

# Building advanced via-hole models for analysis of PCB interconnects



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# Agenda

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- Introduction
- Multiport parameters
- Electromagnetic analysis of via-holes
- Analysis of single via-holes
- Analysis of differential via-holes
- Examples of differential via-hole optimization
- Conclusion

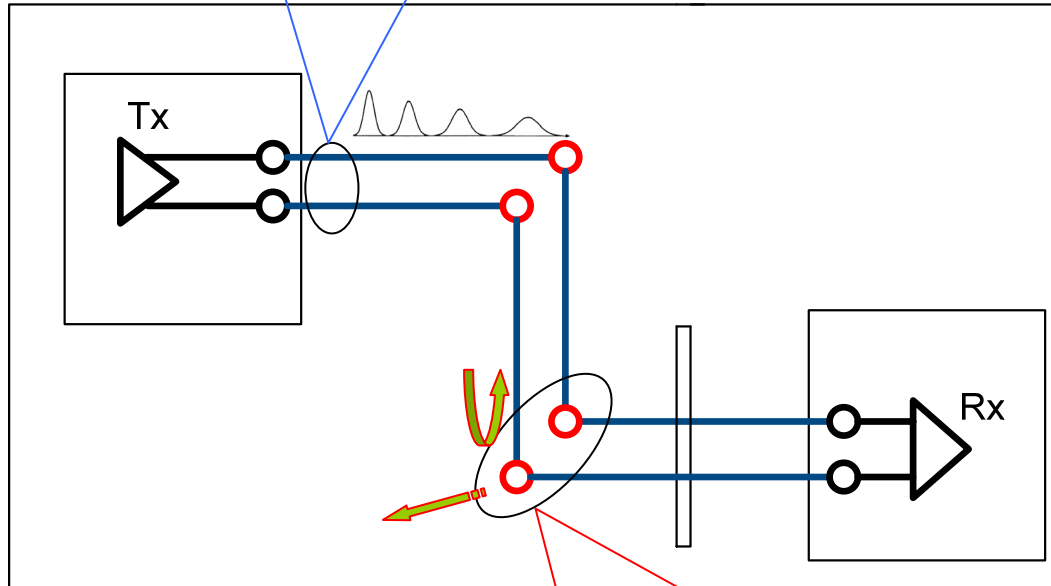
# Introduction

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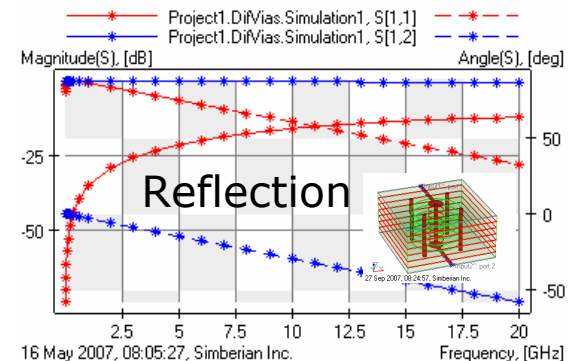
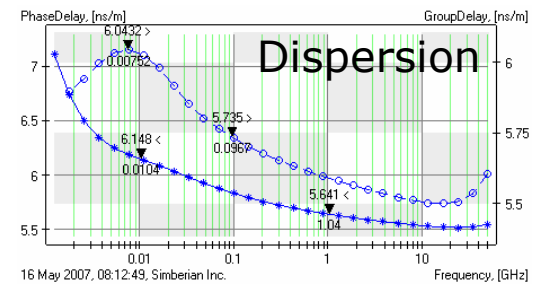
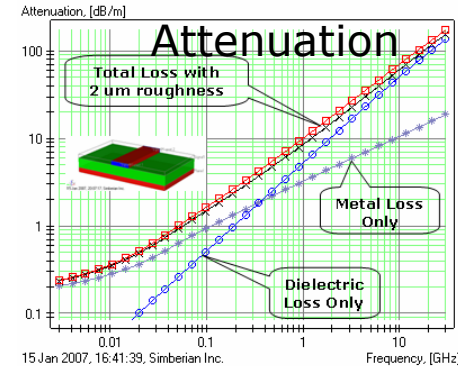
- Faster data rates drive the need for accurate electromagnetic models for multi-gigabit data channels
- Without the electromagnetic models, a channel design may require
  - Test boards, experimental verification, ...
  - Multiple iterations to improve performance
- No models or simplified static models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

# Signal degradation factors in a channel

Transmission lines:  
Attenuation and Dispersion



Via-hole transitions are the major contributors to signal degradation: Reflection, Radiation and Coupling to Power Distribution Networks (PDNs)



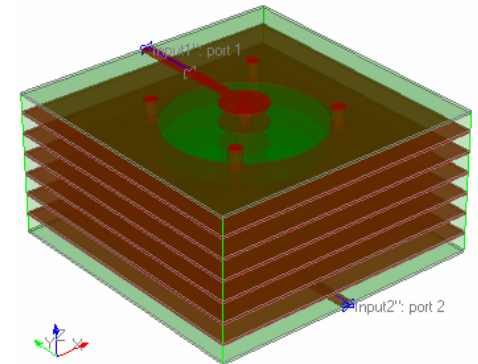
# Via-holes

## □ Via-hole types

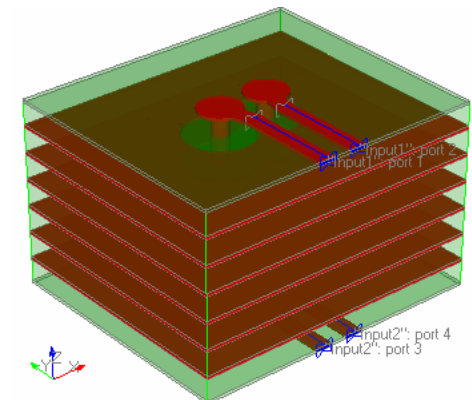
- Single-ended and differential
- Through, blind, buried and back-drilled vias
- Signal vias and stitching vias
- Micro-vias
- ...

## □ Common traits

- There are no simple models for analysis of multi-gigabit signal propagation (especially if vias are going through multiple parallel planes)
- Electromagnetic analysis is required
- All types of vias can be characterized with multiport parameters



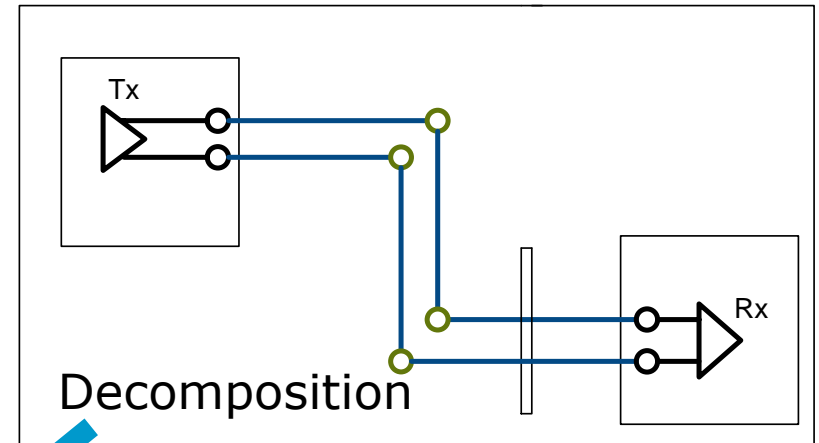
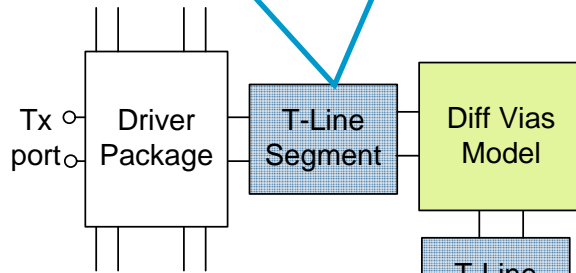
27 Sep 2007, 06:37:29, Simberian Inc.



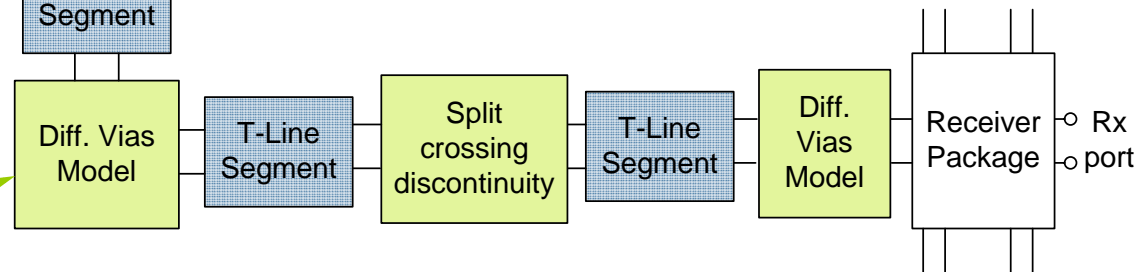
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# De-compositional analysis of a channel

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



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



*Via-hole S-parameter or equivalent distributed models are required for successful analysis of multi-gigabit channels!*

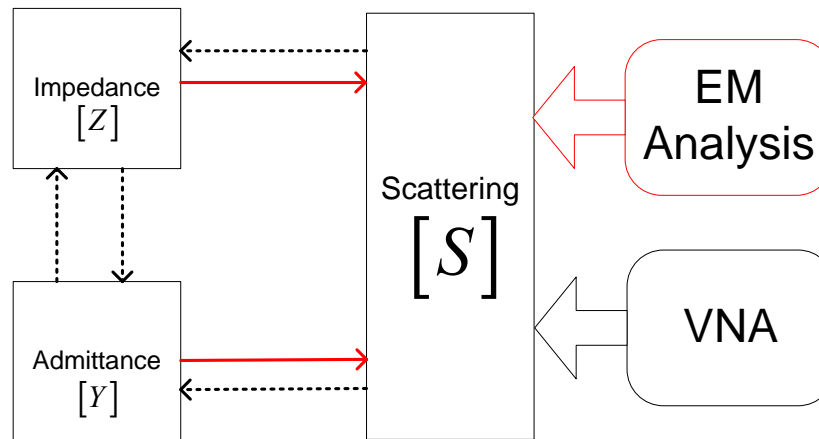
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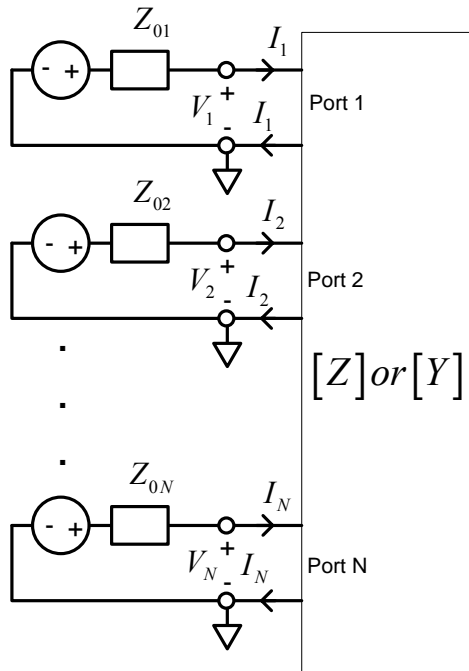
# Multiport parameters

- ❑ Multiport is a natural and scalable black-box description of linear structures comparable with or larger than wavelength
- ❑ Multiport parameters of via-holes and other components are usually united and simulated with non-linear drivers and receivers in time domain
- ❑ Multiport parameters are output of electromagnetic simulators as well as Vector Network Analyzers (VNA)





# Impedance and admittance parameters



Equivalent currents and voltages at ports:  $\bar{I} \in C^{N \times 1}$ ,  $\bar{V} \in C^{N \times 1}$

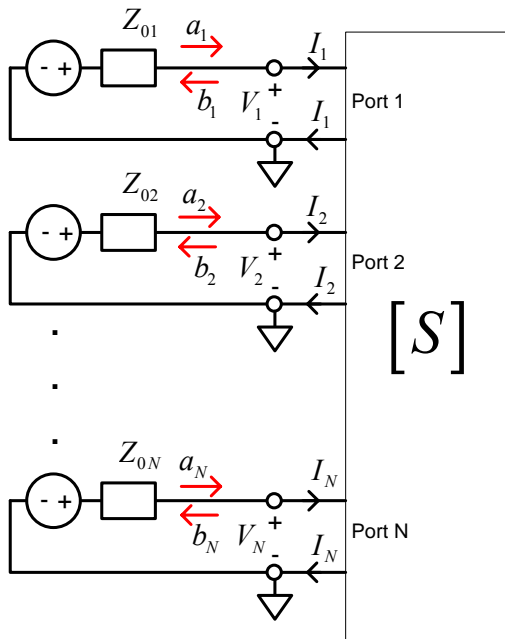
Impedance parameters:  $\bar{V} = Z \cdot \bar{I}$ ,  $Z \in C^{N \times N}$ ,  $Z_{i,j} = \left. \frac{V_i}{I_j} \right|_{I_k=0 \ k \neq j}$

Admittance parameters:  $\bar{I} = Y \cdot \bar{V}$ ,  $Y \in C^{N \times N}$ ,  $Y_{i,j} = \left. \frac{I_i}{V_j} \right|_{V_k=0 \ k \neq j}$

Conversion:  $Y = Z^{-1}$ ,  $Z = Y^{-1}$

- Imittance parameters are convenient for circuit analysis
- Matrix elements may have large dynamic range and singularities
- Can not be measured directly at high frequencies when structure size is comparable with wavelength

# Scattering parameters definition



Incident and reflected waves:

$$\bar{a} = \frac{1}{2\sqrt{Z_0}}(\bar{V} + Z_0 \cdot \bar{I}), \quad \bar{b} = \frac{1}{2\sqrt{Z_0}}(\bar{V} - Z_0 \cdot \bar{I}), \quad \bar{a}, \bar{b} \in C^{N \times 1}$$

$$Z_0 = \text{diag}\{Z_{0i}, i = 1, \dots, N\} \in C^{N \times N}$$

Scattering parameters:

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

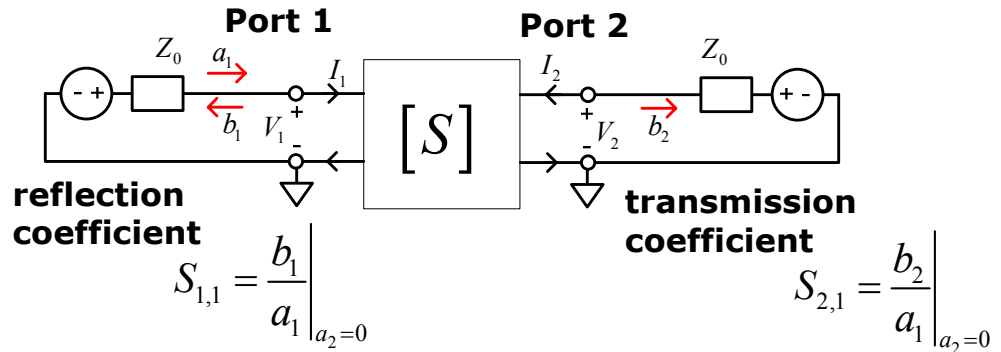
$$S = (U - Y_N) \cdot (U + Y_N)^{-1}, \quad Y_N = Z_0^{1/2} \cdot Y \cdot Z_0^{1/2}$$

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

- S-parameters can be directly extracted from electromagnetic simulation or measured at any frequency including DC
- S-parameters are free from singularities
- S-parameters can be directly used in a SPICE simulators

# S-parameters definitions for 2-port model

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



$$V_i^+ = \sqrt{Z_0} \cdot a_i \quad \text{voltage of incident wave}$$

$$V_i^- = \sqrt{Z_0} \cdot b_i \quad \text{voltage of reflected wave}$$

$$V_i = V_i^+ + V_i^- \quad \text{total voltage}$$

$$I_i = \frac{1}{Z_0} (V_i^+ + V_i^-) \quad \text{total current}$$

$$|S_{i,j}| = \sqrt{\text{Re}(S_{i,j})^2 + \text{Im}(S_{i,j})^2} \quad \text{magnitude}$$

$$|S_{i,j}|_{dB} = 20 \cdot \log(|S_{i,j}|) \quad \text{magnitude in dB}$$

$$P_i^+ = |a_i|^2 \quad \text{power of incident wave}$$

$$P_i^- = |b_i|^2 \quad \text{power of reflected wave}$$

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

$$\angle S_{i,j} = \arctan(\text{Im}(S_{i,j})/\text{Re}(S_{i,j})) \quad \text{phase}$$

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

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# Via-hole modeling

3D Static and Magneto-Quasi-Static Solvers

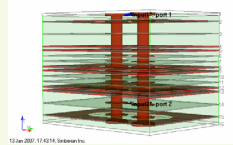
$$\vec{E} = -\nabla\phi - j\omega\vec{A}$$

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

3D Full-Wave Solver

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

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

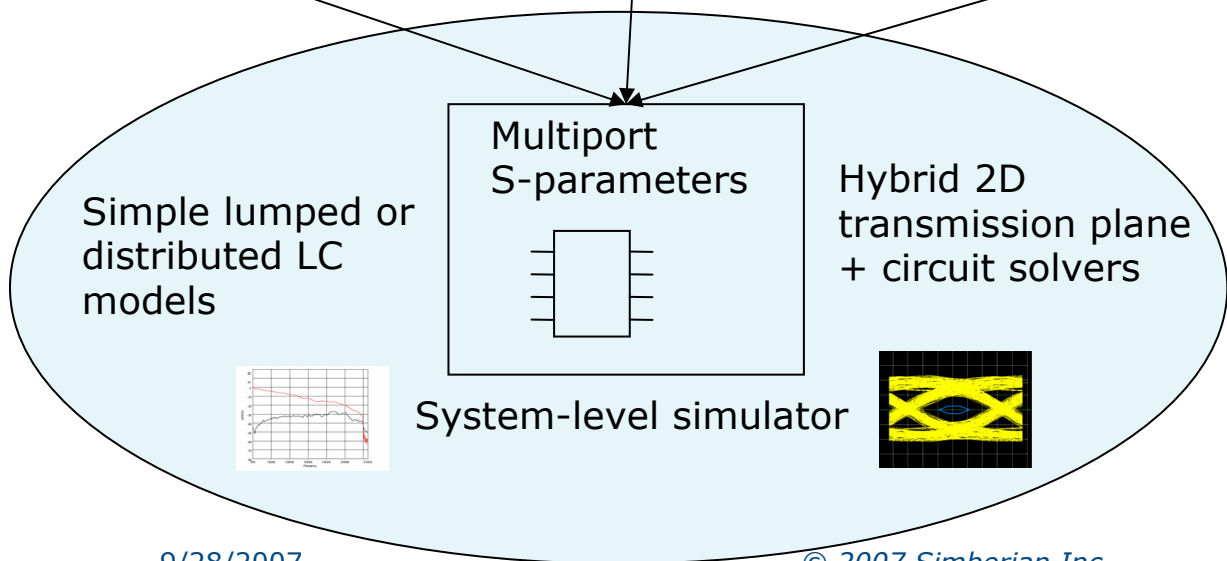


Transmission Plane Solvers

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

$$\frac{\partial V}{\partial x} = -Z_{\square}(\omega)\cdot J_{sx}$$

$$\frac{\partial V}{\partial y} = -Z_{\square}(\omega)\cdot J_{sy}$$



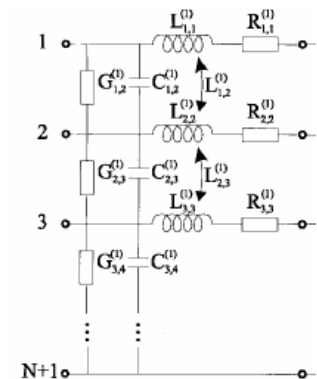
# 3D static or magneto-quasi-static solvers

- The original vector boundary-value problems are reduced to potential problems (no displacement current, no retardation)
  - Static scalar problems are formulated to extract C, G
  - Magneto-static vector problem is formulated to extract frequency-dependent L and R
  - Mixed potential integral equation formulations can be used to extract RLCG simultaneously (Partial Electrical Equivalent Circuit, PEEC)
- The final RLGC or similar circuit can be transformed into S-parameters
- For accurate results, the structure can be electrically long along the transmission direction, but must be considerably smaller than wavelength in the plane perpendicular to the transmission direction

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

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

**Plus boundary conditions**



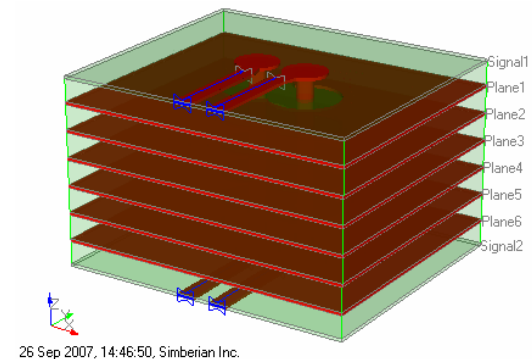
# 3D electromagnetic analysis in frequency-domain

- Time-harmonic or steady-state Maxwell's partial differential equations (PDEs) or integral equations (IEs) derived from them using Green's functions
- PDEs are usually solved with Finite Element Method (FEM)
  - Volume meshing – simple algorithms
- IEs are usually solved with Method of Moments (MoM)
  - Surface metal meshing – more accurate for multilayered problems
  - May be inefficient if multiple planes have to be meshed
- Hybrid techniques such as Method of Lines
  - Combines advantages of FEM and MoM (simple and accurate) and does not mesh planes
- All frequency-dependent dispersion and loss phenomena can be accounted naturally
- **The via-hole analysis problem has to be almost always truncated or localized with some boundary conditions**

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

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

**Plus boundary conditions**



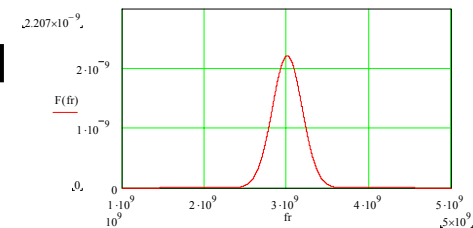
# 3D electromagnetic analysis in time-domain

- Direct volumetric discretization of partial differential equations (PDEs) is usually used
  - Finite Difference Time Domain (FDTD) Method
  - Finite Integration Technique (FIT)
  - Transmission Line Matrix (TLM) Method
- Combined simulation with lumped elements and transmission line segments is possible
  - Basis of some hybrid distributed simulators implementing co-simulation with IBIS, SPICE and transmission line models
- Output is a transient response of the system excited with lumped sources or through wave-ports
  - Discrete Fourier Transform (DFT) can be used to obtain frequency domain parameters (S-matrices)
- **Difficult to account frequency-dependent loss and dispersion effects**

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

$$\nabla \times \vec{H} = \varepsilon \frac{\partial \vec{E}}{\partial t} + \sigma \vec{E} + \vec{J}$$

**Plus boundary conditions**



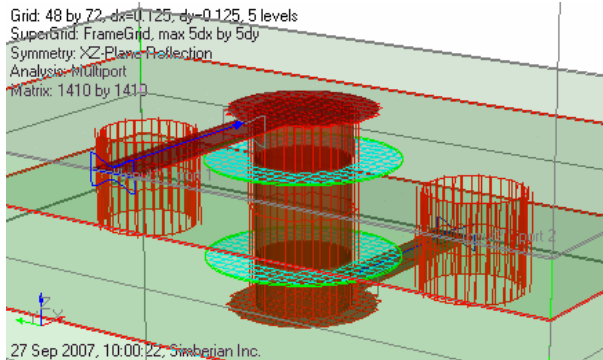
$$\nabla \times \vec{E}(\vec{r}, t) = -z_{\delta}(\vec{r}, t) * \vec{H}(\vec{r}, t)$$

$$\nabla \times \vec{H}(\vec{r}, t) = y_{\delta}(\vec{r}, t) * \vec{E}(\vec{r}, t) + \vec{J}$$



# Simbeor solver combines advantages of MoM and FEM methods

Solves Maxwell's equations to find S-parameters of via-holes and other discontinuities:



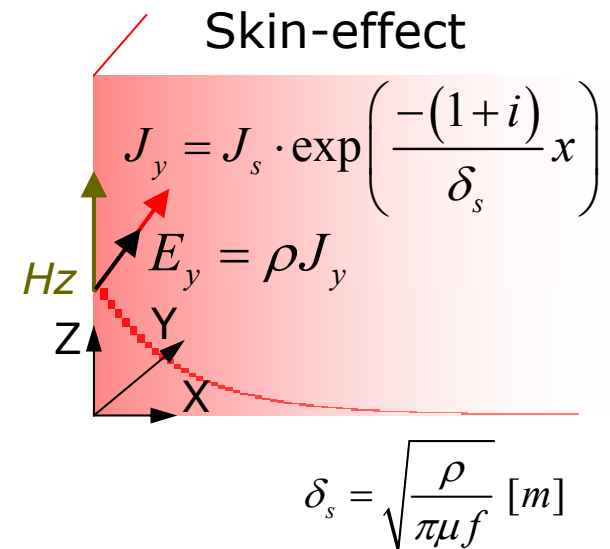
$$\begin{aligned}\nabla \times \vec{E} &= -i\omega\mu\vec{H} \\ \nabla \times \vec{H} &= i\omega\varepsilon\vec{E} + \sigma\vec{E} + \vec{J}\end{aligned} \quad \rightarrow \quad [S]$$

**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 precise de-embedding of extracted multiport S-parameters

# Boundary conditions for all types of electromagnetic analysis

- Reflecting boundary conditions
  - Perfect Electric Conductor (PEC)
  - Perfect Magnetic Conductor (PMC)
  - Surface Impedance Boundary Conditions (SIBC)
- Non-reflecting boundary conditions
  - Absorbing Boundary Conditions (ABC)
  - Perfectly Matched Layer (PML)
- Excitation boundary conditions
  - Lumped or component ports
  - Wave-ports or transmission line ports

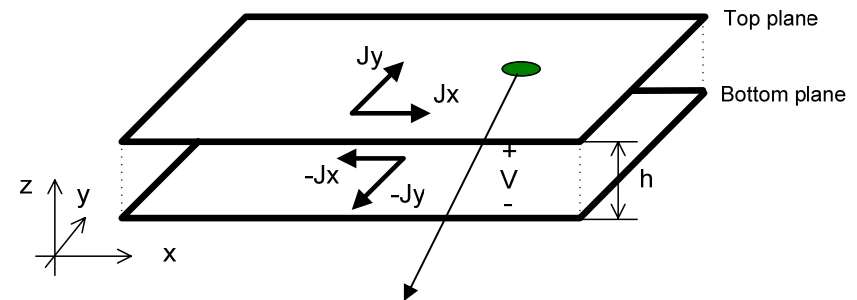


# Transmission plane model as component of a system model for via-holes

2D Telegrapher's equations in frequency domain

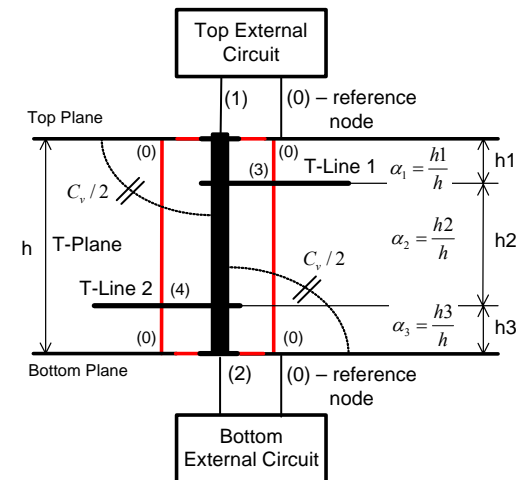
$$\left. \begin{aligned} \frac{\partial J_{sx}}{\partial x} + \frac{\partial J_{sy}}{\partial y} &= -Y_{\square}(\omega) \cdot V + J_z \\ \frac{\partial V}{\partial x} &= -Z_{\square}(\omega) \cdot J_{sx} \\ \frac{\partial V}{\partial y} &= -Z_{\square}(\omega) \cdot J_{sy} \end{aligned} \right\} (x, y) \in \Omega$$

Plus boundary conditions



2D Telegrapher's equations in time domain

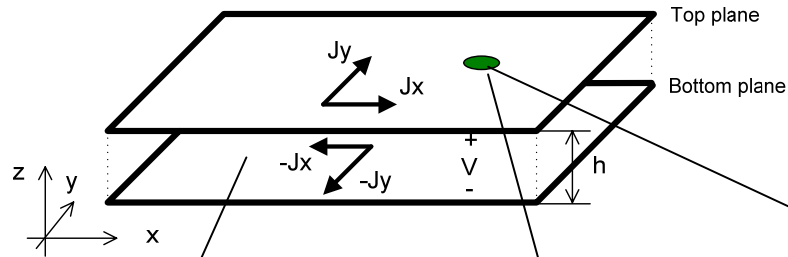
$$\left. \begin{aligned} \frac{\partial J_{sx}(t)}{\partial x} + \frac{\partial J_{sy}(t)}{\partial y} &= -Y_{\square\delta}(t) * V(t) + J_z(t) \\ \frac{\partial V(t)}{\partial x} &= -Z_{\square\delta}(t) * J_{sx}(t) \\ \frac{\partial V(t)}{\partial y} &= -Z_{\square\delta}(t) * J_{sy}(t) \end{aligned} \right\} (x, y) \in \Omega$$



Impedance and admittance per square for broadband models are derived in

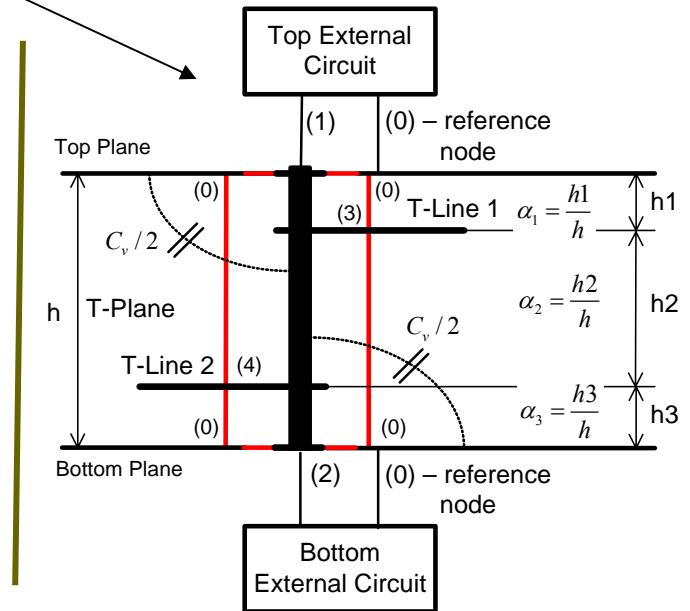
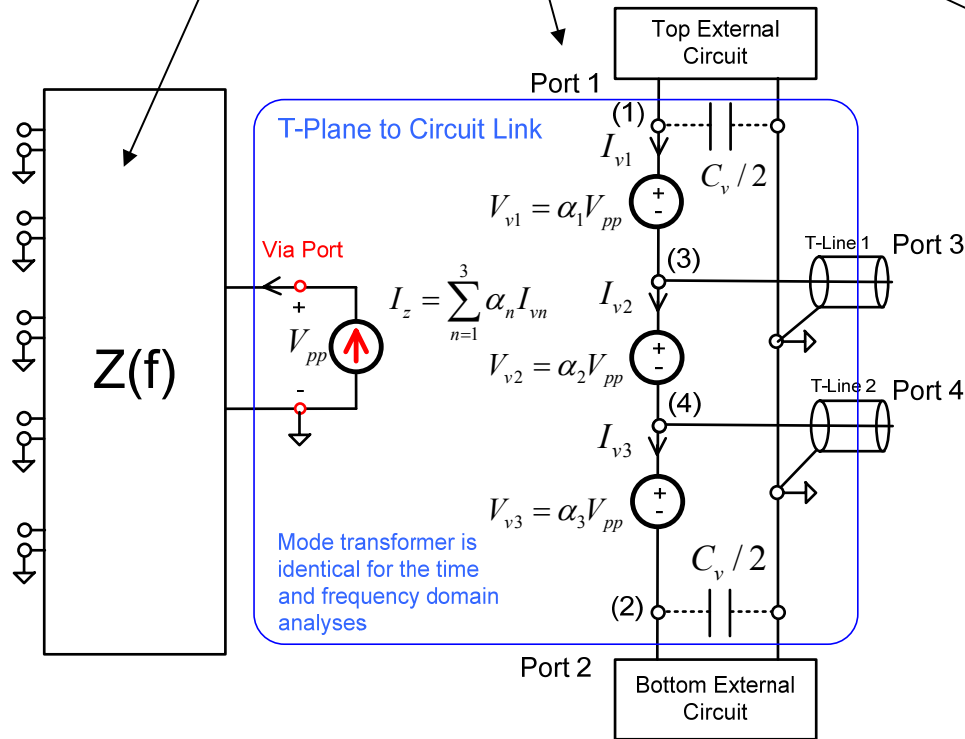
Y. Shlepnev, Transmission plane models for parallel-plane power distribution system and signal integrity analysis, - 22nd Annual Review of Progress in Applied Computational Electromagnetics, 2006, p. 382-389.

# Hybrid distributed via-hole model with signal coupled to transmission plane model



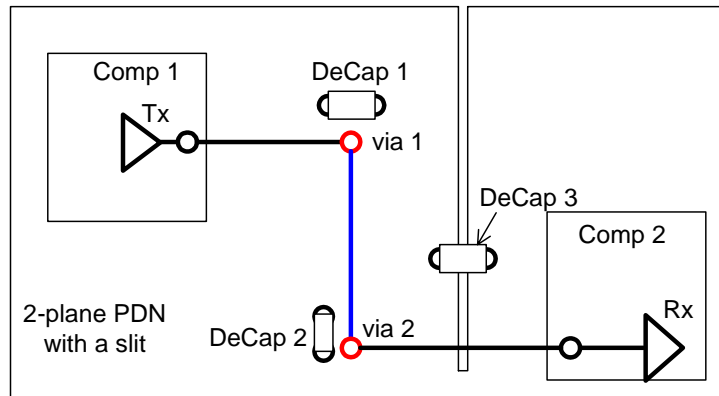
Impedance matrix  $Z(f)$  contains information about the geometry of the planes

T-Plane to Circuit Link is a modal transformer between plane and t-line modes

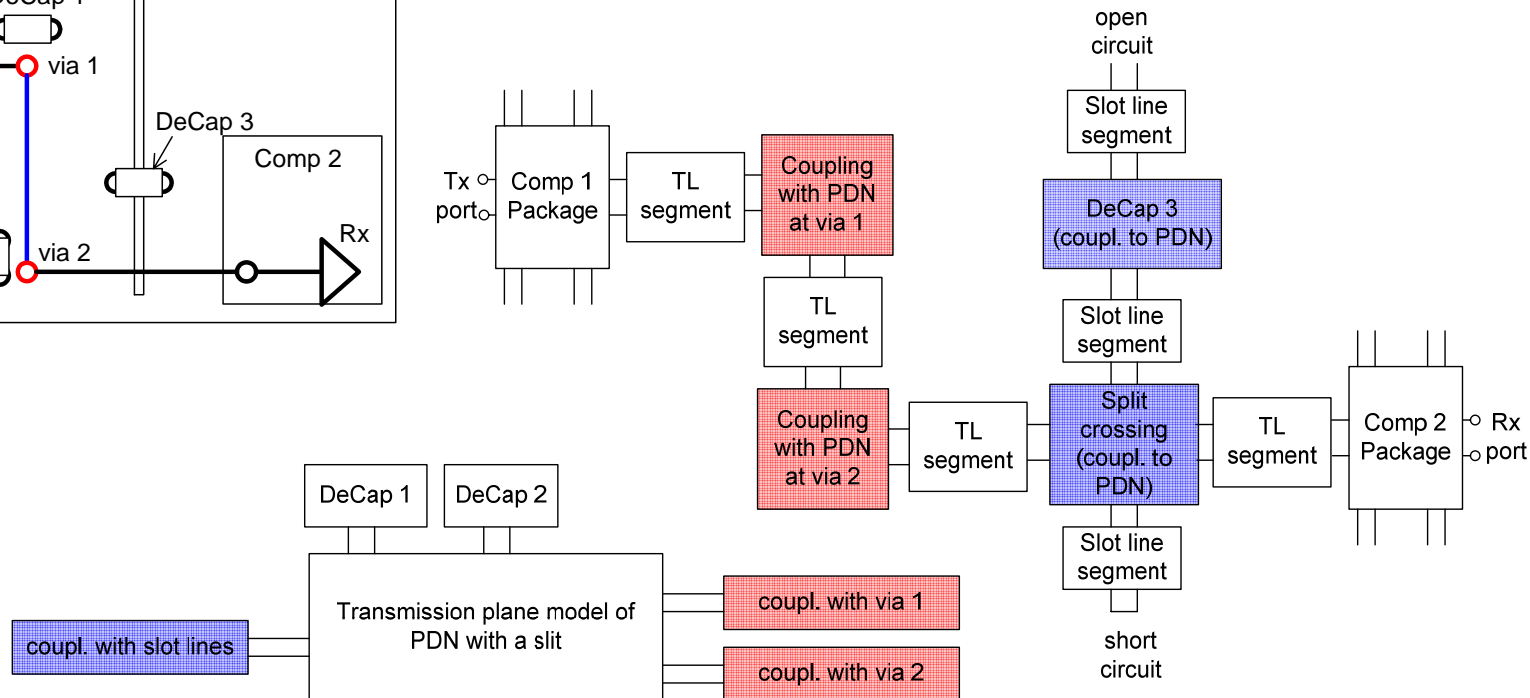


# Hybrid-distributed analysis of circuits coupled to parallel-planes

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



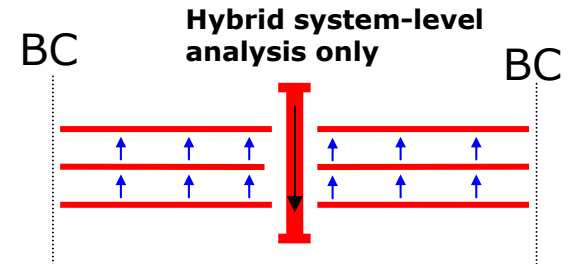
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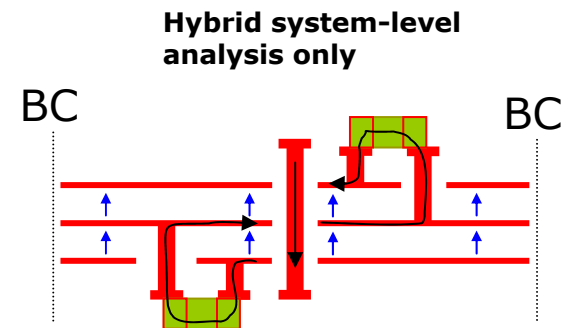
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# Analysis of single vias going through multiple parallel planes

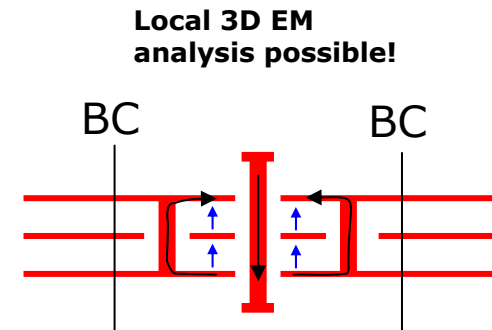
- Planes are not terminated and the return current is the “displacement” current between the planes
  - The problem is **non-localizable** – requires analysis of the whole board



- 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 (**localizable**) and solved with any boundary conditions



# How to model the non-localizable cases (without stitching vias)

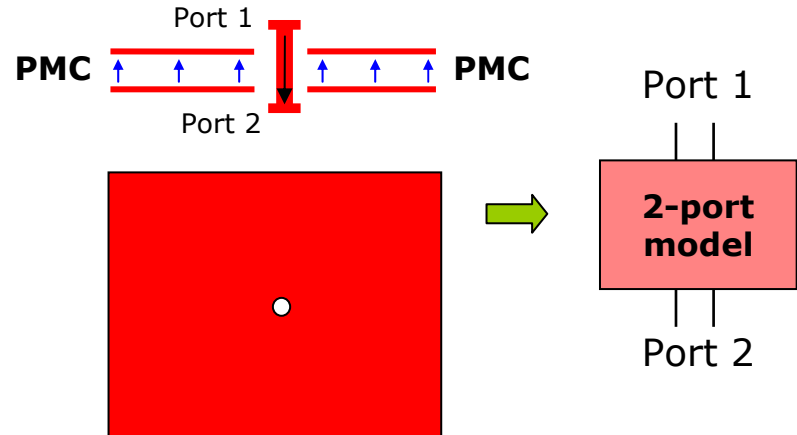
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- S-parameter models become dependent on the simulation area and boundary conditions
  - Independence of the boundary conditions indicates that the problem is localizable
- Any type of locally enforced boundary conditions is not correct for non-localizable problem
  - PEC (electric walls) are equivalent to short-circuiting the planes at a distance from vias – preferable
  - PMC (magnetic walls) are equivalent to open-circuiting the planes at a distance from vias – incorrect low frequency asymptotic of S-parameters
  - PML (perfectly matched layer) or ABC (absorbing boundary conditions) – absorbs energy at a distance from vias
    - Not equivalent to the infinite planes (infinite planes or radial waveguides reflect energy at any location)
    - Absorbed energy is completely lost for the system level analysis (it will appear somewhere)
- Possible way to solve a non-localizable problem is either put the whole board into 3D solver or to use a hybrid simulation technique with transmission plane solver

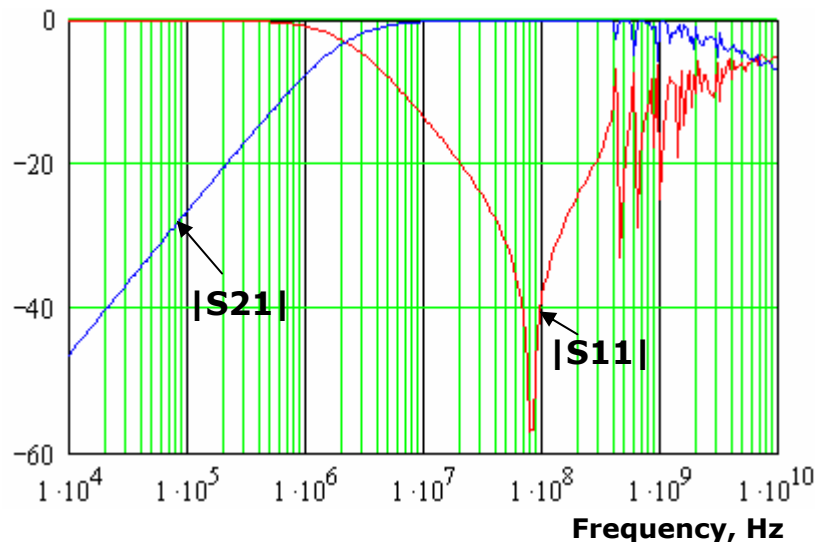


# Example of a single via through not terminated planes (widely spaced planes)

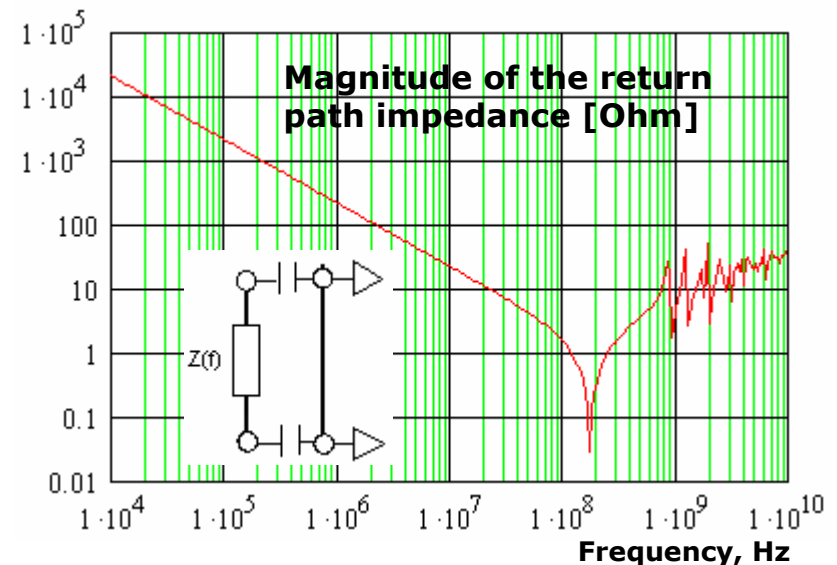
- 165 mm by 165 mm rectangular board with two metal planes
  - Planes thickness 17.8  $\mu\text{m}$ , distance between planes 1.524 mm
  - Dielectric between the planes:  $D_k=4.2$ ,  $L_t=0.018$  at 1 GHz
- Whole board analysis is required!



**S-parameters [dB]  
(50-Ohm normalized)**



9/28/2007

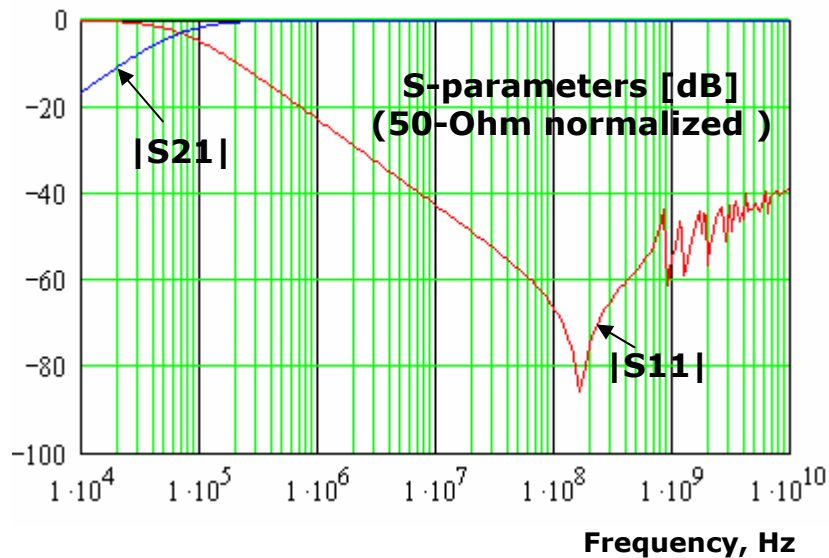
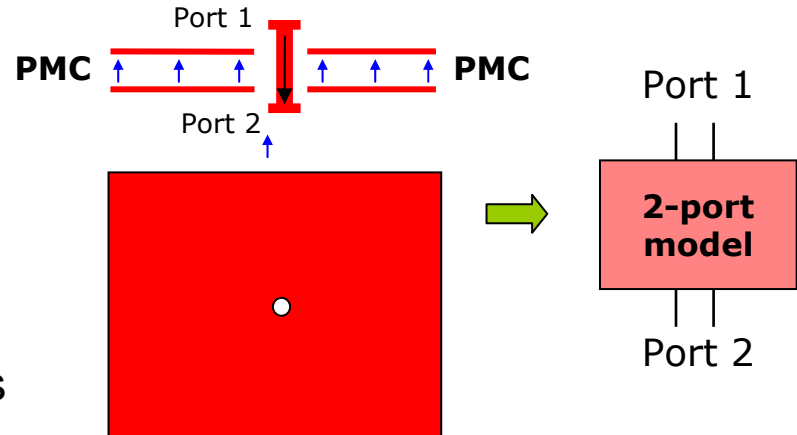


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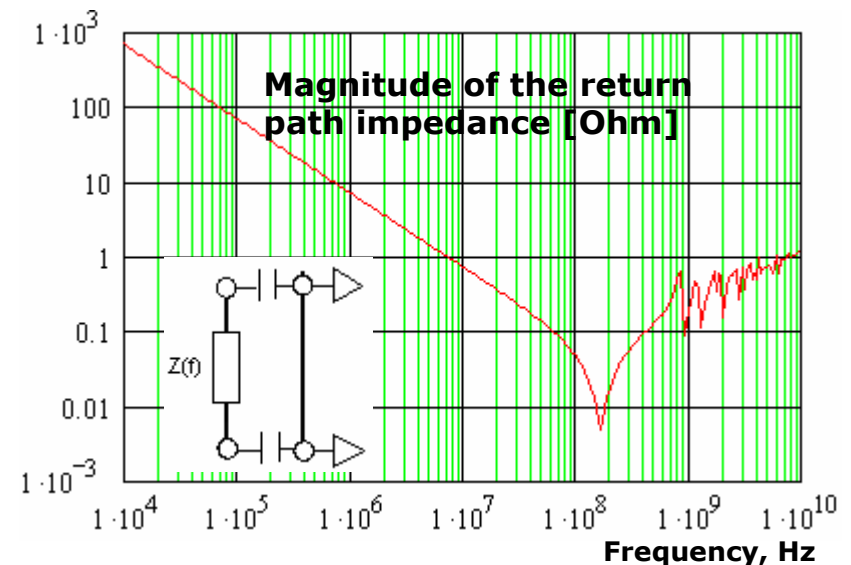
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# Example of a single via going through not terminated planes (buried capacitance)

- 165 mm by 165 mm rectangular board with two metal planes
  - Planes thickness 17.8  $\mu\text{m}$ , distance between planes 50.8  $\mu\text{m}$  (2 mil)
  - Dielectric between the planes:  $D_k=4.2$ ,  $L_t=0.018$  at 1 GHz
- Infinite plane assumption is good approximation – no system-level analysis



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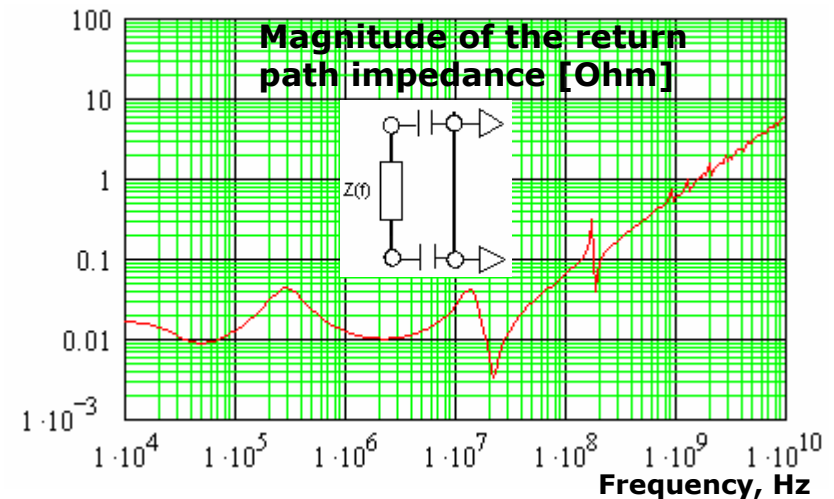
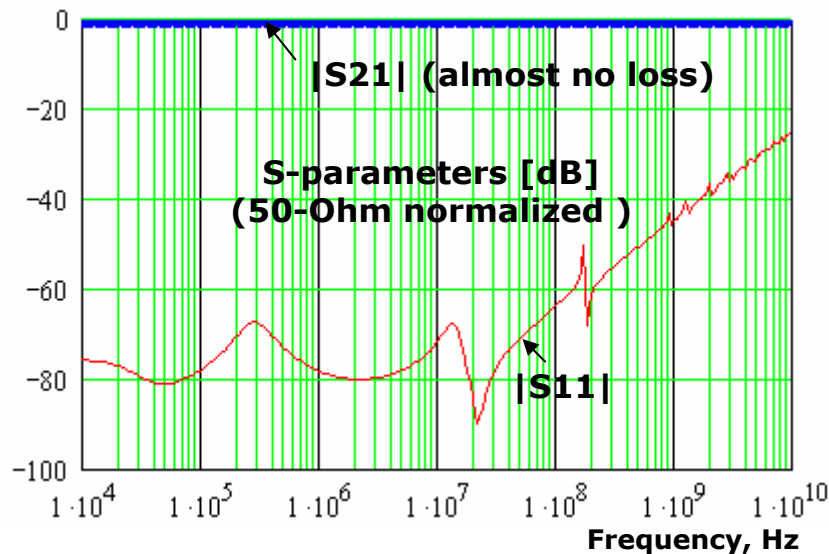
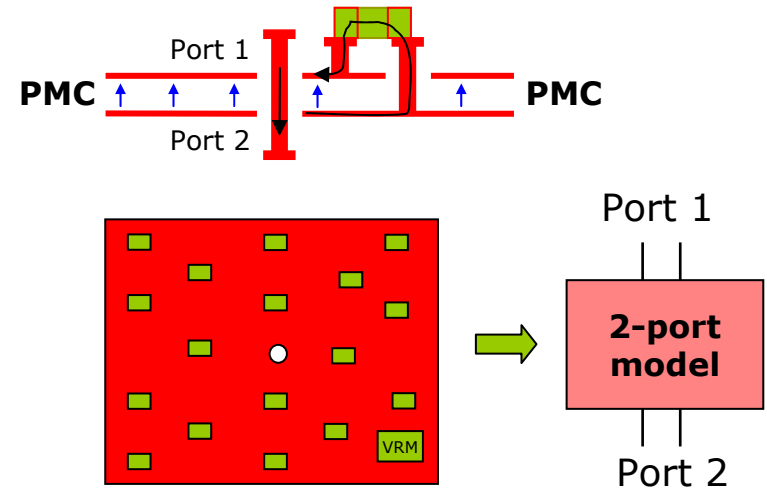


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# Example of a single via through terminated planes (widely spaced planes)

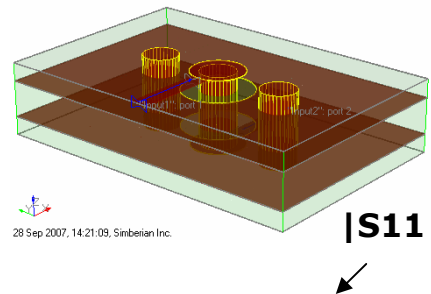
- 165 mm by 165 mm rectangular board with two metal planes
  - Planes thickness 17.8  $\mu\text{m}$ , distance between planes 1.524 mm
  - Dielectric between the planes:  $D_k=4.2$ ,  $L_t=0.018$  at 1 GHz
  - VRM, 5 bulk capacitors and 50 medium and high-frequency decoupling capacitors
- Good PDN design -> good via bypass
- Whole board analysis is required!



# Single via-holes with stitching vias

- All reference planes for connected layers have to be connected in vicinity to the via
- Only this type of vias can be locally simulated with any boundary conditions at a sufficient distance (though PEC are asymptotically correct and preferable)

Example of analysis of a single via-hole crossing two parallel planes with two stitching vias



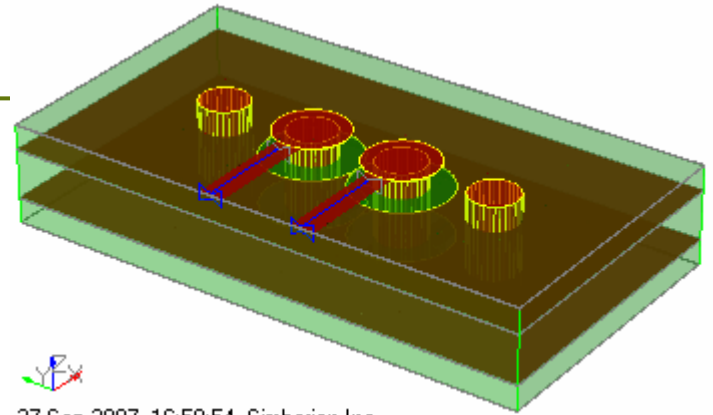
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- Examples of differential via-hole optimization
- Conclusion

# Differential via-holes

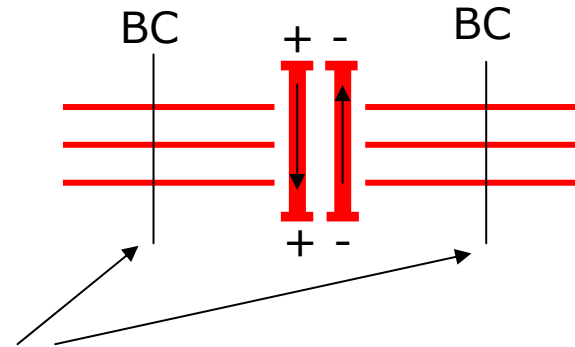
- Differential vias are two-via transitions through multiple parallel planes with possible stitching vias nearby
- Two modes propagate independently through a **symmetrical** pair
  - Differential (+-) – two vias are two conductors:  $I_d = I_1 = -I_2$ ,  $V_d = V_1 - V_2$
  - Common (++) – two vias one conductor and parallel planes with decoupling is another conductor:  $I_c = I_1 + I_2$ ,  $V_c = 0.5(V_1 + V_2)$
- Signal in differential pair always contain differential mode (useful) and may contain common mode induced by asymmetries in driver and discontinuities



27 Sep 2007, 16:50:54, Simberian Inc.

# S-parameter model for differential mode (differential model)

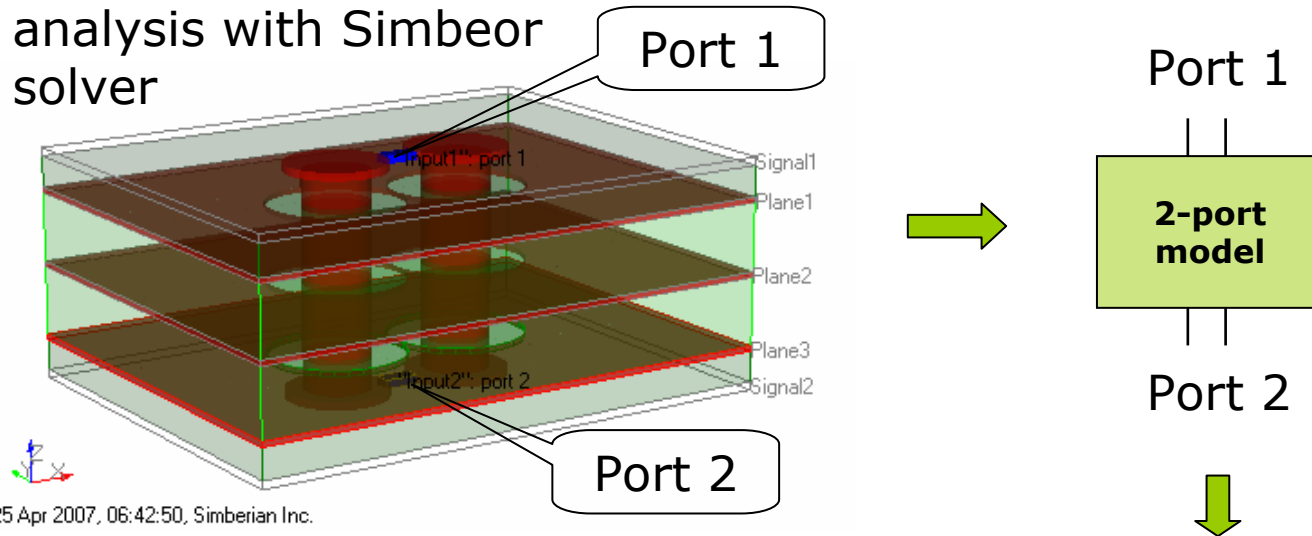
- Differential mode has two identical currents on the via barrels



- 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 larger than the largest distance between the planes to reduce the effect of sidewalls
  - In that case, the differential mode S-parameters are practically independent of the boundary conditions

# Any 3D full-wave solver can be used to generate a differential model

Differential via-holes analysis with Simbeor solver



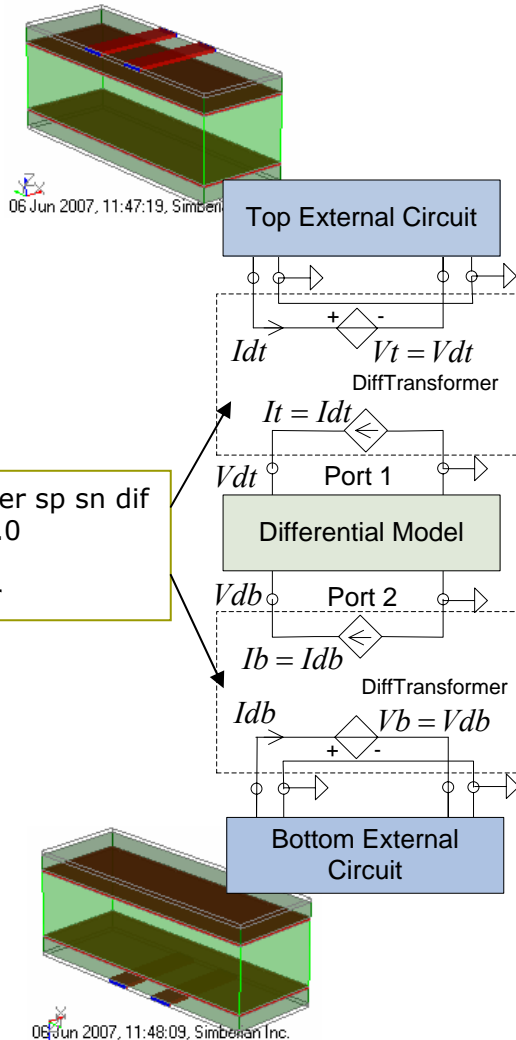
Solver generates Touchstone s2p file with tabulated scattering (S) parameters

```
# Hz S MA R 50
!Touchstone multiport model file <Project1_DifVias_Simulation1.s2p>
!Created with Simbeor 2007.05

!Frequency Hz |S[1,1]| arg(S[1,1]) |S[1,2]| arg(S[1,2]) |S[2,1]| arg(S[2,1]) |S[2,2]| arg(S[2,2])
1e+007 0.00019967204037128 80.2602696805706 0.99996135358717 -0.0323490304593697 0.999961353587184 -0.0323490304593693 0.000199673211473872 80.2582768233478
1.6681e+007 0.000322468838468293 82.6000881350992 0.999950431873029 -0.0533202384129493 0.999950431873026 -0.0533202384194636 0.000322467239270051 82.6017028071304
...
1.9e+010 0.244570464339174 35.2840265413718 0.961777164971986 -55.3104068976763 0.961777164971987 -55.3104068976763 0.244498847955767 35.2671173277359
2e+010 0.248547099289332 32.5237839600036 0.960378206732979 -58.1318674656129 0.960378206732979 -58.1318674656129 0.248469713303658 32.5058580236588
```

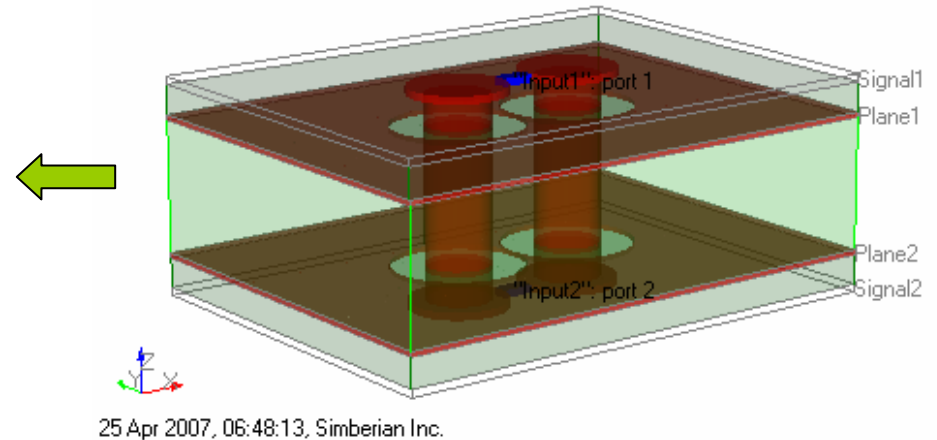


# Differential via-hole model in the system-level analysis



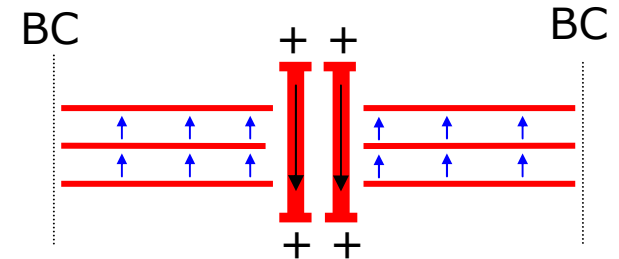
2 CCCS and 2 VCVS can be used to convert the 2-port differential model into 4-port

Common mode is open-circuited (reflected)

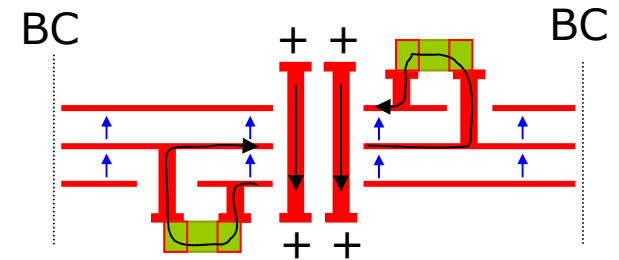


# Analysis of common mode in differential vias going through multiple parallel planes

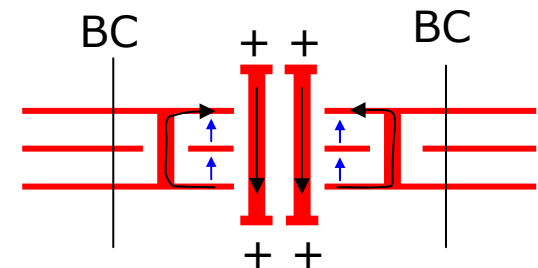
- Planes are not terminated and the return current for common mode is the “displacement” current between the planes
  - The problem is **non-localizable** – may require analysis of the whole board



- 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

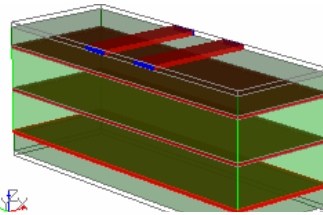


- Stitching vias are used to connect the reference planes for the connected layers and the return current is mostly conductive
  - Problem can be localized (**localizable**) and solved with any boundary conditions

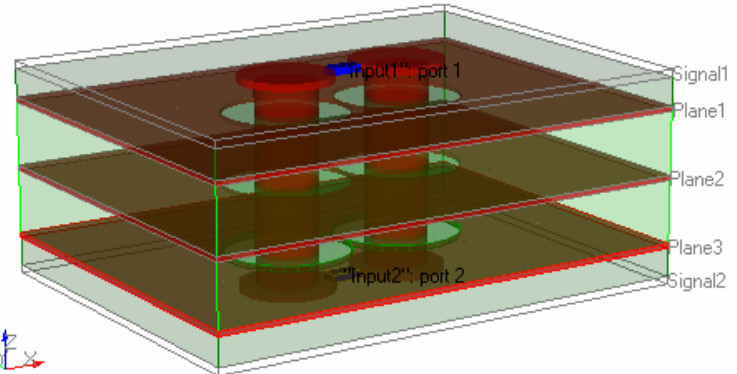
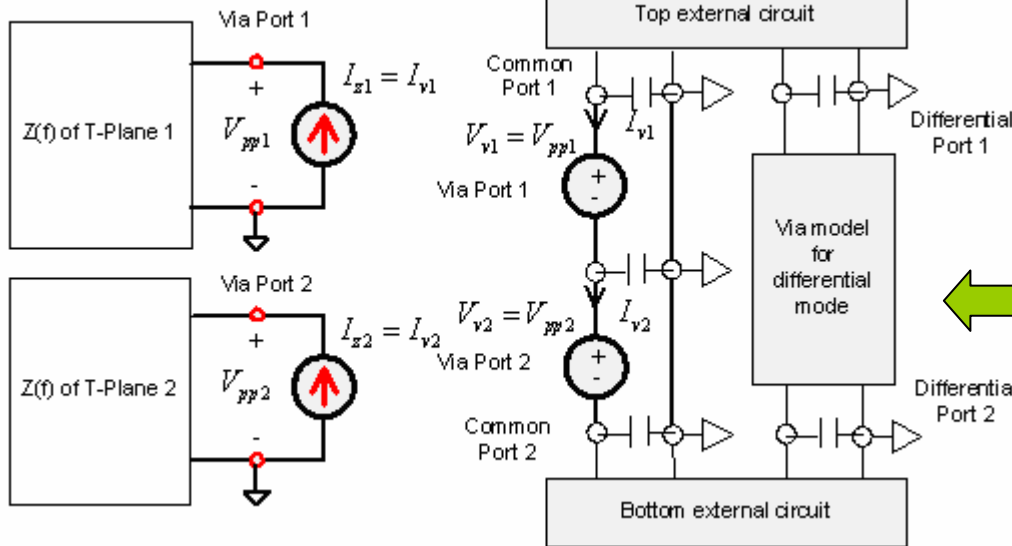


# Example of a hybrid differential via-hole model for a system-level analysis

3 planes require 2 transmission plane (t-plane) models

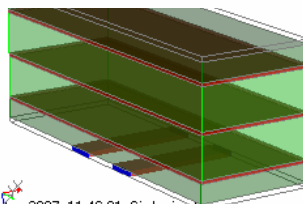


1 CCCS and 1 VCVS per plane pair can be used to link the common mode to the t-plane models



25 Apr 2007, 06:42:50, Simberian Inc.

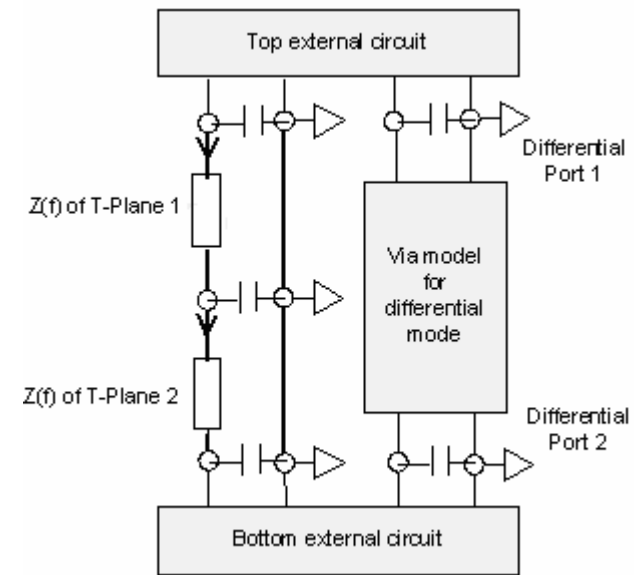
$Z(f)$  of the planes define the return path impedance for the common mode!



06 Jun 2007, 11:46:01, Simberian Inc.

# Return path $Z(f)$ for typical model cases

- ❑ PEC close to vias:  $Z(f) \sim 0$
- ❑ PMC close to vias:  $Z(f) \sim \text{infinity}$
- ❑ ABC close to vias:  $Z(f)$  is frequency-independent plane-wave impedance
- ❑ PML close to vias:  $Z(f)$  is a complex function not related to the board geometry
- ❑ Complete board with all Power Delivery Net termination:  $Z(f)$  is complex function with non-trivial frequency-dependency

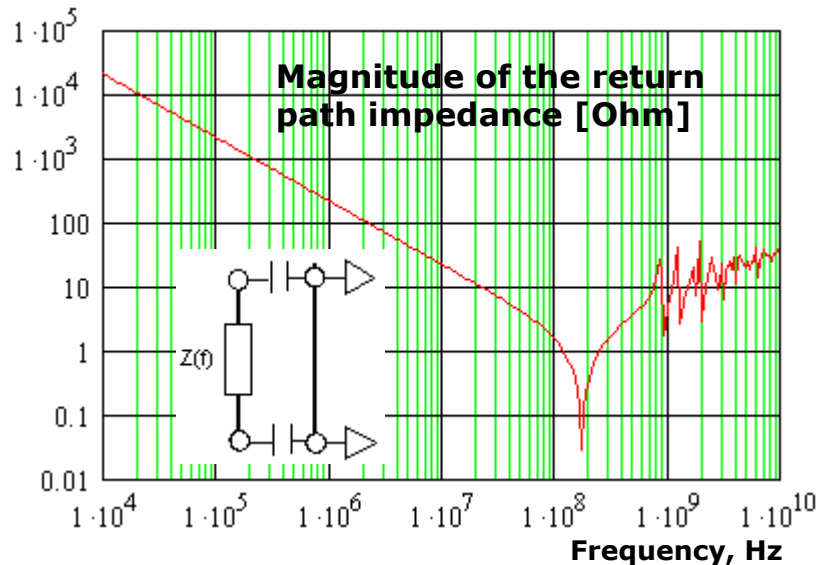


# Common mode parallel-plane $Z(f)$ profiles

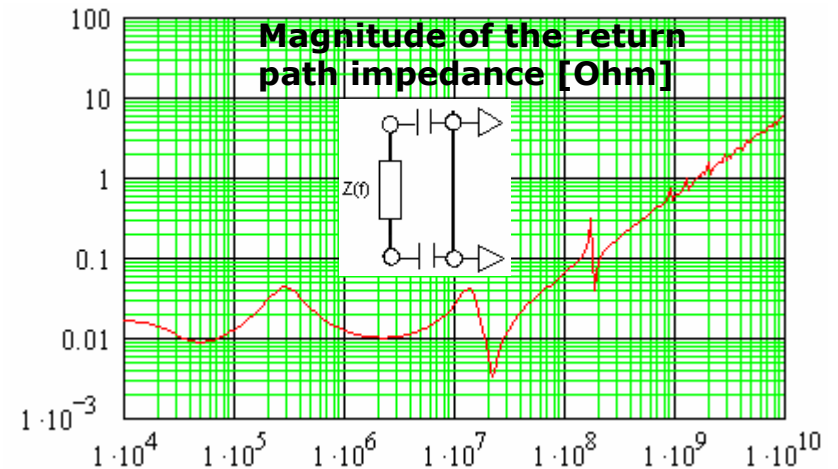
## 165 mm by 165 mm rectangular planes example

Board geometry and decoupling structures defines the impedance of the return path for common mode  $Z(f)$

Bare board with distance between planes 1.524 mm



De-coupled board with distance between planes 1.524 mm



Frequency, Hz

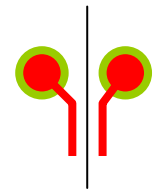
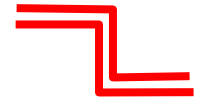
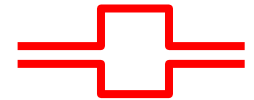
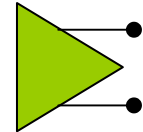
# How to simulate differential vias with the common mode?

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- ❑ Simulate the whole board in a 3D full-wave solver with all plane terminations
  - Hardly ever possible and not practical
- ❑ Use a hybrid solver with 2D parallel-plane models
  - Practical, but accurate only if such solvers include 3D full-wave models for differential mode (no such solvers available so far)
- ❑ Localize the problem with asymptotically correct PEC boundary conditions
  - Influence of the boundary conditions on the performance at the system level may be insignificant in many cases
  - May work if no common mode or it is very small at the connections to the vias

# Avoid conversion from differential to common mode by design!

- ❑ Use symmetrical drivers and receivers that do not generate common mode
- ❑ Keep traces at the same distance and bypass discontinuity symmetrically if absolutely necessary
- ❑ If a discontinuity converts differential mode into common – use mirror discontinuity to convert it back into differential mode
- ❑ Via-holes and transition to the traces have to be symmetrical
- ❑ Common mode analysis might not be necessary if no common mode generated



# Agenda

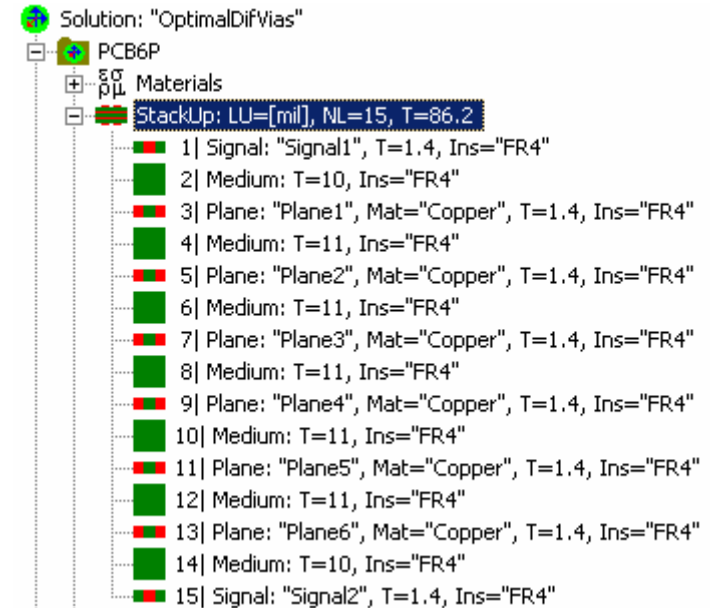
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- Introduction
- Multiport parameters
- Electromagnetic analysis of via-holes
- Analysis of single via-holes
- Analysis of differential via-holes
- Examples of differential via-hole optimization**
- Conclusion



# Differential via-holes design example

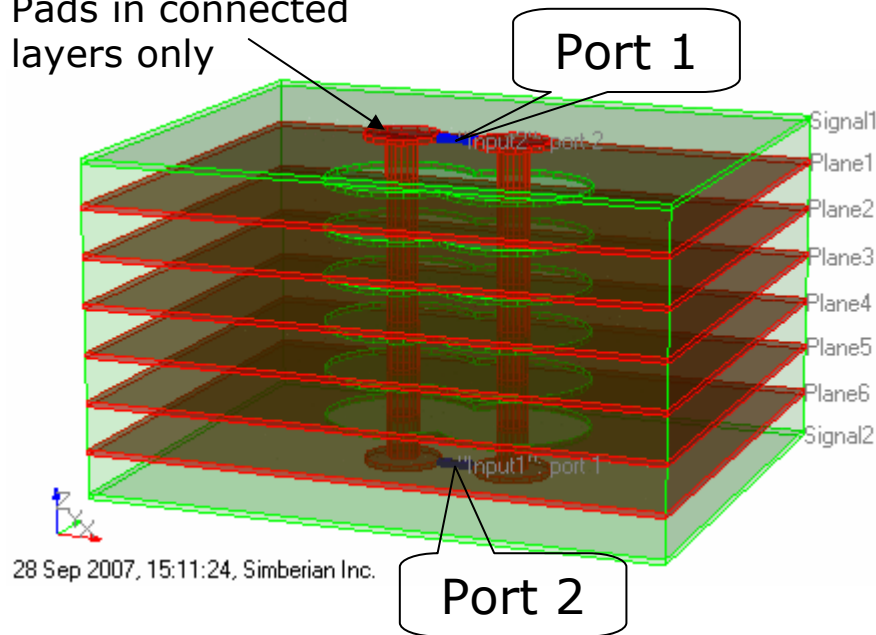
- Stackup with 10 plane layers
- Drill diameter is 8 mil, differential traces are 6 mil wide 10 mil apart
- Design differential through-vias from Signal1 to Signal2



1. Create and optimize via-holes geometry for differential mode only (using lumped ports) by minimizing S11 for 100-Ohm normalized S-parameters
2. Add transition to differential transmission line and generate final model

# Geometry synthesis and electromagnetic analysis with Simbeor 2007

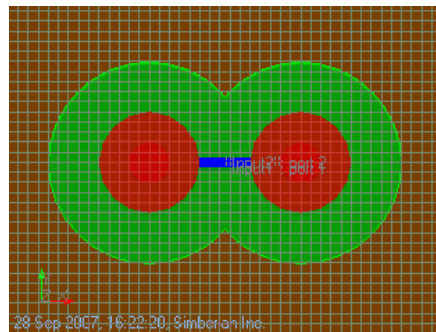
Pads in connected layers only



DifViasOptimal – vias are 30 mil apart with 40 mil anti-pads and 20 mil pads -  $|S_{11}|$  below -20 dB

DifViasOptimal2 – vias are 24 mil apart with 32 mil anti-pads and 20 mil pads -  $|S_{11}|$  below -25 dB

$|S_{11}|$



S-parameters normalized to 100 Ohm

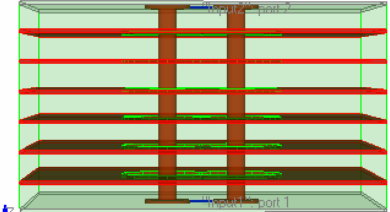
# Simulation in time domain with system-level simulator HyperLynx 7.7 with Eldo

Simple SPICE 100 Ohm differential drivers to generate 10 Gbps pulse train

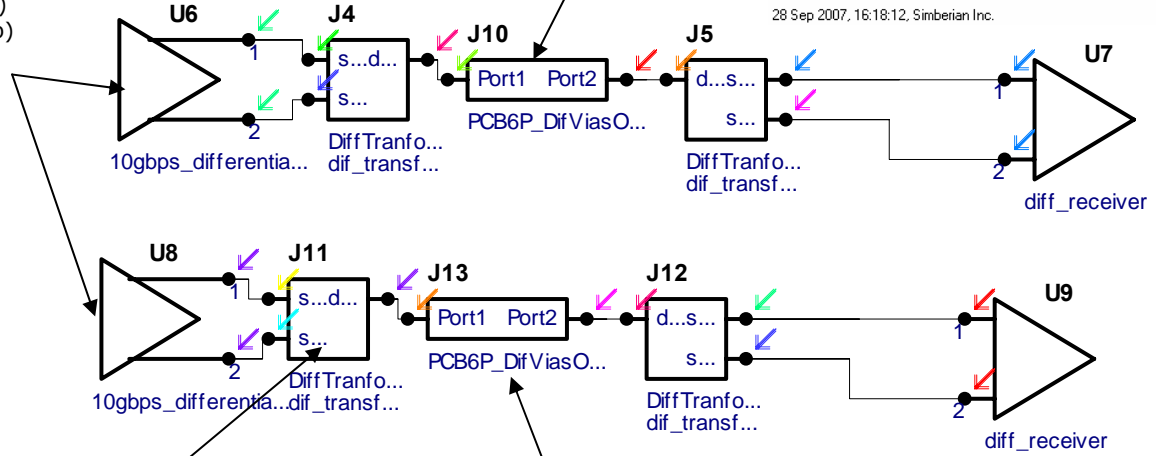
```
.subckt 10gbps_differential_driver outp outm
voutp outp n1 pulse(-1.0 1.0 0 25p 25p 62.5p 200p)
voutm outm n2 pulse(1.0 -1.0 0 25p 25p 62.5p 200p)
routp n1 0 50
routm n2 0 50
.ends differential_driver
```

Design File Optimal Us Differential s...  
HyperLynx 7.7

DifViasOptimal with -20 dB reflection

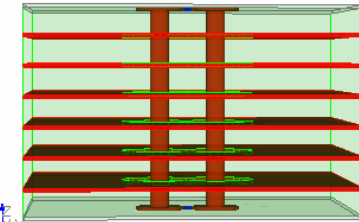


28 Sep 2007, 16:18:12, Simberian Inc.



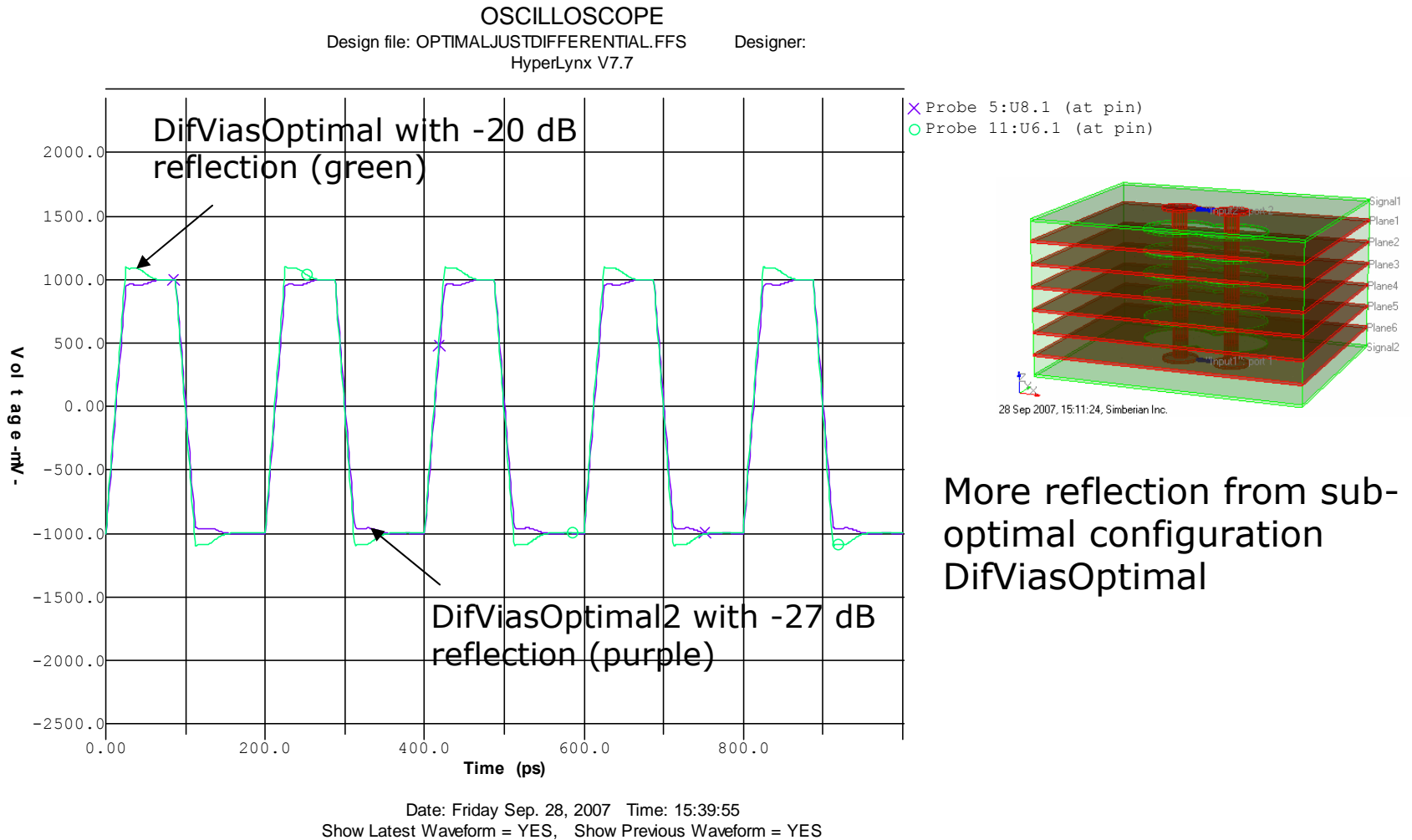
Transformers from terminal space to differential mode and back

DifViasOptimal2 with -27 dB reflection



28 Sep 2007, 16:17:18, Simberian Inc.

# Comparison of reflection in time-domain

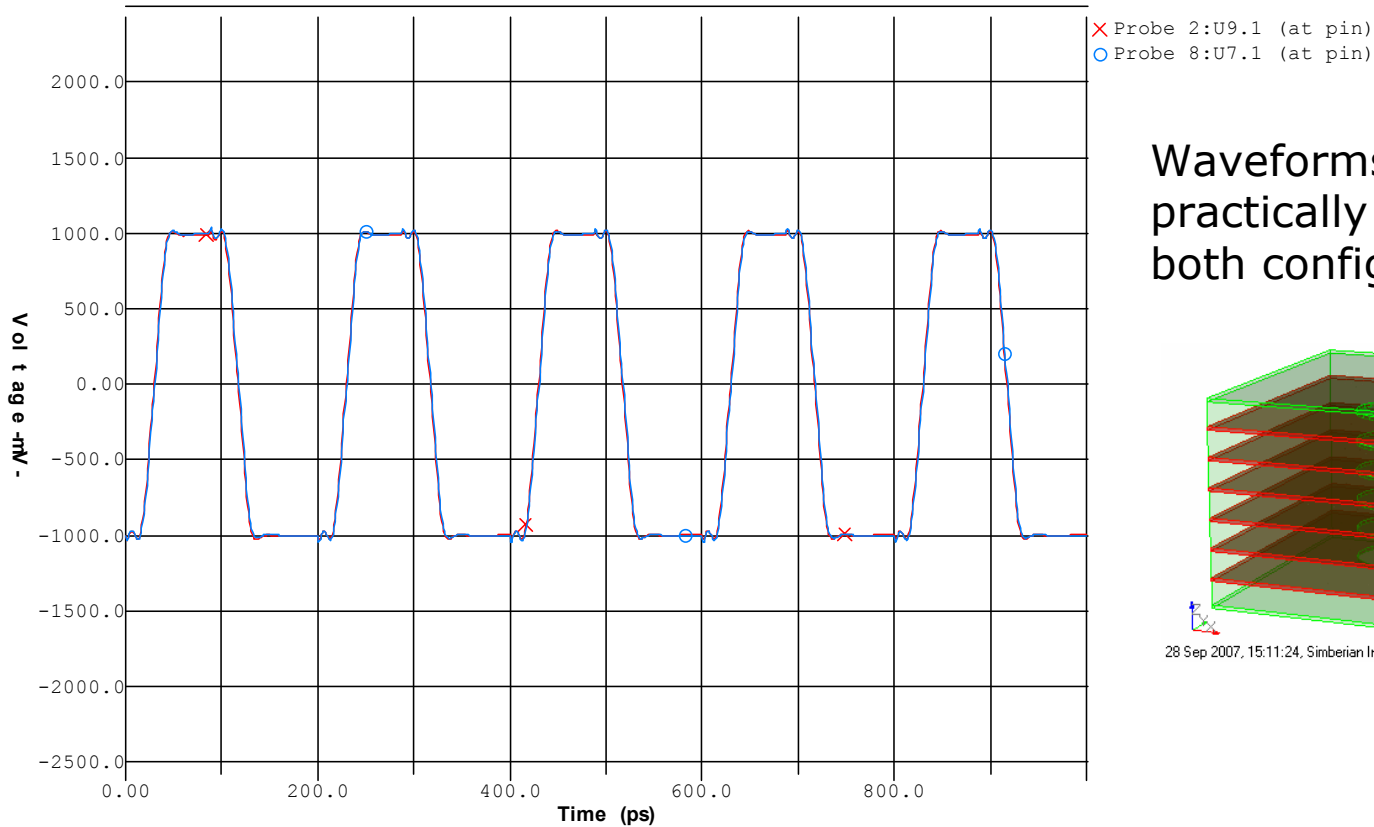


# Comparison of transmission in time-domain

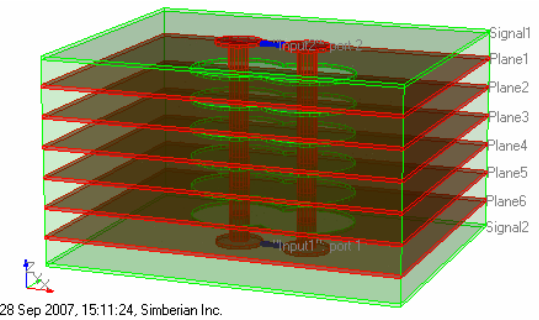
## OSCILLOSCOPE

Design file: OPTIMALJUSTDIFFERENTIAL.FFS  
HyperLynx V7.7

Designer:



Waveforms are practically identical for both configurations

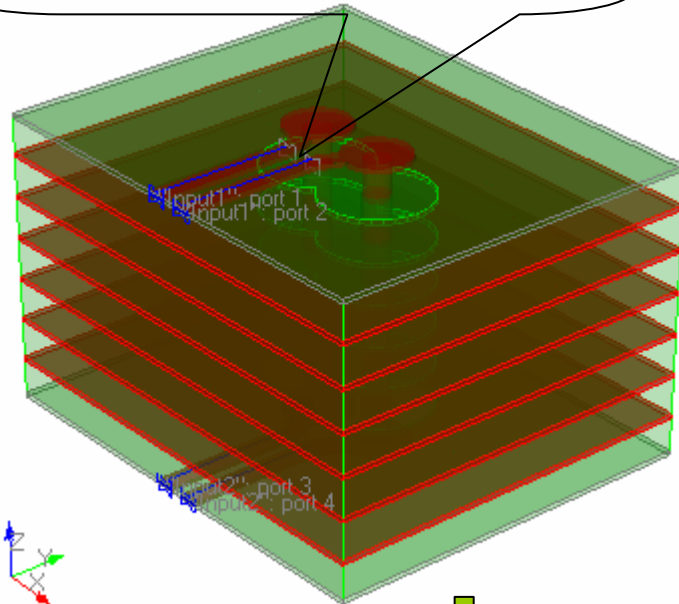


28 Sep 2007, 15:11:24, Simberian Inc.

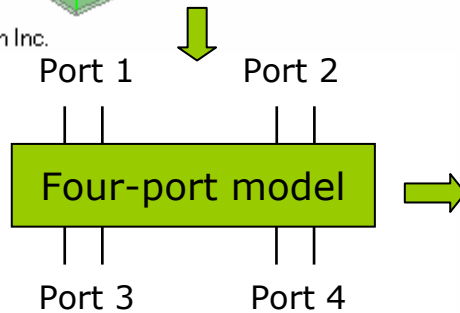
Date: Friday Sep. 28, 2007 Time: 15:30:54  
Show Latest Waveform = YES, Show Previous Waveform = YES

# Final via-holes design and 4-port model

De-embedded wave-ports are used with the phase reference shift to the edges of anti-pads



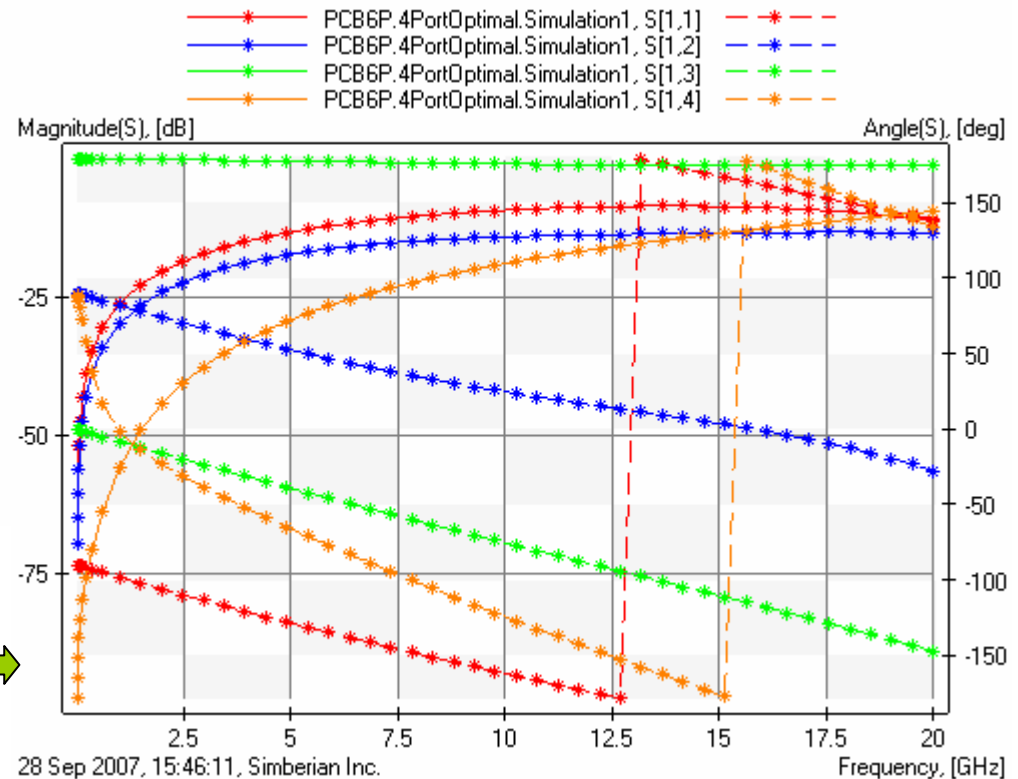
28 Sep 2007, 15:45:33, Simberian Inc.



9/28/2007

Vias are 24 mil apart with 32 mil anti-pads and 20 mil pads, traces are 6 mil wide 10 mil apart

The model can not be used to predict common mode propagation – no stitching vias



28 Sep 2007, 15:46:11, Simberian Inc.

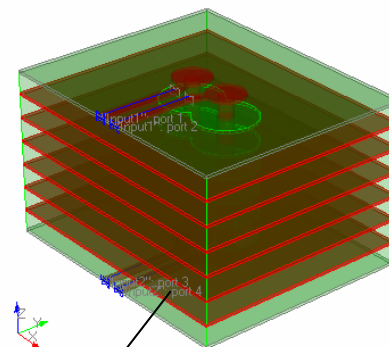
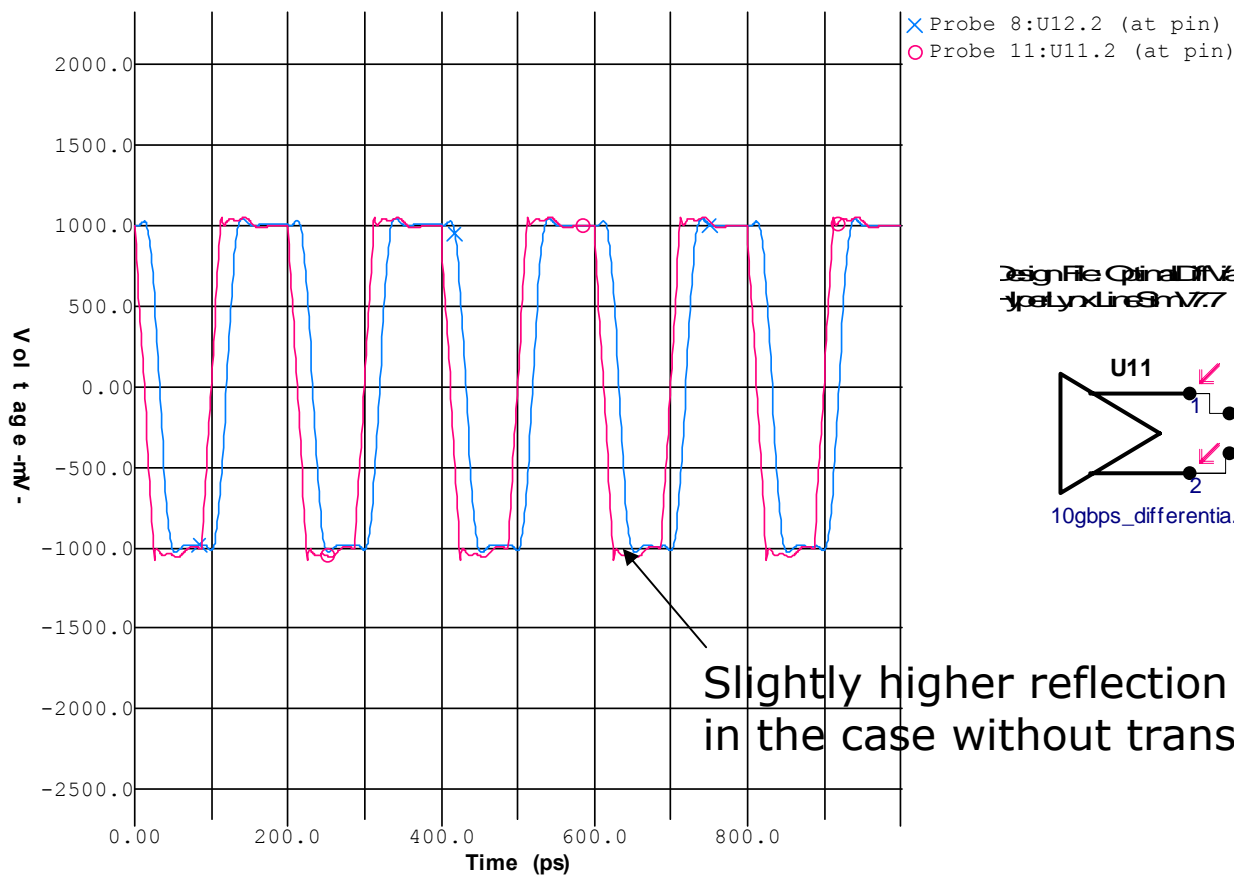
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# Reflection and transmission of differential mode in time domain

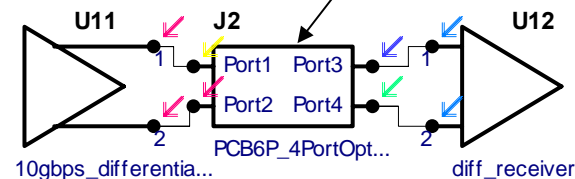
## OSCILLOSCOPE

Design file: OPTIMALDIFFVIAS.FFS Designer:  
HyperLynx V7.7



Design File: OptimalDiffVIAS.ffs  
HyperLynx V7.7

28 Sep 2007, 15:45:33, Simberian Inc.



Slightly higher reflection than  
in the case without transition

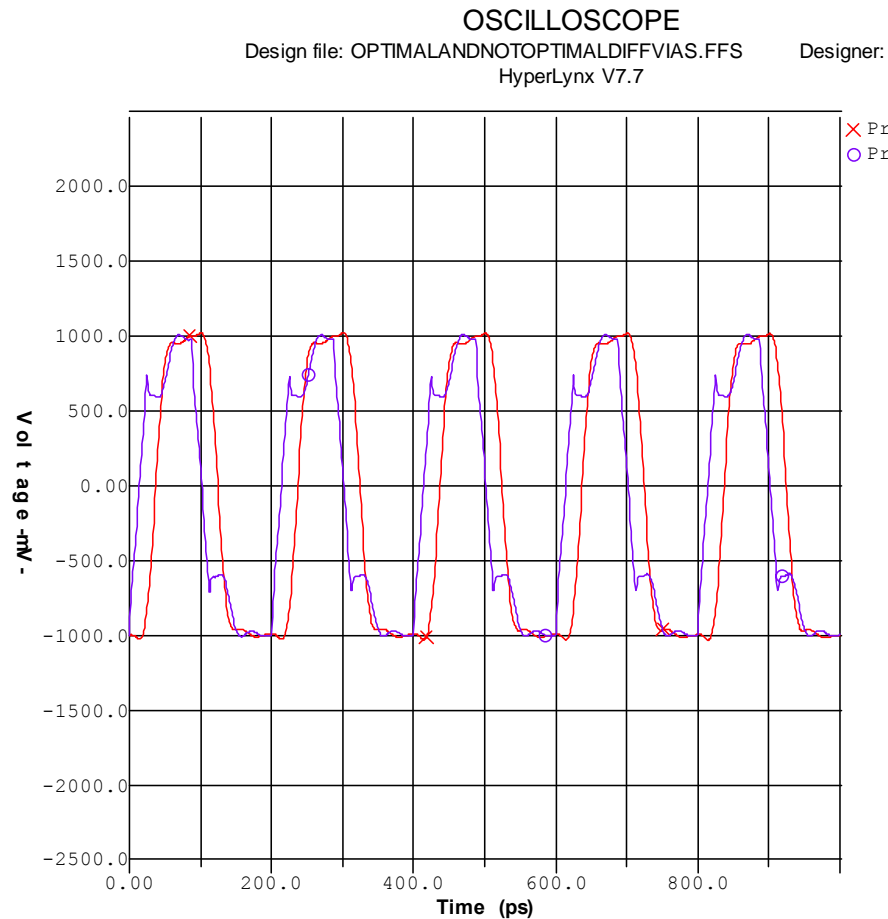
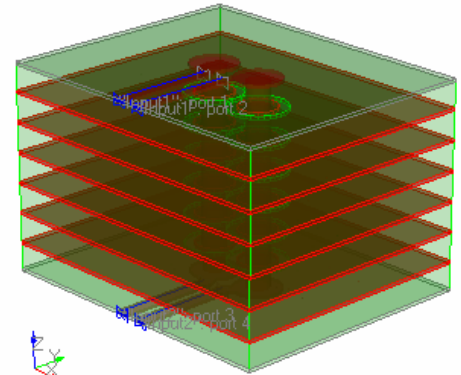
Date: Friday Sep. 28, 2007 Time: 16:00:32  
Show Latest Waveform = YES, Show Previous Waveform = YES

9/28/2007

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# Not optimal via-holes can significantly degrade the signal



10-mil vias are 24 mil apart with 24 mil anti-pads and 16 mil pads in all layers, traces are 6 mil wide 10 mil apart

Pads in plane layers increase the capacitance and decrease the effective impedance of vias to 60 Ohm

Date: Friday Sep. 28, 2007 Time: 16:26:12  
Show Latest Waveform = YES, Show Previous Waveform = YES



# Conclusion – Select the right tools to optimize and build via-hole models

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- Distinguish localizable and non-localizable cases
  - Dependence of S-parameters from the boundary conditions usually shows that the problem is non-localizable
- Only localizable cases can be analyzed with a 3D full-wave solver
  - Single-ended via or differential vias with stitching vias nearby
  - Differential mode propagation in differential vias without stitching vias
- Complete board analysis is required to build accurate models for non-localizable cases
  - Hybrid distributed analysis with transmission plane solver is the best choice for such problems considering performance and accuracy
  - Localized 3D full-wave differential via analysis can be integrated into system-level hybrid solution to increase accuracy

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- Skype *shlepnev*

## □ Biography

- Yuriy Shlepnev is the president and founder of Simberian Inc., where 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.