assuming well-developed skin effect

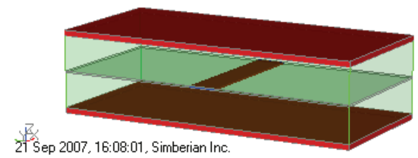
Solver outputs *Lo=Lext, Co=C(f), Ro, Go, Rs=Rs(fo)/sqrt(fo), Gd=G(fo)/fo*

Frequency-dependency is reconstructed

```
.MODEL Model_W001 W MODELTYPE=RLGC N=1
* Lo  (H/m)
+ Lo = 3.25062e-007

* Co  (F/m)
+ Co = 1.3325e-010

* Ro (Ohm/m)
+ Ro = 4.77

* Go (S/m)
+ Go = 0

* Rs (Ohm/m-sqrt(Hz))
+ Rs = 0.00108482

* Gd (S/m-Hz)
+ Gd = 1.6654e-011
```

Frequency-dependent impedance p.u.l. model based on static solution

8-mil strip, 20-mil plane to plane distance. DK=4.2, LT=0.02 at 1 GHz, no dielectric conductivity. Strip is made of copper, planes are ideal, no roughness no high-frequency dispersion.



$$Z(f) = R_{DC} + (1+i)R_s(f_0)\sqrt{\frac{f}{f_0}} + i2\pi f \cdot L_{ext} \quad \left[\frac{Ohm}{m}\right]$$

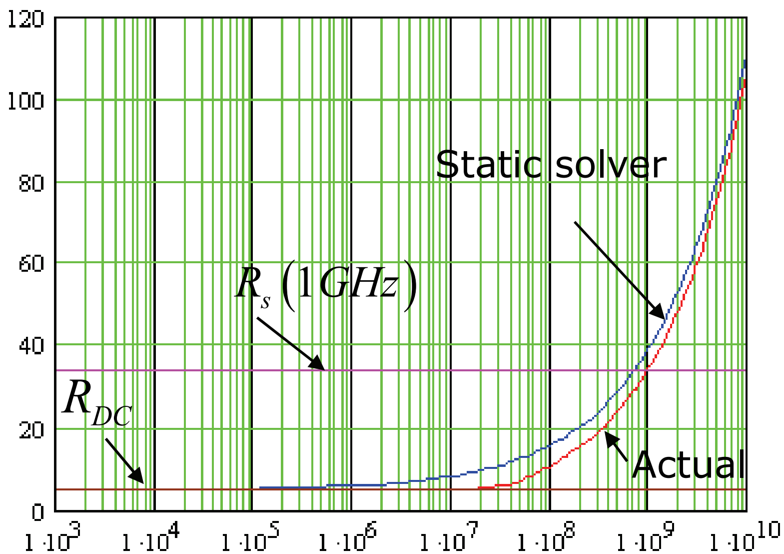
R_{DC} included at high-frequencies

Inductance diverges to infinity at DC

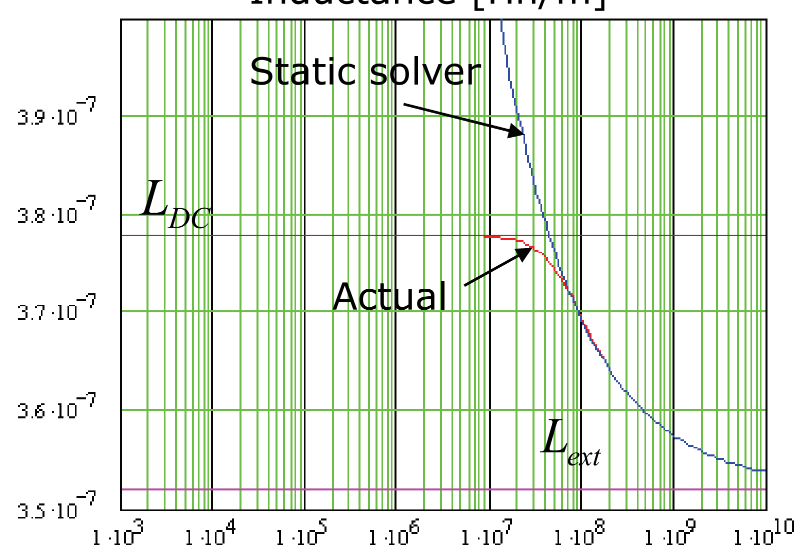
No actual transition to skin-effect

No roughness of metal finish effects

Resistance [Ohm/m]



Inductance [Hn/m]



Frequency-dependent admittance p.u.l. model based on static solution

$$Y(f) = f\frac{G(\varepsilon'', f_0)}{f_0} \cdot + i2\pi f \cdot C(\varepsilon', f_0)$$

Non-causal admittance model (typical)

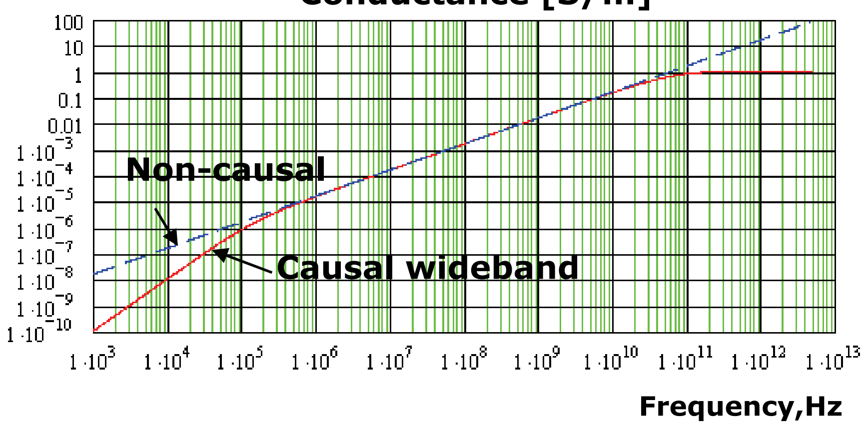
$$Y(f) = i2\pi f \cdot \left[C_\infty + \frac{C_{DC} - C_\infty}{(m_2 - m_1)\cdot \ln(10)} \cdot \ln\left(\frac{10^{m2} + if}{10^{m1} + if}\right)\right]$$

Causal wideband Debye model from Eldo (valid for lines with homogeneous dielectric)

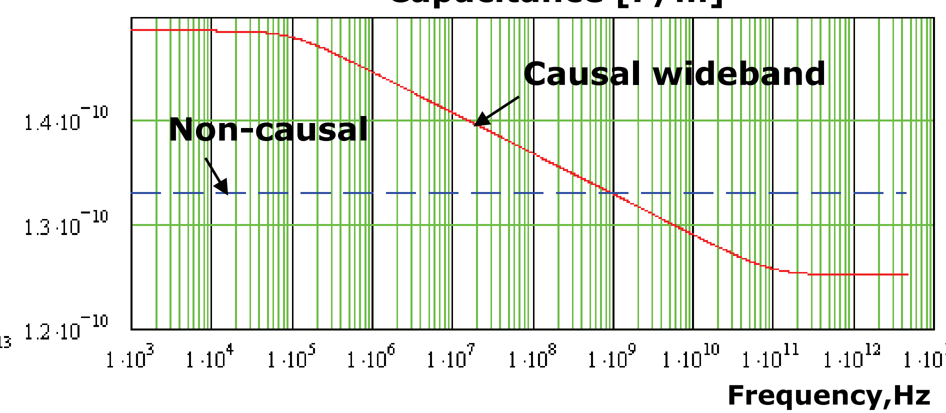
No high-frequency dispersion due to inhomogeneous dielectric

$$C_\infty = C(f_0) + \frac{\mathrm{Re}\left[\ln\left(\frac{10^{m2} + if_0}{10^{m1} + if_0}\right)\right]}{2\pi \cdot \mathrm{Im}\left[\ln\left(\frac{10^{m2} + if_0}{10^{m1} + if_0}\right)\right]} \cdot G_d(f_0) \qquad C_{DC} = C_\infty + \frac{(m_2 - m_1)\cdot \ln(10)}{-2\pi \cdot \mathrm{Im}\left[\ln\left(\frac{10^{m2} + if_0}{10^{m1} + if_0}\right)\right]} \cdot G_d(f_0)$$

Conductance [S/m]

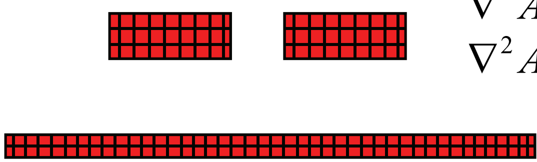


Capacitance [F/m]



2D quasi-static field solvers

Solve Laplace's equations outside of the conductors simultaneously with diffusion equations inside the conductors to find frequency-dependent resistance and inductance p.u.l.



$$\nabla^2 A_z(x,y) = i\omega\sigma\mu\cdot\left(A_z - A_0\right) \ \textit{inside conductors}$$
$$\nabla^2 A_z(x,y) = 0 \ \textit{outside conductors}$$

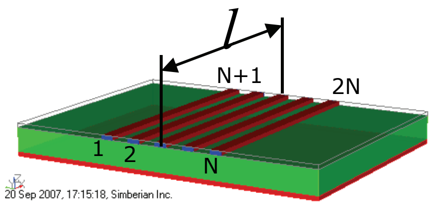
$$\Rightarrow \quad R(f), \ L(f)$$

Plus additional boundary conditions at the conductor surfaces

- Finite Element Method meshes whole cross-section of t-line including the metal interior
- Integral Equation Method can be used to mesh just the interior of the strips and planes
- Both approaches have significant numerical complexity (despite on being 2D):
    - To extract parameters of a line up to 10 GHz, the element or filament size near the metal surface has to be at least ¼ or skin depth that is about 0.16 um
    - It would be required about 236000 elements to mesh interior of 10 mil by 1 mil trace for instance and in addition interior of one or two planes have to be meshed too (another two million elements may be required)
- If element size is larger than skin-depth, effect of saturation of R can be observed (R does not grow with frequency)
- In addition, there is no influence of dielectric on the extracted R and L

3D full-wave solvers

Solve Maxwell's equations for a transmission line segment to find S-parameters:



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

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

$$\Rightarrow \quad S(\omega, l)$$

Plus additional boundary conditions such as Surface Impedance Boundary Conditions (SIBC)

Finite Element Method (FEM) meshes space and possibly interior of metal, but more often uses SIBC at the metal surface

Finite Integration Method (FIT) or Finite Difference Time Domain (FDTD) Method mesh space and usually use narrow-band approximation of SIBC

Method of Moments (MoM) meshes the surface of the strips and uses SIBC

No RLGC parameters per unit length as output

Approximate roughness models based on adjustment of conductor resistivity

No models for multilayered metal coating

No broadband dielectric models (1 or 2-poles Debye models in some solvers)

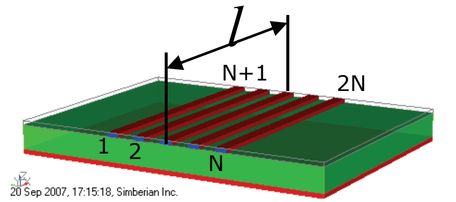
Simbeor: 3D full-wave hybrid solver

Solve Maxwell's equations for a transmission line segment to find S-parameters and frequency-dependent matrix RLGC per unit length parameters:

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

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

$$\Rightarrow \quad S(\omega, l) \quad R(\omega),\ L(\omega) \quad G(\omega),\ C(\omega)$$



Plus additional boundary conditions at the metal and dielectric surfaces

- Method of Lines (MoL) for multilayered dielectrics
  - High-frequency dispersion in multilayered dielectrics
  - Losses in metal planes
  - Causal wideband Debye dielectric polarization loss and dispersion models
- Trefftz Finite Elements (TFE) for metal interior
  - Metal interior and surface roughness models to simulate proximity edge effects, transition to skin-effect and skin effect
- Method of Simultaneous Diagonalization (MoSD) for lossy multiconductor line and multiport S-parameters extraction
  - Advanced 3-D extraction of modal and RLGC(f) p.u.l. parameters of lossy multi-conductor lines

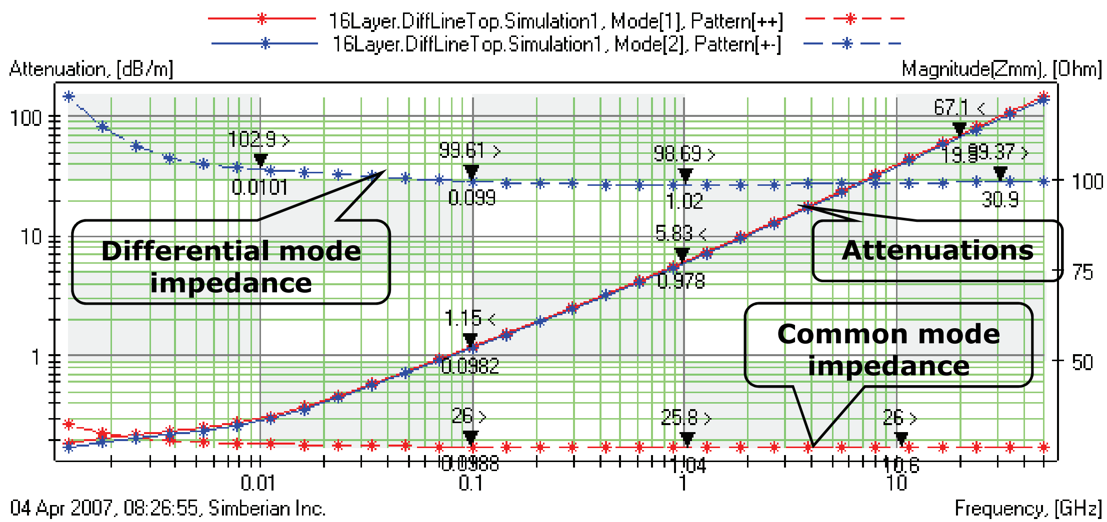
Comparison of field solvers technologies

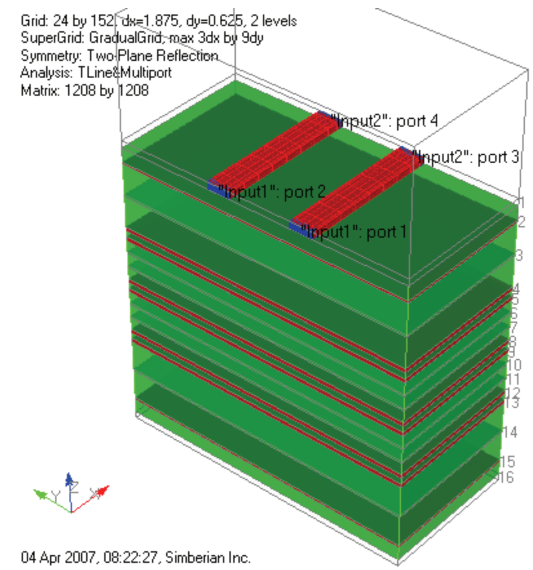
Feature \ Field Solver	Static Field Solvers	Quasi-Static Field Solvers	3D EM with intra-metal models (Simbeor 2007)
Output parameters	C, L, Ro, Go, Rs, Gs	L(f), R(f)	R(f), L(f), G(f), C(f)
Thin dielectric layers	Difficult	No	Yes
Transition to skin-effect in planes and traces	No	Yes	Yes
Skin and proximity effects	Yes	Yes with high-frq saturation effect	Yes
Metal surface roughness	No	No	Yes
Dispersion	No	No	Yes
3D characteristic impedance	No	No	Yes

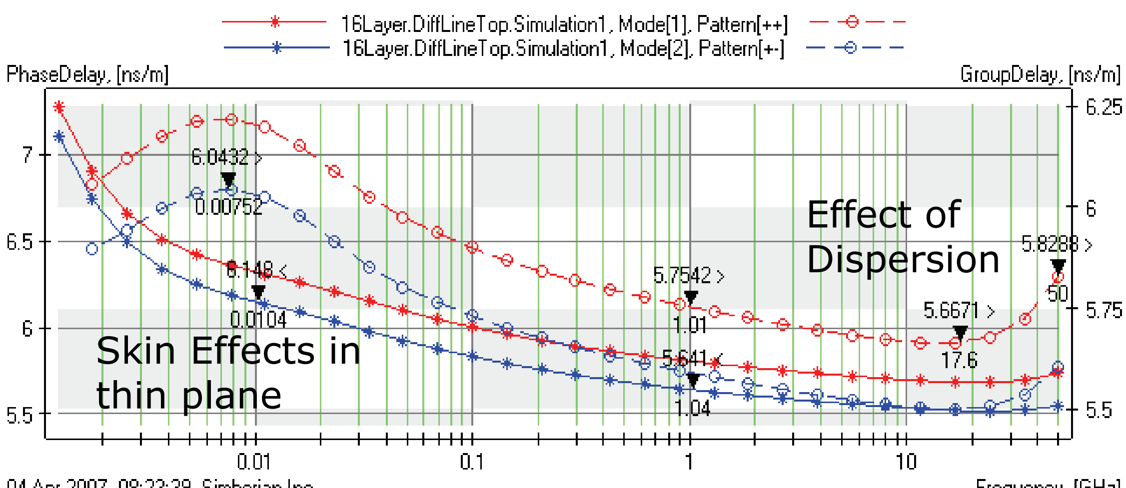
Agenda

- Introduction
- Broadband transmission line theory
  - Modal superposition
  - S-parameters of t-line segment
- Signal degradation factors
  - Conductor effects
  - Dielectric effects
- RLGC parameters extraction technologies
  - 2D static field solvers
  - 2D magneto-static field solvers
  - 3D full-wave solvers
- Examples of broadband parameters extraction
- Conclusion

Effect of skin-effect in thin plane on a PCB differential microstrip line



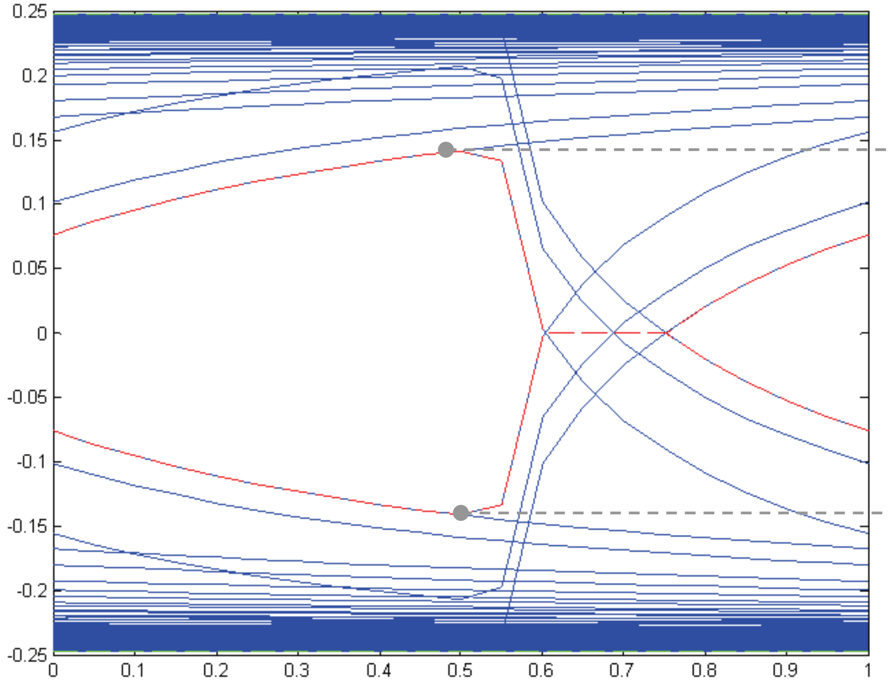




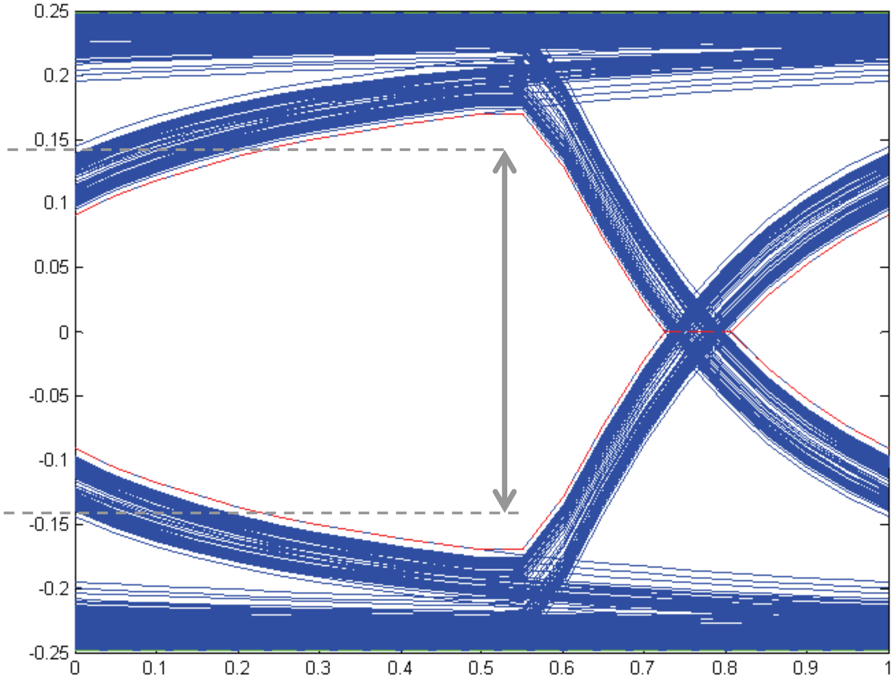
7.5 mil wide 2.2 mil thick strips 20 mil apart. Dielectric substrate with Dk=4.1 and LT=0.02 at 1 GHz. Substrate thickness 4.5 mil, plane thickness 0.594 mil, metal surface roughness 0.5 um

Eye diagram comparison for 5-inch differential micro-strip line segment with 20 Gbs data rate

Two 7.5 mil traces 20 mil apart on 4.5 mil dielectric and 0.6 mil plane, 0.5 um roughness.
Worst case eye diagram for 50 ps bit interval – May affect channel budget!



Worst case eye diagram computed with W-element defined with tabulated RLGC parameters extracted with Simbeor



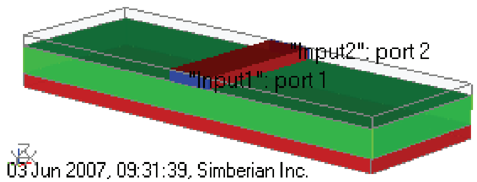
Worst case eye diagram computed with W-element defined with t-line parameters extracted with a static solver

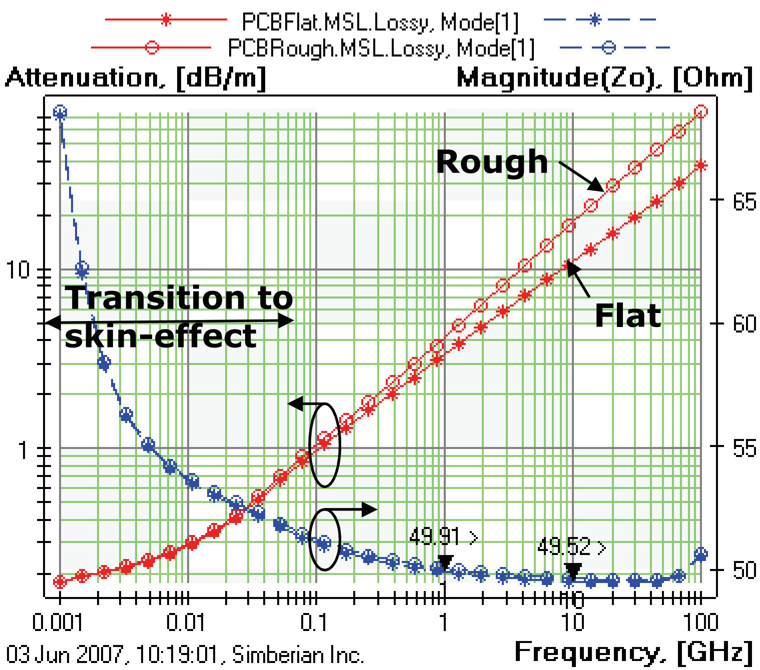
Computed by V. Dmitriev-Zdorov, Mentor Graphics

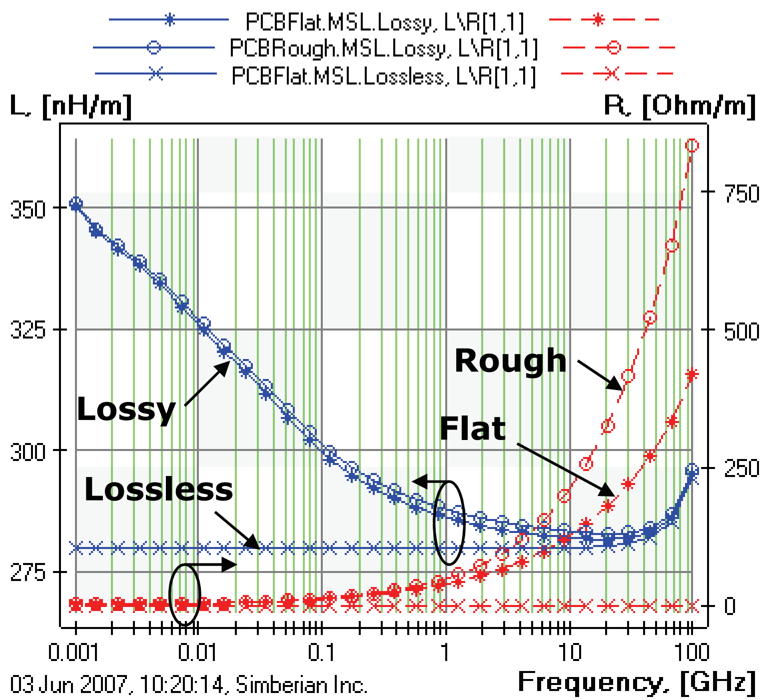
Effect of roughness on a PCB microstrip line

7 mil wide and 1.6 mil thick strip, 4 mil substrate, Dk=4,
2-mil thick plane. Strip and plane is copper.
Metal surface RMS roughness 1 um, rms roughness factor 2
No dielectric losses

25 % loss increase at 1 GHz and 65% at 10 GHz



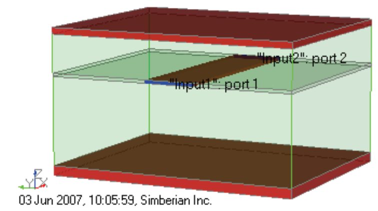


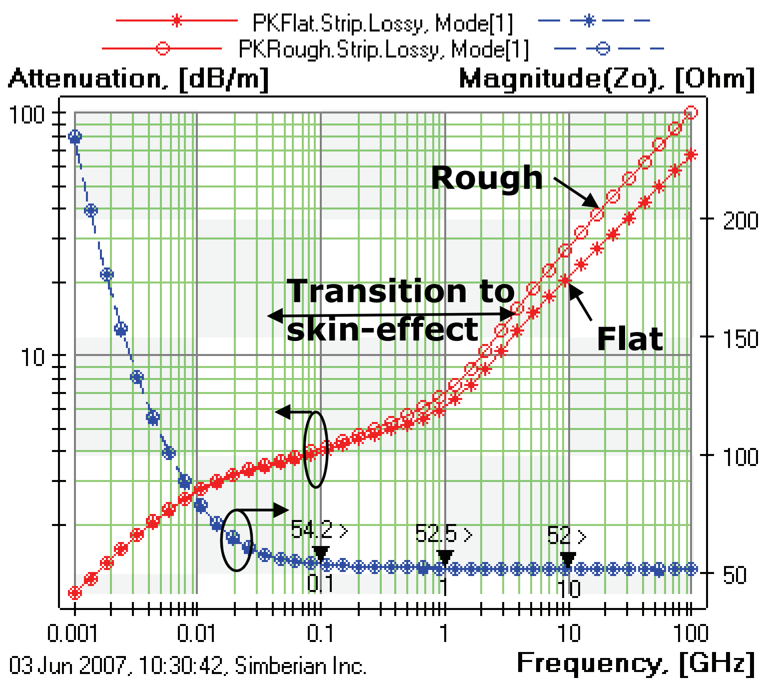


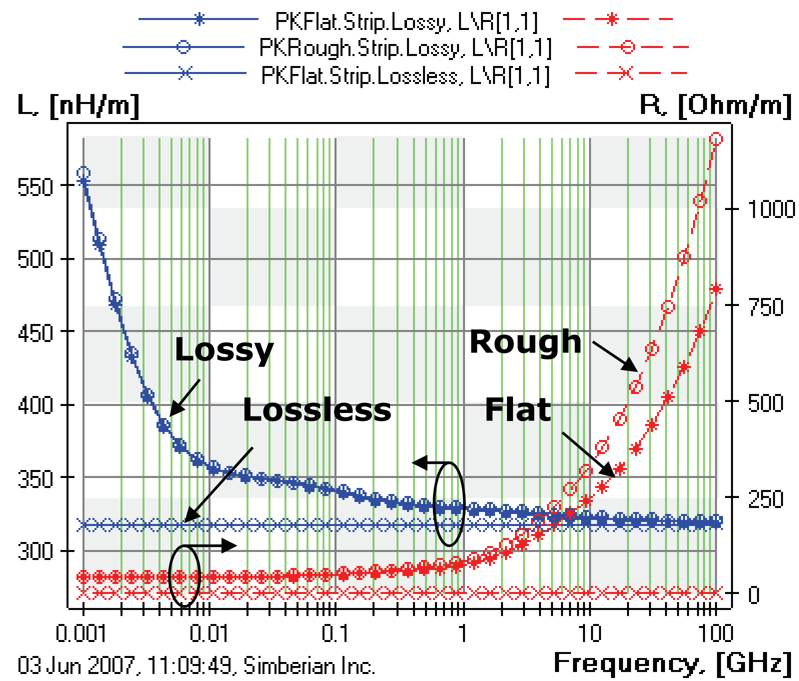
Transition to skin-effect and roughness in a package strip-line

79 um wide and 5 um thick strip in dielectric with Dk=3.4. Distance from strip to the top plane 60 um, to the bottom plane 138 um. Top plane thickness is 10 um, bottom 15 um. RMS roughness is 1 um on bottom surface and almost flat on top surface of strip, RMS roughness factor is 2.

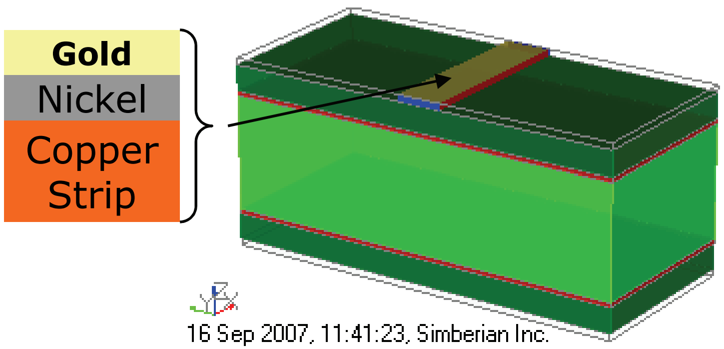
33% loss increase at 10 GHz



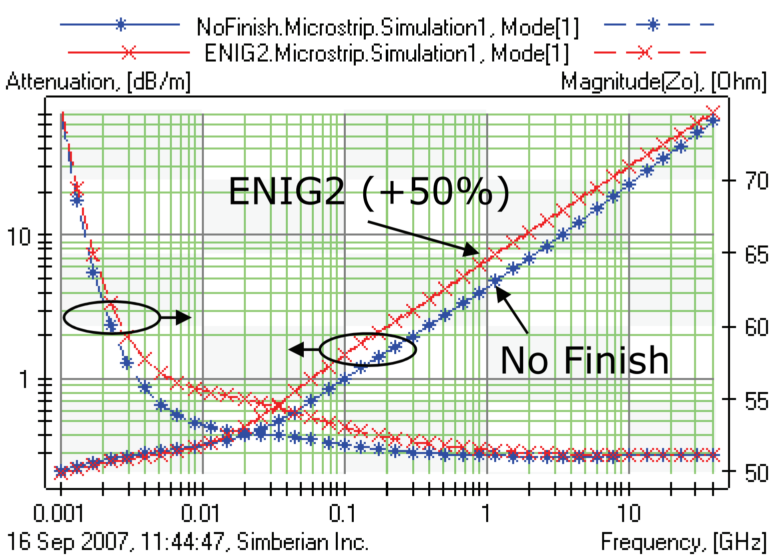


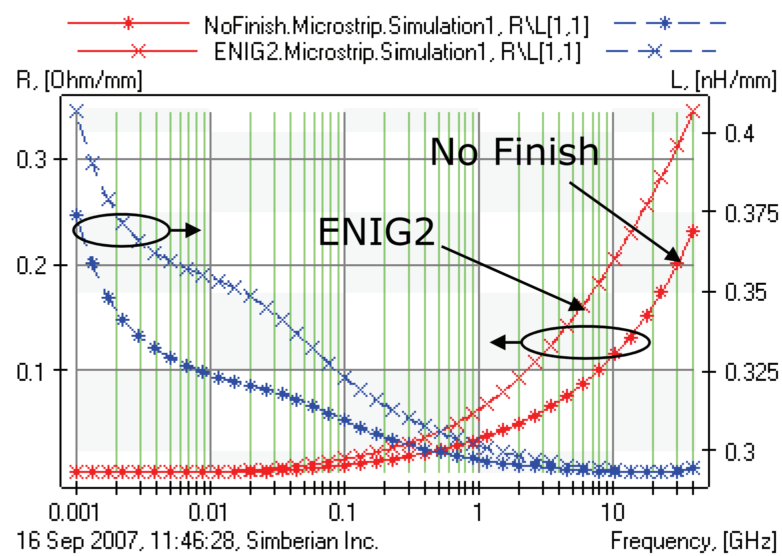


Effect of metal surface finish on a PCB microstrip line parameters

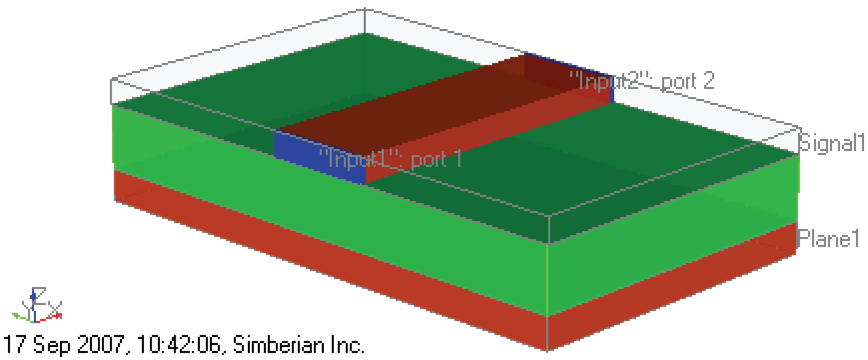


NoFinish – 8 mil microstrip on 4.5 mil dielectric with Dk=4.2, LT=0.02 at 1 GHz.
ENIG2 - microstrip surface is finished with 6 um layer of Nickel and 0.1 um layer of gold on top
Nickel resistivity is 4.5 of copper, mu is 10





Effect of dielectric models on a PCB microstrip line parameters

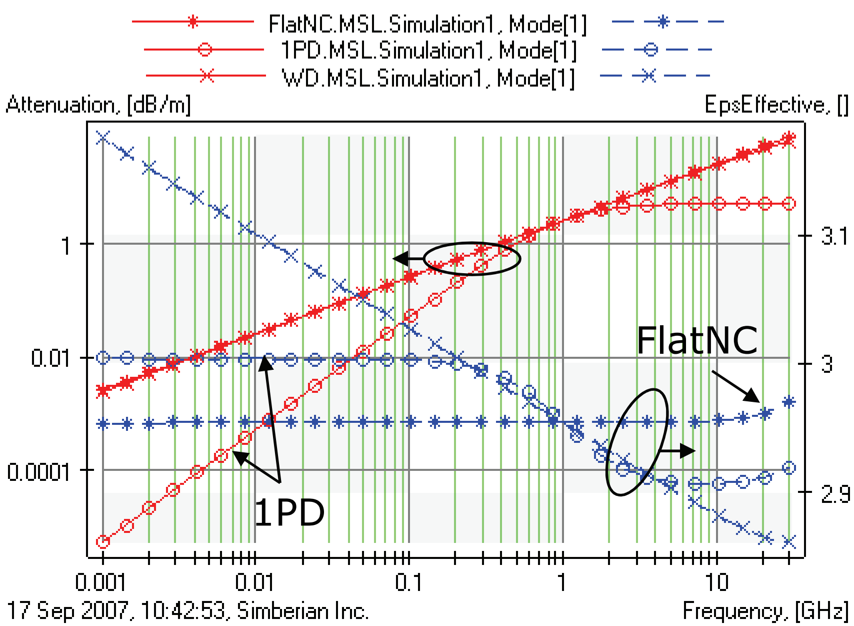


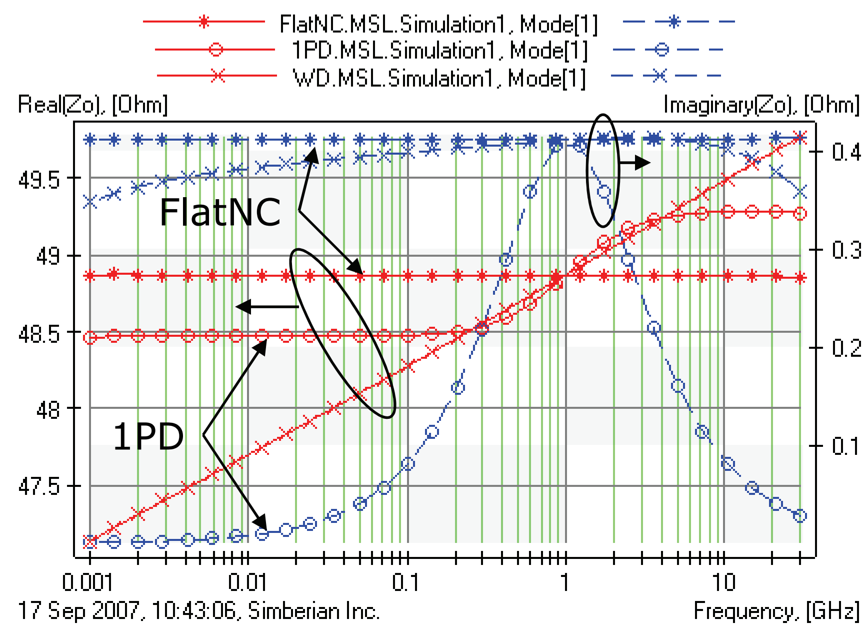
FlatNC – 7 mil microstrip on 4.0 mil dielectric with Dk=4.2, LT=0.02 and without dispersion

1PD – same line with Dk=4.2, LT=0.02 at 1 GHz and 1-pole Debye dispersion model

WD – same line with Dk=4.2, LT=0.02 at 1 GHz and wideband Debye dispersion model

No metal losses to highlight the effect





Conclusion – Select the right tool to build broadband transmission line models

- Use broadband and causal dielectric models
- Simulate transition to skin-effect, shape and proximity effects at medium frequencies
- Account for skin-effect, dispersion and edge effect at high frequencies
- Have conductor models valid and causal over 5-6 frequency decades in general
- Account for conductor surface roughness and finish
- Automatically extract frequency-dependent modal and RLGC matrix parameters per unit length for W-element models of multiconductor lines

Author: Yuriy Shlepnev

- Contact
  - E-mail shlepnev@simberian.com
  - Tel. +1-206-409-2368
  - Skype *shlepnev*
- Biography
  - Yuriy Shlepnev is the president and founder of Simberian Inc., were he develops electromagnetic software for electronic design automation. He received M.S. degree in radio engineering from Novosibirsk State Technical University in 1983, and the Ph.D. degree in computational electromagnetics from Siberian State University of Telecommunications and Informatics in 1990. He was principal developer of a planar 3D electromagnetic simulator for Eagleware Corporation. From 2000 to 2006 he was a principal engineer at Mentor Graphics Corporation, where he was leading the development of electromagnetic software for simulation of high-speed digital circuits. His scientific interests include development of broadband electromagnetic methods for signal and power integrity problems. The results of his research published in multiple papers and conference proceedings.