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Conference

January 28–30, 2025

Santa Clara Convention Center

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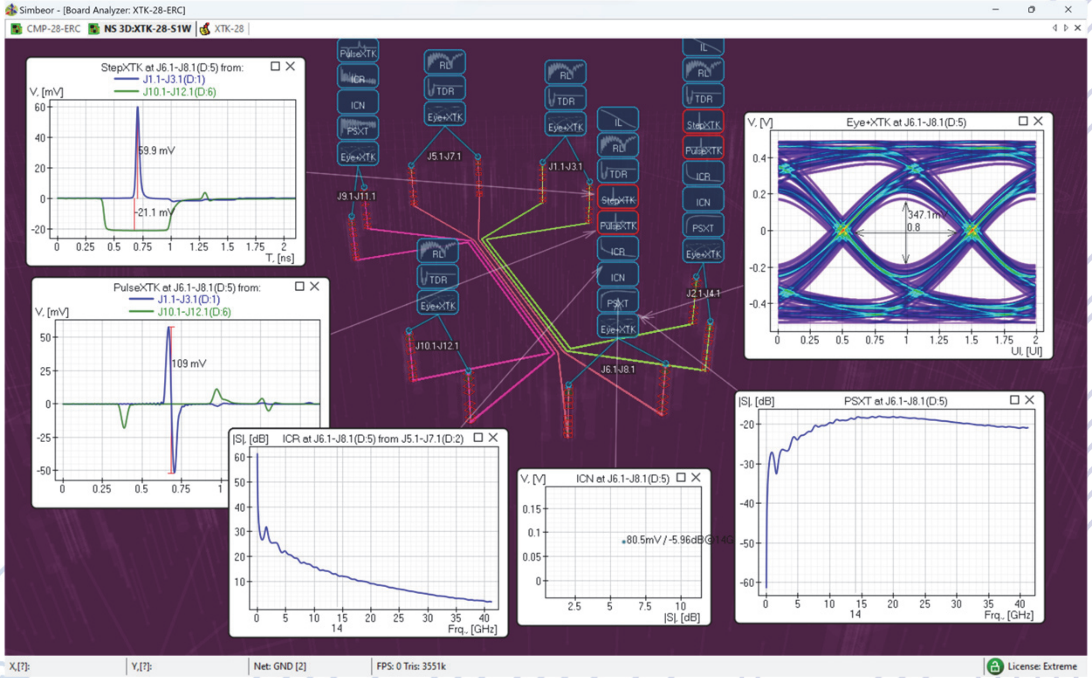
January 29–30, 2025

How Interconnects Work: Crosstalk Anatomy and Quantification

Yuriy Shlepnev, Simberian Inc.



Date: Tuesday, January 28
9:00 AM -11:30 AM Pacific Time
Location: Ballroom D

Update of this presentation and **some solutions** are at
https://www.simberian.com/TechnicalPresentations.php

OUTLINE

- Introduction
- Basics: Fields and S-parameters
- Crosstalk Anatomy - Qualitative Analysis
- Crosstalk Quantification
- Distant Crosstalk - Sources and Mitigation
- Conclusion

Signal Degradation Factors

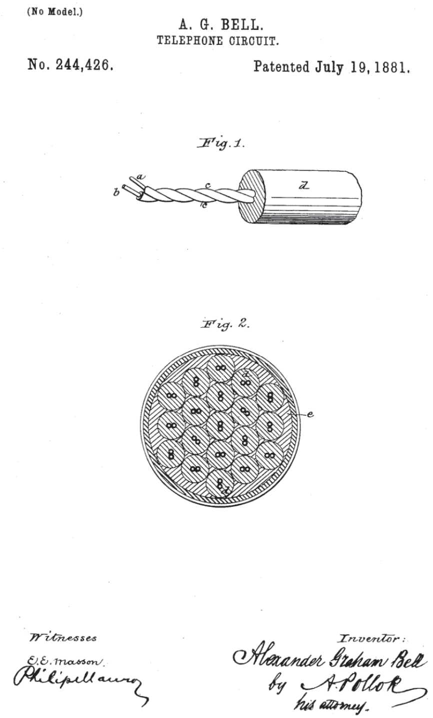
- **Absorption** (dispersion) in dielectrics and conductors (app notes #2016_01, #2021_10)
- **Reflections** from impedance mismatches and discontinuities (app notes #2021_11, #2022_01)
- **Leaks or dissipation** into other interconnects, common mode, power distribution networks (PDNs), and free space (radiated) – source of crosstalk and EMC/EMI
- **Crosstalk** – interference from signals leaked from the other interconnects (app notes #2023_04, #2024_01)

Crosstalk is unwanted noise from the coupled structures caused by signal leaks that degrade the useful signal and may reduce the data transmission rate and even cause complete link failure

App Notes are at https://www.simberian.com/AppNotes.php

Crosstalk History

Crosstalk problem – solved before theoretical formulation



Theory: The Telegrapher's Equations

ELECTROMAGNETIC INDUCTION AND ITS PROPAGATION. 39

XXXV. ELECTROMAGNETIC INDUCTION AND ITS PROPAGATION. (SECOND HALF.)

[*The Electrician*, 1886-7. Section XXV., April 23, 1886, p. 469; XXVI., May 14, p. 8 (vol. 17); XXVII., June 11, p. 88; XXVIII., June 25, p. 128; XXIX., July 23, p. 212; XXX., August 6, p. 252; XXXI., August 20, p. 296; XXXII., August 27, p. 316; XXXIII., November 12, p. 10 (vol. 18); XXXIV., December 24, 1886, p. 143; XXXV., January 14, 1887, p. 211; XXXVI., February 4, p. 281; XXXVII., March 11, p. 390; XXXVIII., April 1, p. 457; XXXIX*a*., May 13, p. 5 (vol. 19); XXXIX*b*., May 27, p. 50; XL., June 3, p. 79; XLI., June 17, p. 124; XLII., July 1, p. 163; XLIII., July 15, p. 206; XLIV., August 12, p. 295; XLV., August 26, p. 340; XLVI., October 7, p. 459; XLVII., December 30, 1887, p. 189 (vol. 20).]

Now let R, L, S and K be the resistance, inductance, permittance and leakage-conductance per unit length of a circuit; and let V and C be the potential-difference (an awkward term) and current at distance x. We have the following fundamental equations of connection:—

$$-\frac{dV}{dx} = (R + Lp)C, \qquad -\frac{dC}{dx} = (K + Sp)V, \quad \ldots\ldots\ldots\ldots (1d)$$

p standing for d/dt. Observe that the space-variation of C is related to V in the same manner (formally) as the space-variation of V is related to C, so that we can translate solutions in an obvious manner by exchanging V and C, R and K, L and S, which are reciprocally related, in a manner.

ELECTRICAL PAPERS

BY

OLIVER HEAVISIDE

IN TWO VOLUMES

VOL. II.

New York
MACMILLAN AND CO.
AND LONDON
1894

Why Crosstalk is Important?

- There is always crosstalk in PCB and Packaging interconnects – they are **open waveguiding structures**
- Crosstalk interference **cannot be corrected at the receiver** end in general
- Neglecting crosstalk can result in **system failures that are hard to diagnose and fix**
- Crosstalk is deterministic, but **difficult to predict in many cases** – too many variables and uncertainties – it is **bounded uncorrelated noise**
- A direct analysis may be not possible or **costly and inefficient**
- Worst-case analysis may lead to **overdesign**

Understanding the sources of crosstalk and mitigation techniques is very important!

Crosstalk Types

- **Local coupling:**
  - Coupling in closely routed signal traces – the most common source of crosstalk
  - Local coupling between viaholes and between viaholes and traces due to proximity
  - Local couplings of traces through cutouts in reference planes
  - Common to differential mode interference and crosstalk due to modal transformations in differential pairs (bends, asymmetry in routing, fiber weave effect,…)
- **Distant coupling and multipath propagation:**
  - Through parallel planes (PDNs)
  - Through split-planes (slots)
  - Through surface dielectric layers (surface and leaky waves)
  - Through PCB enclosure (box resonances)

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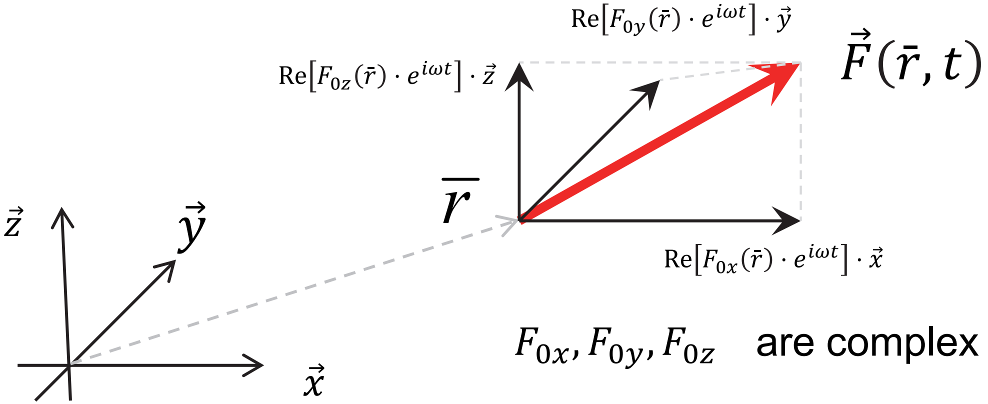
Time-Harmonic Vector Fields

$$\vec{F}(\bar{r},t) = \mathrm{Re}\left[\vec{F}_0(\bar{r}) \cdot e^{i\omega t}\right] = \mathrm{Re}\left[F_{0x}(\bar{r}) \cdot e^{i\omega t}\right] \cdot \vec{x} + \mathrm{Re}\left[F_{0y}(\bar{r}) \cdot e^{i\omega t}\right] \cdot \vec{y} + \mathrm{Re}\left[F_{0z}(\bar{r}) \cdot e^{i\omega t}\right] \cdot \vec{z}$$



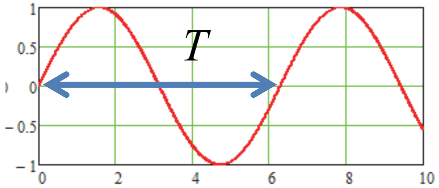
Instantaneous field or current values: amplitude and direction are periodic functions of time

F_{0x}, F_{0y}, F_{0z} are complex numbers

Peak value: $\vec{F}_{peak}(\bar{r}) = \vec{F}(\bar{r}, t_{peak})$ $\quad \left|\vec{F}(\bar{r}, t_{peak})\right| \geq \left|\vec{F}(\bar{r},t)\right|, 0 \leq t < T$

Sources:

$$V(t) = V_0 \cdot \sin(\omega t + \varphi)$$



$$\omega = 2\pi \cdot f = \frac{2\pi}{T}$$

frequency

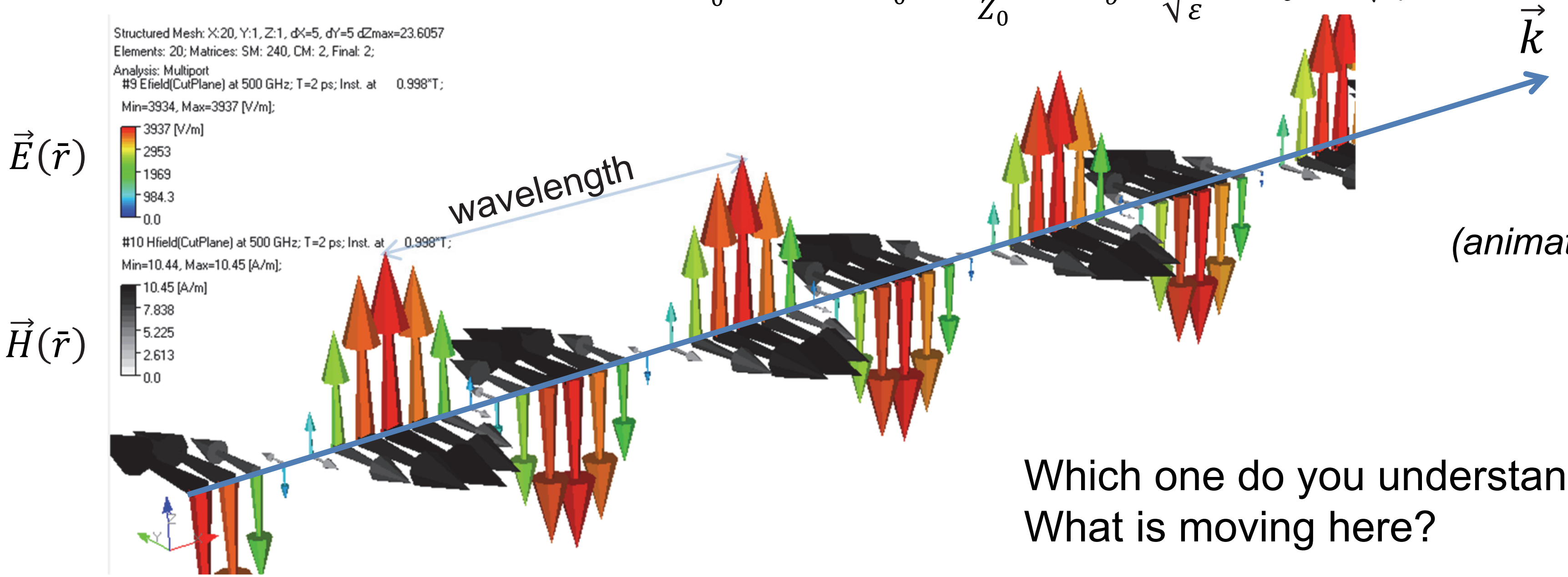
period

Fields: Equations or Pictures of Fields?

Plane wave solution of Maxwell's Equations

$$\begin{pmatrix} \vec{E}(\vec{r}) \\ \vec{H}(\vec{r}) \end{pmatrix} = \begin{pmatrix} \vec{E}_0 \\ \vec{H}_0 \end{pmatrix} \cdot e^{-ik_o \vec{k}\cdot\vec{r}} \qquad \vec{P} = \vec{E} \times \vec{H}$$

$$\vec{k} \cdot \vec{E}_0 = 0 \qquad \vec{H}_0 = \frac{\vec{k} \times \vec{E}_0}{Z_0} \qquad Z_o = \sqrt{\frac{\mu}{\varepsilon}} \qquad k_o = \omega\sqrt{\varepsilon\mu}$$

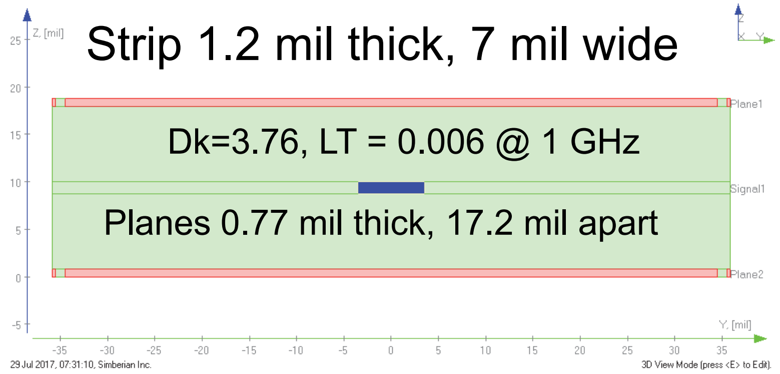


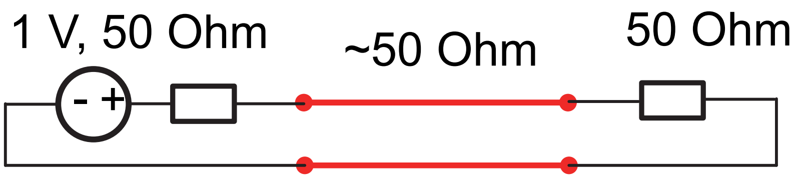
(animated)

Which one do you understand?
What is moving here?

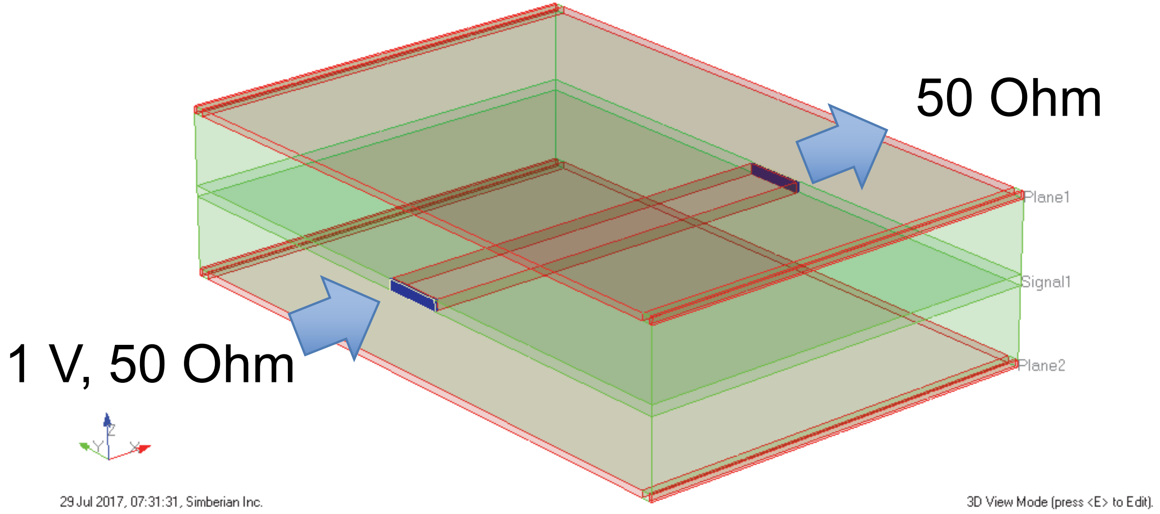
More on field visualization in Webinar #7

Fields, Currents and Power Flow in Stripline





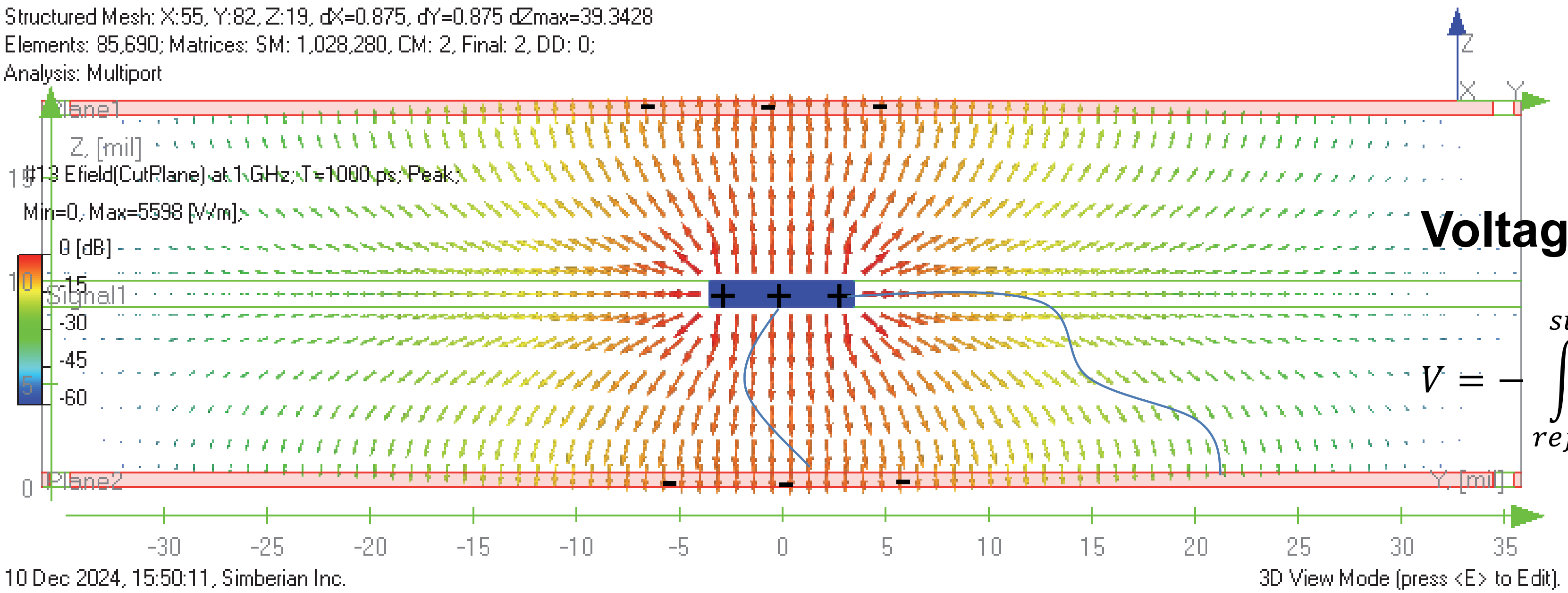
Vpeak ~ 0.5 V
Ipeak ~ 0.01 A



•#2016_03: **How Interconnects Work™**: EM field, current and power flow in strip line, 10 min
YouTube: https://youtu.be/iys0de3Xq4E
Simbeor Solution: StripLineFields_2016_03

Electric Field at 1 GHz

Electric Field of Quasi-TEM wave (V/m, peak values in dB)



$\bar{E}$

Voltage

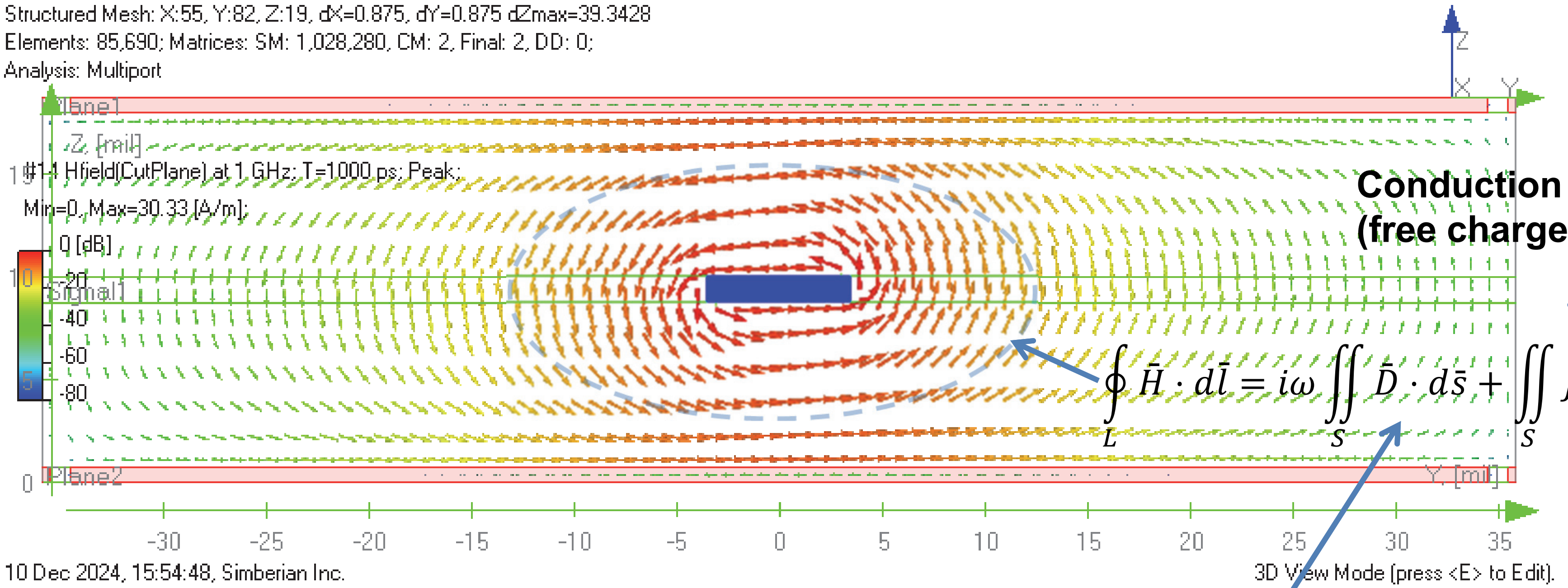
$$V = -\int_{ref}^{sig} \bar{E} \cdot d\bar{l} \approx 0.5V$$

Will integration path matter? Hint: Faraday prevents it☺

$$\oint_L \bar{E} \cdot d\bar{l} = -i\omega\mu \iint_S \bar{H} \cdot d\bar{s}$$

Magnetic Field at 1 GHz

Magnetic Field of Quasi-TEM wave at 1 GHz (A/m, peak values in dB)



$\bar{H}$

Conduction current (free charges)

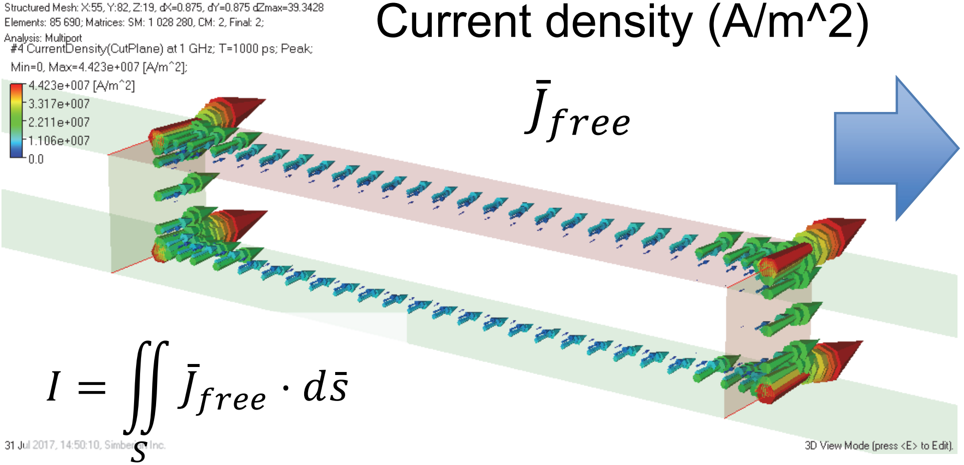
$$\oint_L \bar{H} \cdot d\bar{l} = i\omega \iint_S \bar{D} \cdot d\bar{s} + \iint_S \bar{J}_{free} \cdot d\bar{s}$$

Displacement current (bound charges)

Currents in Stripline at 1 GHz

Current density (A/m^2)

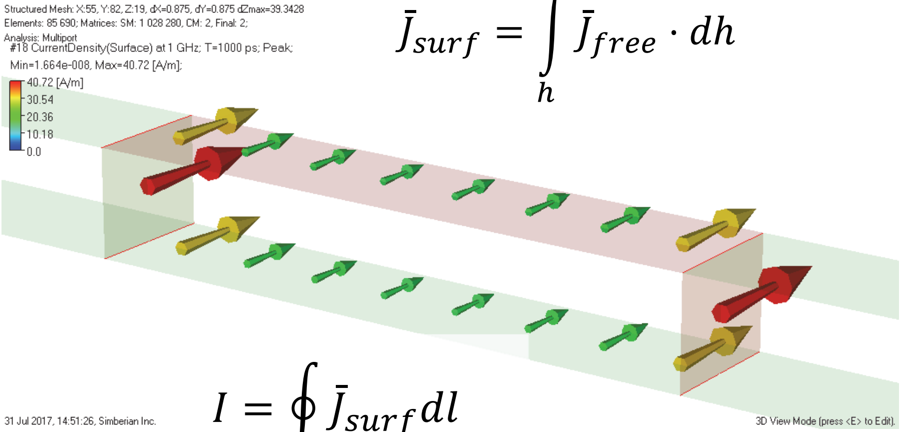
$\bar{J}_{free}$



$$I = \iint_S \bar{J}_{free} \cdot d\bar{s}$$

Surface current density (A/m)

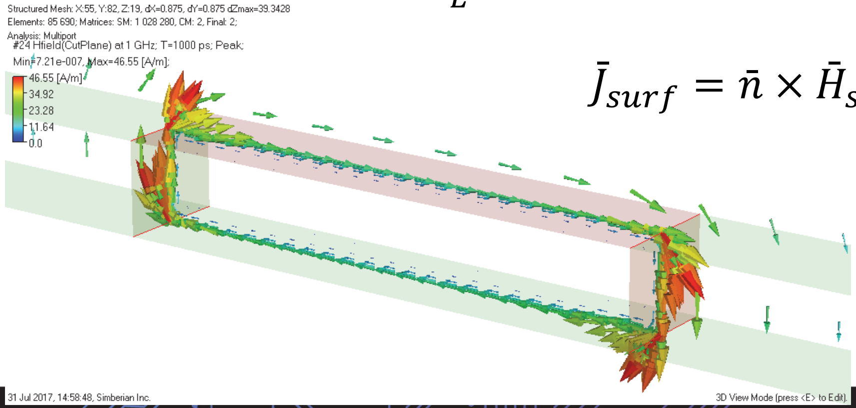
$$\bar{J}_{surf} = \int_h \bar{J}_{free} \cdot dh$$



$$I = \oint_L \bar{J}_{surf} dl$$

Magnetic field intensity -> current in circuit theory

$\bar{H}$

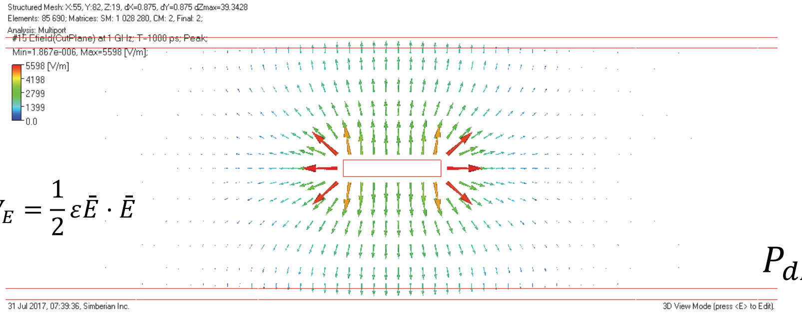


$$\bar{J}_{surf} = \bar{n} \times \bar{H}_{surf}$$

Power Flow Density (PFD) at 1 GHz

Power flow (Poynting vector) is energy passing through unit area in 1 sec

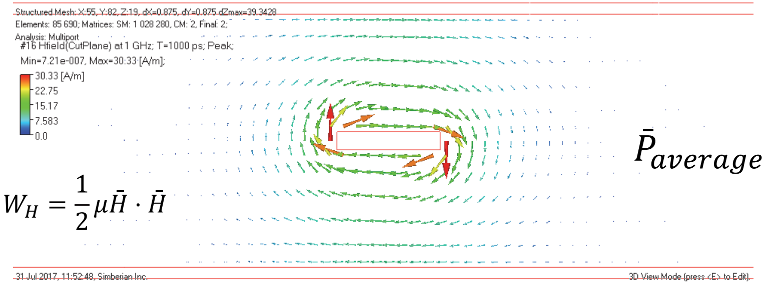
$\bar{E}$



$$W_E = \frac{1}{2}\varepsilon\bar{E}\cdot\bar{E}$$

$$\bar{P}_{flow} = \bar{E}\times\bar{H} \quad \text{W/m^2}$$

$\bar{H}$

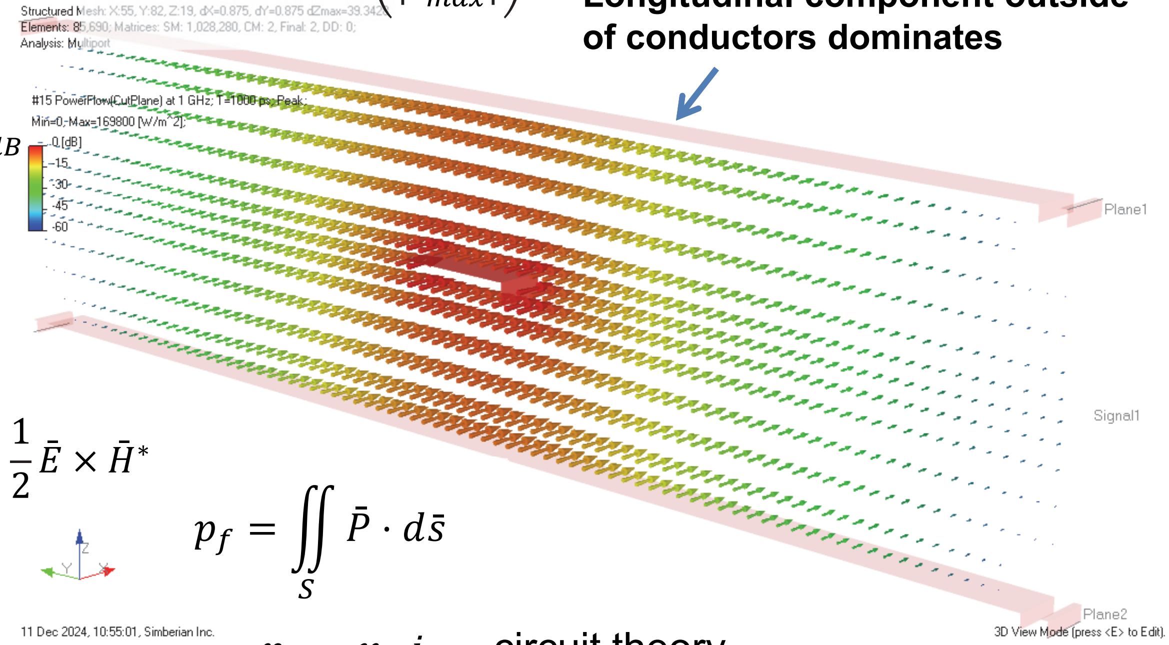


$$W_H = \frac{1}{2}\mu\bar{H}\cdot\bar{H}$$

$$P_{dB} = 10\cdot\log_{10}\left(\frac{|P_{flow}|}{|P_{max}|}\right)$$

Longitudinal component outside of conductors dominates

P_{dB}



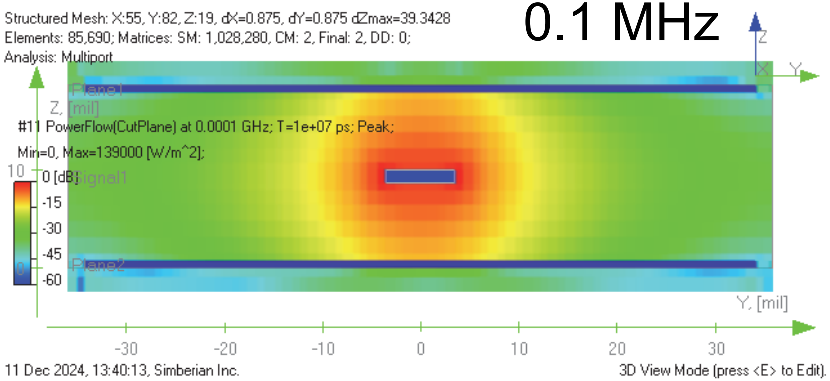
$$\bar{P}_{average} = \frac{1}{2}\bar{E}\times\bar{H}^*$$

$$p_f = \iint_S \bar{P}\cdot d\bar{s}$$

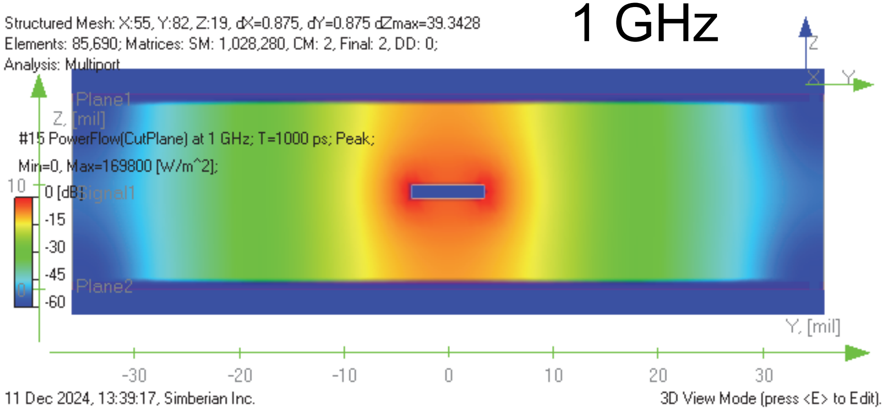
$$p_f = v\cdot i$$ - circuit theory

PFD in Stripline –Signal Space

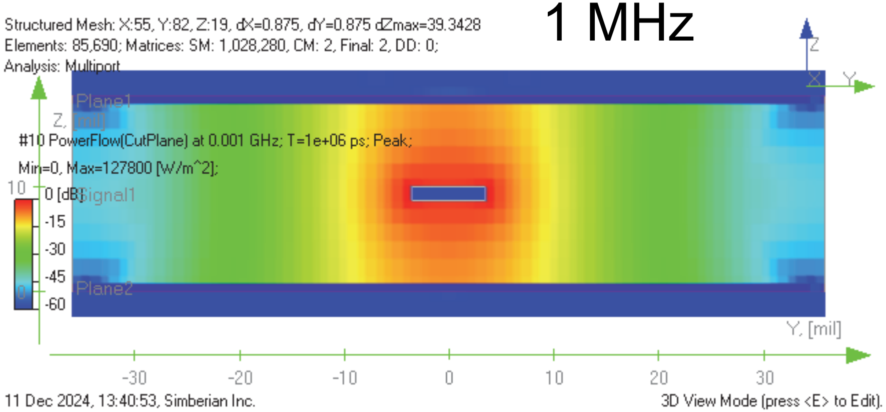
0.1 MHz



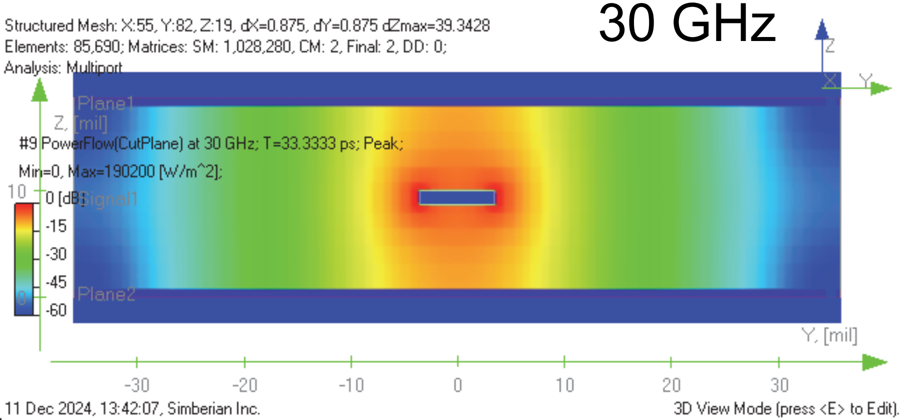
1 GHz



1 MHz

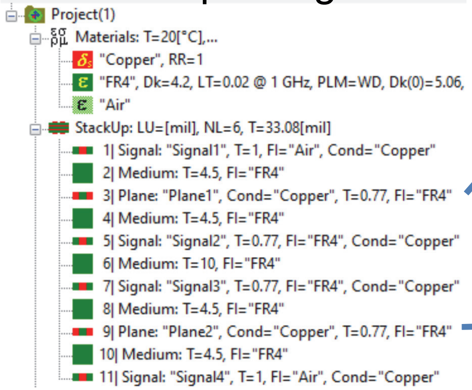


30 GHz

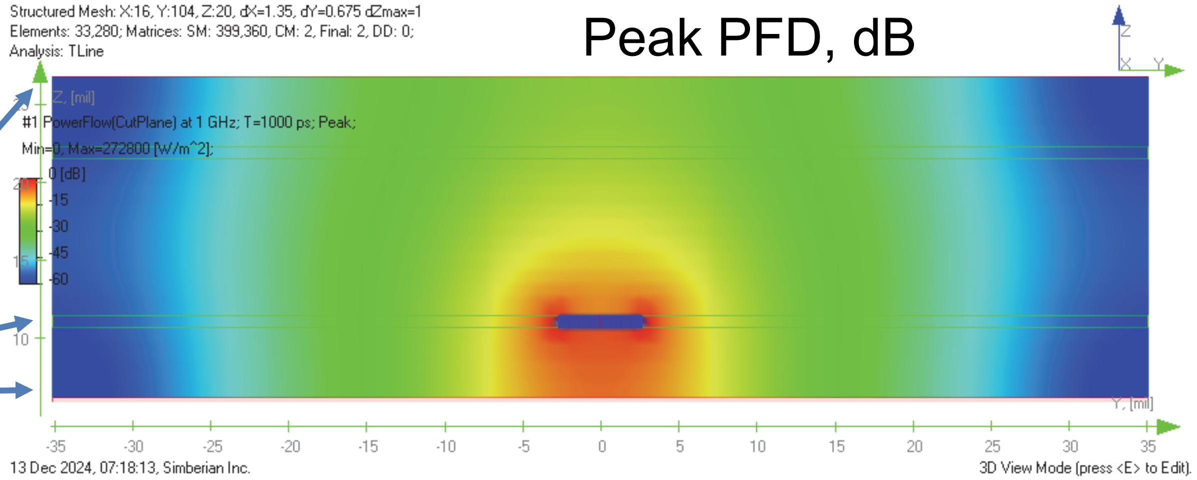


Signal Space of Asymmetric Striplines at 1 GHz

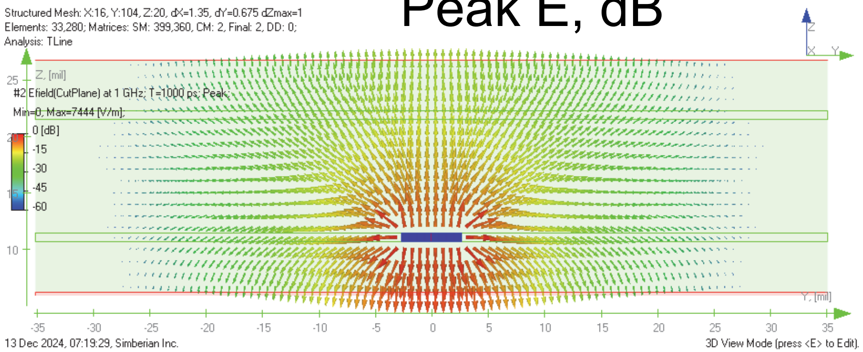
5.4mil strip in Signal3



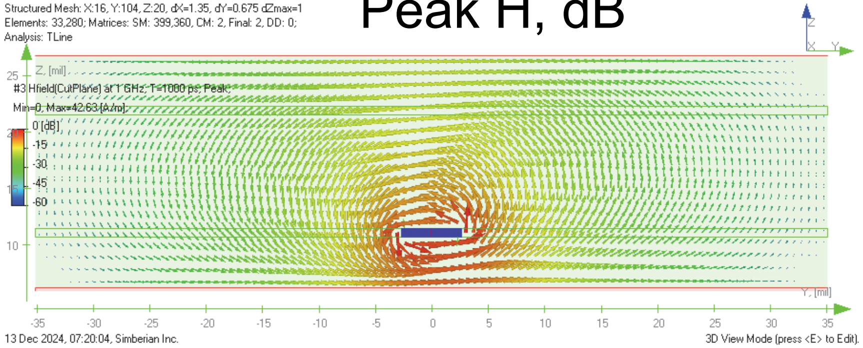
Peak PFD, dB



Peak E, dB



Peak H, dB



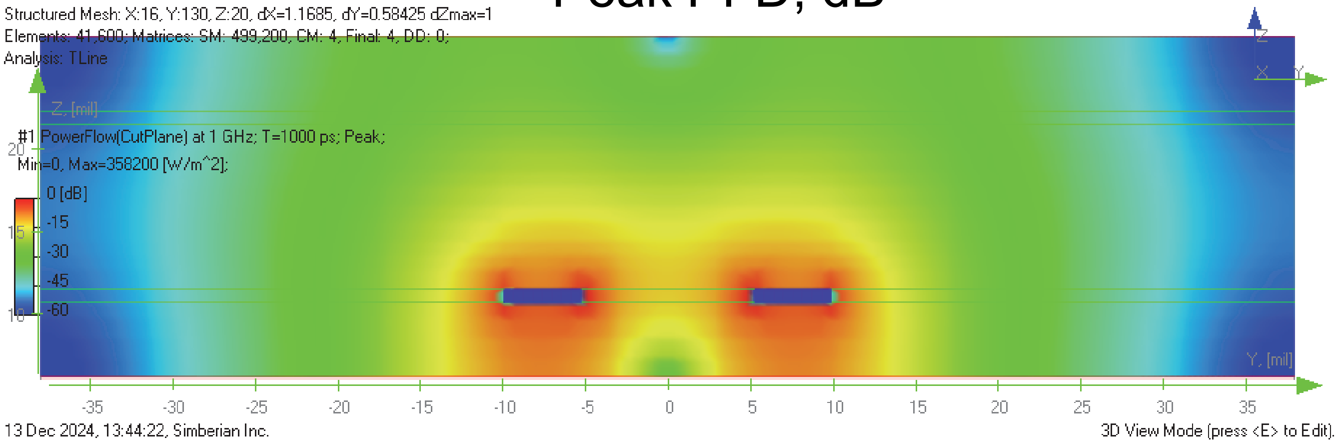
Simbeor Solution: TLinePFD_AN2023_04

Signal Space of Diff. Asymmetric Stripline at 1 GHz

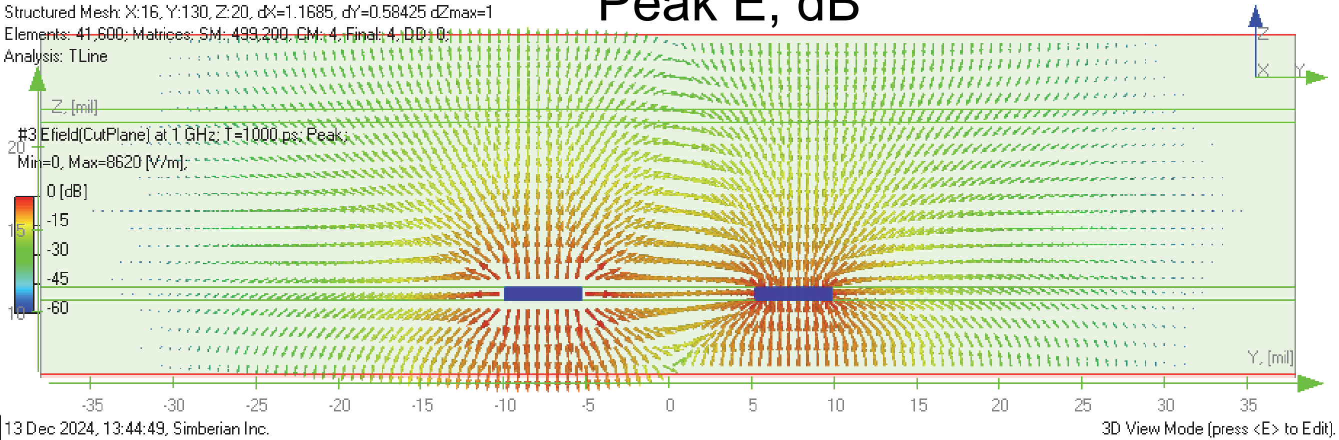
4.7mil strips in Signal3
10.5mil separation

- Project(1)
  - Materials: T=20[°C],...
    - "Copper", RR=1
    - "FR4", Dk=4.2, LT=0.02 @ 1 GHz, PLM=WD, Dk(0)=5.06,
    - "Air"
  - StackUp: LU=[mil], NL=6, T=33.08[mil]
    - 1| Signal: "Signal1", T=1, Fl="Air", Cond="Copper"
    - 2| Medium: T=4.5, Fl="FR4"
    - 3| Plane: "Plane1", Cond="Copper", T=0.77, Fl="FR4"
    - 4| Medium: T=4.5, Fl="FR4"
    - 5| Signal: "Signal2", T=0.77, Fl="FR4", Cond="Copper"
    - 6| Medium: T=10, Fl="FR4"
    - 7| Signal: "Signal3", T=0.77, Fl="FR4", Cond="Copper"
    - 8| Medium: T=4.5, Fl="FR4"
    - 9| Plane: "Plane2", Cond="Copper", T=0.77, Fl="FR4"
    - 10| Medium: T=4.5, Fl="FR4"
    - 11| Signal: "Signal4", T=1, Fl="Air", Cond="Copper"

Peak PFD, dB

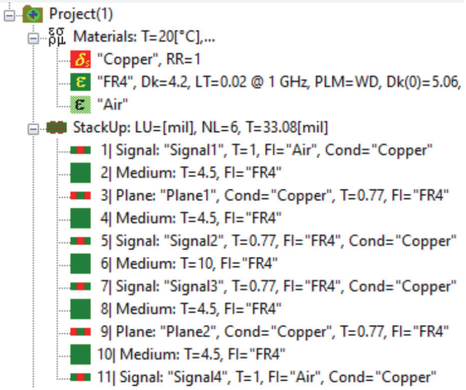


Peak E, dB

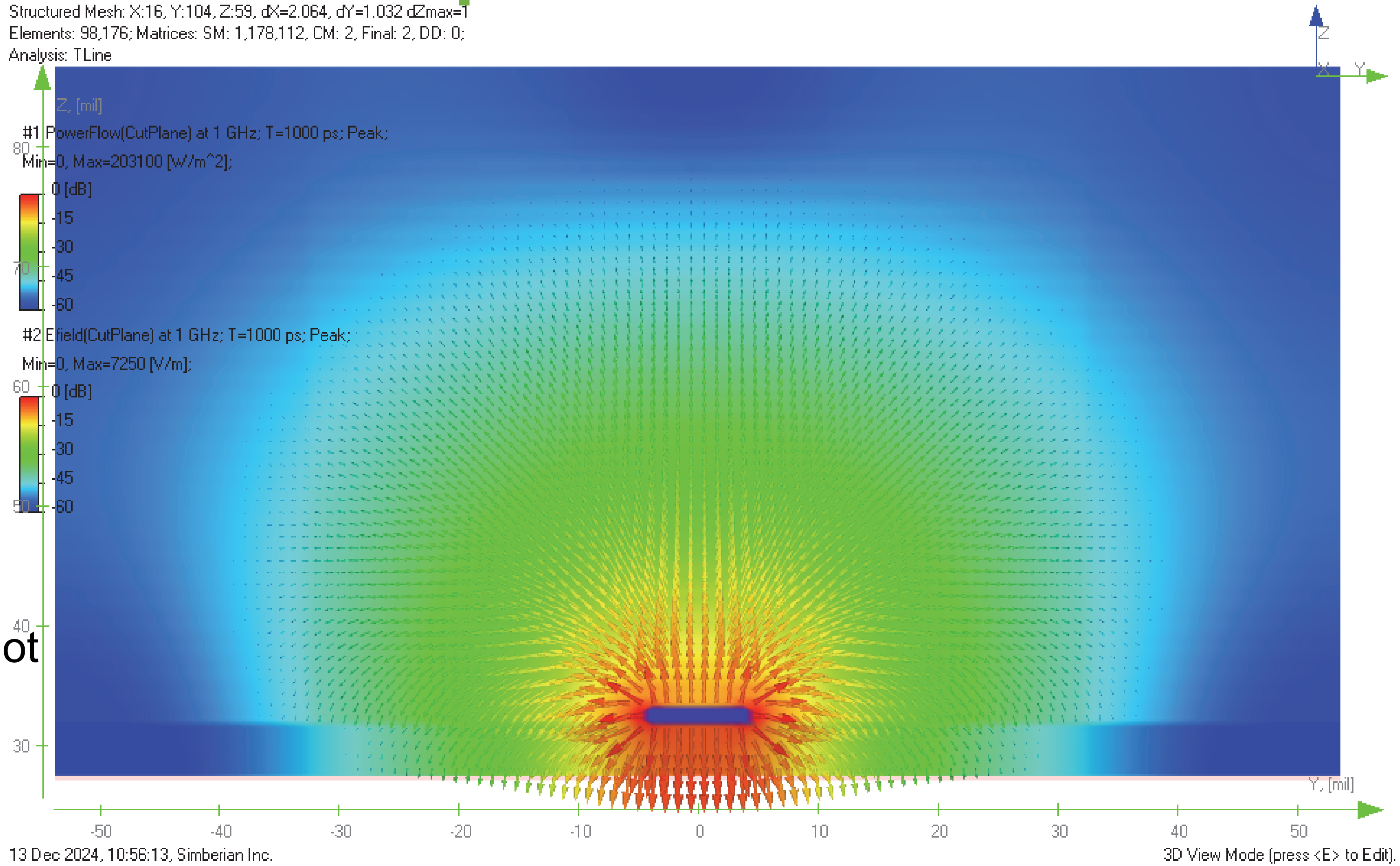


Signal Space of Microstrip Line at 1 GHz

8.26mil strip in Signal1

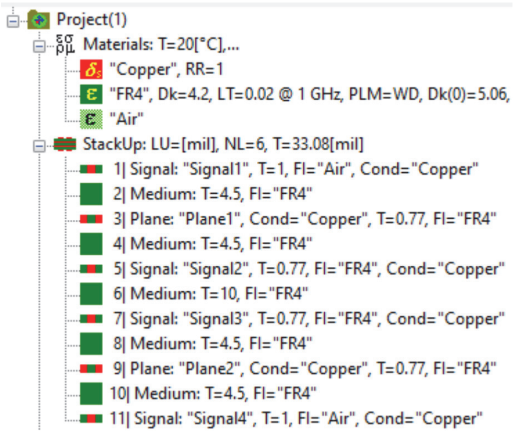


Peak PFD, dB – color plot
Peak E, dB - arrows

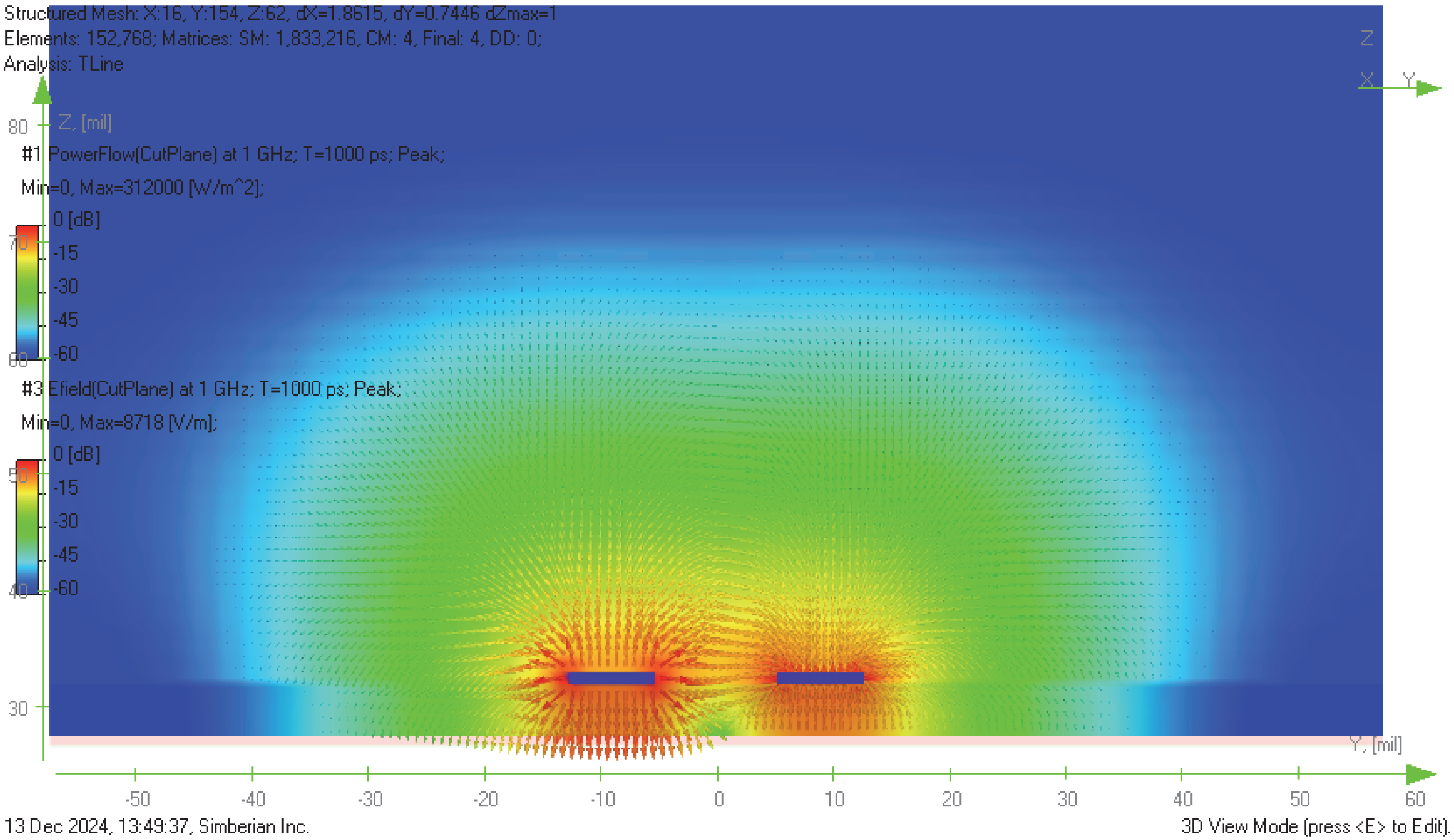


Signal Space of Diff. Microstrip Line at 1 GHz

7.45mil strips in Signal1
10.5mil separation

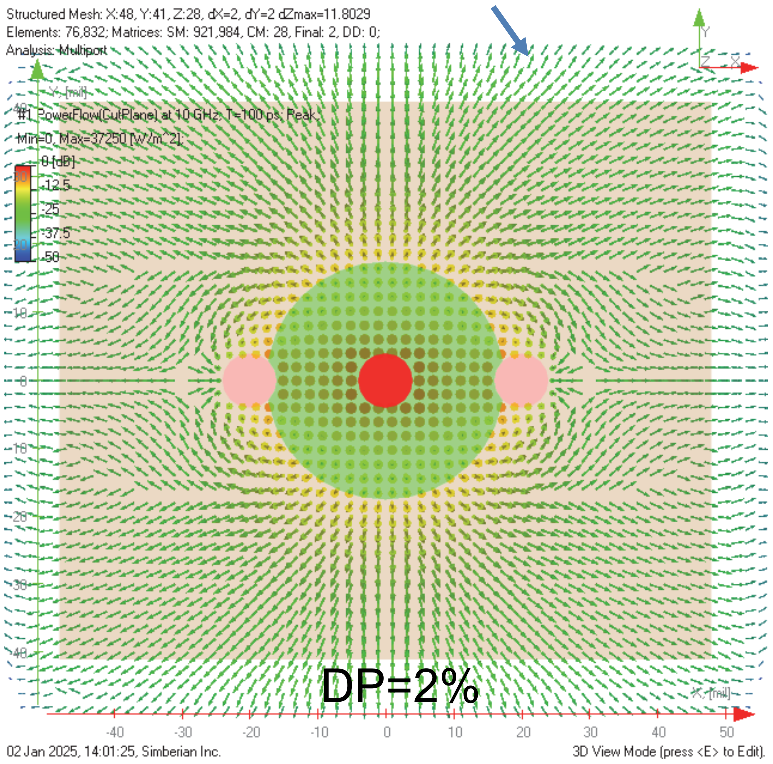


Peak PFD, dB – color plot
Peak E, dB - arrows

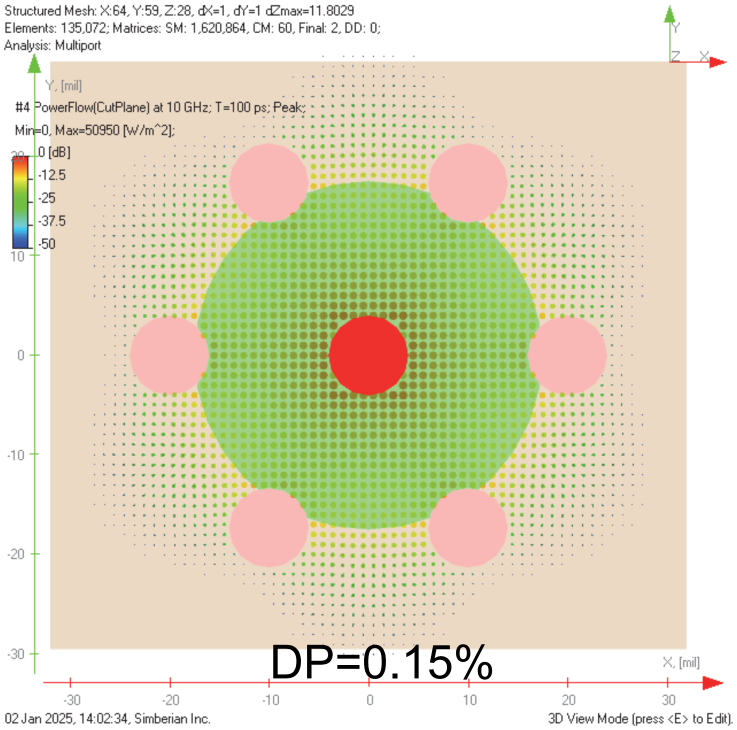


Signal Space for Single-Ended Vias

Signal energy goes sideway – not localized!



Signal space is localized by reference vias

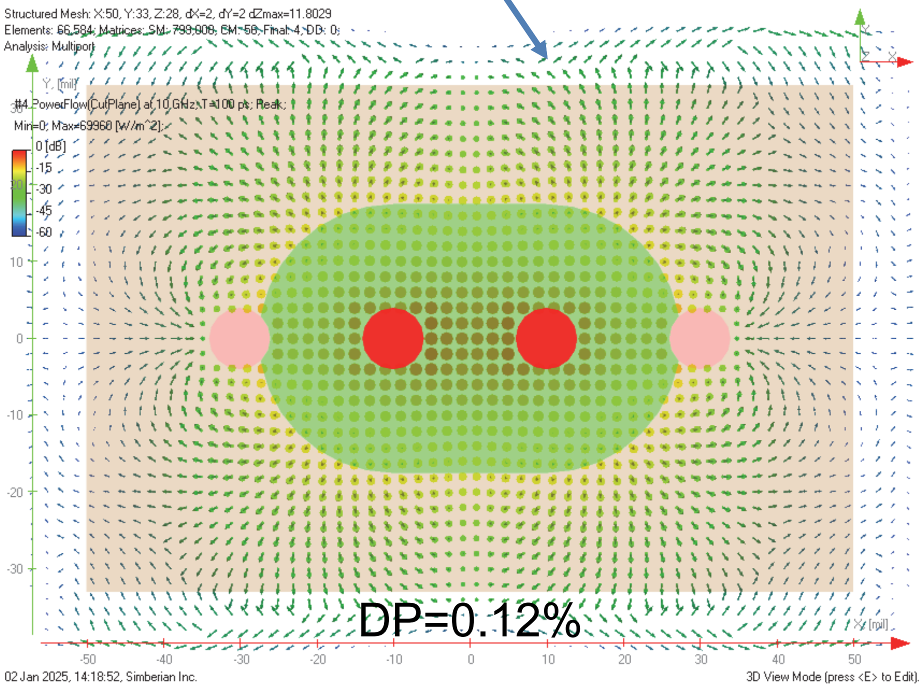


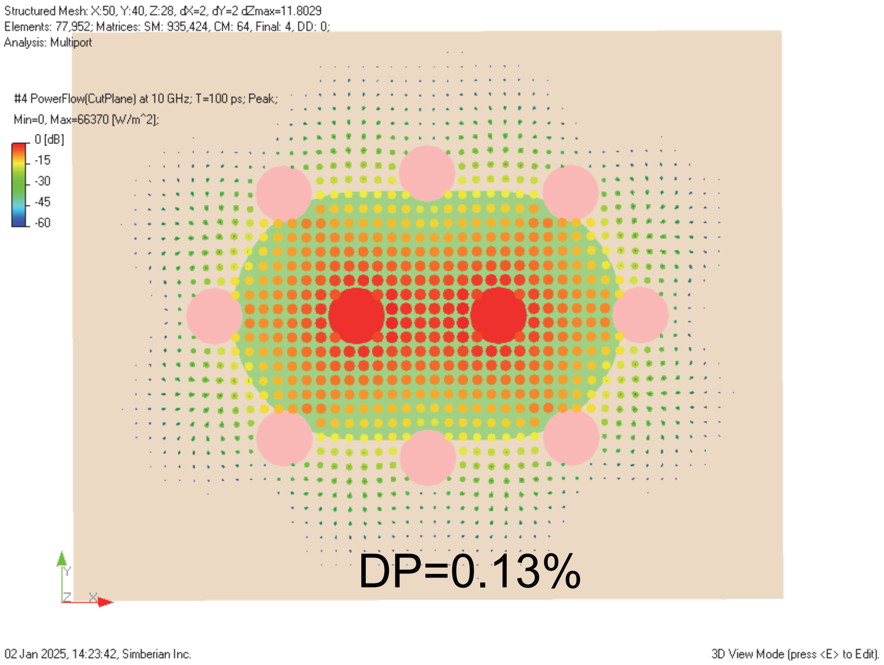
Single-ended via through two parallel planes: Peak PFD at 10 GHz (example from "Distant Crosstalk")

Signal Space for Diff. Vias (Diff. Mode)

Signal energy spreads, though it does not go sideway (diff. mode is localized)

Signal space of diff. mode is well localized by reference vias – spreading is reduced





Differential vias through two parallel planes: Peak PFD at 10 GHz (example from "Distant Crosstalk")

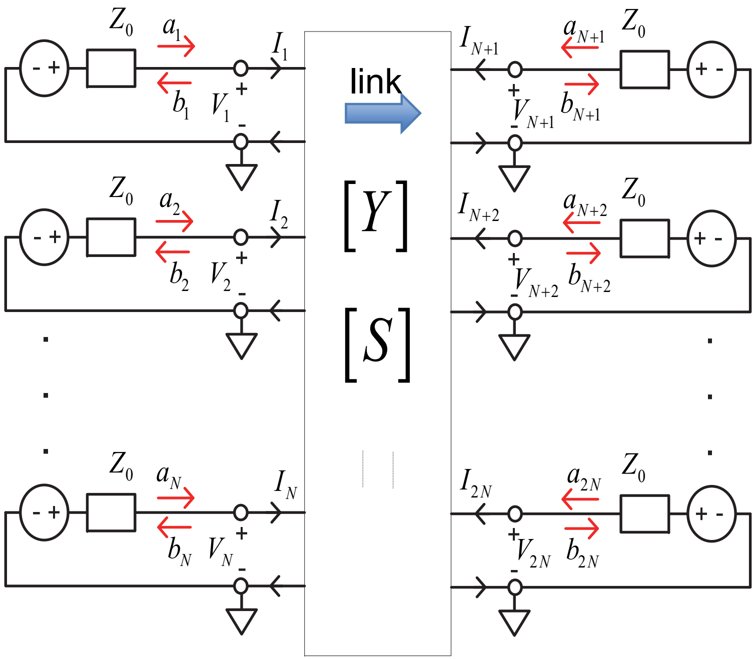
Multiports and S-parameters: From I & V to a & b

$\bar{V} = (V_1, V_2, \ldots, V_N)^t$ - vector of port voltages

$\bar{I} = (I_1, I_2, \ldots, I_N)^t$ - vector of port currents

$\bar{I} = Y \cdot \bar{V}$ - admittance matrix

$\bar{V} = Z \cdot \bar{I}$ - impedance matrix



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

$\bar{a} = \frac{1}{2} Z_0^{-1/2} \cdot (\bar{V} + Z_0 \cdot \bar{I})$ - vector of incoming waves

$\bar{b} = \frac{1}{2} Z_0^{-1/2} \cdot (\bar{V} - Z_0 \cdot \bar{I})$ - vector of outgoing waves

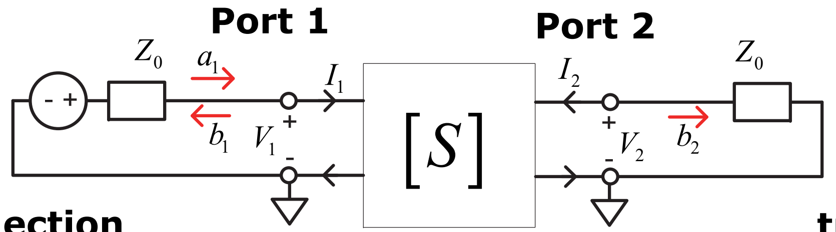
Scattering matrix (exists always):

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

More at P. J. Pupalaikis, S-parameters for Signal Integrity, Cambridge University Press, 2020.

Waves and Power

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



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

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

$v_i^+ = \sqrt{Z_0} \cdot a_i$ **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**

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

$P_i^+ = |a_i|^2$ **power of incoming wave**

$P_i^- = |b_i|^2$ **power of outgoing wave**

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

Magnitude is limited by 1 for passive systems!

S-parameters: dB and Voltages

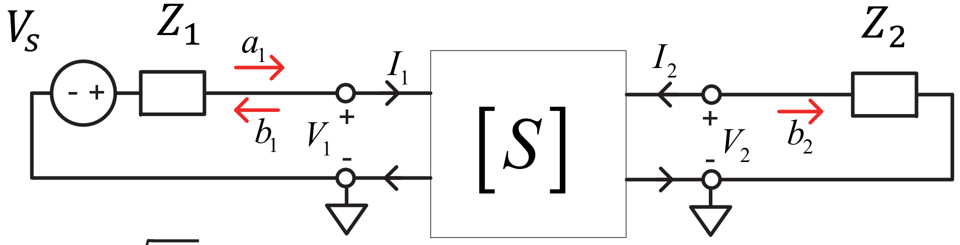
$$|S_{i,j}|_{dB} = 20 \cdot \log(|S_{i,j}|)$$ **magnitude in dB**

$$|S_{i,j}| = 10^{\frac{|S_{i,j}|_{dB}}{20}}$$

$a_j = 1\sqrt{Wt}$ $\quad V_{sj} = 1V$

$\vert S_{i,j}\vert_{dB}$	$\vert b_i\vert, \sqrt{\mathbf{Wt}}$	V_i, **V**
0dB	1.0	0.5
-3dB	0.708	0.354
-6dB	0.5	0.25
-20dB	0.1	0.05
-40dB	0.01	0.005
-60dB	0.001	0.0005

$$V_i = \sqrt{Z_i}(a_i + b_i) \qquad I_i = \frac{1}{\sqrt{Z_i}}(a_i - b_i)$$



$$V_1 = \sqrt{Z_1}(a_1 + b_1)$$

$$V_2 = \sqrt{Z_2} \cdot b_2 = \sqrt{Z_2} \cdot |S_{21}| a_1$$

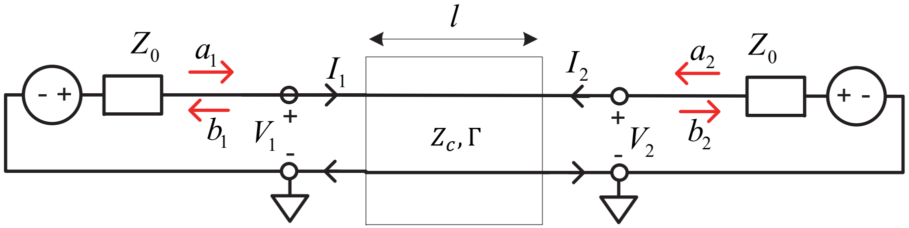
$$a_1 = \frac{V_s}{2\sqrt{Z_1}}$$

$$V_2 = \sqrt{\frac{Z_2}{Z_1}} \frac{V_s}{2} |S_{21}|$$

$$Z_1 = Z_2$$

$$V_2 = \frac{V_s}{2} |S_{21}| \quad \Rightarrow \quad V_2 = \frac{V_s}{2} 10^{\frac{|S_{21}|_{dB}}{20}}$$

S-parameters of Transmission Line Segment



$$S_{i,j} = \left.\frac{b_i}{a_j}\right|_{a_k=0k\neq j}$$

$$\Gamma = \sqrt{Z \cdot Y} = \alpha + i\frac{2\pi}{\Lambda}$$

$$Z_c = \sqrt{\frac{Z}{Y}} \qquad \Lambda = \frac{c}{f \cdot \sqrt{\varepsilon_{ef}}}$$

Solution is **superposition of two waves** propagating from port 1 to port 2 and back

$$S(f,l) = \begin{bmatrix} \frac{\left(Z_c^2 - Z_0^2\right)}{D} & \frac{2 \cdot Z_c \cdot Z_0 \cdot csh(\Gamma \cdot l)}{D} \\ \frac{2 \cdot Z_c \cdot Z_0 \cdot csh(\Gamma \cdot l)}{D} & \frac{\left(Z_c^2 - Z_0^2\right)}{D} \end{bmatrix}$$

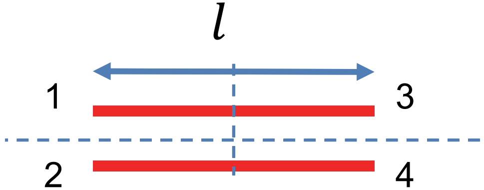
$$D = Z_c^2 + Z_0^2 + 2 \cdot Z_c \cdot Z_0 \cdot cth(\Gamma \cdot l)$$

For N+1 conductor lines the solution is **superposition of N waves** propagating in opposite directions

$$Z_c = Z_0 \;\Rightarrow\; S(f,l) = \begin{bmatrix} 0 & \exp(-\Gamma \cdot l) \\ \exp(-\Gamma \cdot l) & 0 \end{bmatrix}$$

More on S-parameters in Technical Presentation #2010_01… and on t-lines in App Note #2021_11

Even-Odd Mode Decomposition for Symmetric 2+1-Condictor T-Line Segment



$$S = \frac{1}{2}\begin{bmatrix} S_{11}^{ev}+S_{11}^{od} & S_{12}^{ev}+S_{12}^{od} & S_{11}^{ev}-S_{11}^{od} & S_{12}^{ev}-S_{12}^{od} \\ S_{12}^{ev}+S_{12}^{od} & S_{11}^{ev}+S_{11}^{od} & S_{12}^{ev}-S_{12}^{od} & S_{11}^{ev}-S_{11}^{od} \\ S_{11}^{ev}-S_{11}^{od} & S_{12}^{ev}-S_{12}^{od} & S_{11}^{ev}+S_{11}^{od} & S_{12}^{ev}+S_{12}^{od} \\ S_{12}^{ev}-S_{12}^{od} & S_{11}^{ev}-S_{11}^{od} & S_{12}^{ev}+S_{12}^{od} & S_{11}^{ev}+S_{11}^{od} \end{bmatrix}$$

p.u.l.: $Y = \begin{bmatrix} Y_{11} & Y_{12} \\ Y_{11} & Y_{11} \end{bmatrix}$ $Z = \begin{bmatrix} Z_{11} & Z_{12} \\ Z_{11} & Z_{11} \end{bmatrix}$

$$S^{ev} = \begin{bmatrix} \frac{\left(Z_{ev}^2 - Z_0^2\right)}{D_{ev}} & \frac{2 \cdot Z_{ev} \cdot Z_0 \cdot csh(\Gamma_{ev} \cdot l)}{D_{ev}} \\ \frac{2 \cdot Z_{ev} \cdot Z_0 \cdot csh(\Gamma_{ev} \cdot l)}{D_{ev}} & \frac{\left(Z_{ev}^2 - Z_0^2\right)}{D_{ev}} \end{bmatrix}$$

$$S^{od} = \begin{bmatrix} \frac{\left(Z_{od}^2 - Z_0^2\right)}{D_{od}} & \frac{2 \cdot Z_{od} \cdot Z_0 \cdot csh(\Gamma_{od} \cdot l)}{D_{od}} \\ \frac{2 \cdot Z_{ev} \cdot Z_0 \cdot csh(\Gamma_{od} \cdot l)}{D_{od}} & \frac{\left(Z_{od}^2 - Z_0^2\right)}{D_{od}} \end{bmatrix}$$

$$D_{ev} = Z_{ev}^2 + Z_0^2 + 2 \cdot Z_{ev} \cdot Z_0 \cdot cth(\Gamma_{ev} \cdot l)$$

$$D_{od} = Z_{od}^2 + Z_0^2 + 2 \cdot Z_{od} \cdot Z_0 \cdot cth(\Gamma_{od} \cdot l)$$

$$\Gamma_{ev} = \sqrt{(Z_{11} + Z_{12})(Y_{11} + Y_{12})} \qquad Z_{ev} = \sqrt{\frac{Z_{11} + Z_{12}}{Y_{11} + Y_{12}}}$$

$$\Gamma_{od} = \sqrt{(Z_{11} - Z_{12})(Y_{11} - Y_{12})} \qquad Z_{od} = \sqrt{\frac{Z_{11} - Z_{12}}{Y_{11} - Y_{12}}}$$

Even-odd decomposition in FD: R. Mongia, I. Bahl, P. Bhartia, RF and Microwave Coupled-Line Circuits, 1999.

Takeouts

- **Energy travels in the spaces, not on the traces (*)**
- Signal energy is distributed in space between traces and reference conductors - anything that gets into that space is coupled and may have xtalk

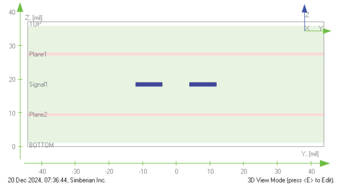
- **Interconnects are multiports – not circuits!**
- S-parameters in frequency domain is the most fundamental way to characterize multiports – learn it!

More on energy concept – (*) R. Morrison, Fast Circuit Boards: Energy Management, 2018

OUTLINE

- Introduction
- Basics: Fields and S-parameters
- Crosstalk Anatomy - Qualitative Analysis (almost)
- Crosstalk Quantification
- Distant Crosstalk - Sources and Mitigation
- Conclusion

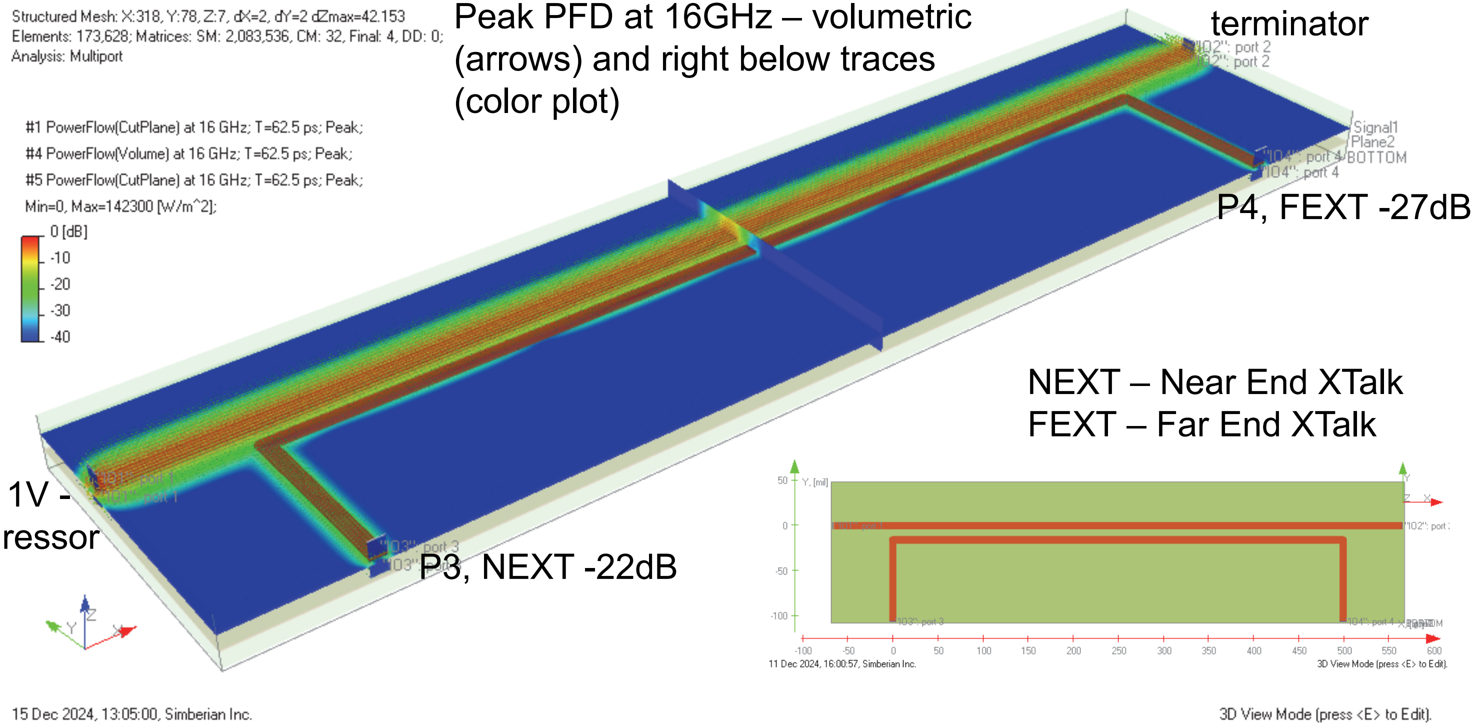
XTalk in Striplines (Same Layer)



8 mil traces (about 50 Ohm) coupled over 0.5 inch with 8 mil separation

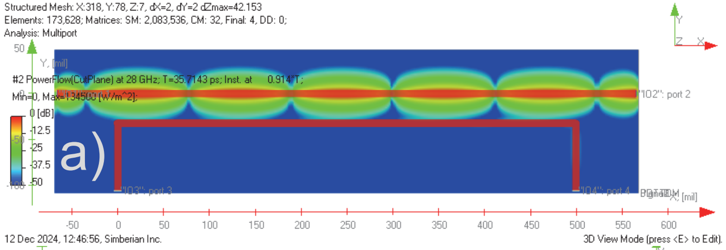
Project(1)
Materials: T=20[°C],...
"Copper", RR=1
"FR4_1", Dk=3.2, LT=0.002 @ 1 GHz, PLM=WD, Dk(0)=3.27,
"FR4_2", Dk=3.8, LT=0.002 @ 1 GHz, PLM=WD, Dk(0)=3.88,
"Air"
"SM", Dk=4, LT=0.002 @ 1 GHz, PLM=WD, Dk(0)=4.08, Dk(i
StackUp: LU=[mil], NL=5, T=37.14[mil]
1| Signal: "TOP", T=1.2, Fl="Air", Cond="Copper"
2| Medium: T=8, Fl="FR4_1"
3| Plane: "Plane1", Cond="Copper", T=0.77, Fl="FR4_1"
4| Medium: T=8, Fl="FR4_2"
5| Signal: "Signal1", T=1.2, Fl="FR4_1", Cond="Copper"
6| Medium: T=8, Fl="FR4_1"
7| Plane: "Plane2", Cond="Copper", T=0.77, Fl="FR4_1"
8| Medium: T=8, Fl="FR4_1"
9| Signal: "BOTTOM", T=1.2, Fl="Air", Cond="Copper"

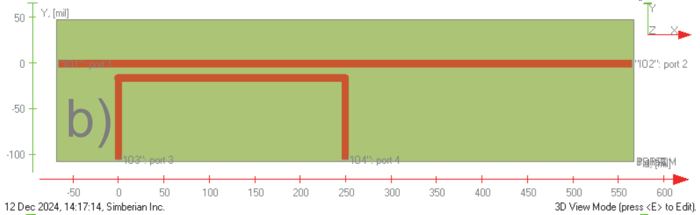
Simbeor Solution:
XTalk_FEXT_NEXT_SL_2019_03a

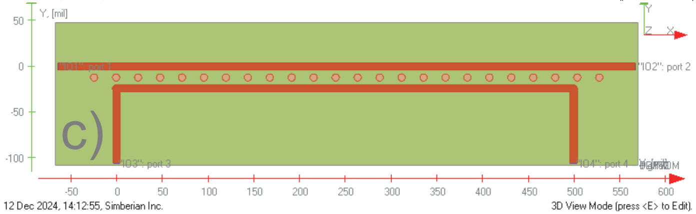


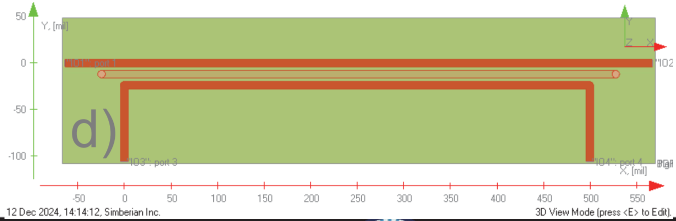
•#2019_03: **How Interconnects Work™**: Crosstalk in striplines and how to reduce it - visualization of coupling with power flow density, electric and magnetic fields and current density, 17 min; YouTube: https://youtu.be/0VFmVp8LbTQ

XTalk in Striplines – How to Reduce IT?









a) Increase separation?

b) Reduce coupling length?

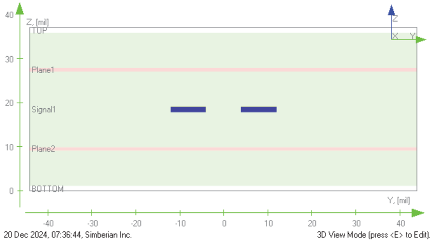
c) Use via fence?

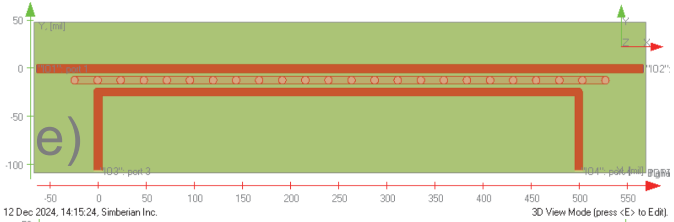
d) Use guarding trace?

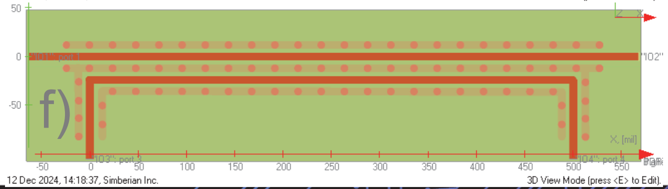
e) Use guarding trace with via fence?

f) Use coplanar strips with via fence?

What else?

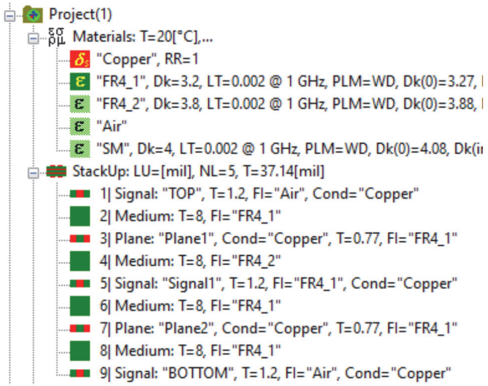


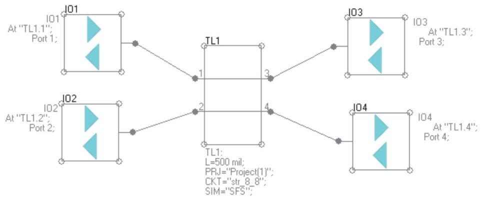


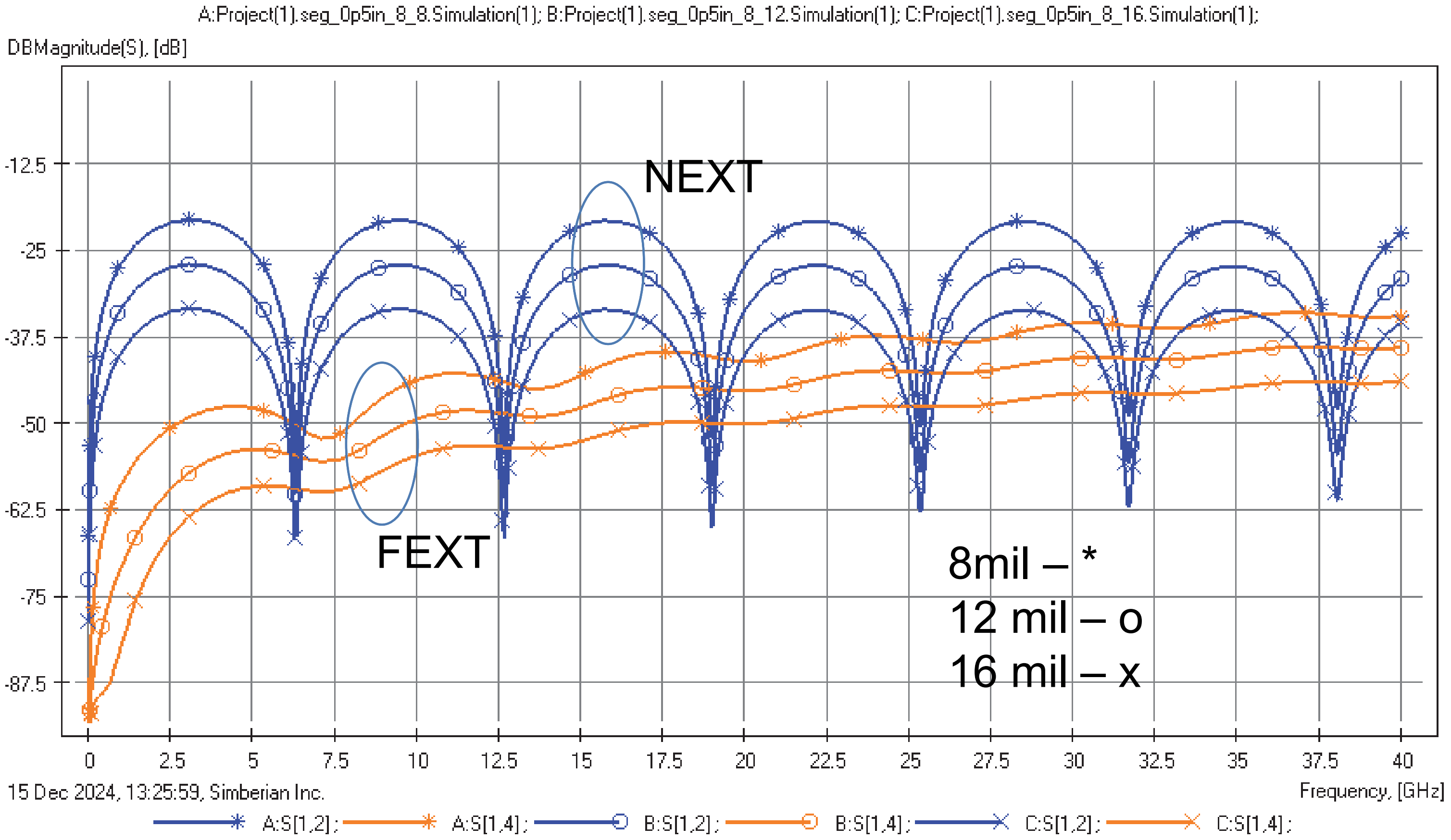


Stripline – XTalk Reduction by Distance

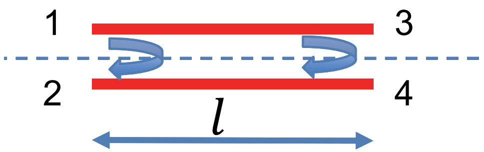
8 mil traces (about 50 Ohm) coupled over 0.5 inch with 8 , 12 and 16 mil separations
Straight segments - No Bends!







NEXT Modal Decomposition for Symmetric Strips



$\varepsilon_{ef} = \varepsilon_{od} \approx \varepsilon_e$

Modal decomposition: odd (+-) and even (++) modes – exact solution:

$$S_{11}^{ev} = \frac{Z_{ev}^2 - Z_0^2}{Z_{ev}^2 + Z_0^2 + 2 \cdot Z_{ev} \cdot Z_0 \cdot cth(\Gamma_{ev} \cdot l)}$$

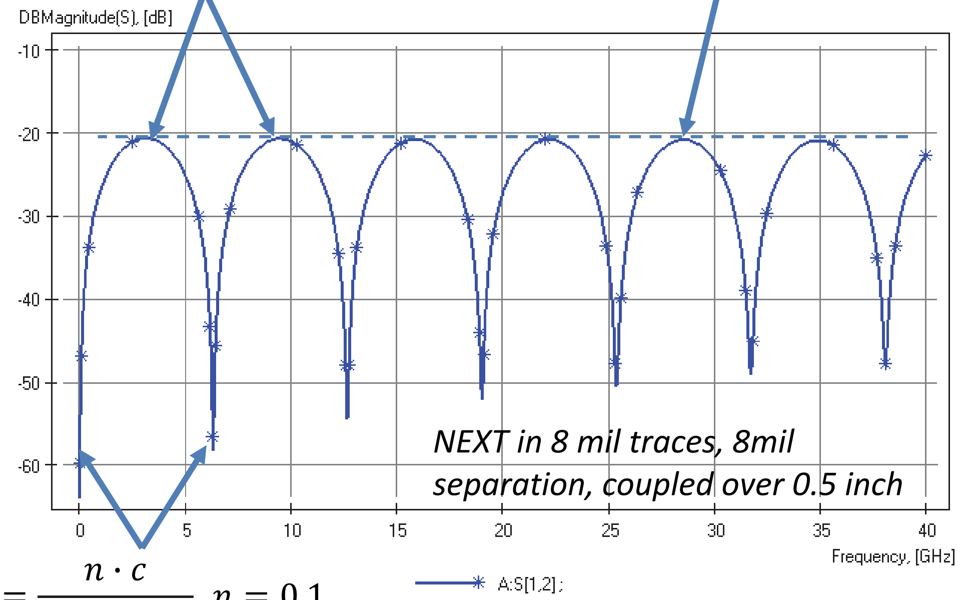
$$S_{11}^{od} = \frac{Z_{od}^2 - Z_0^2}{Z_{od}^2 + Z_0^2 + 2 \cdot Z_{od} \cdot Z_0 \cdot cth(\Gamma_{od} \cdot l)}$$

Exact equation for NEXT:

$$S_N = \frac{S_{11}^{ev} - S_{11}^{od}}{2}$$

$$f_{max} = \frac{(1+2n) \cdot c}{4 \cdot l \cdot \sqrt{\varepsilon_{ef}}}, \text{n} = 0,1 \ldots$$

$$|S_N|_{max} = 20log\left(\frac{1}{2}\left[\frac{Z_{ev}^2 - Z_0^2}{Z_{ev}^2 + Z_0^2} - \frac{Z_{od}^2 - Z_0^2}{Z_{od}^2 + Z_0^2}\right]\right)$$



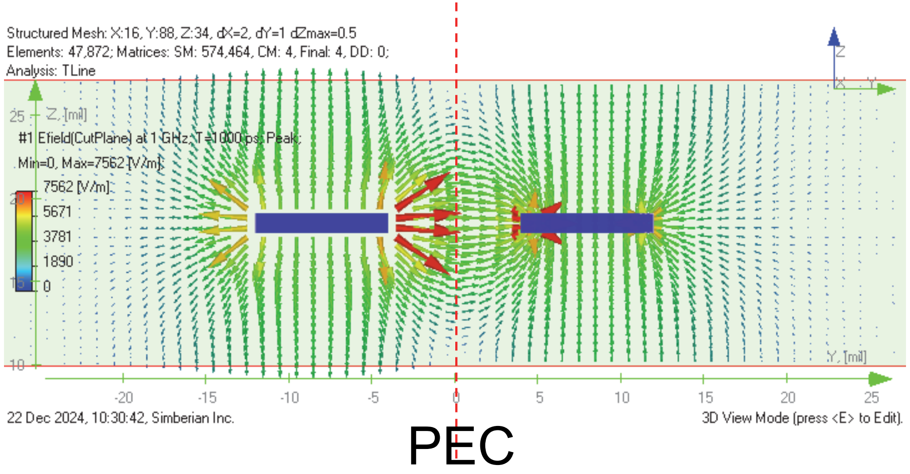
$$f_{min} = \frac{n \cdot c}{2 \cdot l \cdot \sqrt{\varepsilon_{ef}}}, n = 0,1 \ldots$$

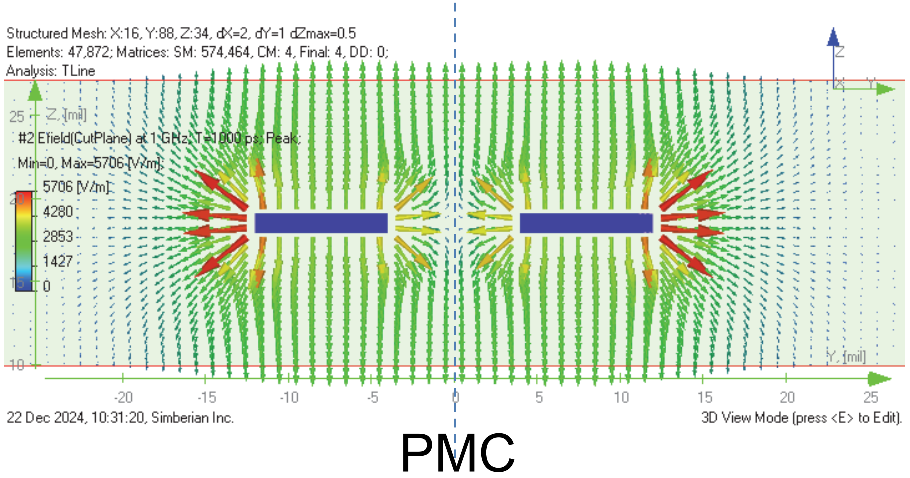
Even-odd decomposition in FD: R. Mongia, I. Bahl, P. Bhartia, RF and Microwave Coupled-Line Circuits, 1999.

Waves or Modes in Symmetric Stripline

Odd Mode $\begin{pmatrix} \frac{+1}{\sqrt{2}} & \frac{-1}{\sqrt{2}} \end{pmatrix}$

Even Mode $\begin{pmatrix} \frac{+1}{\sqrt{2}} & \frac{+1}{\sqrt{2}} \end{pmatrix}$





Which one has lower impedance? (Hint: PEC is ideal conductor)

Modal Impedance and NEXT Minimization

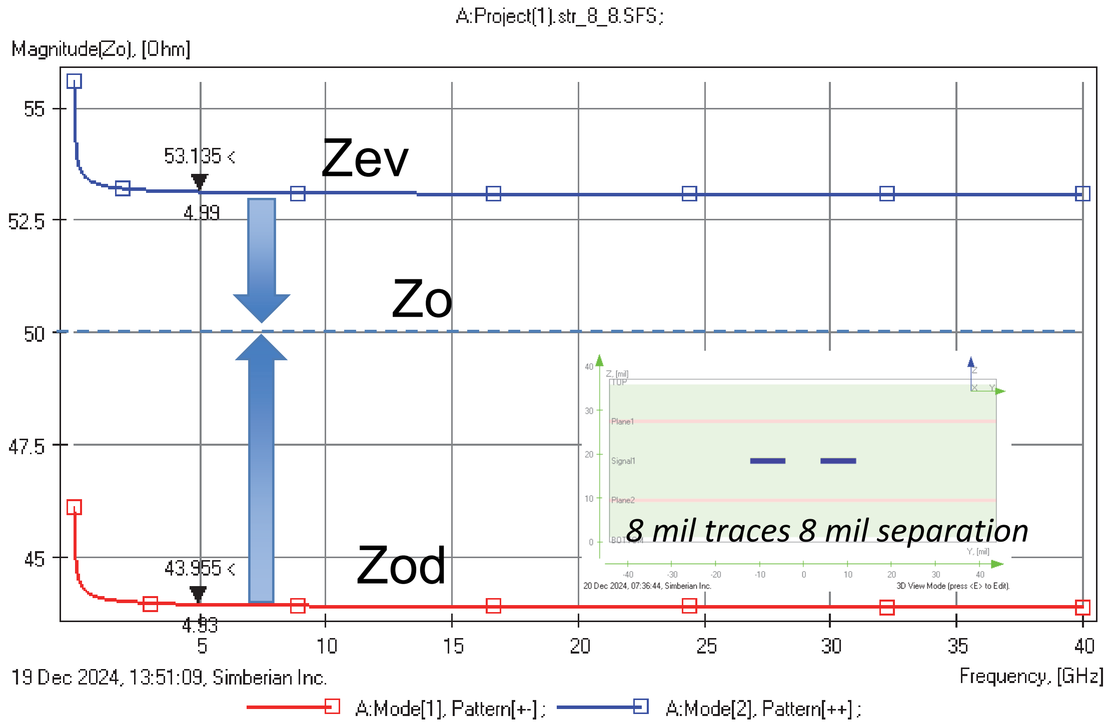
$$|S_N|_{max} = \frac{1}{2}\left[\frac{Z_{ev}^2 - Z_0^2}{Z_{ev}^2 + Z_0^2} - \frac{Z_{od}^2 - Z_0^2}{Z_{od}^2 + Z_0^2}\right]$$

$$Z_{od} = \sqrt{\frac{Z_{11} - Z_{12}}{Y_{11} - Y_{12}}} \qquad Z_{ev} = \sqrt{\frac{Z_{11} + Z_{12}}{Y_{11} + Y_{12}}}$$

Z and Y are p.u.l. 2x2 impedance and admittance matrices

NEXT is zero if Zev = Zod

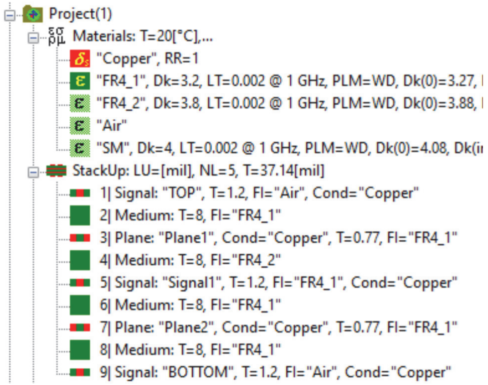
How to achieve that?

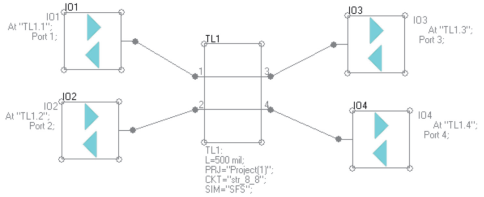


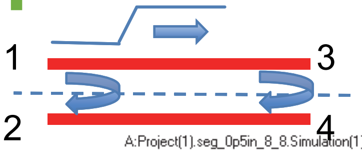
#2007_01: Y. Shlepnev, Broadband transmission line models for analysis of serial data channel interconnects, PCB Design Conference East, Durham NC, October 23, 2007.

NEXT in Time Domain

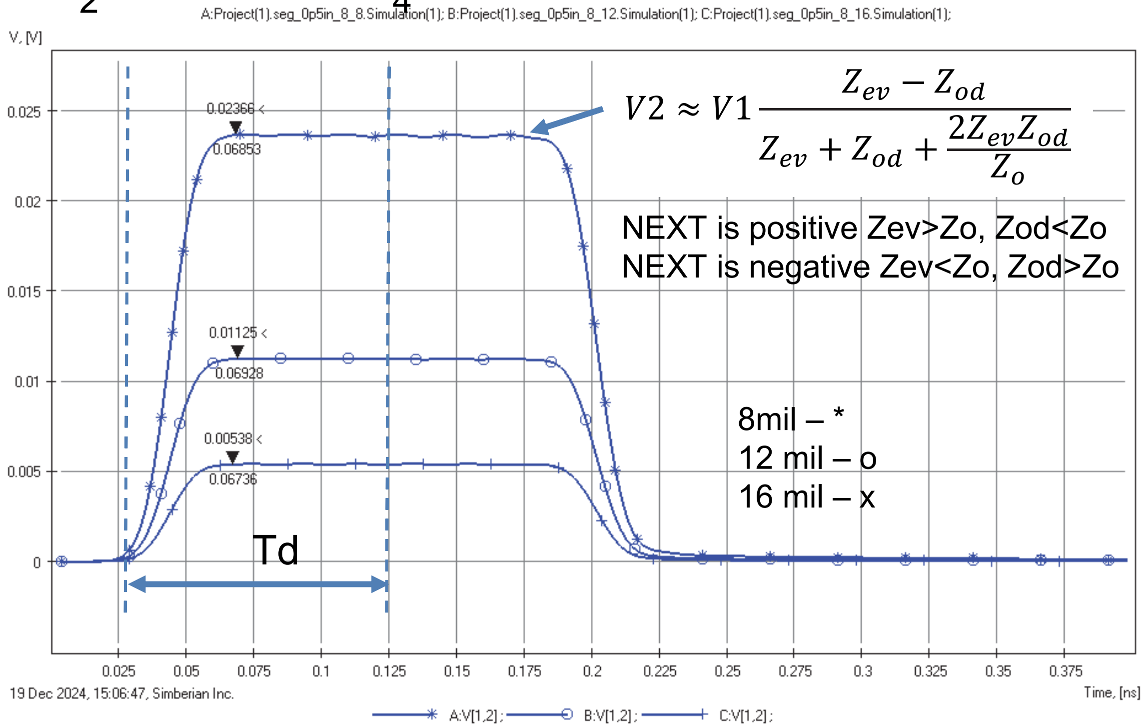
8 mil traces (about 50 Ohm) coupled over 0.5 inch with 8 , 12 and 16 mil separations
1V in series with 50 Ohm – V1=0.5V
20ps (10-90%) Gaussian step response







At what length NEXT is reduced?

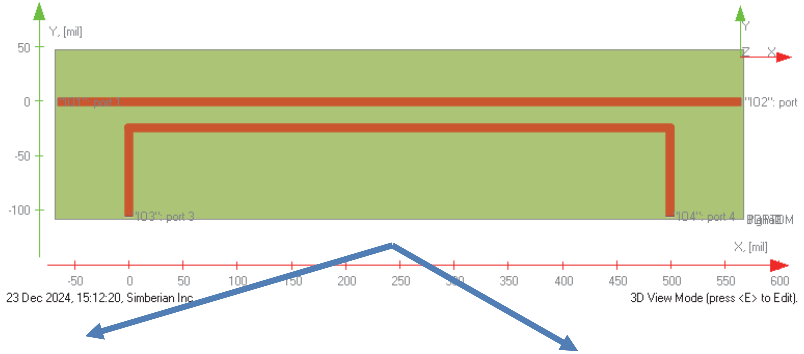


$$V2 \approx V1 \frac{Z_{ev} - Z_{od}}{Z_{ev} + Z_{od} + \frac{2Z_{ev}Z_{od}}{Z_o}}$$

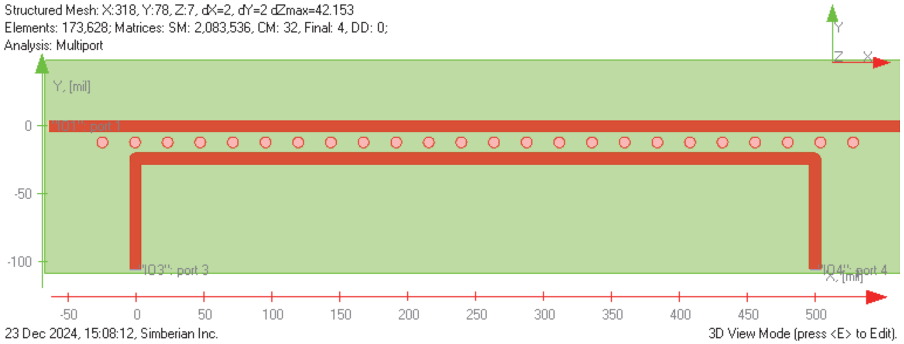
NEXT is positive Zev>Zo, Zod<Zo
NEXT is negative Zev<Zo, Zod>Zo

Even-odd decomposition in TD: B. Young, Digital Signal Integrity – Modeling and Simulation with Interconnects and Packages, 2000

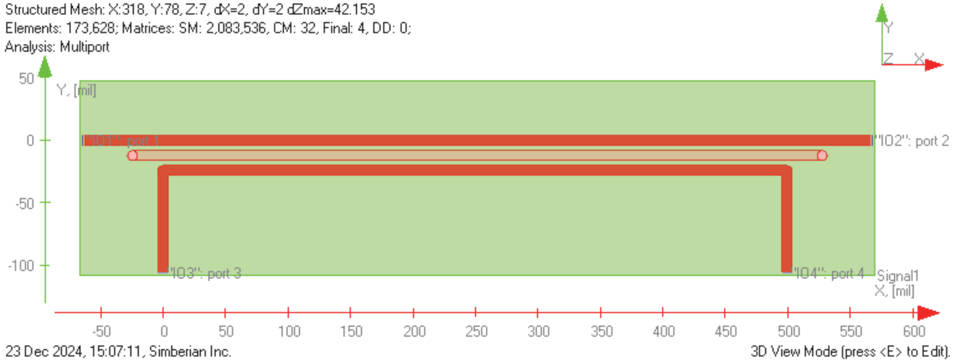
Will This Work?



c) Just stitching vias?

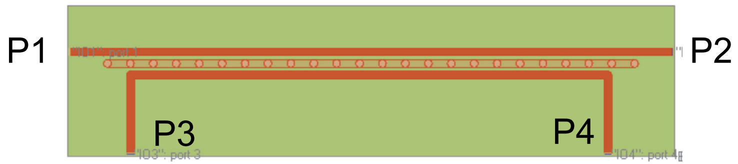


d) Guard trace with just 2 stitching vias?

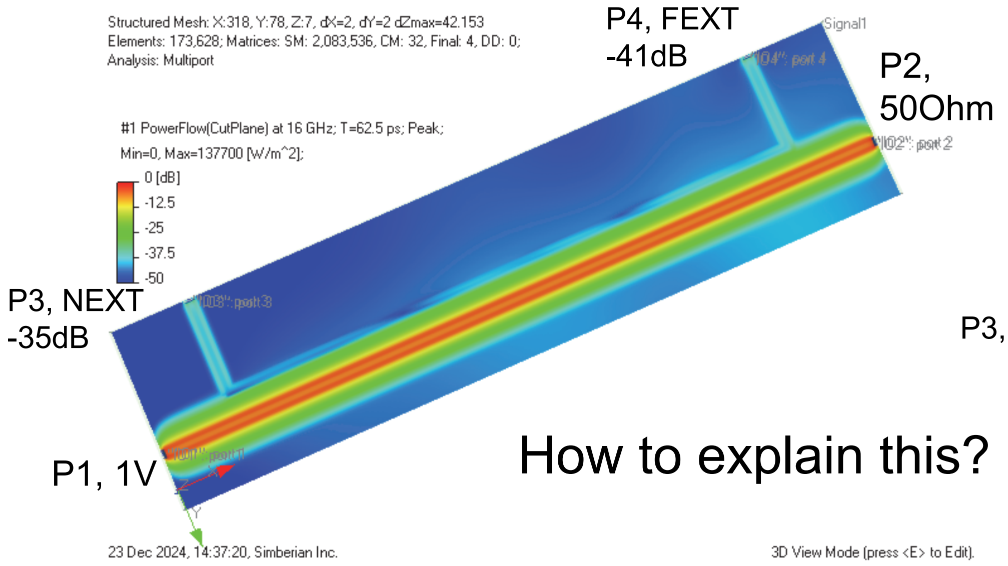


Will it reduce coupling, comparing to the separation? – Hint: NOPE

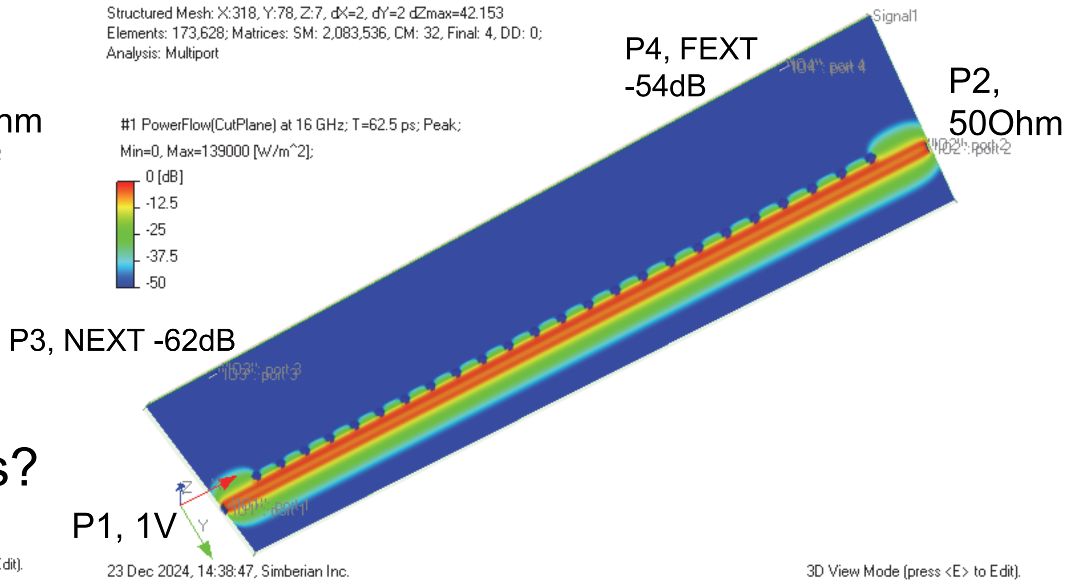
Striplines XTalk Reduction with Proper Guarding



Peak PFD at 16 GHz right below traces

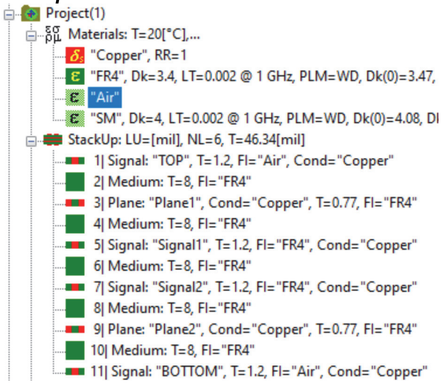


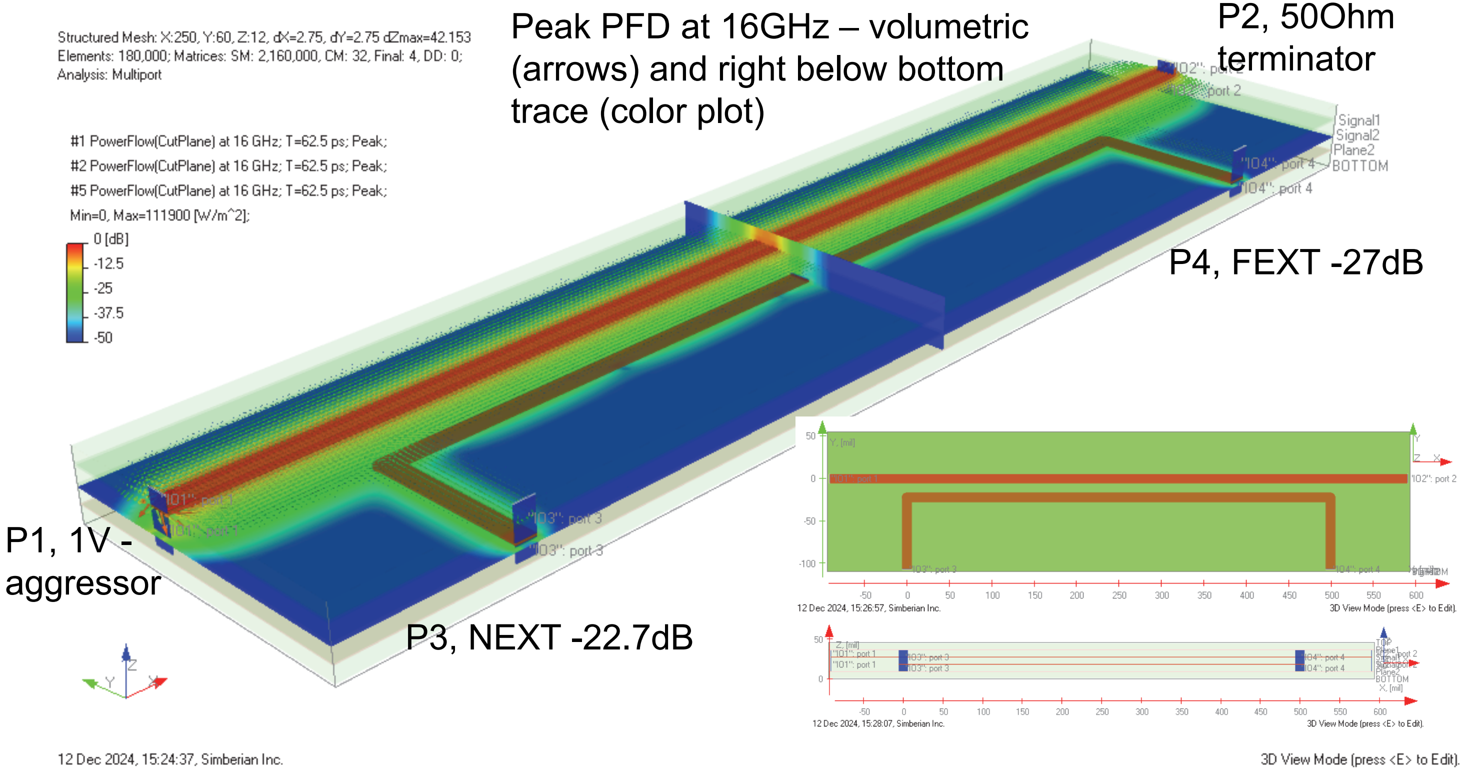
How to explain this?



XTalk in Broadside Coupled Striplines

11 mil traces (about 50 Ohm) in Signal1 and Signal2 coupled over 0.5 inch with 11 mil edge to edge separation





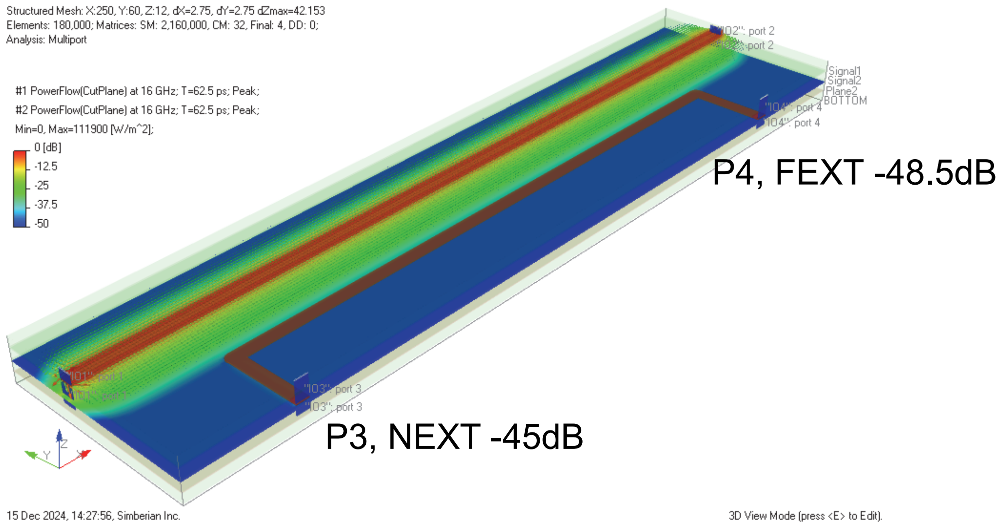
Simbeor Solution:
XTalk_FEXT_NEXT_BSC_2019_05

#2019_05: **How Interconnects Work™**: Crosstalk in adjacent striplines and how to reduce it - visualization with power flow density, 14 min; YouTube: https://youtu.be/7t5WYyf8tss

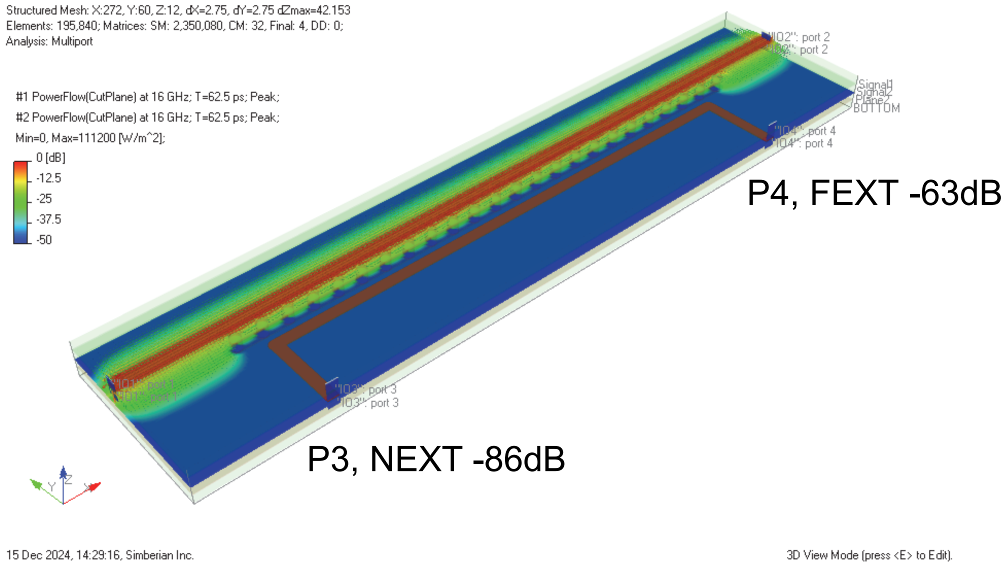
Broadside Coupled Traces - XTalk Reduction

Peak PFD at 16GHz – volumetric (arrows) and right below bottom trace (color plots)

11 mil traces in Signal1 and Signal2 coupled over 0.5 inch with 33 mil edge to edge separation



11 mil traces in Signal1 and Signal2 coupled over 0.5 inch with 22 mil edge to edge separation and two guard traces with via fence



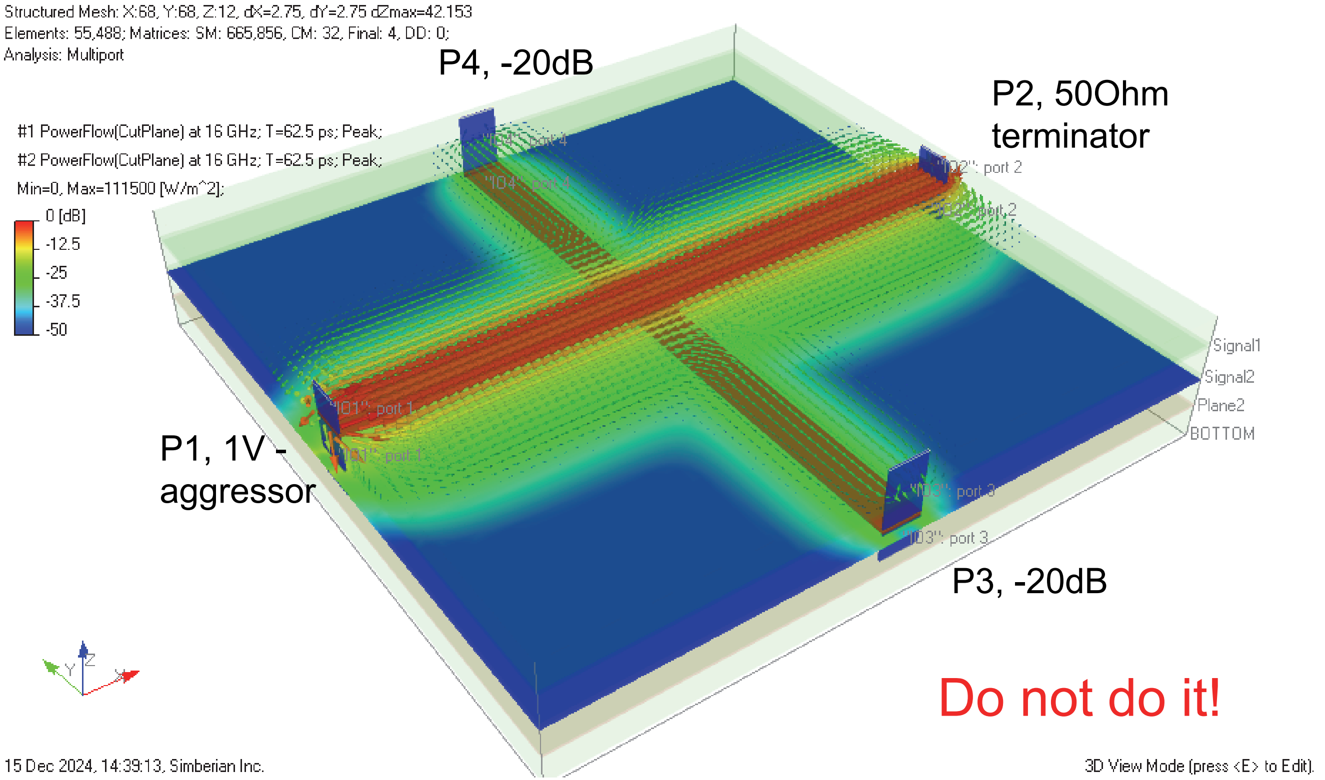
Orthogonal Routing – Does It Reduce XTalk?

Tow orthogonal 11 mil traces (about 50 Ohm) in Signal1 and Signal2

- Project(1)
  - Materials: T=20[°C],...
    - "Copper", RR=1
    - "FR4", Dk=3.4, LT=0.002 @ 1 GHz, PLM=WD, Dk(0)=3.47,
    - "Air"
    - "SM", Dk=4, LT=0.002 @ 1 GHz, PLM=WD, Dk(0)=4.08, DI
  - StackUp: LU=[mil], NL=6, T=46.34[mil]
    - 1| Signal: "TOP", T=1.2, Fl="Air", Cond="Copper"
    - 2| Medium: T=8, Fl="FR4"
    - 3| Plane: "Plane1", Cond="Copper", T=0.77, Fl="FR4"
    - 4| Medium: T=8, Fl="FR4"
    - 5| Signal: "Signal1", T=1.2, Fl="FR4", Cond="Copper"
    - 6| Medium: T=8, Fl="FR4"
    - 7| Signal: "Signal2", T=1.2, Fl="FR4", Cond="Copper"
    - 8| Medium: T=8, Fl="FR4"
    - 9| Plane: "Plane2", Cond="Copper", T=0.77, Fl="FR4"
    - 10| Medium: T=8, Fl="FR4"
    - 11| Signal: "BOTTOM", T=1.2, Fl="Air", Cond="Copper"

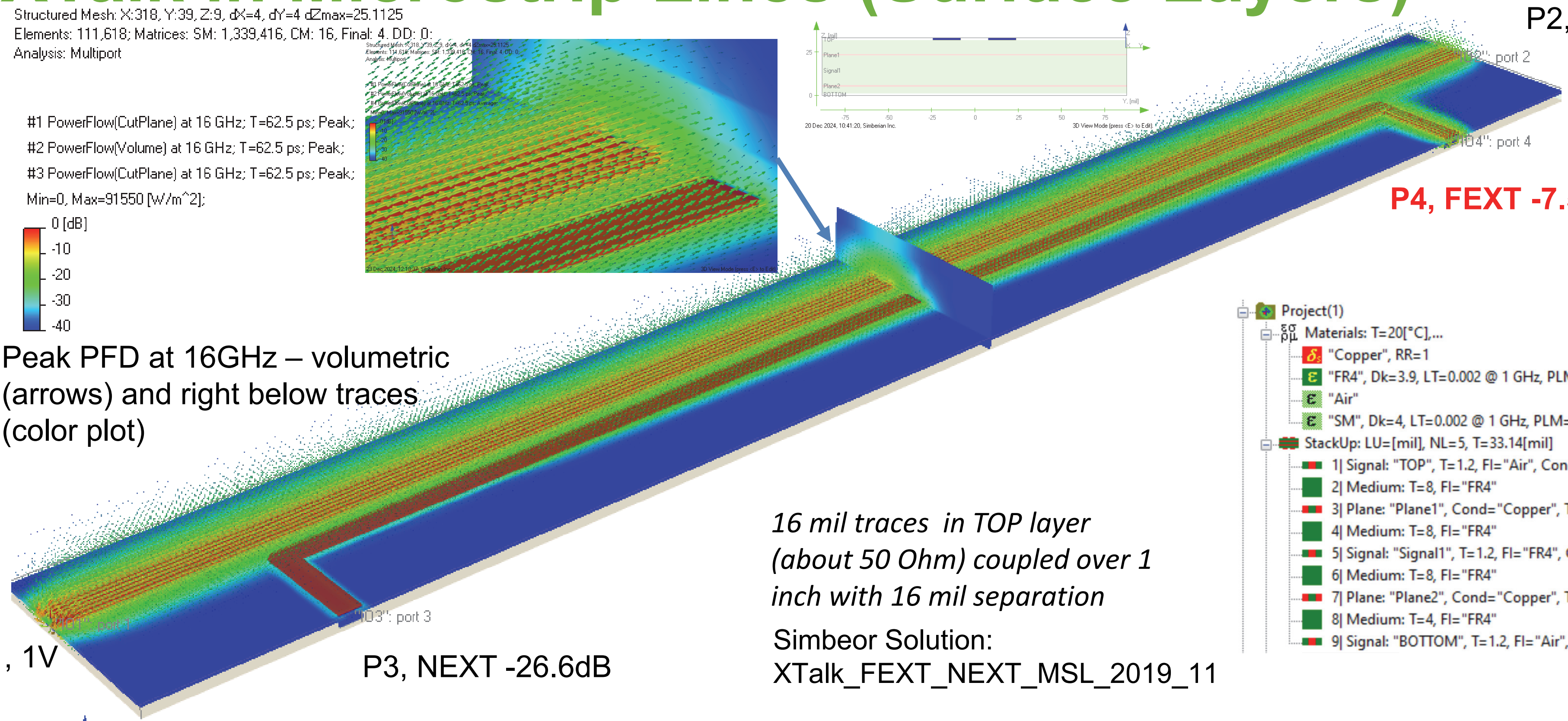
Simbeor Solution: XTalk_FEXT_NEXT_BSC_2019_05

Peak PFD at 16GHz – volumetric (arrows) and right below bottom trace (color plots)



Do not do it!

XTalk in Microstrip Lines (Surface Layers)



P2, 50Ohm

P4, FEXT -7.5dB

Peak PFD at 16GHz – volumetric (arrows) and right below traces (color plot)

- Project(1)
  - Materials: T=20[°C],...
    - "Copper", RR=1
    - "FR4", Dk=3.9, LT=0.002 @ 1 GHz, PLM=WD, Dk(0)=3.98, I
    - "Air"
    - "SM", Dk=4, LT=0.002 @ 1 GHz, PLM=WD, Dk(0)=4.08, Dk
  - StackUp: LU=[mil], NL=5, T=33.14[mil]
    - 1| Signal: "TOP", T=1.2, Fl="Air", Cond="Copper"
    - 2| Medium: T=8, Fl="FR4"
    - 3| Plane: "Plane1", Cond="Copper", T=0.77, Fl="FR4"
    - 4| Medium: T=8, Fl="FR4"
    - 5| Signal: "Signal1", T=1.2, Fl="FR4", Cond="Copper"
    - 6| Medium: T=8, Fl="FR4"
    - 7| Plane: "Plane2", Cond="Copper", T=0.77, Fl="FR4"
    - 8| Medium: T=4, Fl="FR4"
    - 9| Signal: "BOTTOM", T=1.2, Fl="Air", Cond="Copper"

16 mil traces in TOP layer (about 50 Ohm) coupled over 1 inch with 16 mil separation

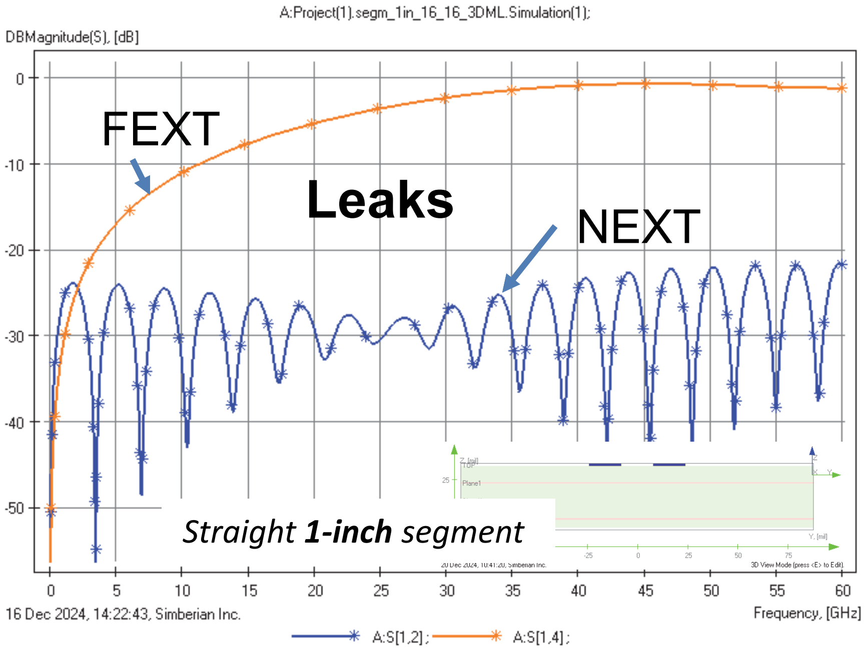
Simbeor Solution: XTalk_FEXT_NEXT_MSL_2019_11

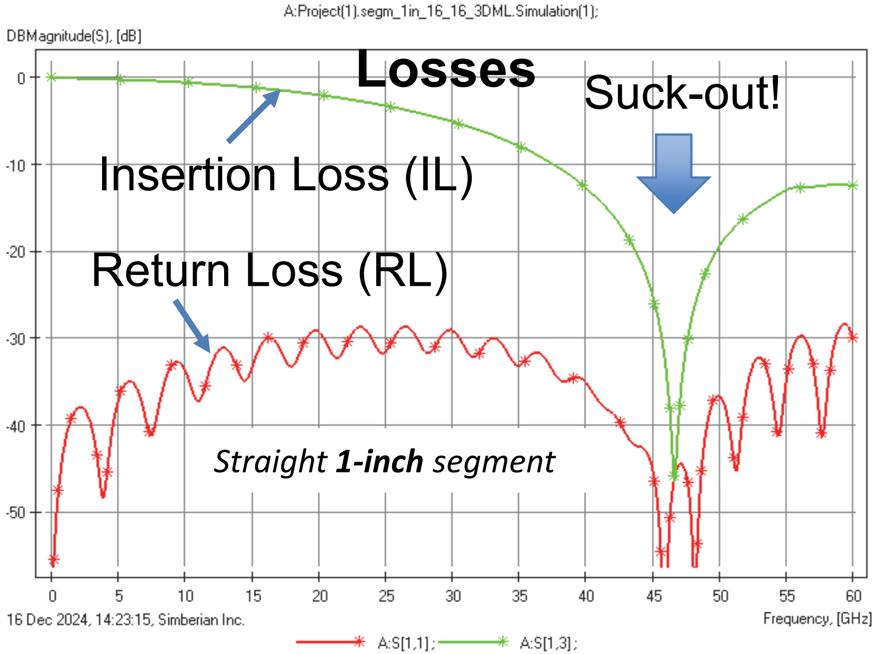
P1, 1V

P3, NEXT -26.6dB

- #2019_11: **How Interconnects Work™**: Crosstalk in microstrip lines and how to reduce it, 12 min
- YouTube: https://youtu.be/OQx3habvfgM

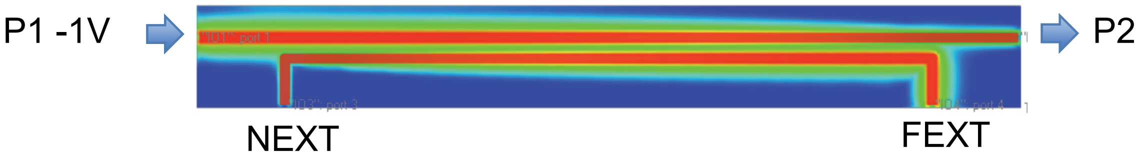
Leak and XTalk in 1-inch MSL – “Suck-out”





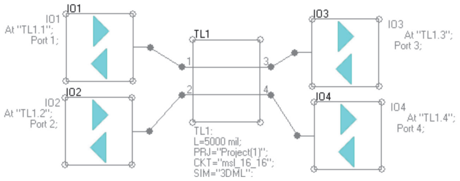
*16 mil traces in TOP layer (about 50 Ohm) coupled over **1 inch** with 16 mil separation*

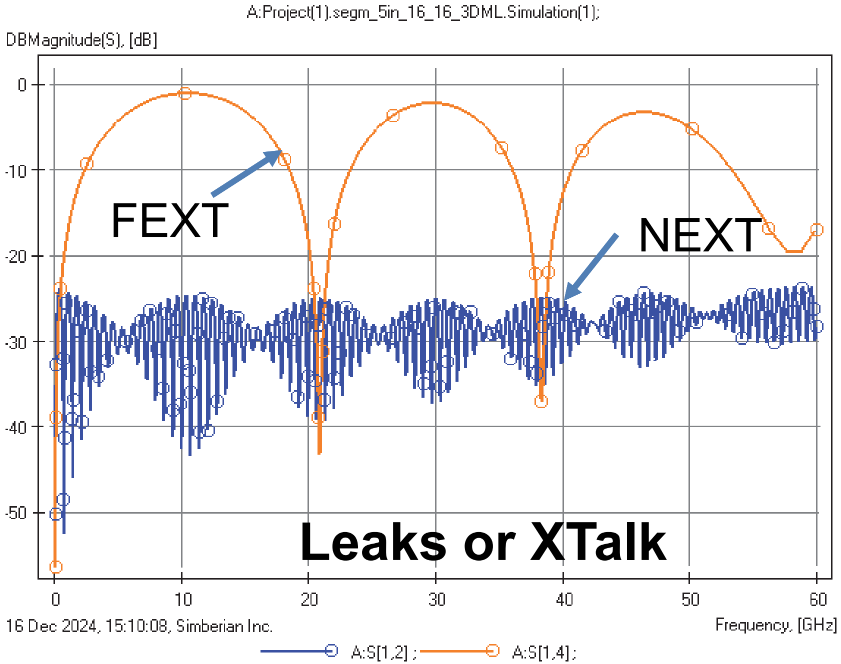
Peak PFD at 47GHz

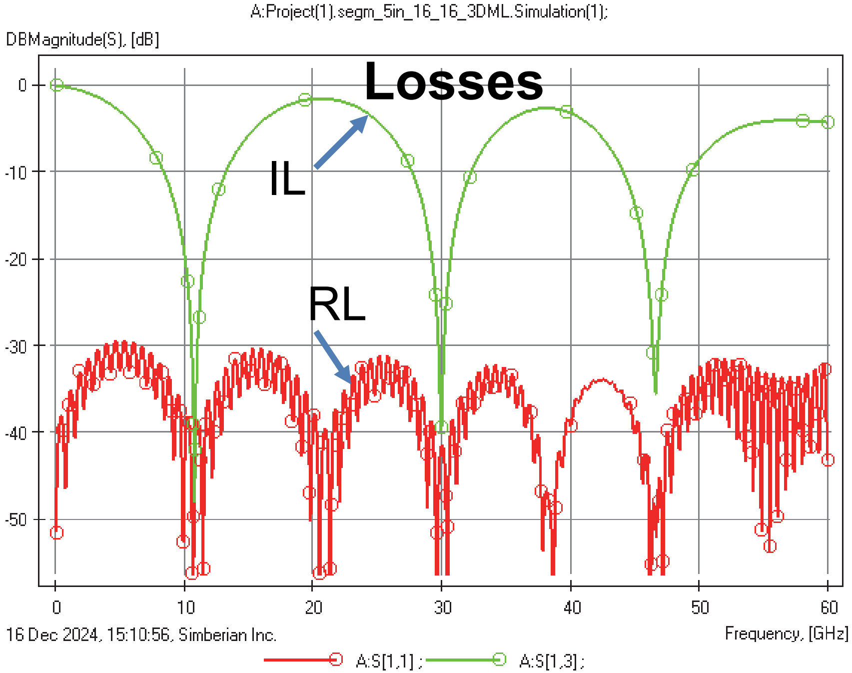


Leak and XTalk in 5-inch MSL

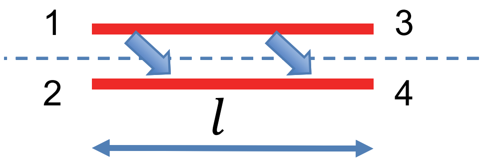
*16 mil traces in TOP layer (about 50 Ohm) separated by 16 mil and coupled over straight **5-inch segment***

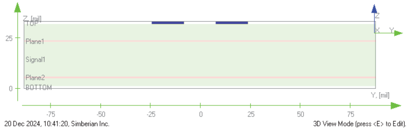






FEXT Modal Decomposition for Symmetric Strips





*16 mil traces in TOP layer (about 50 Ohm) coupled over **5 inch** with 16 mil separation (straight segment – no bends)*

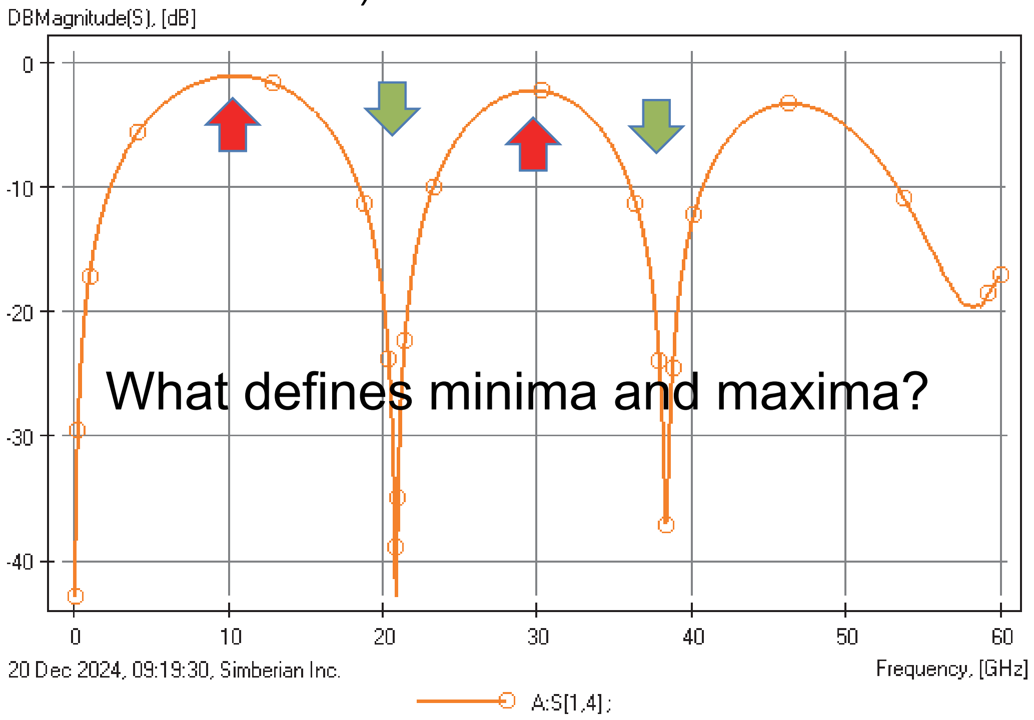
Modal decomposition: odd (+-) and even (++) modes – exact solution:

$$S_{12}^{ev} = \frac{2 \cdot Z_{ev} \cdot Z_0 \cdot csh(\Gamma_{ev} \cdot l)}{Z_{ev}^2 + Z_0^2 + 2 \cdot Z_{ev} \cdot Z_0 \cdot cth(\Gamma_{ev} \cdot l)}$$

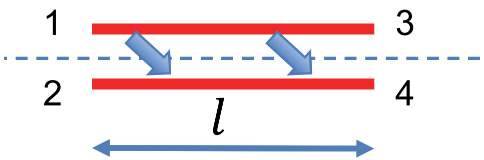
$$S_{11}^{od} = \frac{2 \cdot Z_{od} \cdot Z_0 \cdot csh(\Gamma_{od} \cdot l)}{Z_{od}^2 + Z_0^2 + 2 \cdot Z_{od} \cdot Z_0 \cdot cth(\Gamma_{od} \cdot l)}$$

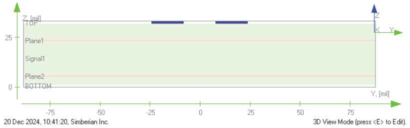
Exact equation for FEXT:

$$S_F = \frac{S_{12}^{ev} - S_{12}^{od}}{2}$$



FEXT Minima and Maxima





$$f_{max} = \frac{(1+2n)\cdot c}{2\cdot l\cdot\left|\sqrt{\varepsilon_{ev}}-\sqrt{\varepsilon_{od}}\right|}, \text{n} = 0,1\ldots$$

FEXT in case of **no reflections (no NEXT)**:

$$S_F = \frac{S_{12}^{ev} - S_{12}^{od}}{2} \approx \frac{e^{-\Gamma_{ev}} - e^{-\Gamma_{od}}}{2}$$

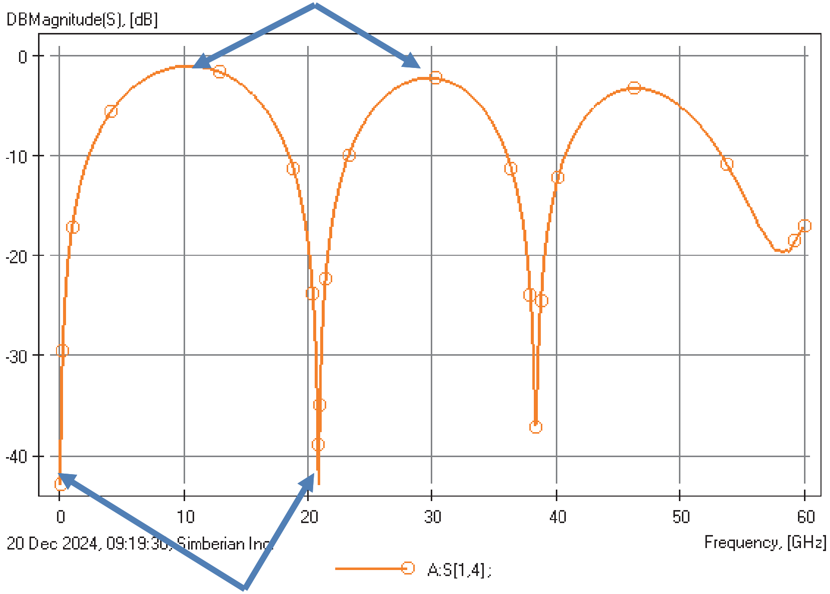
FEXT in case of **no losses**:

$$S_F \approx -i\cdot \underline{sin\left(\pi f l\left(\sqrt{\varepsilon_{ev}}-\sqrt{\varepsilon_{od}}\right)/c\right)}\cdot e^{i\pi f l\left(\sqrt{\varepsilon_{ev}}-\sqrt{\varepsilon_{od}}\right)/c}$$

$$\Gamma_{ev} = \sqrt{(Z_{11}+Z_{12})(Y_{11}+Y_{12})} \approx \frac{i2\pi f\sqrt{\varepsilon_{ev}}}{c}$$

$$\Gamma_{od} = \sqrt{(Z_{11}-Z_{12})(Y_{11}-Y_{12})} \approx \frac{i2\pi f\sqrt{\varepsilon_{od}}}{c}$$

Z and Y are 2x2 p.u.l. impedance and admittance matrices

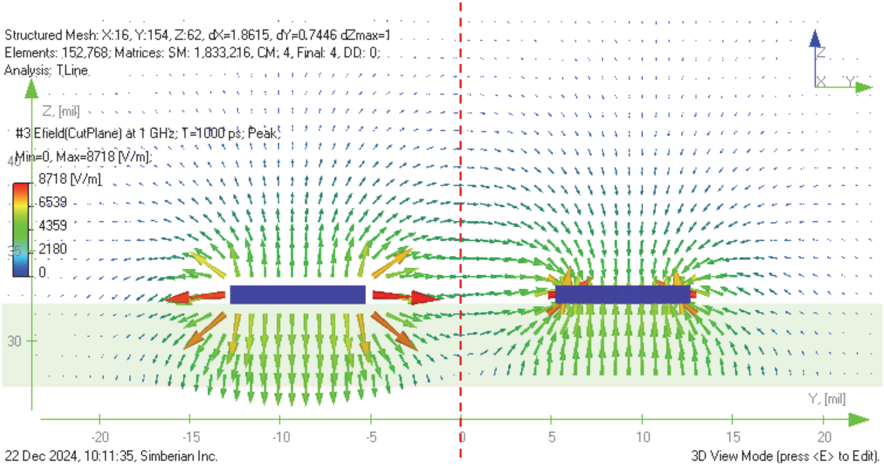


$$f_{min} = \frac{n\cdot c}{l\cdot\left|\sqrt{\varepsilon_{ev}}-\sqrt{\varepsilon_{od}}\right|}, n = 0,1\ldots$$

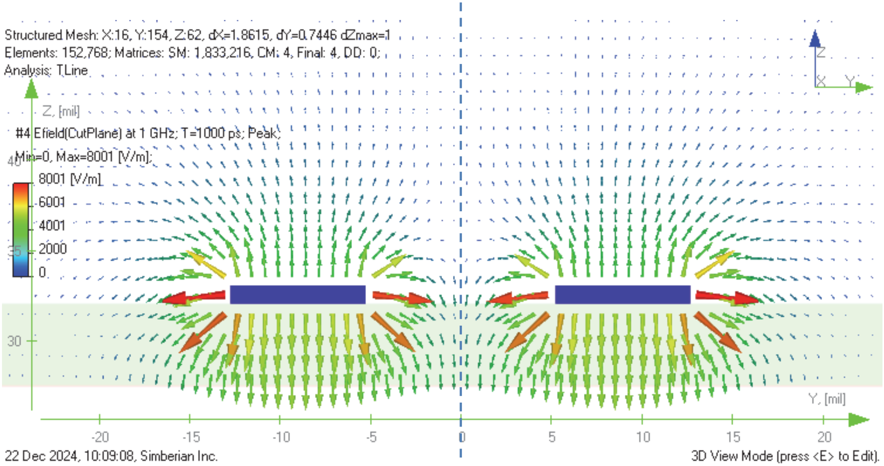
Waves or Modes in Symmetric MSL

Odd Mode $\begin{pmatrix} \frac{+1}{\sqrt{2}} & \frac{-1}{\sqrt{2}} \end{pmatrix}$

Even Mode $\begin{pmatrix} \frac{+1}{\sqrt{2}} & \frac{+1}{\sqrt{2}} \end{pmatrix}$



PEC



PMC

Which one is faster or have lower ε_{eff} ?

Hint: Look at the energy…

FEXT Modal Decomposition and Minimization

x

$$\begin{pmatrix}1\\0\end{pmatrix} = \begin{pmatrix}\frac{+1}{2}\\\frac{-1}{2}\end{pmatrix} e^{-\Gamma_{od}\cdot x} + \begin{pmatrix}\frac{+1}{2}\\\frac{+1}{2}\end{pmatrix} e^{-\Gamma_{ev}\cdot x}$$

l

$$\Gamma_{ev} \approx \frac{i2\pi f\sqrt{\varepsilon_{ev}}}{c}$$

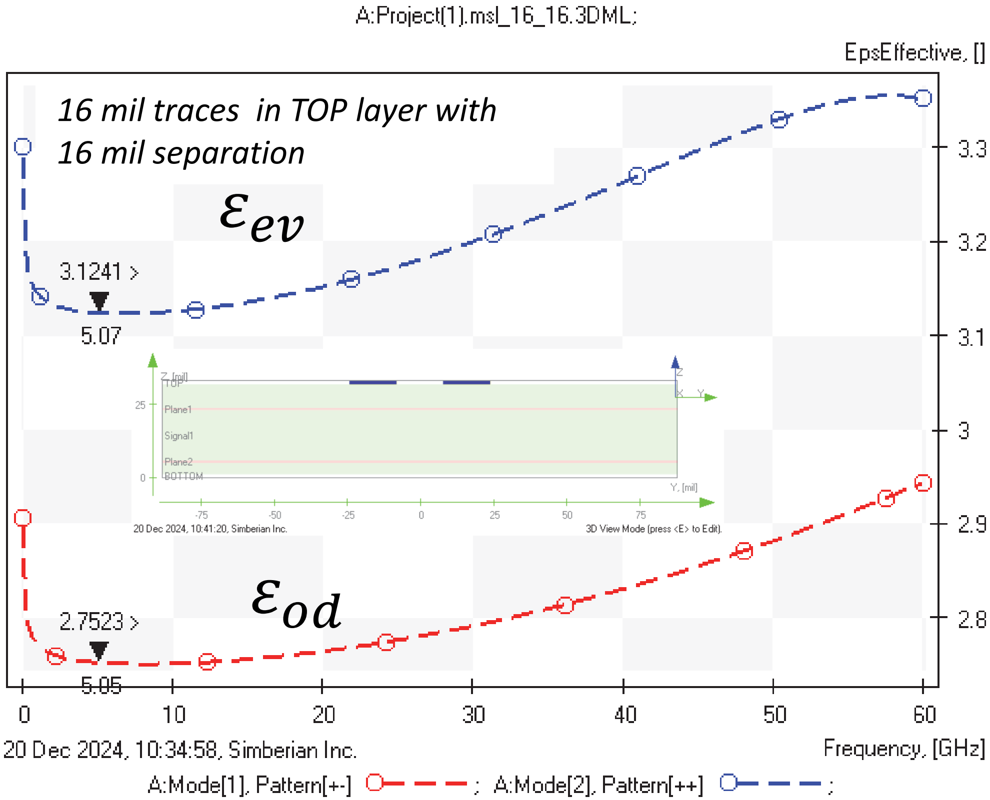
$$\Gamma_{od} \approx \frac{i2\pi f\sqrt{\varepsilon_{od}}}{c}$$

$$S_F \approx -i\cdot sin(\pi f l(\sqrt{\varepsilon_{ev}}-\sqrt{\varepsilon_{od}})/c)\cdot e^{i\pi f l(\sqrt{\varepsilon_{ev}}-\sqrt{\varepsilon_{od}})/c}$$

FEXT is zero if phase delay or velocity of even and odd modes are equal

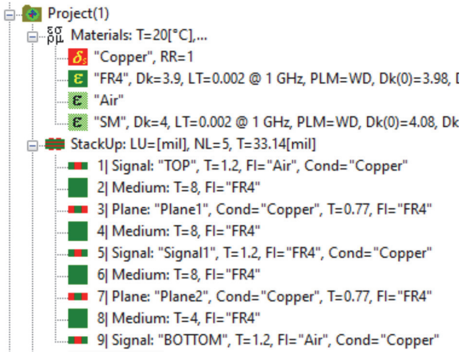
$$\varepsilon_{ev} = \varepsilon_{od}$$

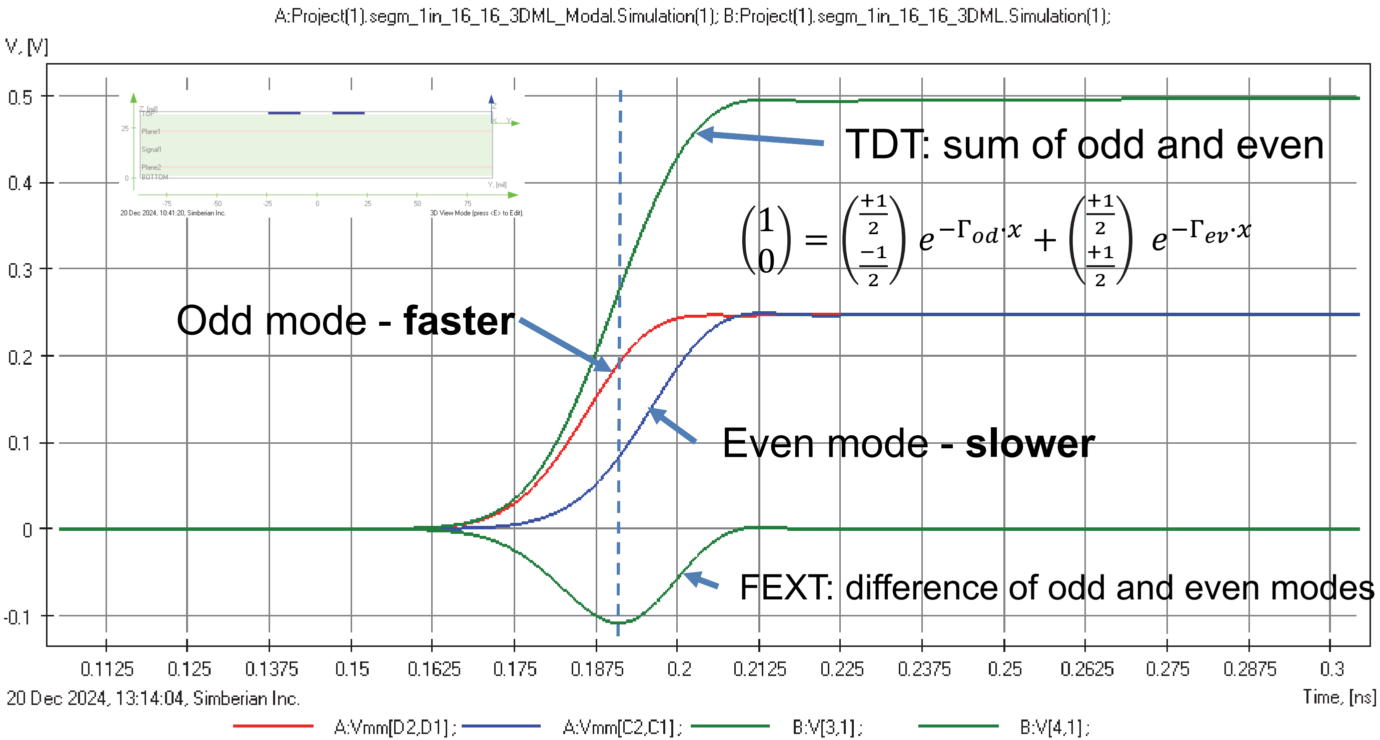
How to achieve that?



FEXT in Time Domain

*16 mil traces in TOP layer (about 50 Ohm) coupled over **1 inch** with 16 mil separation*
1V in series with 50 Ohm – V1=0.5V
20ps (10-90%) Gaussian step response



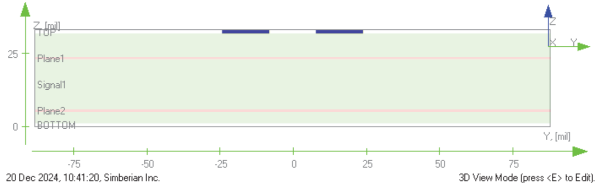


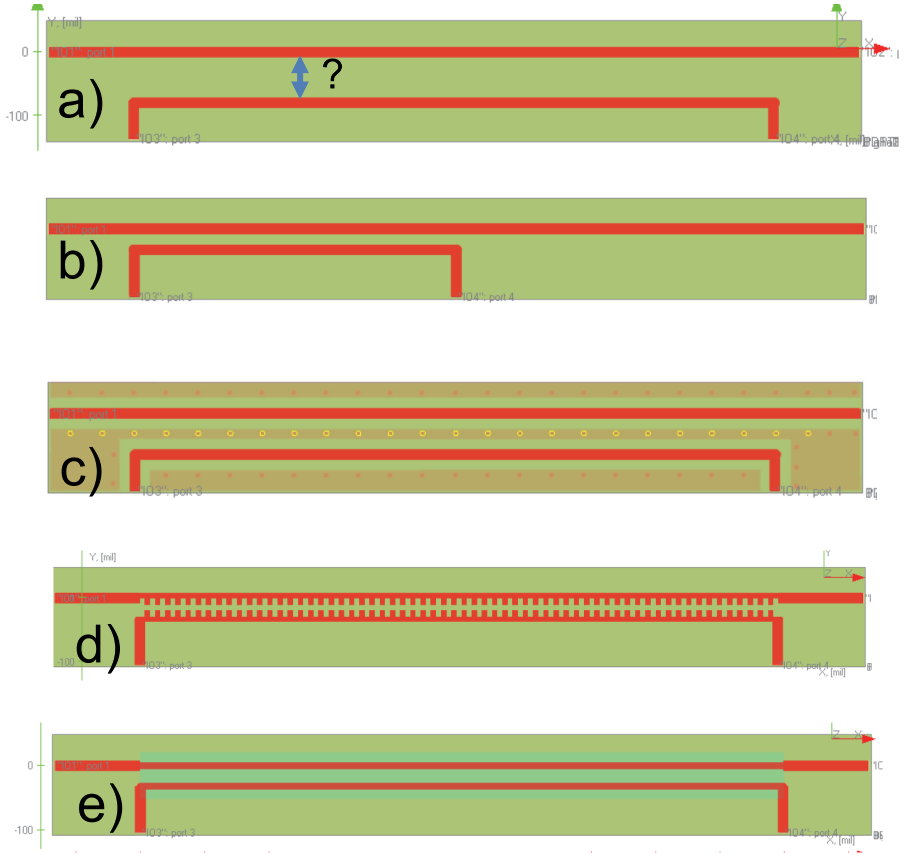
$$\begin{pmatrix}1\\0\end{pmatrix} = \begin{pmatrix}\frac{+1}{2}\\ \frac{-1}{2}\end{pmatrix} e^{-\Gamma_{od}\cdot x} + \begin{pmatrix}\frac{+1}{2}\\ \frac{+1}{2}\end{pmatrix} e^{-\Gamma_{ev}\cdot x}$$

FEXT is negative if odd mode is faster (usual case) $\varepsilon_{od} < \varepsilon_{ev}$
FEXT is positive if odd more is slower $\varepsilon_{od} > \varepsilon_{ev}$

More in B. Young, Digital Signal Integrity – Modeling and Simulation with Interconnects and Packages, 2000

XTalk in Microstrips: How to Reduce IT?



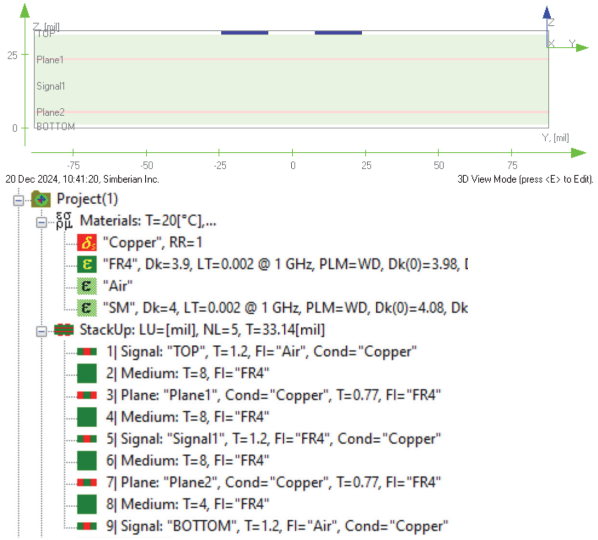


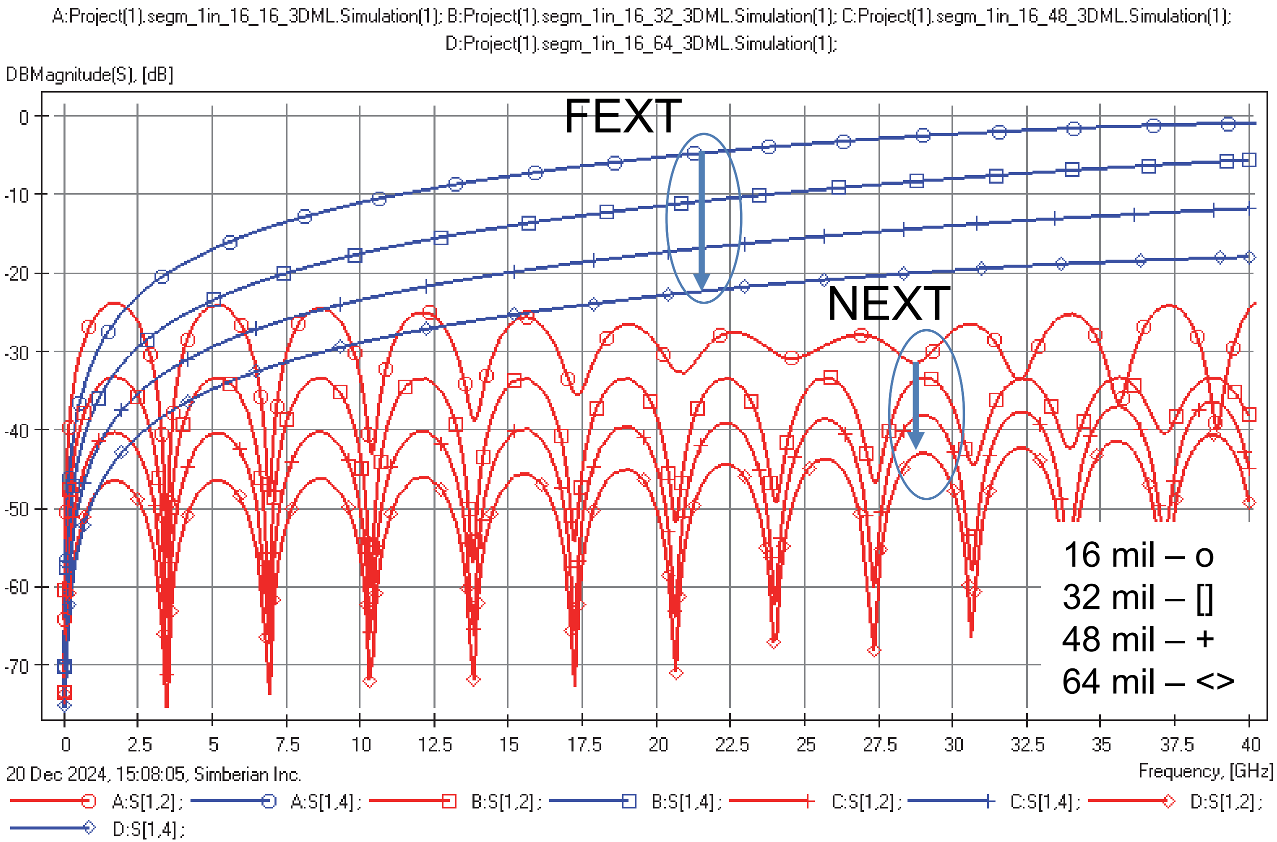
a) Increase separation – how much ?
b) Reduce coupling length?
c) Use guarding trace with via fence?
d) Use tabbed traces?
e) Use embedded microstrips (overlay)?
f) Switch to striplines?
…

What else?

MSL XTalk Reduction by Separation

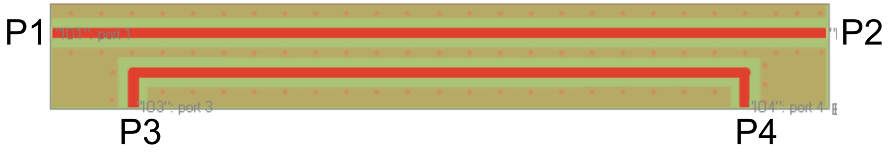
*16 mil traces in TOP layer (about 50 Ohm) coupled over **1 inch** with 16, 32, 48 and 64 mil separation*



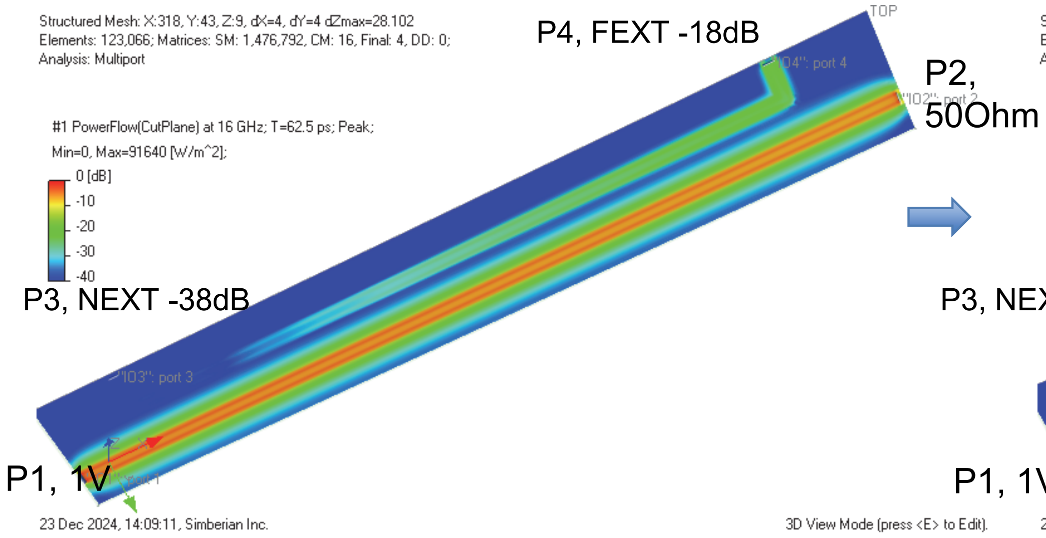


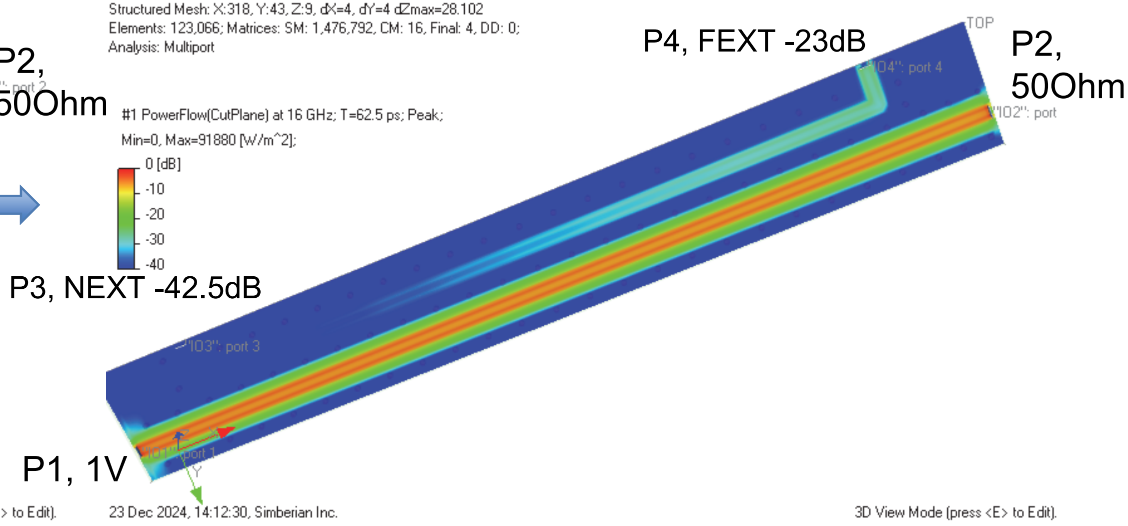
Very difficult to achieve high isolation!

MSL Guard Traces with Via Fence (Coplanar)



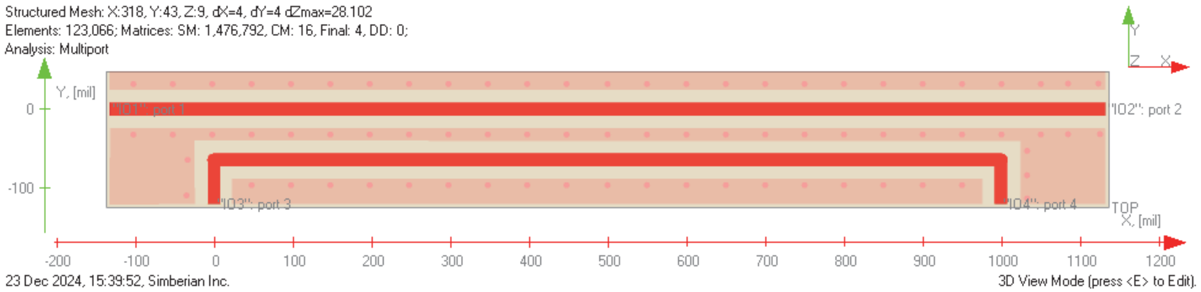
Peak PFD at 16GHz right below traces (in substrate dielectric)

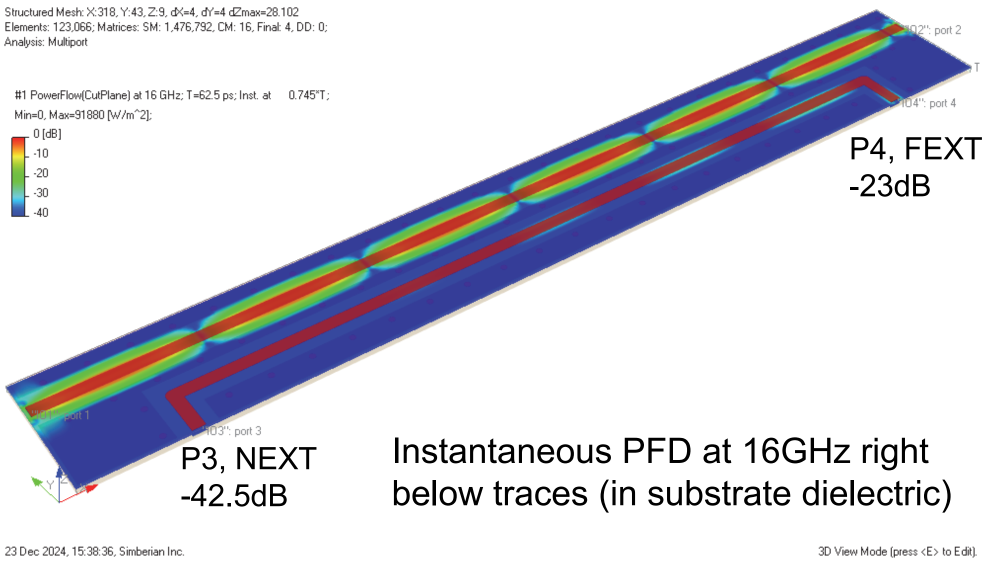




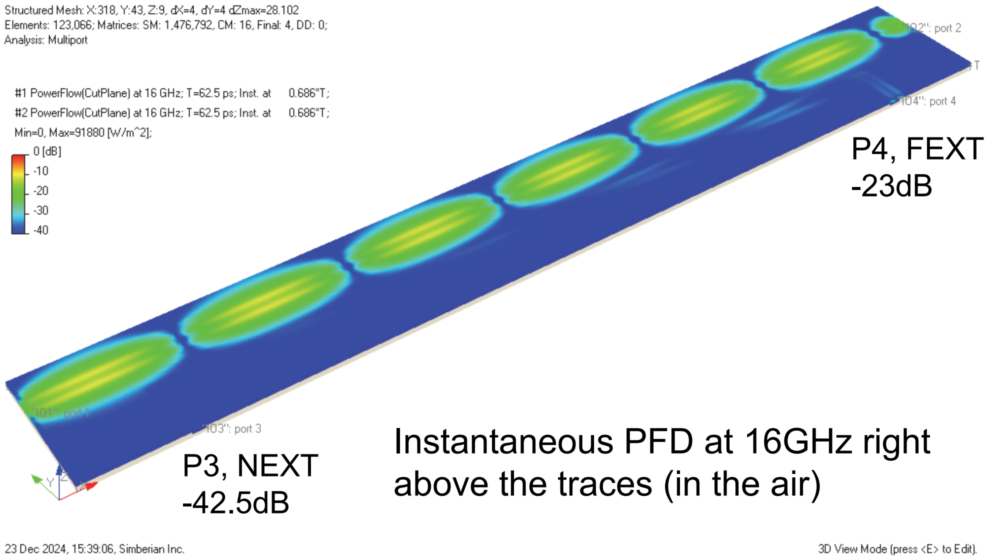
Why no dramatic xtalk reduction as in case of stripline?

Why NO Dramatic XTalk Reduction…?





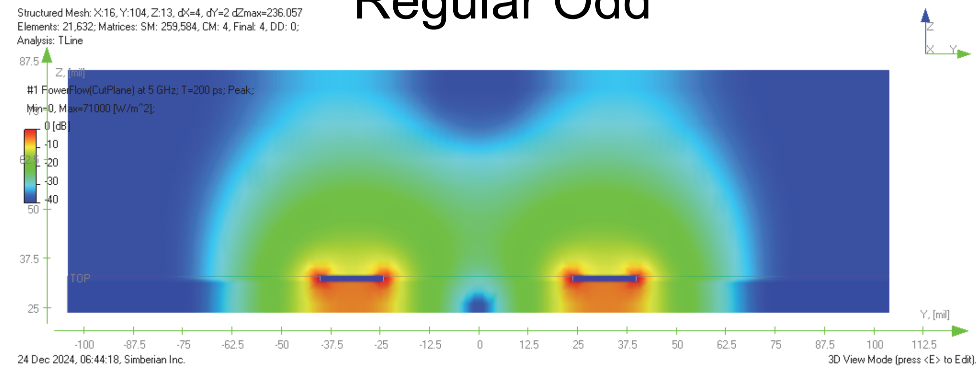
Instantaneous PFD at 16GHz right below traces (in substrate dielectric)



Instantaneous PFD at 16GHz right above the traces (in the air)

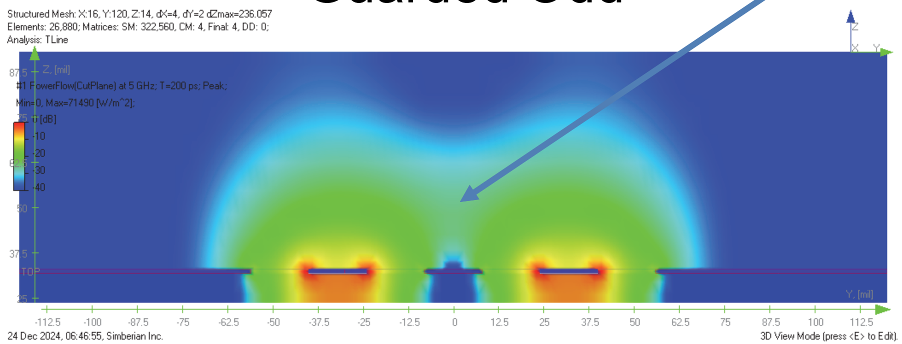
MSL vs Guarded MSL: Peak PFD at 5 GHz

Regular Odd

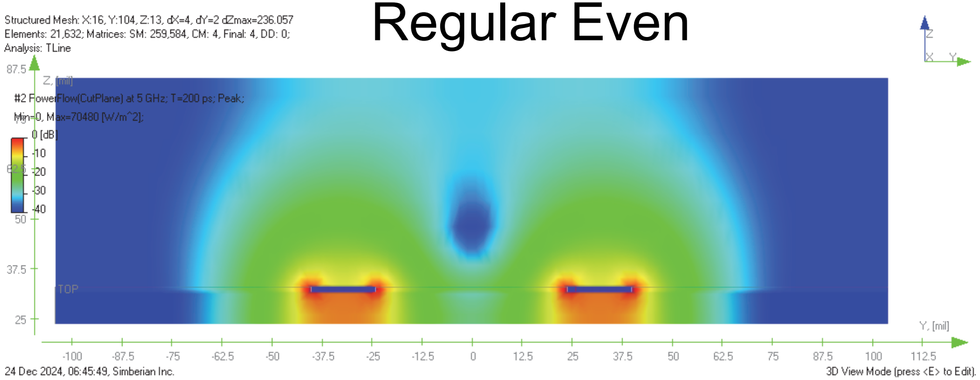


Guarded Odd

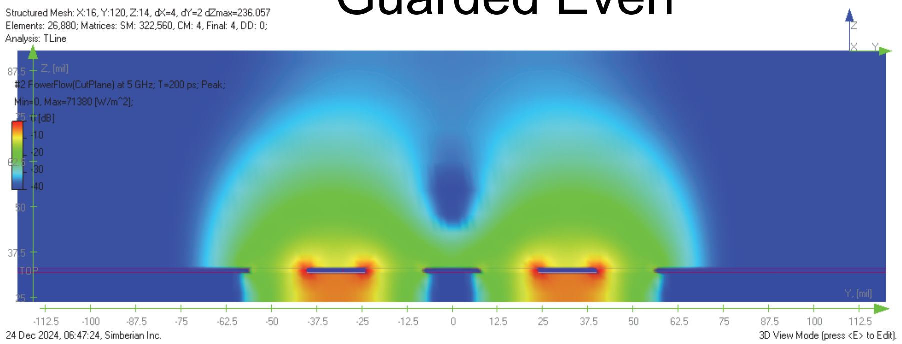
Not localized



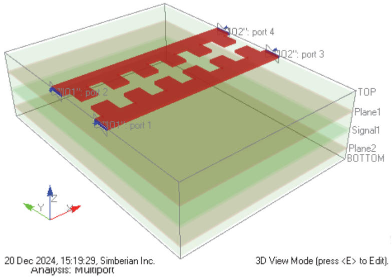
Regular Even



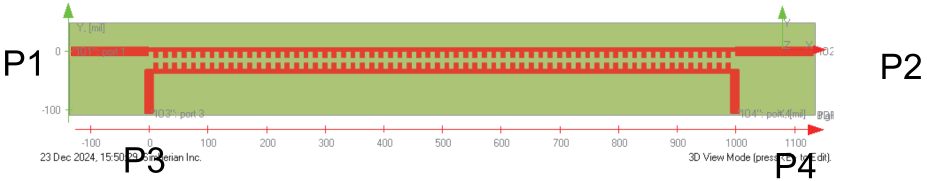
Guarded Even

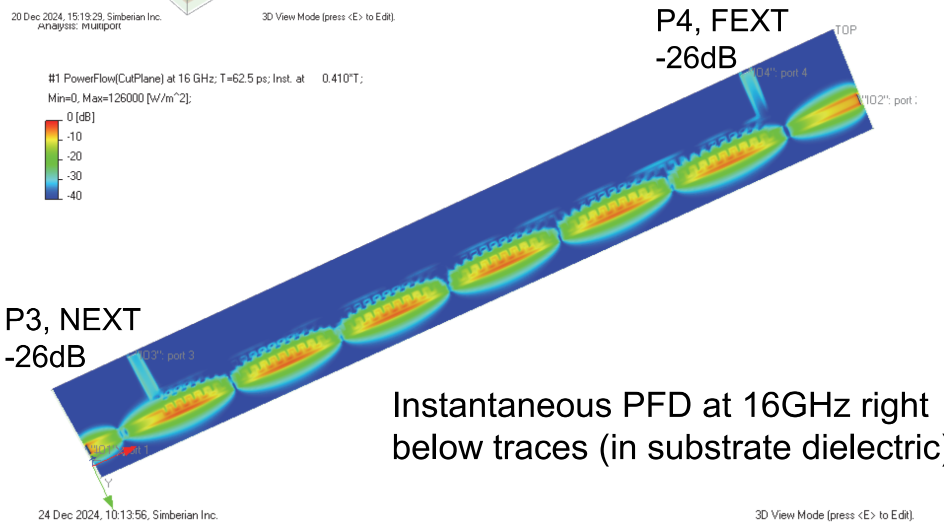


MSL with Tabbed Traces

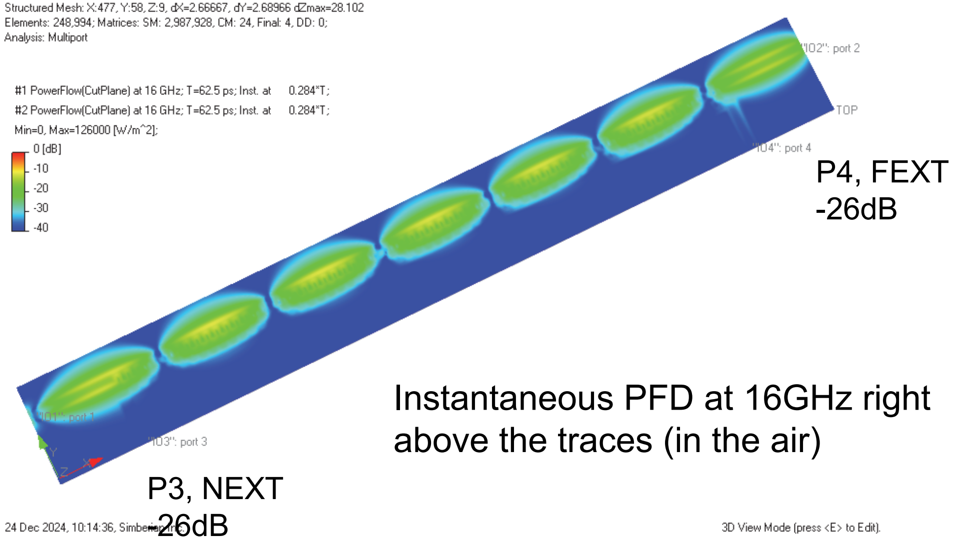


Tabbed: 8 mil traces, 24 mil apart, 8.5 by 7.5 mil tabs (40 mil total width)





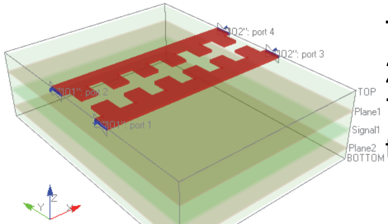
Instantaneous PFD at 16GHz right below traces (in substrate dielectric)



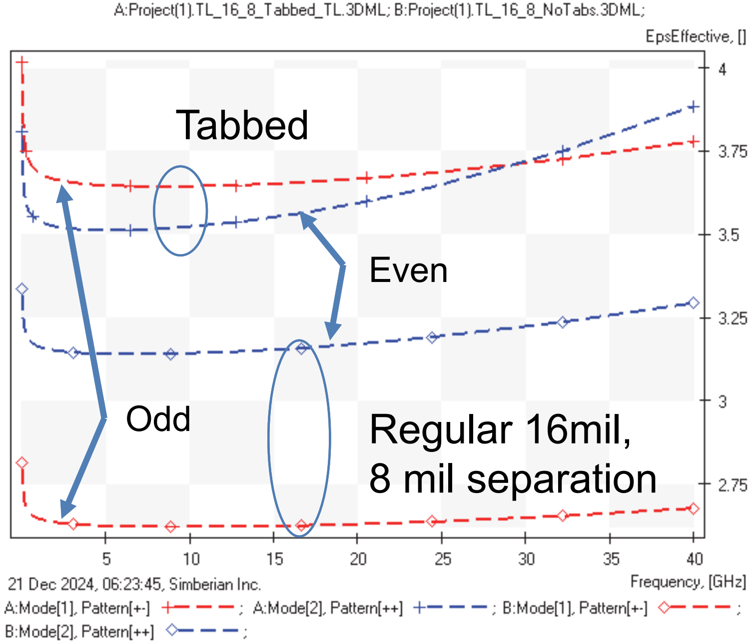
Instantaneous PFD at 16GHz right above the traces (in the air)

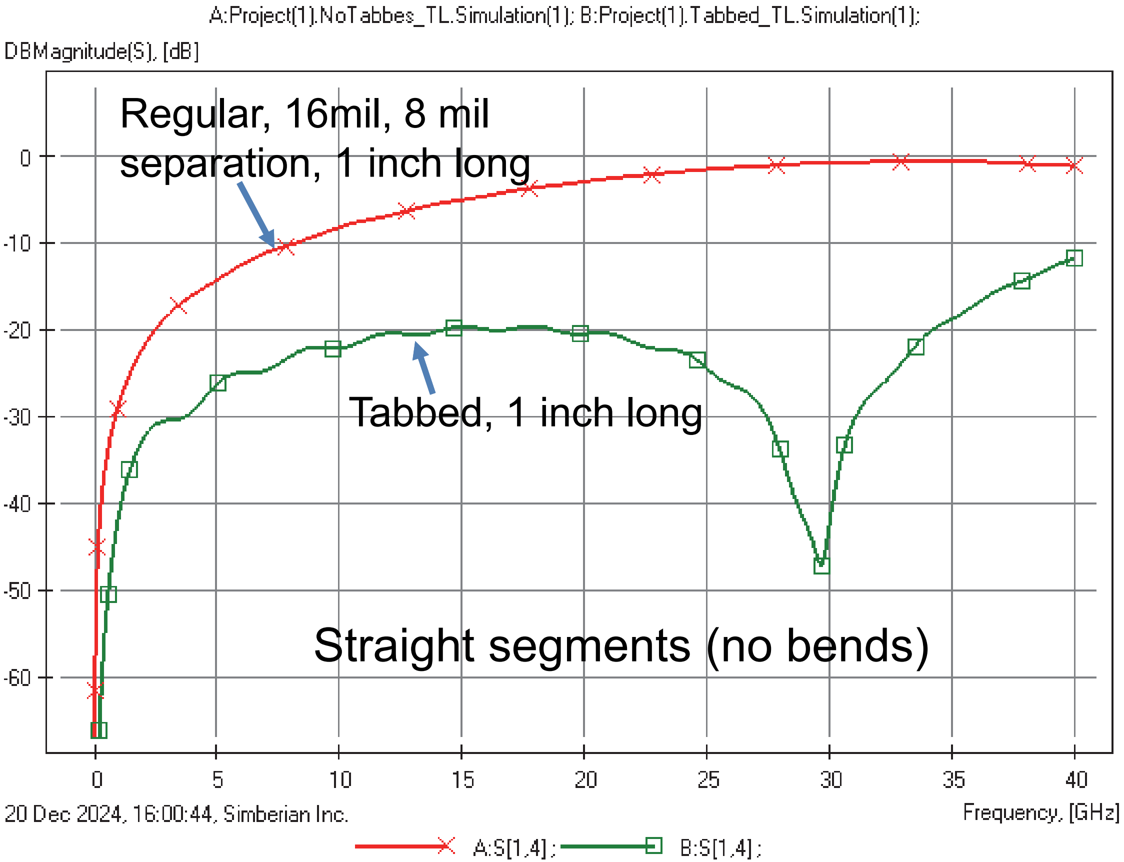
Is this reliable solution? (Hint: plating and solder mask uncertainties)

MSL XTalk Reduction with Periodic Tabs

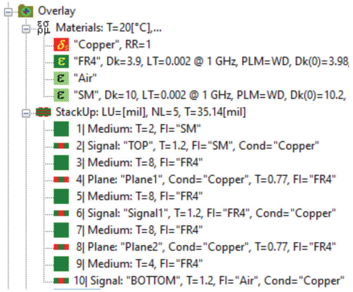


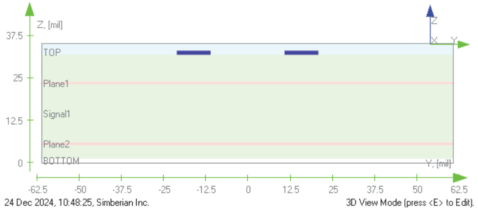
Tabbed: 8 mil traces, 24 mil apart, 8.5 by 7.5 mil tabs (40 mil total width)

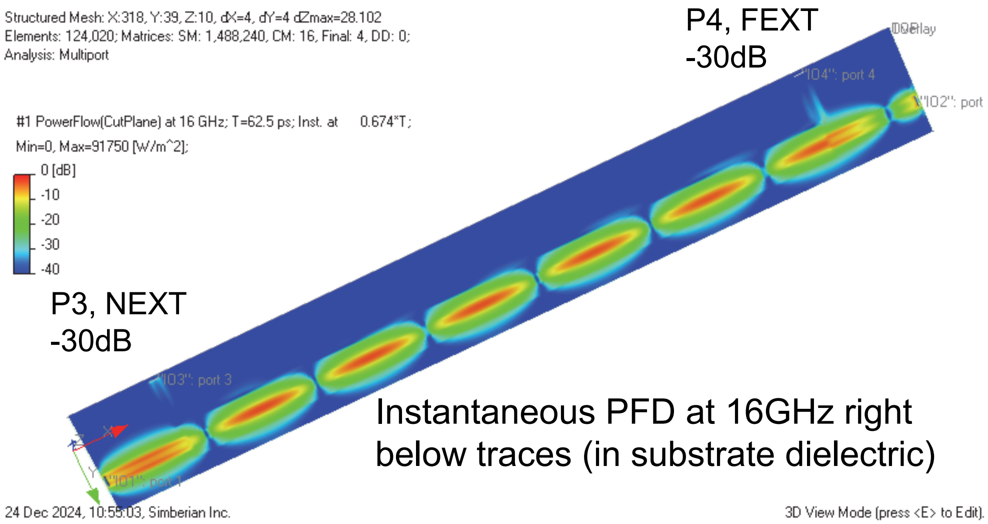


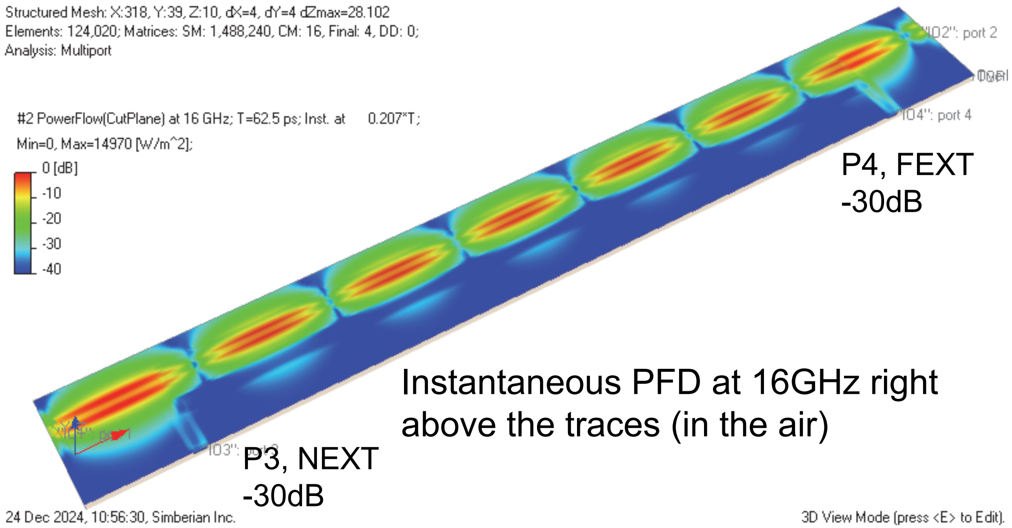


MSL with Overlay (Superstrate or Embedded)



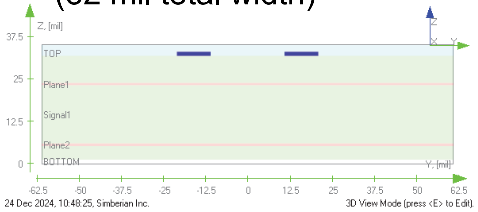


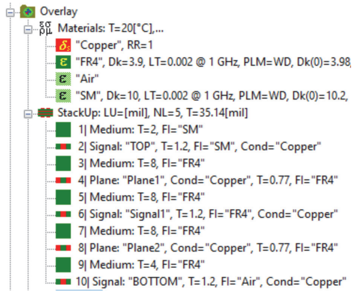


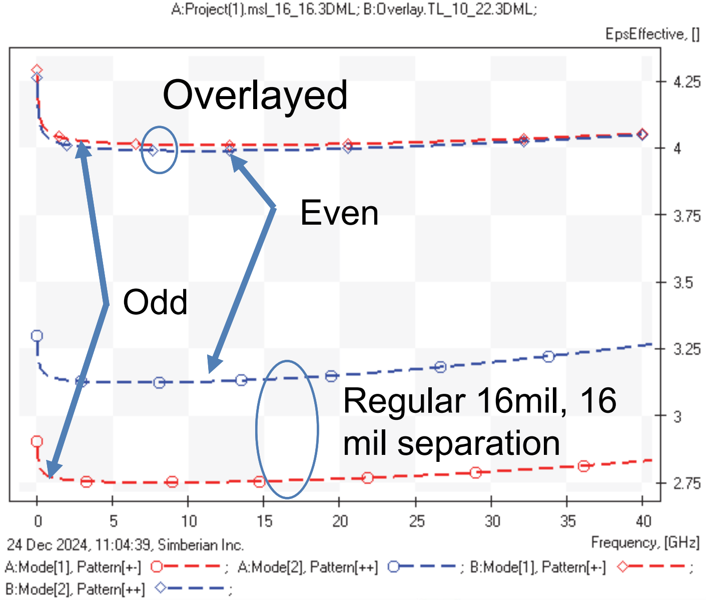


MSL XTalk Reduction with Overlay

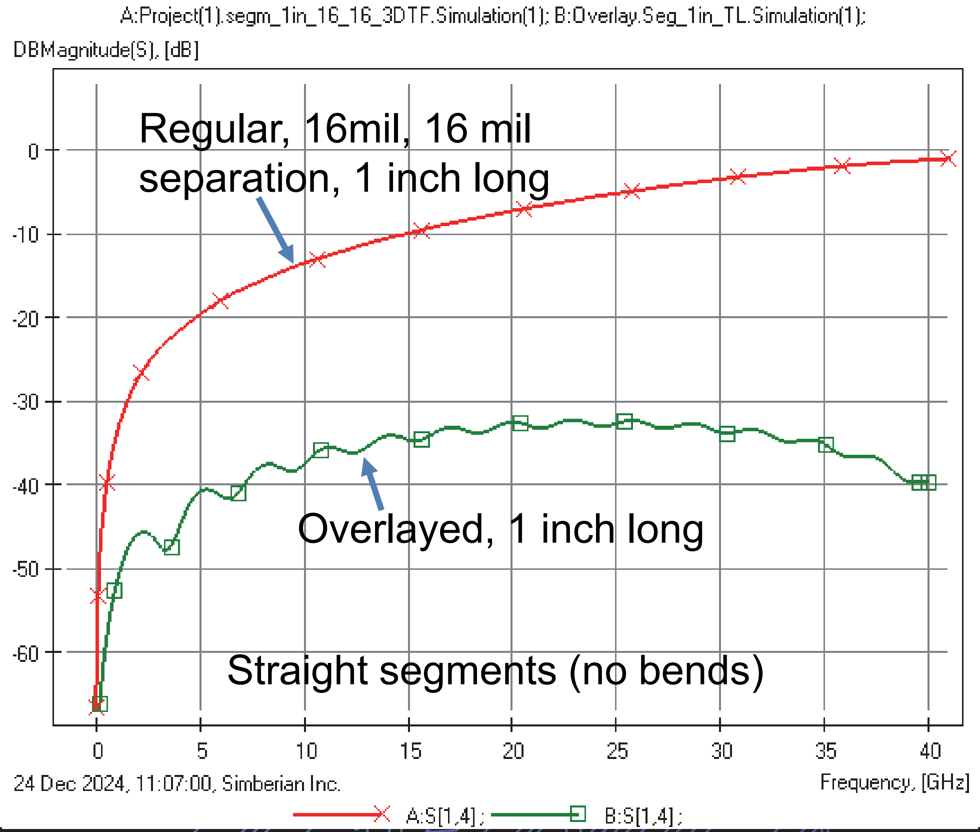
Overlayed: 10 mil traces, 22 mil apart (32 mil total width)





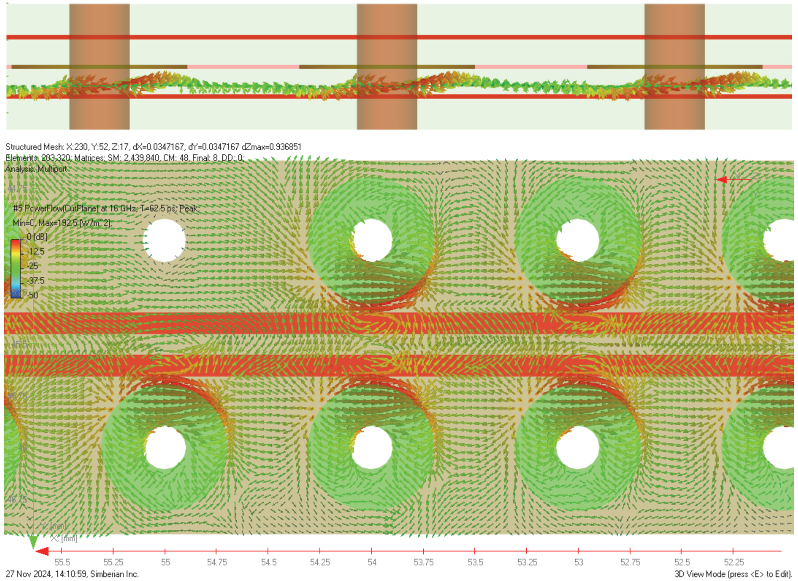


This is the second-best way to reduce XTalk in MSL



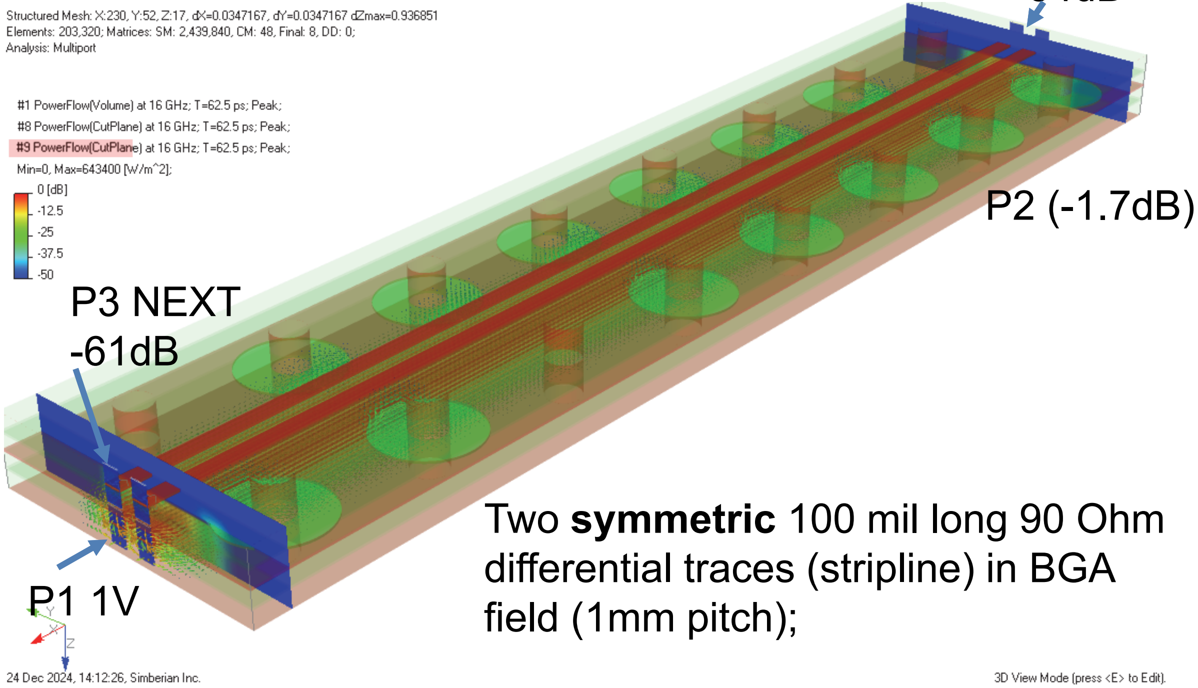
Traces in BGA Area – XTalk Through Antipads

Peak PFD right above the victim traces at 16GHz



Simbeor Solution: Xtalk_BGA_ViaField_2019_09

Peak PFD at 16 GHz - volume and at the ends

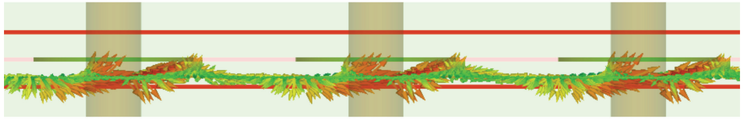


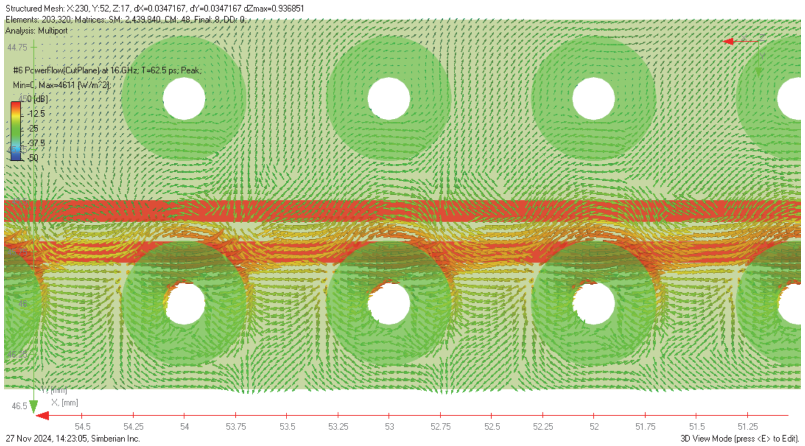
Two **symmetric** 100 mil long 90 Ohm differential traces (stripline) in BGA field (1mm pitch);

#2019_09: **How Interconnects Work™**: Where crosstalk may come from - case of stripline coupling through antipads in BGA breakout areas, 12 min – YouTube: https://youtu.be/gTjvG3sUzI4

Traces in BGA Area – XTalk Trough Antipads

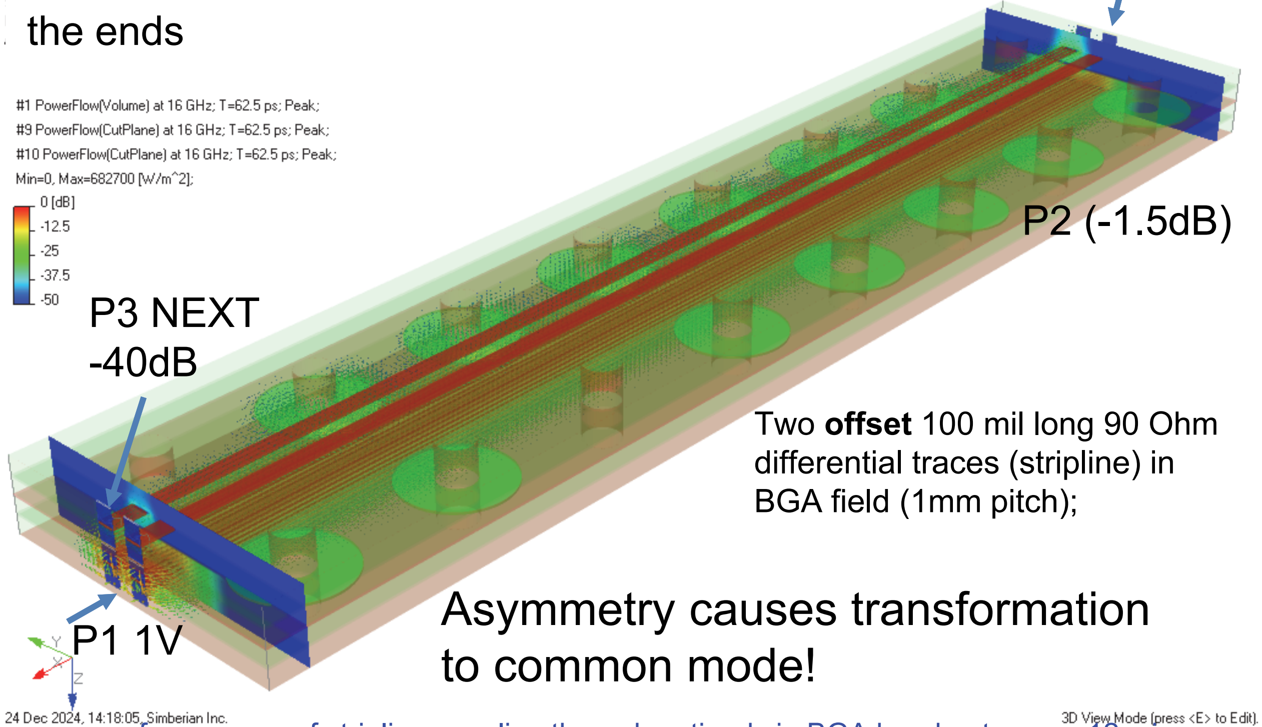
Peak PFD right above the victim traces at 16GHz





Simbeor Solution:
Xtalk_BGA_ViaField_2019_09

Peak PFD at 16GHz – volume and at the ends



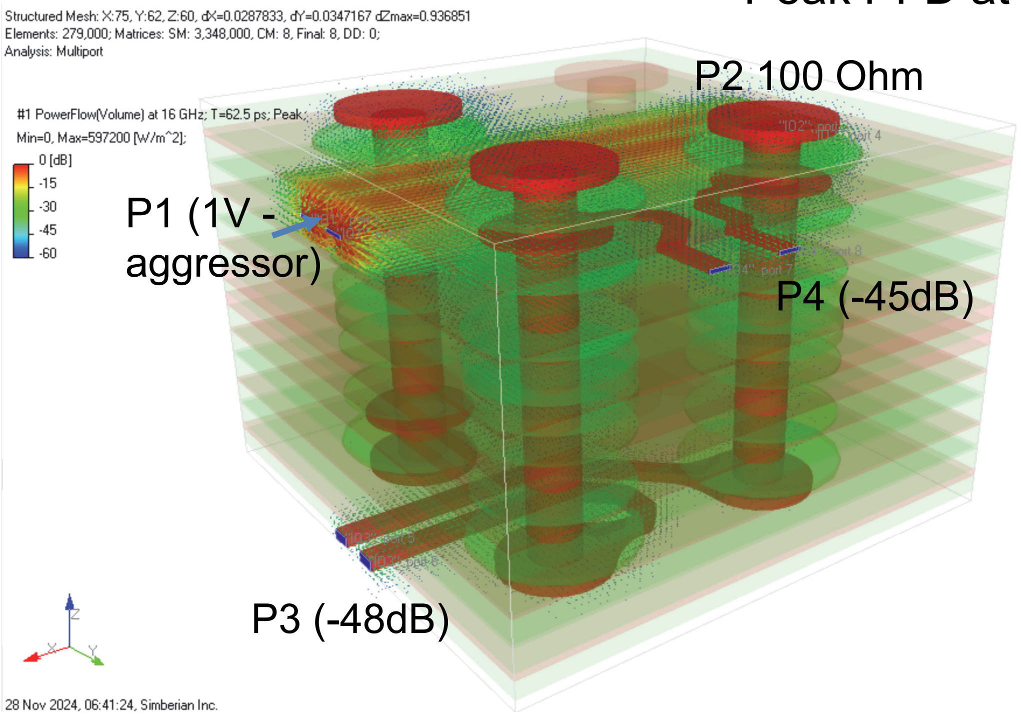
Two **offset** 100 mil long 90 Ohm differential traces (stripline) in BGA field (1mm pitch);

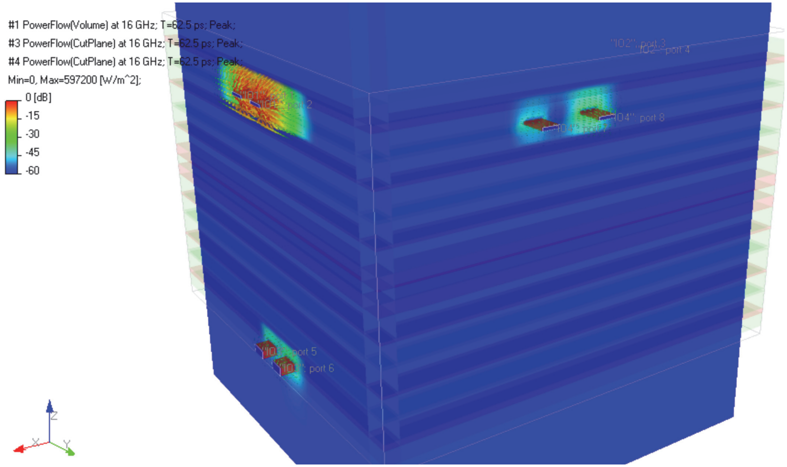
Asymmetry causes transformation to common mode!

#2019_09: **How Interconnects Work™**: Where crosstalk may come from - case of stripline coupling through antipads in BGA breakout areas, 12 min – YouTube: https://youtu.be/gTjvG3sUzI4

Trace to Via XTalk Close To Via Pads

Peak PFD at 16GHz (volume & cut planes)

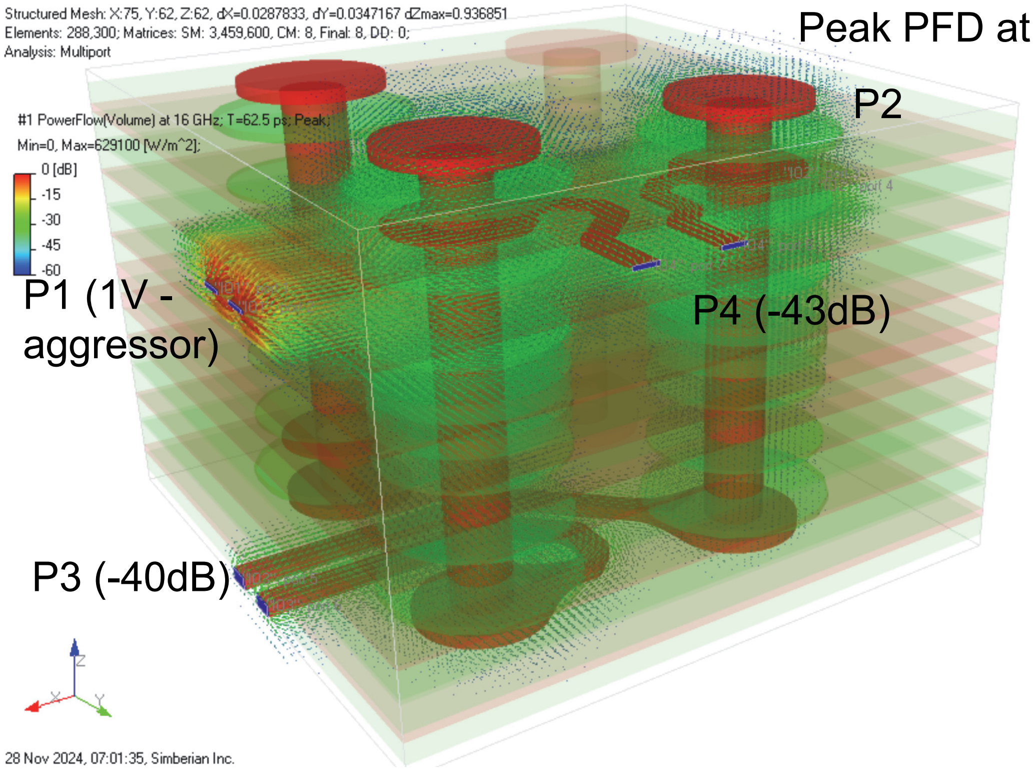




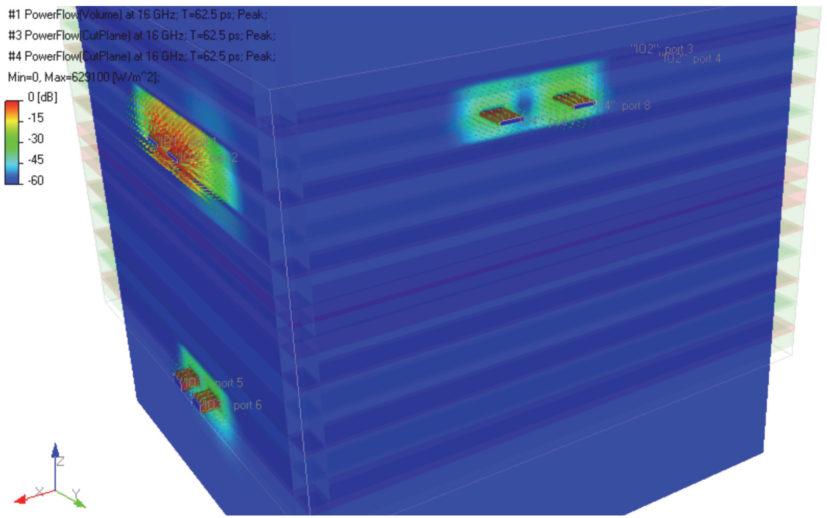
Simbeor Solution:
Xtalk_Strip_To_Vias_2019_10

#2019_10: **How Interconnects Work™**: Where crosstalk may come from - case of coupling between differential striplines and vias, 8 min - https://youtu.be/zKXTUVP8Wnc

XTalk in Trace Over Via Antipads



Peak PFD at 16GHz (volume & cut planes)



Simbeor Solution:
Xtalk_Strip_To_Vias_2019_10

#2019_10: **How Interconnects Work™**: Where crosstalk may come from - case of coupling between differential striplines and vias, 8 min - https://youtu.be/zKXTUVP8Wnc

XTalk Through Meshed Planes (Flex Interconnects)

#2018_06: Y. Shlepnev, Design insights from electromagnetic analysis: Effects of meshed reference planes on interconnects, July 20, 2018.

Cut outs may cause significant cross-talk as well as EMI/EMC issues.

Simbeor Solution:
FlexMeshedPlane_2017_02_03

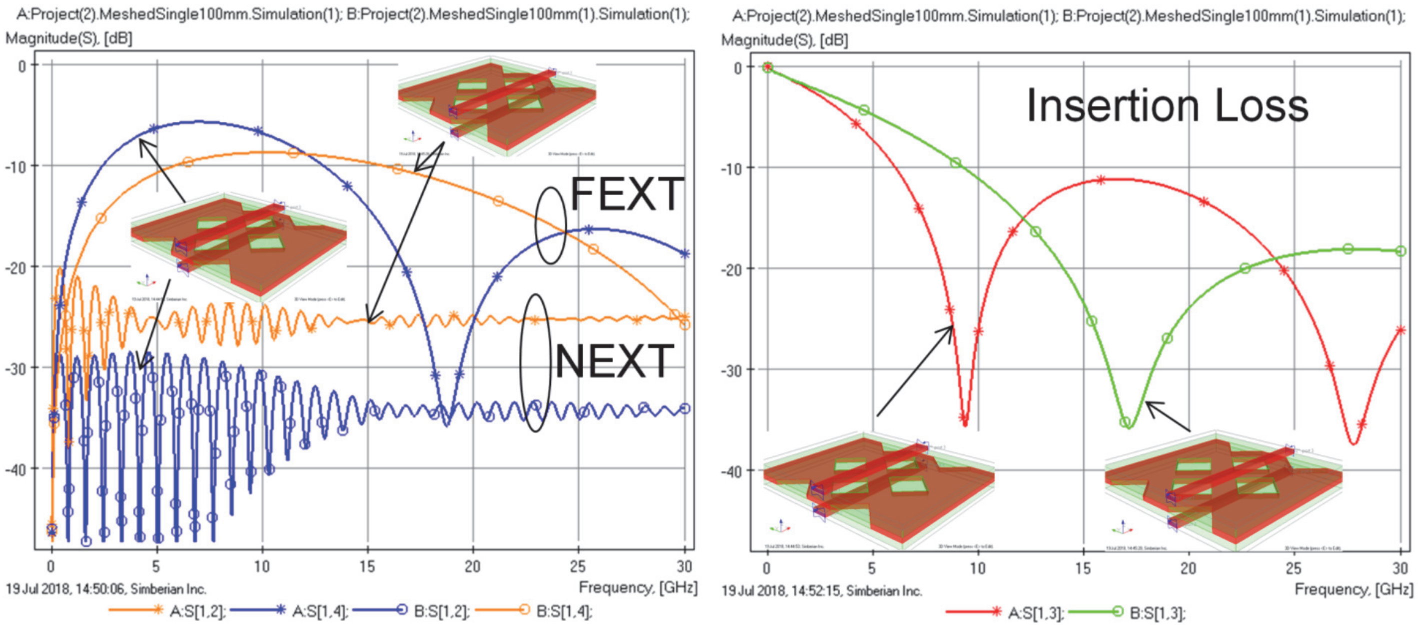
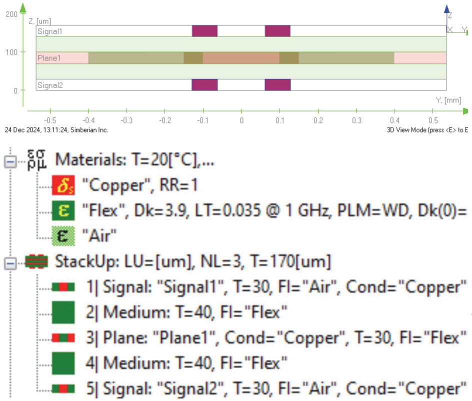
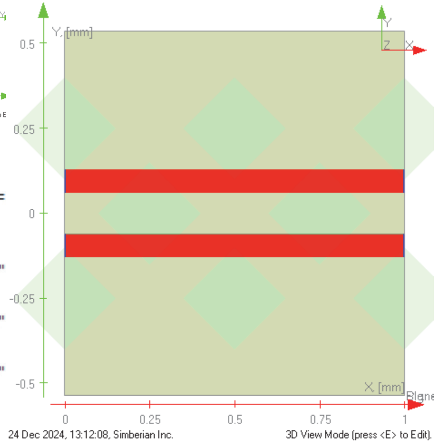


Fig. 8. Near-end (NEXT) and far-end (FEXT) crosstalk (left) and corresponding insertion loss (right) for 10 cm traces over meshed plane for traces over the cutouts and mostly over conductor.

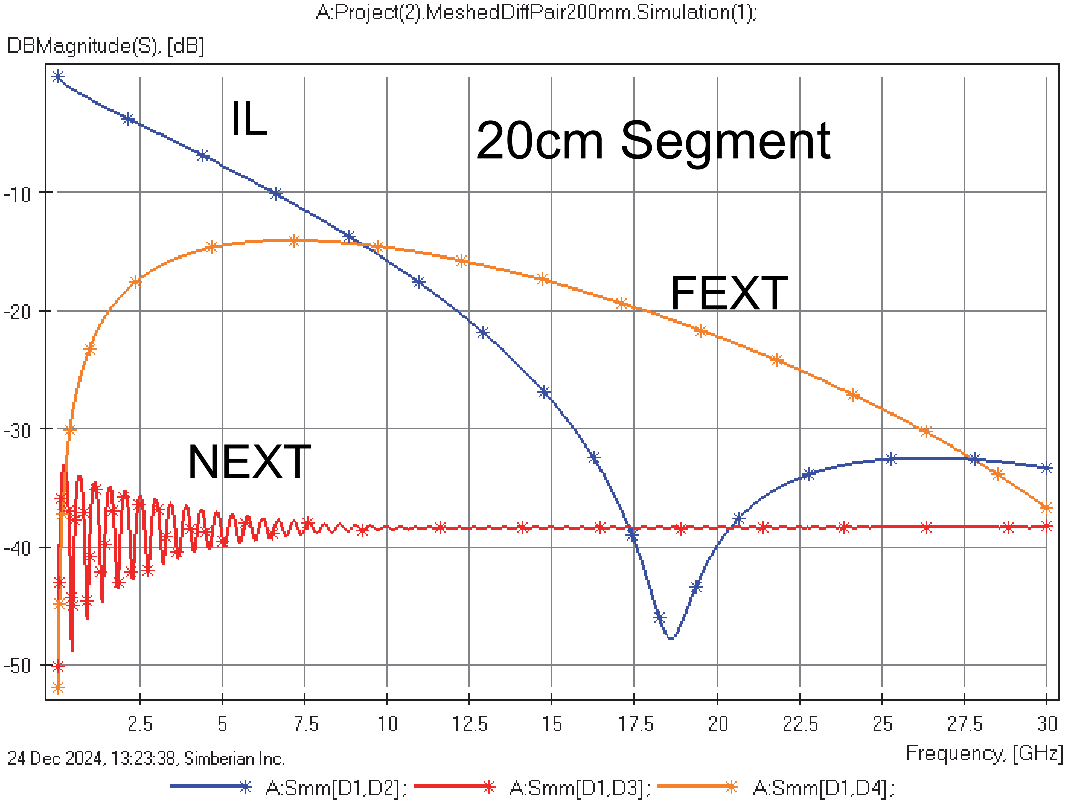
• #2017_02: **How Interconnects Work™**: Microstrip over meshed reference plane in flex interconnects, 16 min – YouTube https://youtu.be/6q8FP1fPyOQ

Differential XTalk in Flex Interconnects





“Suck-out” in interconnects with high losses



Simbeor Solution:
FlexMeshedPlane_2017_02_03

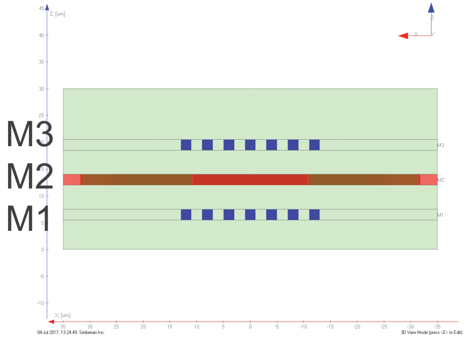
Look at the surface currents…

#2017_03: **How Interconnects Work™**: Differential microstrip over meshed reference plane in flex interconnects, 15 min – YouTube https://youtu.be/zyx1mh0w0t8

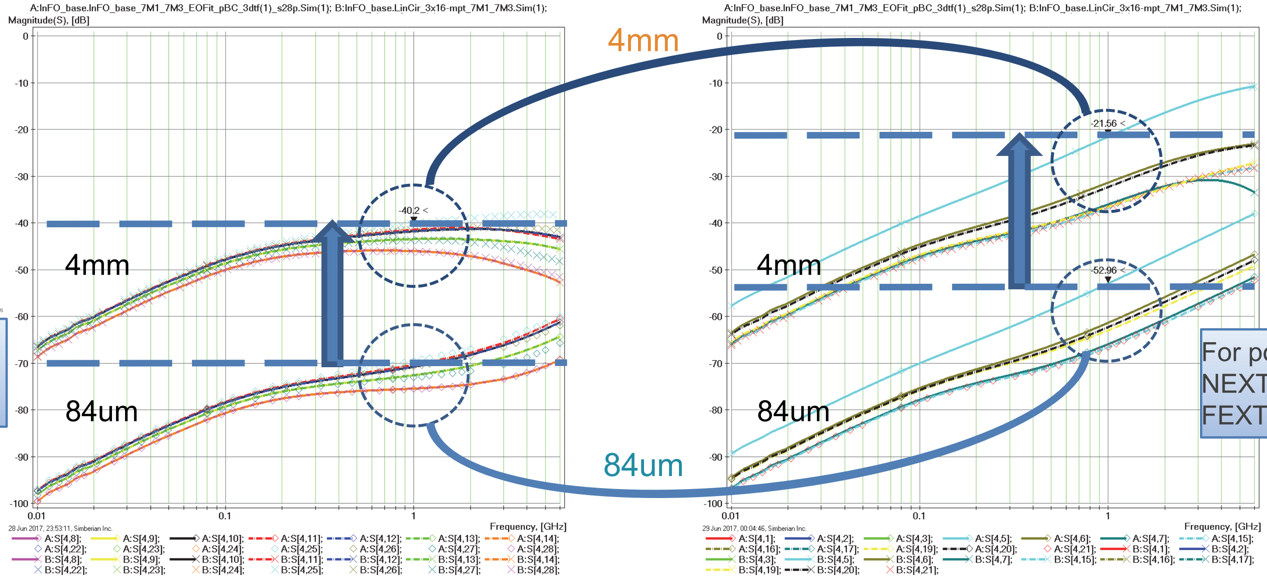
XTalk Across M1/M3 Layers in HBM2

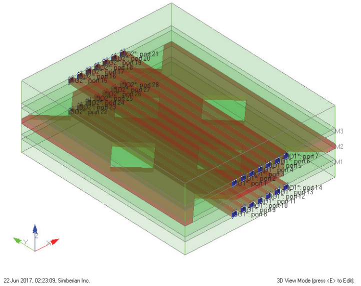
Y. Shlepnev, V. Heyfitch, Tutorial – Design Insights from Electromagnetic Analysis & Measurements of PCB & Packaging Interconnects Operating at 6- to 112-Gbps & Beyond, DesignCon 2020

M3/M1 through M2 mesh is < -40dB @1GHz   *Within M3* is < -21dB @1GHz



For port4 (in M3), ports 8-14 are NEXT and ports 22-28 are FEXT across M2 GND.

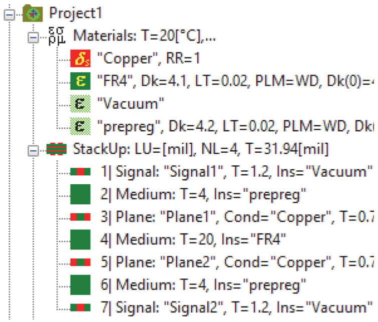


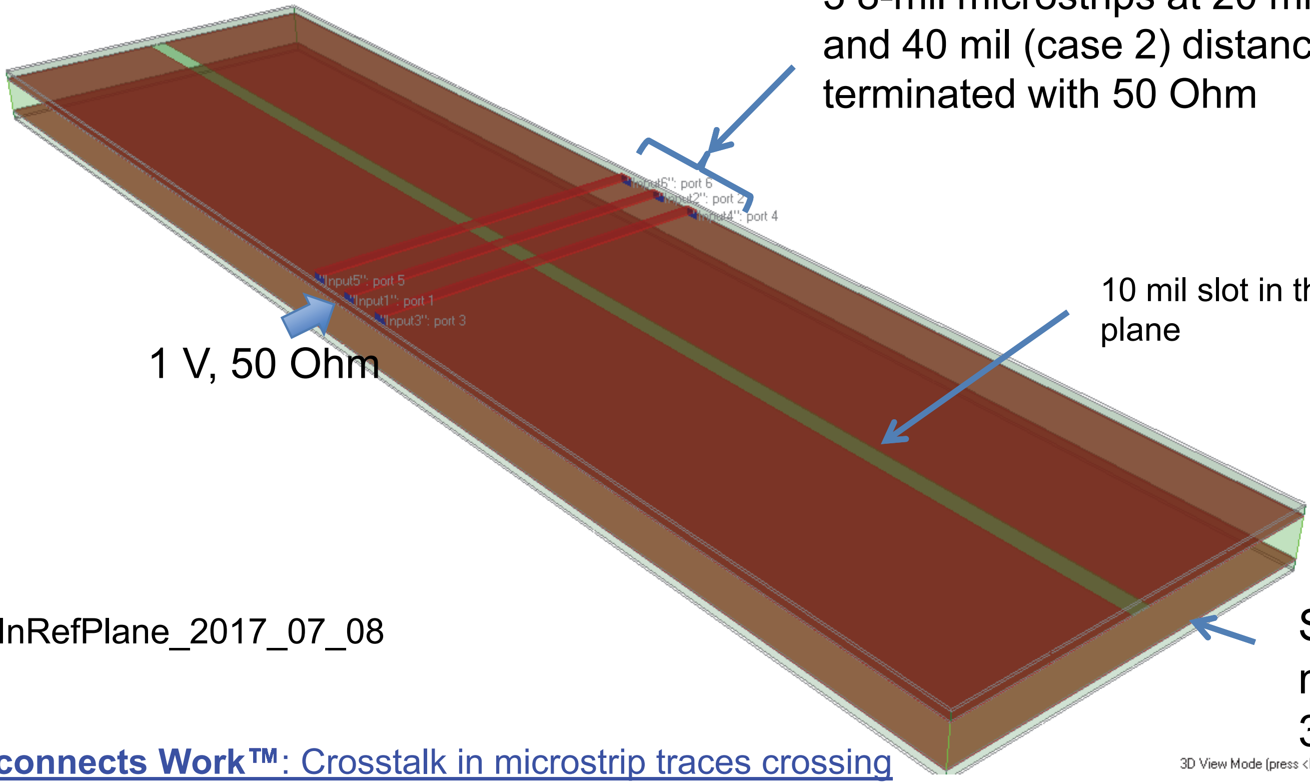


For port4 (in M3), ports 1-7 are NEXT and ports 15-21 are FEXT within M2 GND.

Analysis of crosstalk is relatively easy – accurate analysis of losses is more challenging…

Microstrips Crossing Split-Panes or Slot





3 8-mil microstrips at 20 mil (case 1) and 40 mil (case 2) distances – all terminated with 50 Ohm

10 mil slot in the reference plane

1 V, 50 Ohm

Solid plane at 20 mil (case 1) and 3 mil (case 2)

Solution
XTalk_MSL_OverSlotsInRefPlane_2017_07_08

- #2017_08: **How Interconnects Work™**: Crosstalk in microstrip traces crossing split planes, 10 min – YouTube https://youtu.be/M5mngJ4ntNQ

MSLs Crossing Split-Planes or Slot

MSL crossing split in reference plane is **coupled to slotline and to parallel planes**

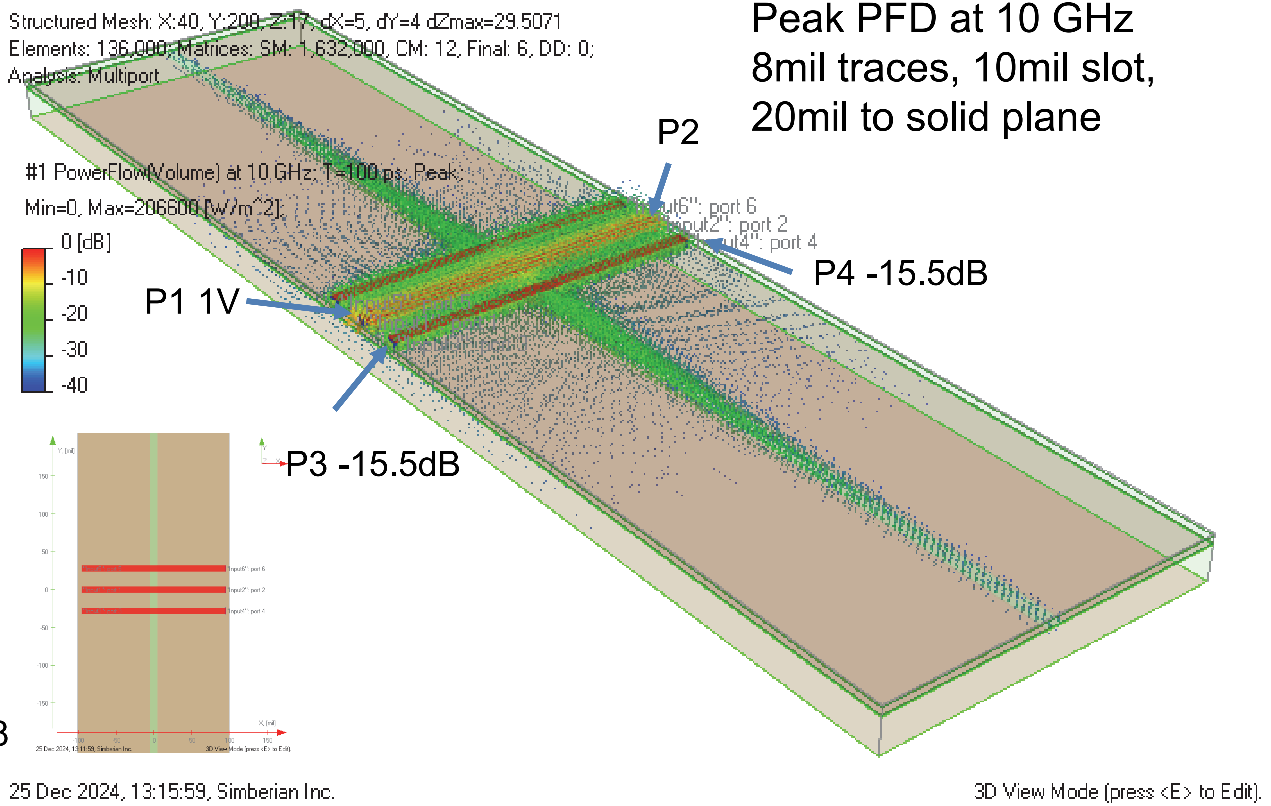
Energy propagates along the slot and between the planes and may be coupled to:

- Other traces crossing the same split (major)
- Vias going through parallel planes (minor)

Requires system-level analysis with PDN structures

AVOID IT!

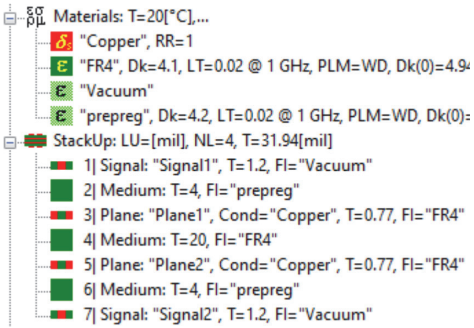
Peak PFD at 10 GHz
8mil traces, 10mil slot,
20mil to solid plane



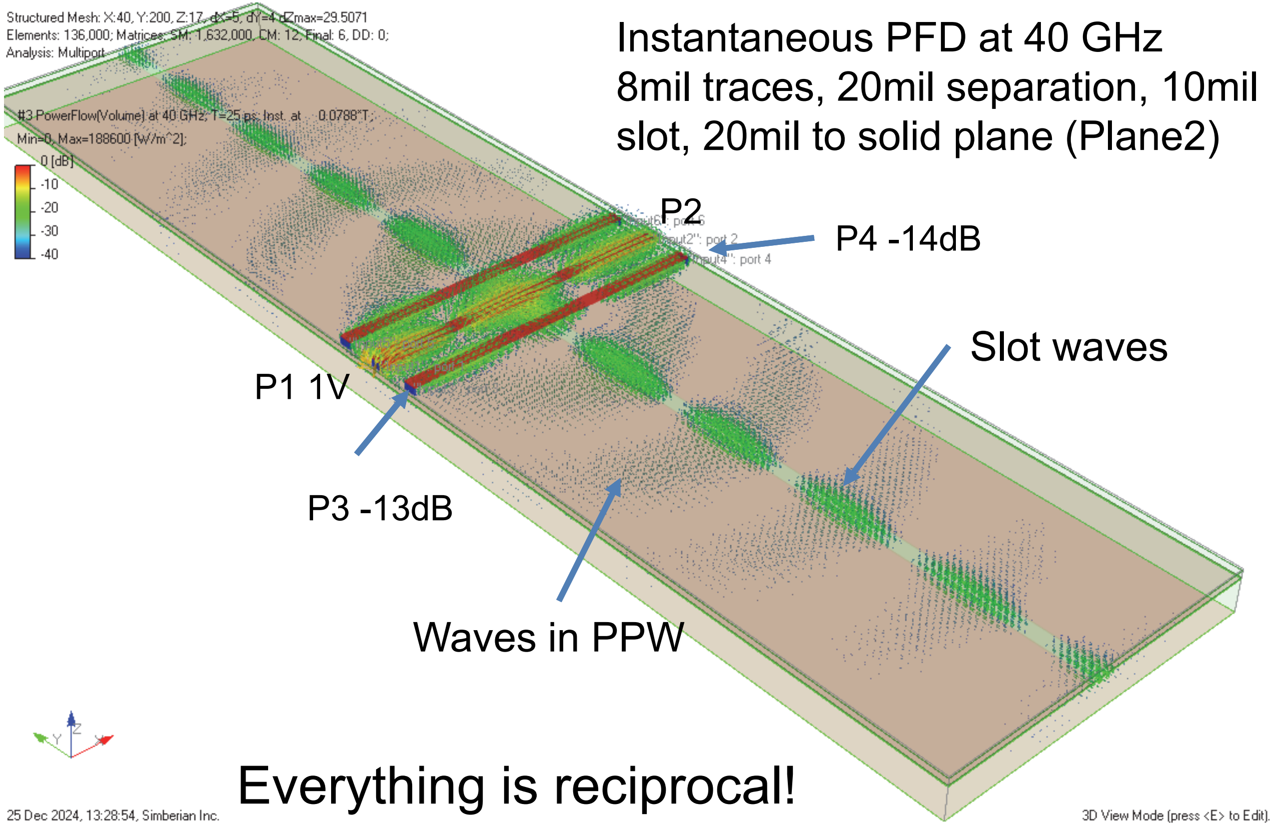
Solution
XTalk_MSL_OverSlotsInRefPlane_2017_07_08

•[#2017_08: **How Interconnects Work™**: Crosstalk in microstrip traces crossing split planes, 10 min](https://youtu.be/M5mngJ4ntNQ) – YouTube https://youtu.be/M5mngJ4ntNQ

MSL Crossing Slot or Split-Planes



Instantaneous PFD at 40 GHz
8mil traces, 20mil separation, 10mil slot, 20mil to solid plane (Plane2)

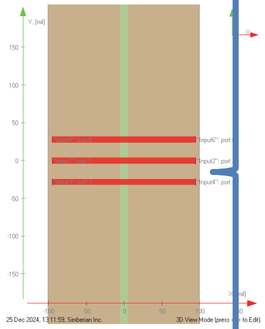


How to reduce the coupling?

a) Increase trace separation?
b) Make slot narrower?
c) "Stitch" planes with AC caps?
d) Place solid plane closer?
e) Switch to differential lines?

Everything is reciprocal!

XTalk in MSL Over Slot - How to Reduce It?



a) Increase trace separation? – Nope

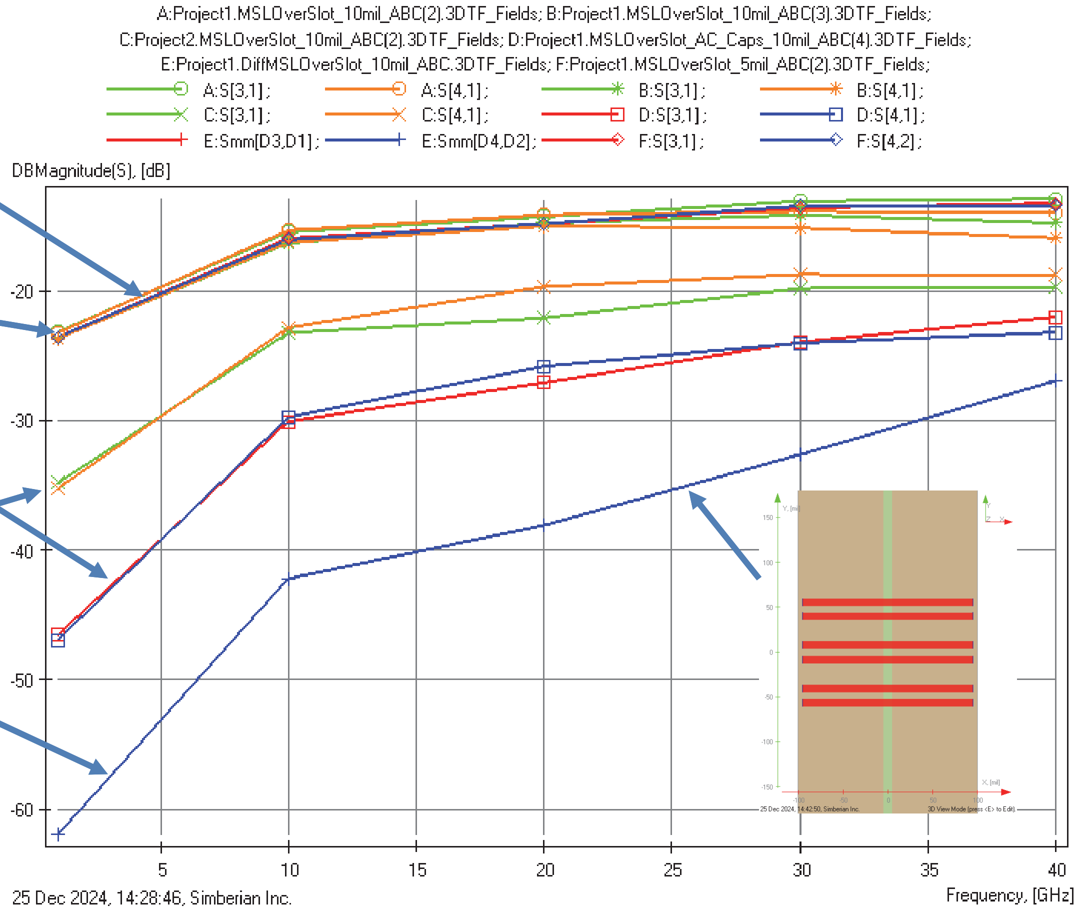
b) Make slot narrower? – Nope

c) “Stitch” planes with AC caps? – Maybe

d) Place solid plane closer? – Yes

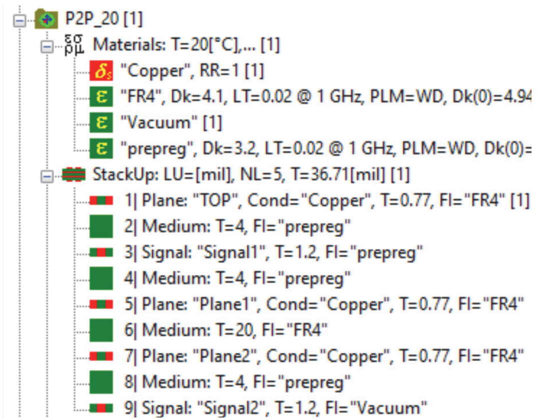
e) Switch to differential lines? – Yes

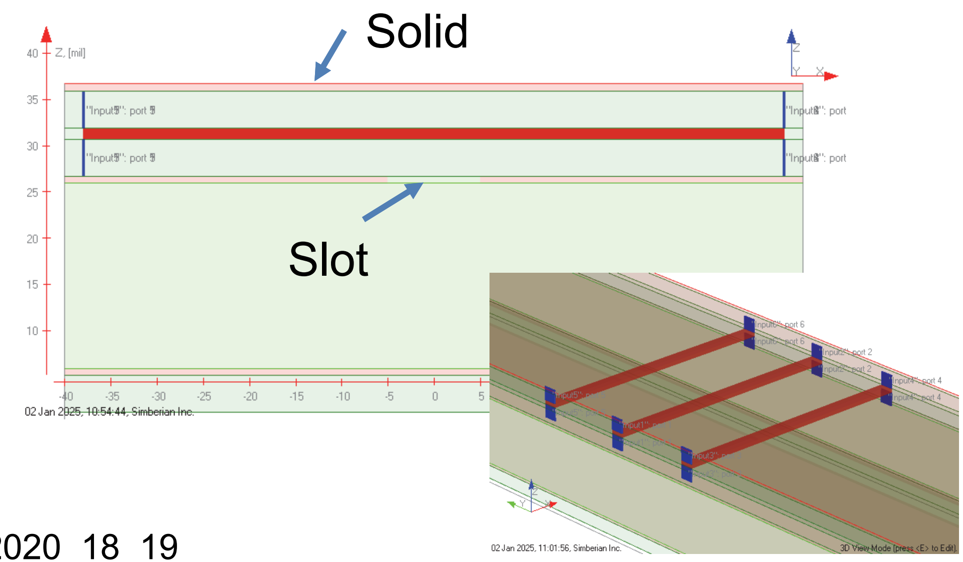
f) Keep plane solid – The Best!

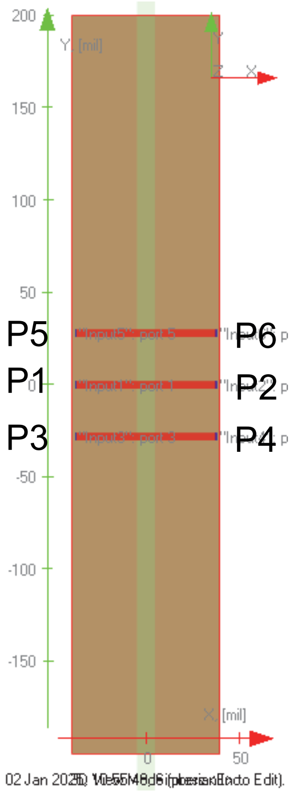


Stripline Crossing Split-Planes or Slot

Keeping one solid reference plane – could it help?







Solution
XTalk_StripsOverSlotsInRefPlane_2020_18_19

- #2020_18: **How Interconnects Work™**: Crosstalk in Single-Ended Striplines Over Split Plane, 5 min – YouTube: https://youtu.be/zdzetSQ2dMk
- #2020_19: **How Interconnects Work™**: Crosstalk in Multiple Striplines Over Split Plane With Closely Spaced Third Solid Plane, 13 min – YouTube: https://youtu.be/p6Y-JhLG3wA
- L. Simonovich, What Happens When Stripline Signals Cross Split Power Planes, SI Journal, June 16, 2020.

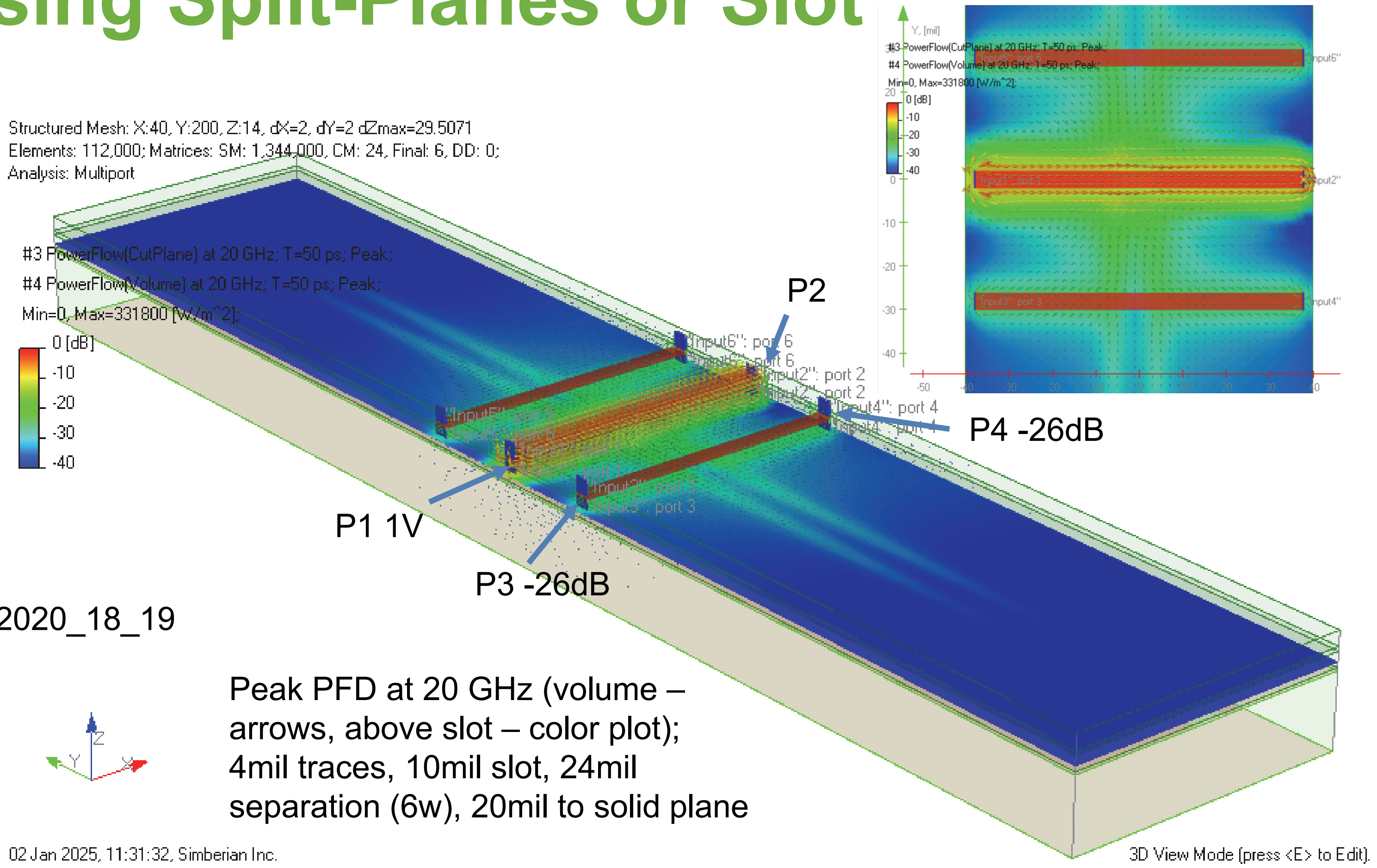
Stripline Crossing Split-Planes or Slot

Stripline crossing split in reference plane is **coupled to slotline and to parallel planes**

Requires system-level analysis with PDN structures

AVOID IT!

Solution XTalk_StripsOverSlotsInRefPlane_2020_18_19

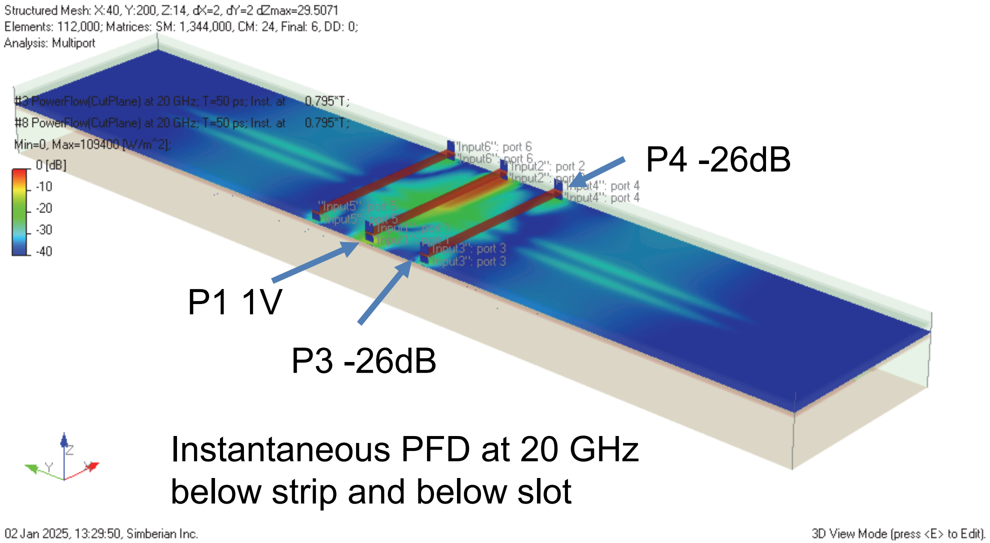


Peak PFD at 20 GHz (volume – arrows, above slot – color plot); 4mil traces, 10mil slot, 24mil separation (6w), 20mil to solid plane

Stripline Crossing Split-Planes or Slot

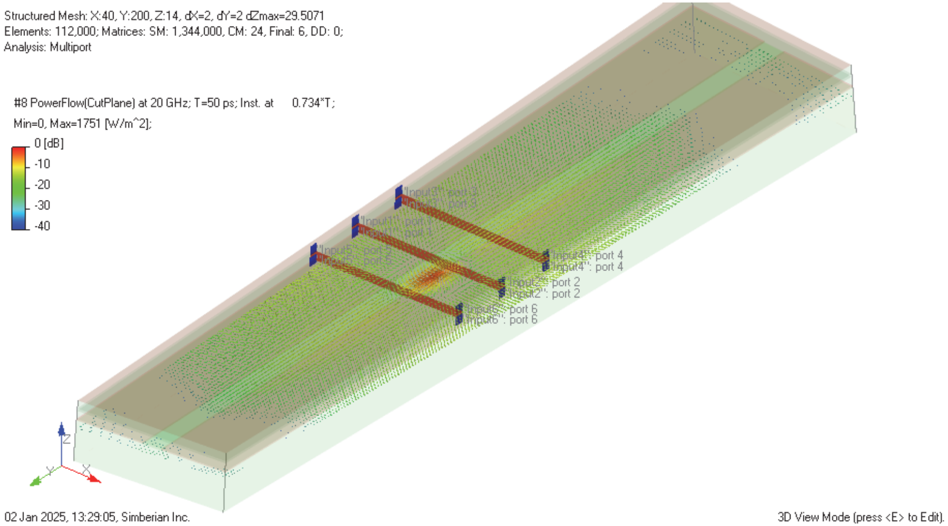
Energy propagates **along the slot and between the planes** and may be coupled to:

- Other traces crossing the same split (major)
- Vias going through parallel planes (minor)



Instantaneous PFD at 20 GHz below strip and below slot

Instantaneous PFD at 20 GHz below slot

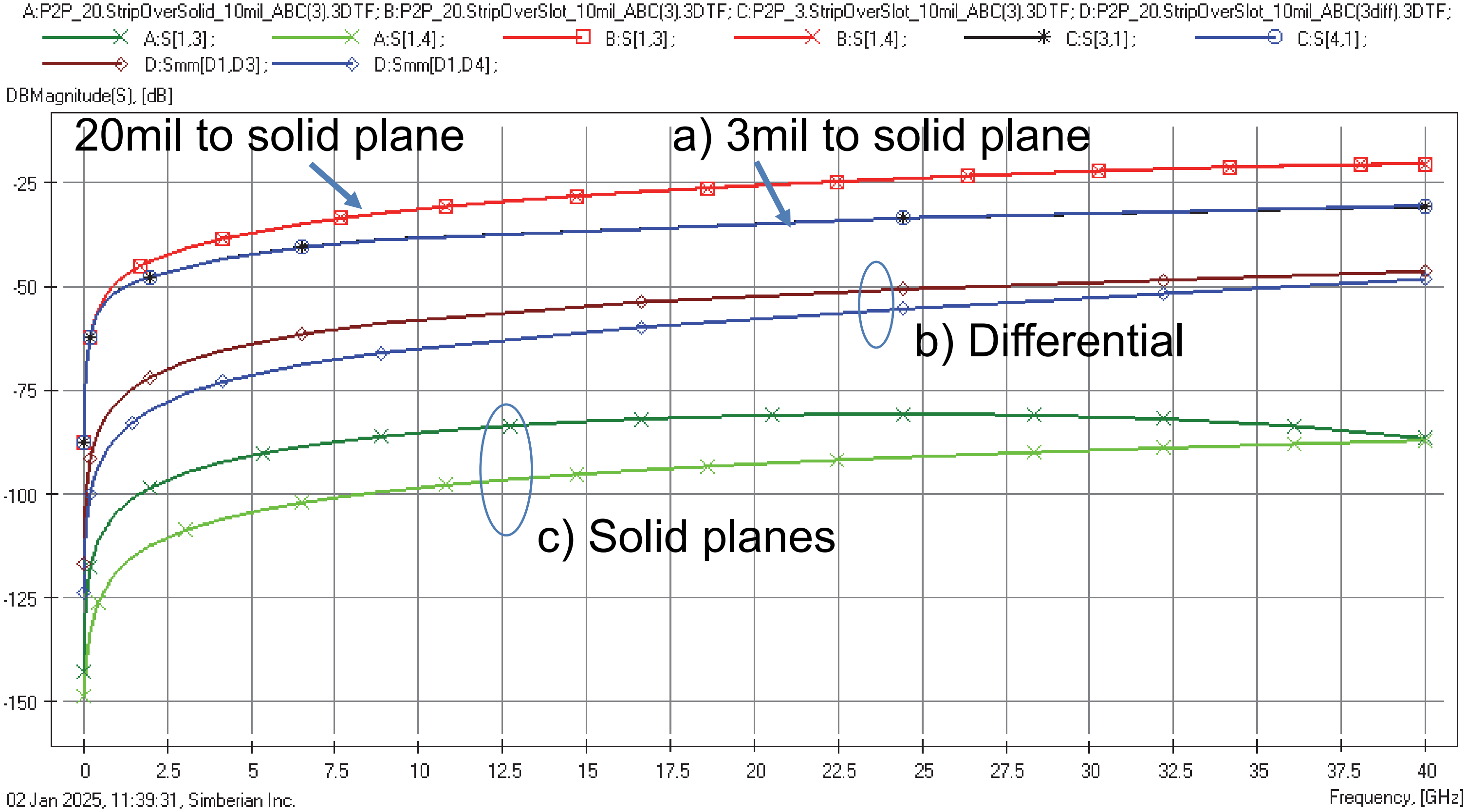


XTalk in Stripline Over Slot - How to Reduce It?

a) Place solid plane closer? – Marginal

b) Switch to differential lines? – Yes

c) Keep planes solid – **The Best Way!**

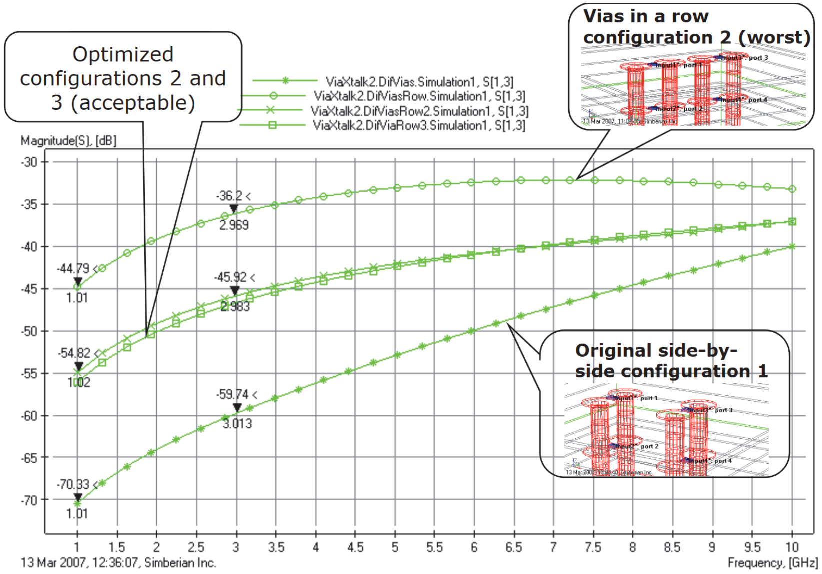


Vias – Coupling and XTalk

- Vias or viaholes are any vertical transitions in PCB or Packaging interconnects
- Vias are not transmission lines in general (can be designed as waveguiding structures)
- Vias are coupled to parallel planes (PDNs) – the effect is similar to trace crossing split-plane
- Microstrip to via transition are coupled to surface waves and external space - radiation
- **Local coupling or xtalk between vias can be evaluated with analysis in isolation**
- Via coupling through PDN depends on via localization (amount of energy that goes sideway or comes from PDN) – **covered in "Distant Crosstalk"…**

#2018_03: Y. Shlepnev, Life beyond 10 Gbps: Localize or Fail!, April 13, 2018.

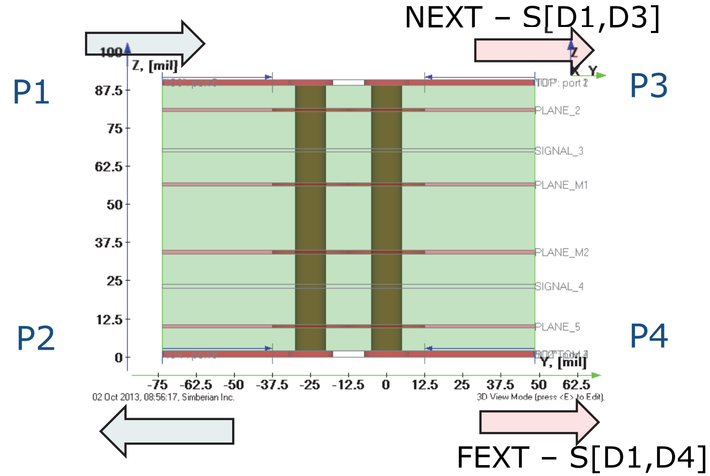
The first use of Simbeor software in 2007 – evaluation of xtalk in vias

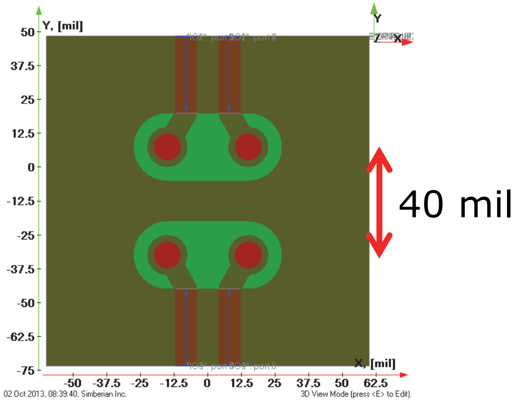


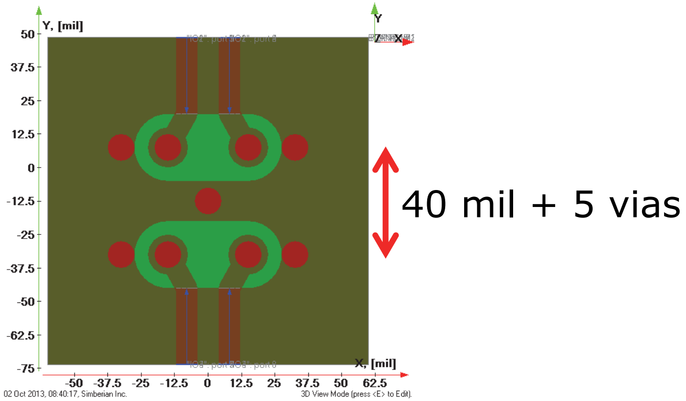
#2007_01: Analysis of via-hole cross-talk and reflection loss for a BGA break-out.

Local Coupling of Differential Vias

Two coupled differential vias in a 120 x 120 mil area caged with PEC wall
Vias are 30 mil apart, antipad 25x55 mil, traces 8 mil MSL, 8 mil separation
The first cage resonance is at about 12 GHz (half wavelength in dielectric)
Stackup from CMP-28 board, Wild River Technology http://wildrivertech.com

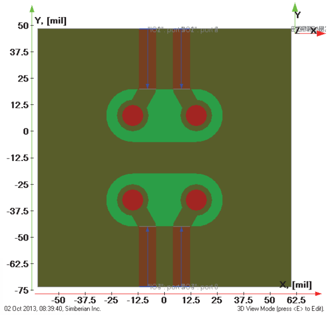






•#2016_13: **How Interconnects Work™**: Crosstalk power flow in differential vias, 10 min - https://youtu.be/IyuQIl8T_uE

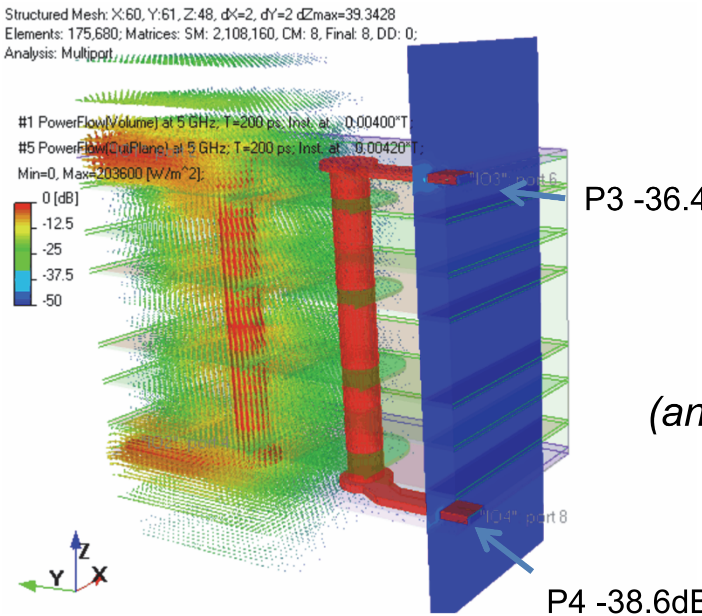
PFD in Differential Vias



10mil diff. vias (30 mil via-to-via), 40mil between diff. pairs (solution XTalk_Vias_CPM28_FRSI)

5 GHz

$\overline{P}_{dB}$

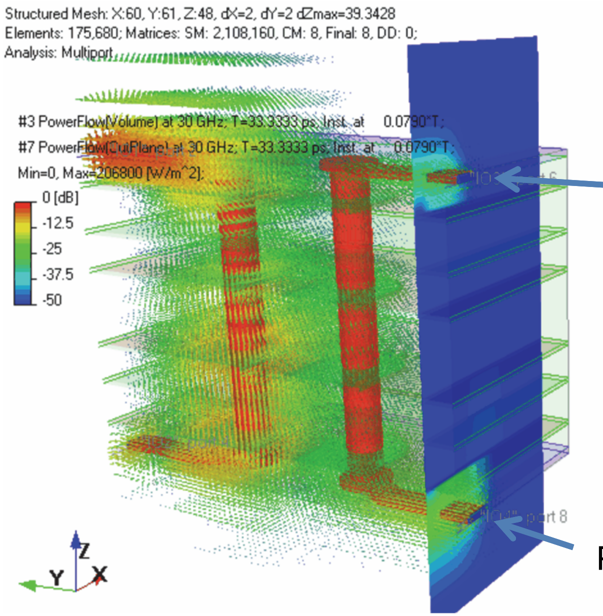


P3 -36.4dB

P4 -38.6dB

30 GHz

$\overline{P}_{dB}$



P3 -23.4dB

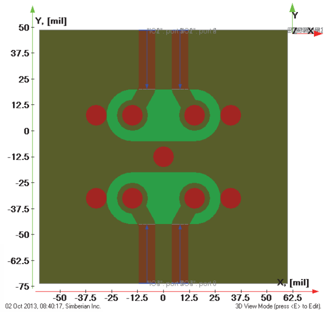
P4 -16.7dB

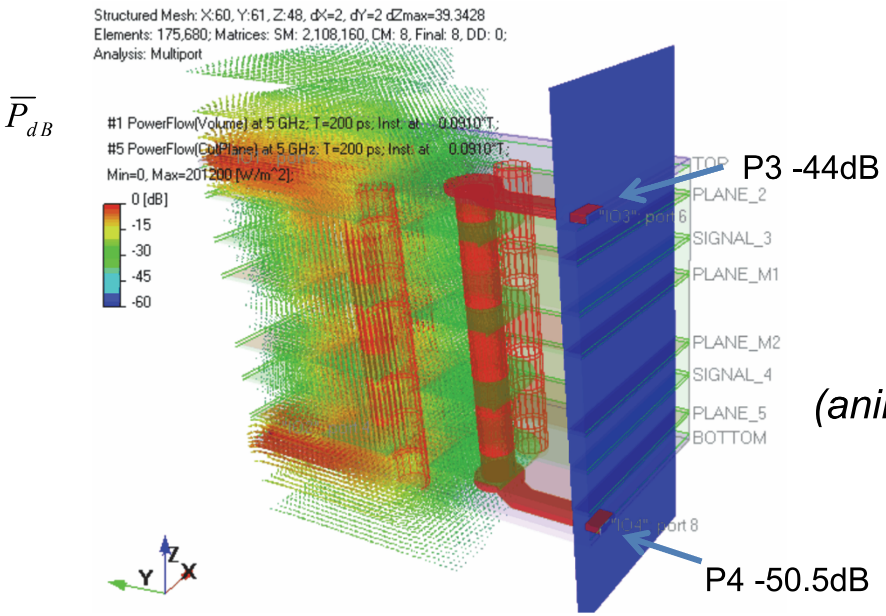
(animated)

Differential excitation, half of the structure is shown

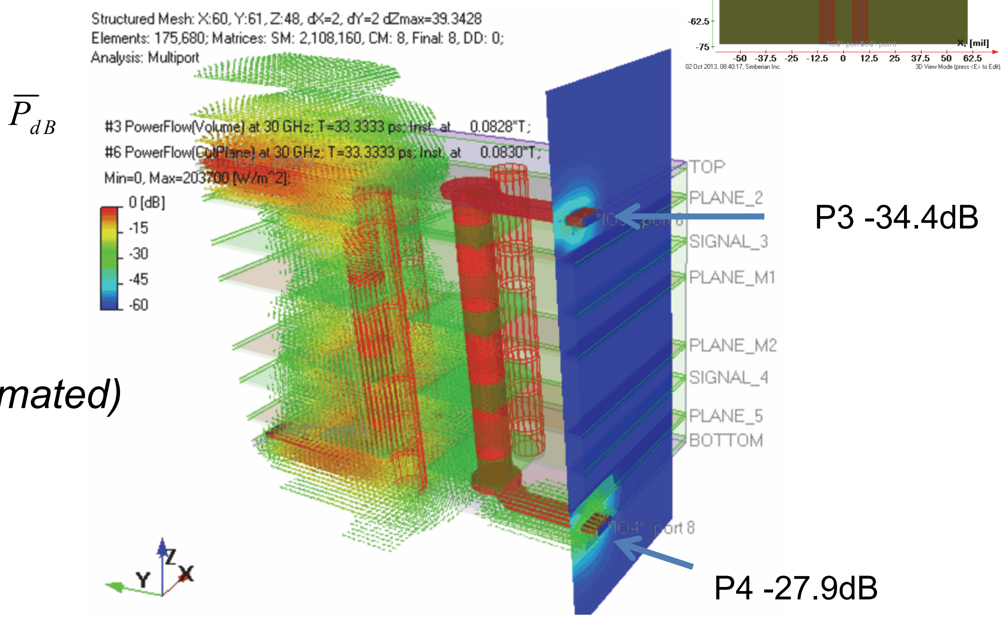
PFD in Differential Vias with Stitching

10mil diff. vias (30 mil via-to-via), 40mil between diff. pairs (solution XTalk_Vias_CPM28_FRSI)





(animated)



Differential excitation, half of the structure is shown

Takeouts

- PCB/Packaging interconnects are open waveguiding structures
  - Energy is distributed in space between traces and reference conductors - anything that gets into that space is coupled
  - Getting out of this space and maintaining reference integrity is a universal way to reduce xtalk
  - Proper guarding and stitching is the future of interconnects
- Modal decomposition is a useful tool to understand and reduce FEXT and NEXT in parallel interconnects
- Only local xtalk in vias can be evaluated in isolation – more on vias in the “Distant Crosstalk” section

OUTLINE

XTalk in Balance of Power

$$P_{out} = P_{in} - P_{reflected} - P_{dissipated} - P_{leaked} + P_{coupled}$$

$$P_{in} = |a_1|^2\ [Wt],\ \ a_2 = 0$$

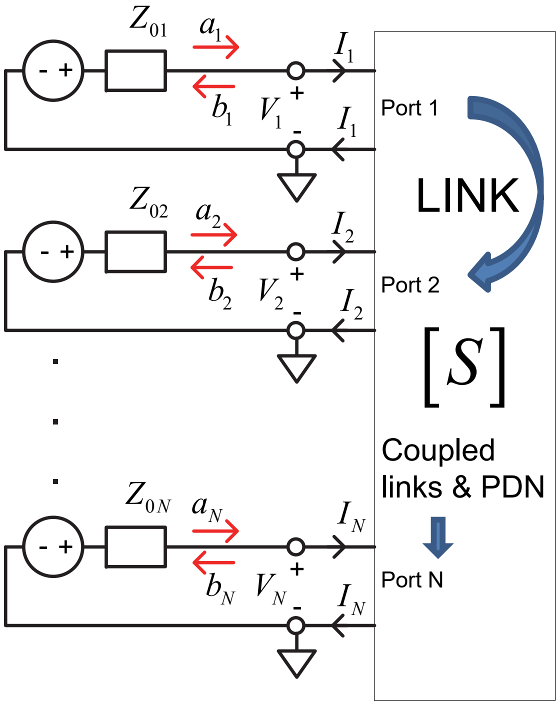
$$P_{out} = |S_{2,1}|^2\ P_{in}$$

$$P_{reflected} = |S_{1,1}|^2\ P_{in}$$

$$P_{dissipated} = \left(1 - \sum_k |S_{k,1}|^2\right) P_{in}$$

$$P_{leaked} = \left(\sum_{k\neq 1,2} |S_{k,1}|^2\right) P_{in}$$

XTalk ⇨ $$P_{coupled} = \sum_{k\neq 1,2} |S_{2,k}|^2 P_{ink}$$



$$\bar{a} = \frac{1}{2} Z_0^{-1/2} \cdot (\bar{V} + Z_0 \cdot \bar{I})$$

$$\bar{b} = \frac{1}{2} Z_0^{-1/2} \cdot (\bar{V} - Z_0 \cdot \bar{I})$$

$\bar{a}, \bar{b} \in C^{N\times 1}$

$Z_0 = diag\{Z_{0i}\} \in R^{N\times N}$

Scattering parameters:

$\bar{b} = S \cdot \bar{a} \quad S \in C^{N\times N}$

$$S_{i,j} = \left.\frac{b_i}{a_j}\right|_{a_k = 0 k\neq j}$$

XTalk quantification is the evaluation of $P_{coupled}$ - it requires signals from other links P_{ink}

XTalk Quantification Approaches

1. Coupling Coefficients (CC): analysis of transmission line cross-sections at one frequency point and use of approximate equations for backward and forward coupling (Kb and Kf)

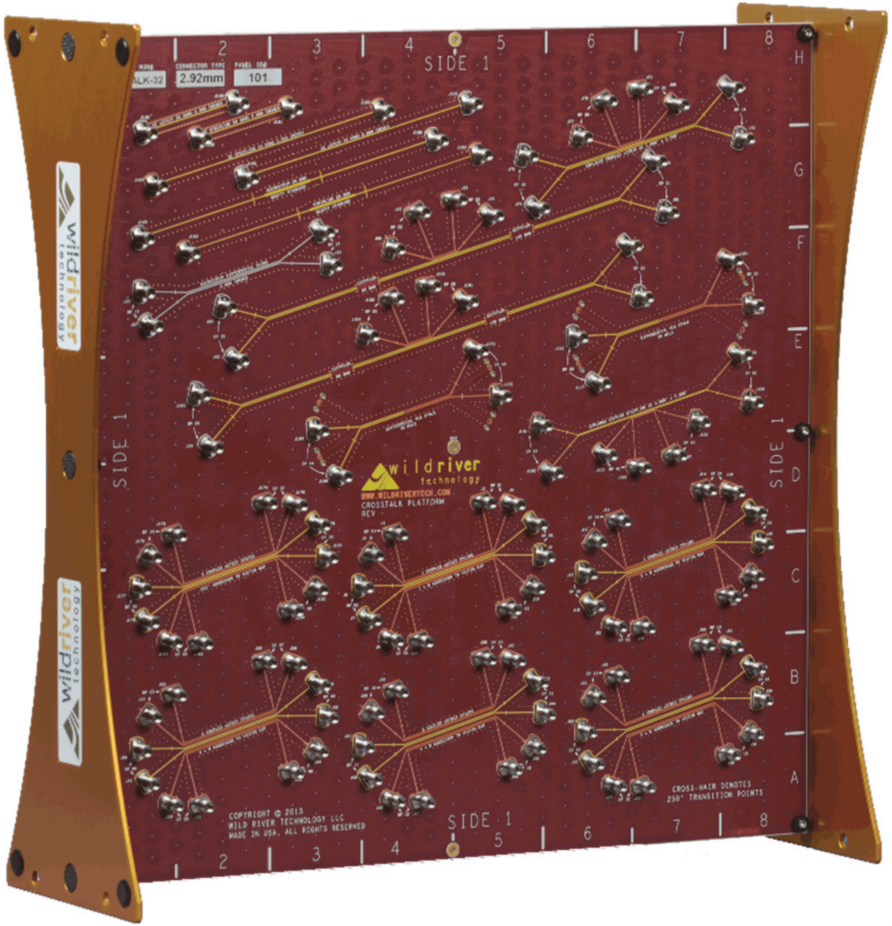
2. Frequency Domain (FD): extraction of S-parameters with coupling in frequency domain and use of crosstalk metrics PSXT, ICR, ICN,…

3. Time Domain (TD): simulation in time domain with step, pulse or PRBS excitation signals (peak voltages or eye distortion)

4. Probabilistic Domain (PD): statistical evaluation of crosstalk impact on channel operating margin (COM) and on bit error rate (BER)

XTalk Validation: XTALK-28/32 Platform

XTALK-28/32 Validation Platform
from Wild River Technology
https://wildrivertech.com/



Prototype of this platform (CMP-08) was used in the extensive analysis to measurement validation project:
#2011_02: J. Bell, S. McMorrow, M. Miller, A. P. Neves, Y. Shlepnev, Unified Methodology of 3D-EM/Channel Simulation/Robust Jitter Decomposition, DesignCon 2011.

XTalk Evaluation with Coupling Coefficients (CC)

- Simulate a cross-section of coupled traces with a field solver at one frequency point
- Use approximate equations for evaluation of forward and backward coupling

$$K_B = \frac{l}{2(T_1 + T_2)} \cdot \left(\frac{L_{21}}{\zeta_1} - \zeta_2 C_{21}\right) \cdot \min(1, \frac{T_1 + T_2}{t_r})$$

$$K_F = -\frac{l}{2 \cdot \max(|T_1 - T_2|, t_r)} \left(\frac{L_{21}}{\zeta_1} + \zeta_2 C_{21}\right)$$

C_{21}, L_{21} - mutual capacitance (Maxwell) and inductance
T_1, T_2 - flight times
ζ_1, ζ_2 - impedances of coupled traces or diff. pairs

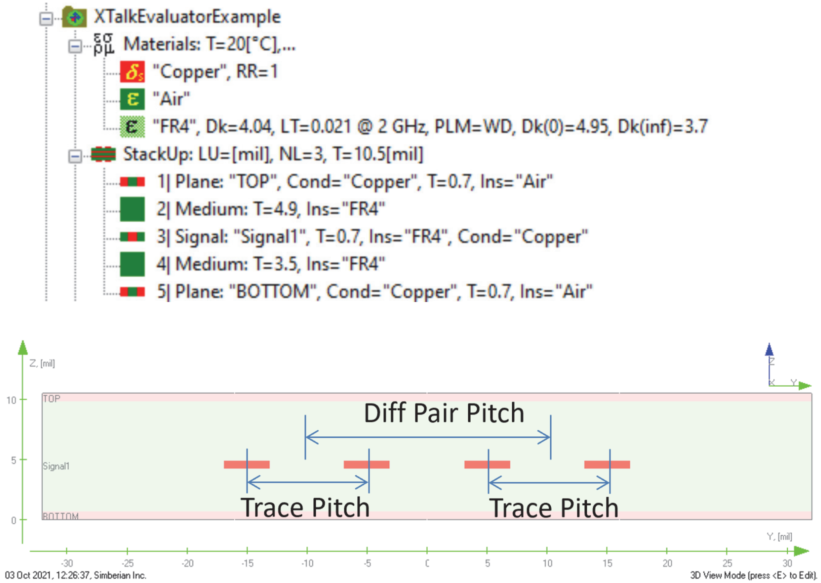
1V step response with rise time tr and **no reflections and no losses** at near end (Kb) and far end(Kf);
For single segment Kb is NEXT and Kf is FEXT;

Can be used as an estimate of the maximal possible step cross-talk in pre- and post-layout analysis;
Under-estimate peak-to-peak xtalk up to 2 times (6dB);

J. E. Bracken, Improved Formulas for Crosstalk Coefficients, DesignCon 2016.
D. B. Jarvis, The Effect of Interconnections on High-Speed Logic Circuits, IEEE Trans. On Electronic Computers, vol. EC-12, 1963, N. 5, pp. 476-487.

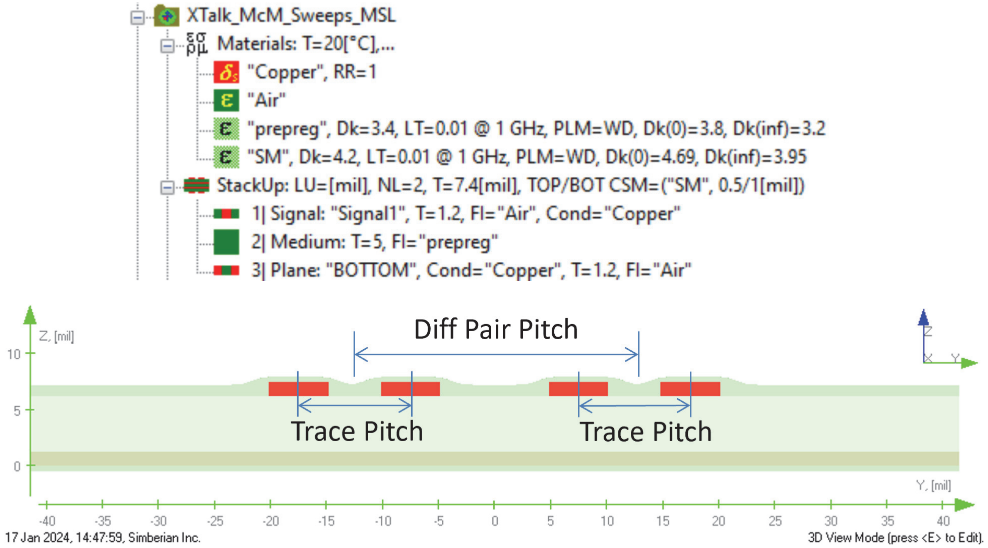
Examples of Xtalk Planning – Pre-Layout

Coupled Differential Striplines



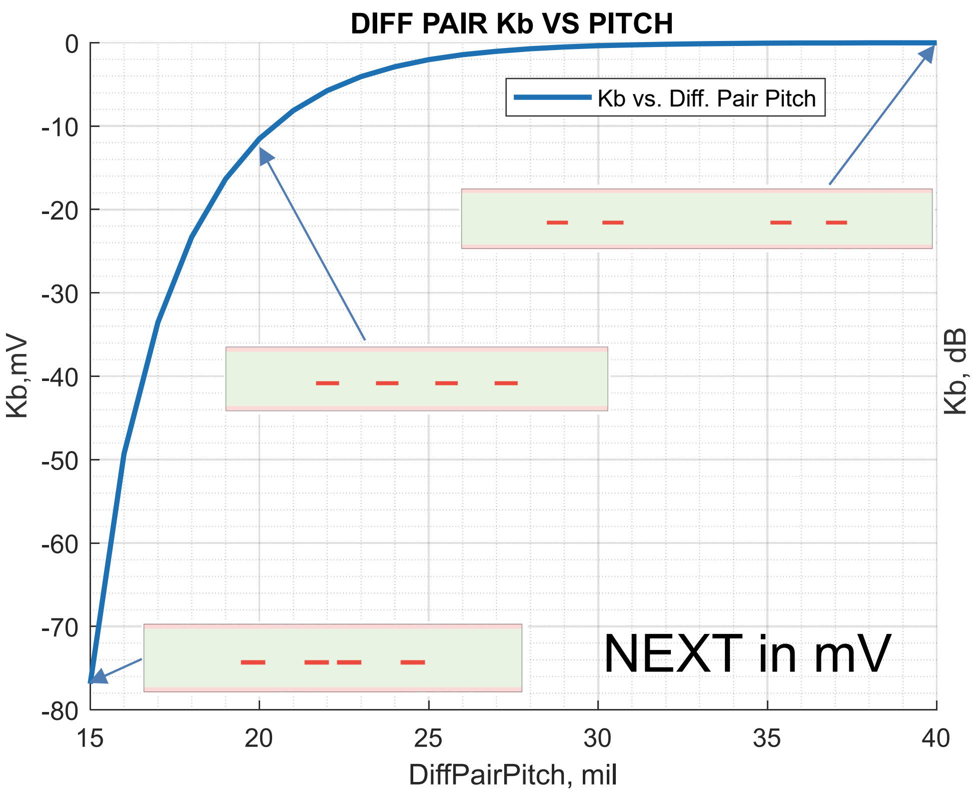
Analysis with Simbeor SFS solver at 1 GHz
Examples from Simbeor SDK AdvXTalkKit:
How to Build Advanced Crosstalk Models for PCB & PKG Traces with Simbeor SDK, Simbeor SDK, 2021.

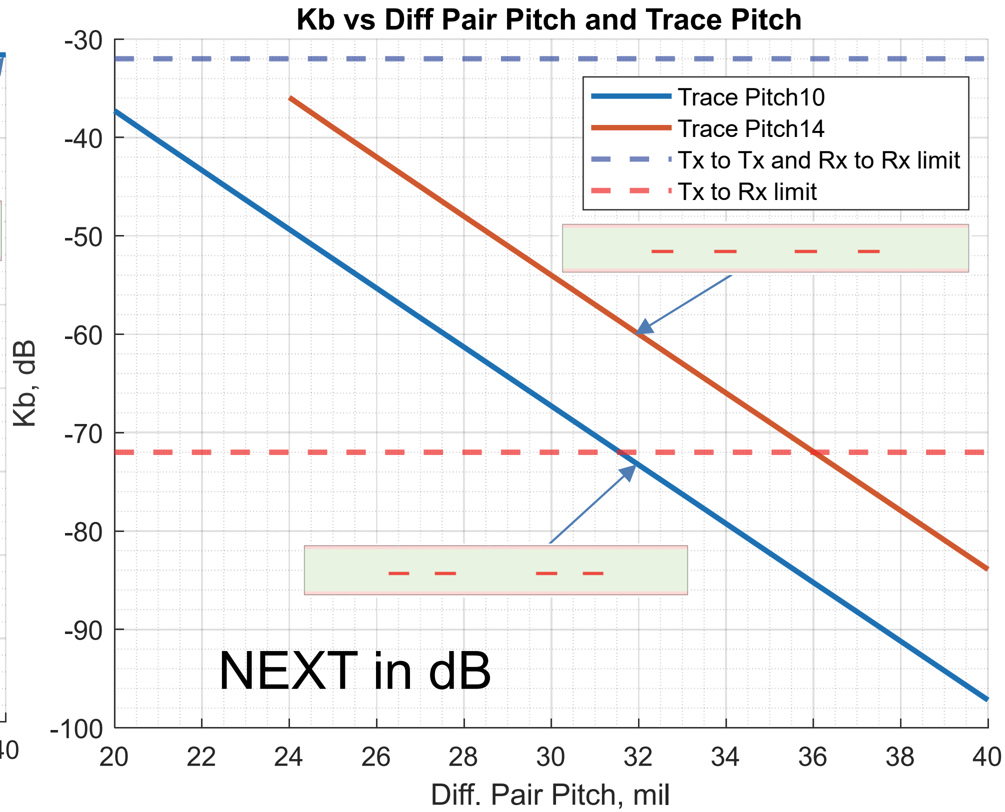
Coupled Differential Microstrip Lines



Inspired by: S. McMorrow, Trace Design For Crosstalk Reduction, Samtek, gEEk spEEk, 2020.

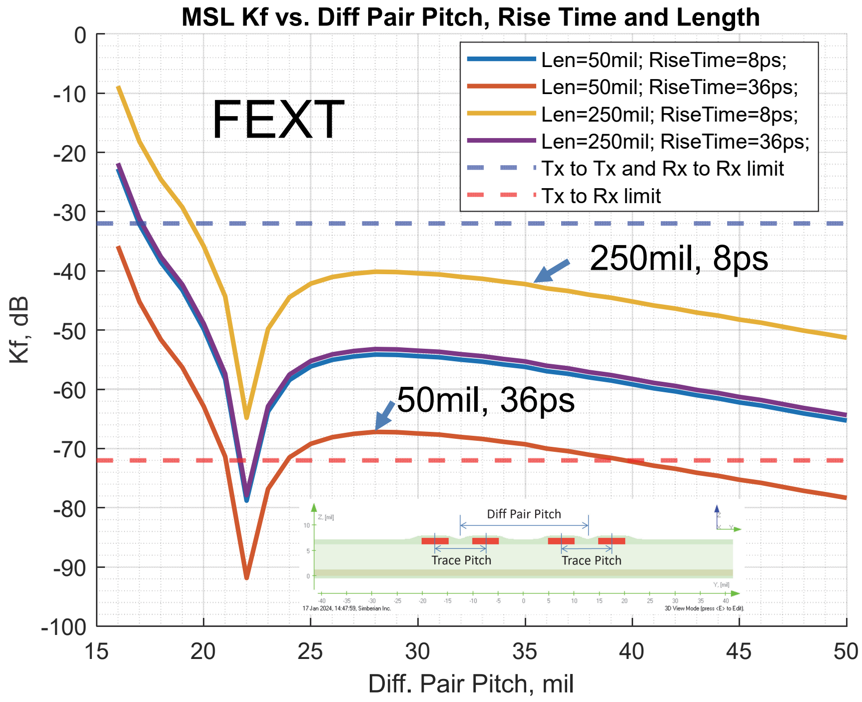
Stripline XTalk Analysis with CC

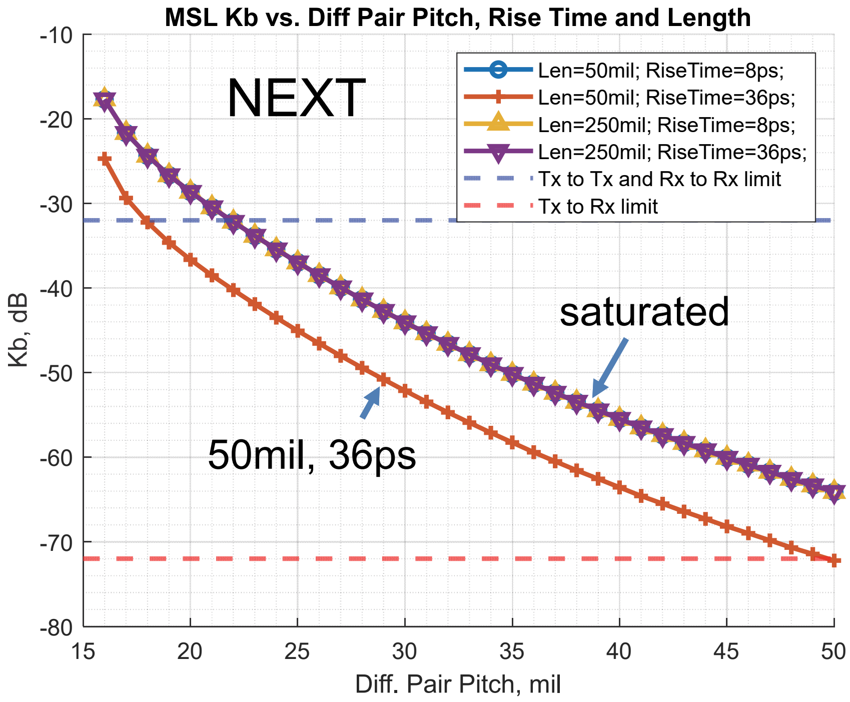




80 Ohm differential striplines (auto-adjusted trace width), 1 inch, 20ps rise time

Microstrip XTalk Analysis with CC





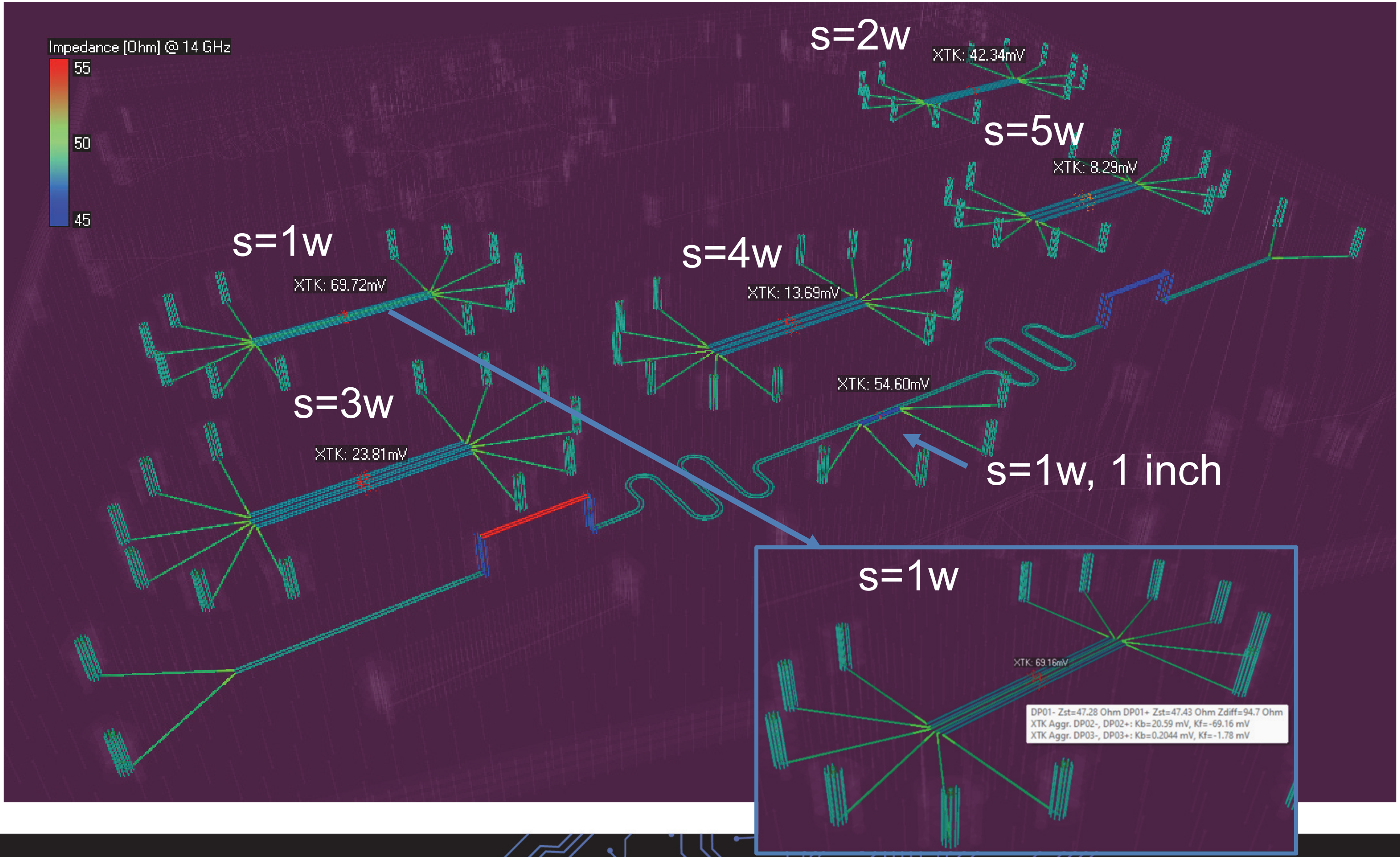
100 Ohm differential microstrips (auto-adjusted trace width), 50mil and 250mil segments, 8ps and 36ps rise time;

Post-Layout Analysis with CC

XTALK-28/32 platform from Wild River Technology

2 inch coupling segments, surface traces w=13.5mil, separation between diff. pairs from 1w to 5w
Rise time 25ps (10-90%)

Impedance of coupled segments is changed



XTalk in Frequency Domain

- Digital signal is a sequence of bits transmitted through PCB or packaging interconnects as sequence of pulses modulated by amplitude (**signal**)
- Signal degradation in interconnects are caused by absorption, dispersion, reflections, coupling and radiation (dissipation)
- **The best way to model those effect is in FREQUENCY DOMAIN**
- In frequency domain all fields, voltages, currents and power are real parts of time-harmonic complex vectors

$$\mathrm{Re}\left[\vec{F}_0(\bar{r}) \cdot e^{i\omega t}\right] \quad \mathrm{Re}\left[\vec{P}_0(\bar{r}) \cdot e^{i2\omega t}\right] \quad \omega = 2\pi \cdot f = \frac{2\pi}{T}$$

- **What is the bandwidth of the signal in the frequency domain and over what bandwidth we have to model or measure it?**

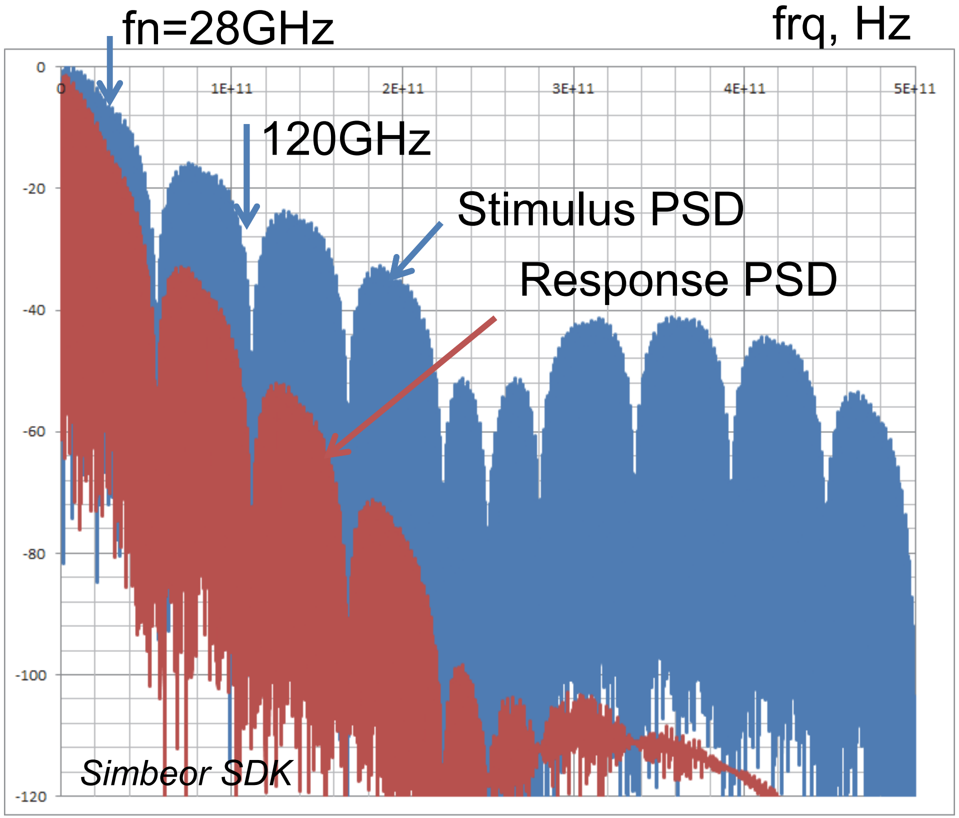
Signal Bandwidth: Power Spectral Density

Example: **112Gbps PAM4, 4ps rise time**
10-inch strip line, W=12mil, H=20mil
Meg7 – Wideband Debye: Dk=3.17, LT=0.0011 @ 1 GHz
Copper: RR=1.4, Roughness – Huray- Bracken Model:
SR=0.14 um, RF=8.7

Bandwidth for IL and FEXT depends on link length – the longer the link, the smaller bandwidth may be used
Only 2-3 Nyquist may be required for accurate analysis
Formal pulse analysis can be used for better estimate (*)

Bandwidth for NEXT (crosstalk in vias) is practically the same as the stimulus ☹ - **maximal possible bandwidth should be used**

(*) Y. Shlepnev, How Interconnects Work: ***Bandwidth for Modeling and Measurements****, Signal Integrity Journal, April 12, 2022*



Bandwidth for Common Signaling Standards

Ethernet, OIF, PCIe 5 and 6, GDDR 5 and 6,…

DR, Gbps	BT or ST, ps	RT, ps	Fn, GHz	BW, GHz
10, NRZ	100	50	5	20
16, NRZ	62.5	30	8	32
28, NRZ	35.7	18	14	54
32, NRZ	31.25	16	16	61
56, PAM4	35.7	18	14	54
64, PAM4	31.25	16	16	61
112, PAM4	17.8571	10	32	101
128, PAM4	15.625	10	32	108
224, PAM4	8.92857	8	64	120
448, PAM4	4.46429	4	112	239

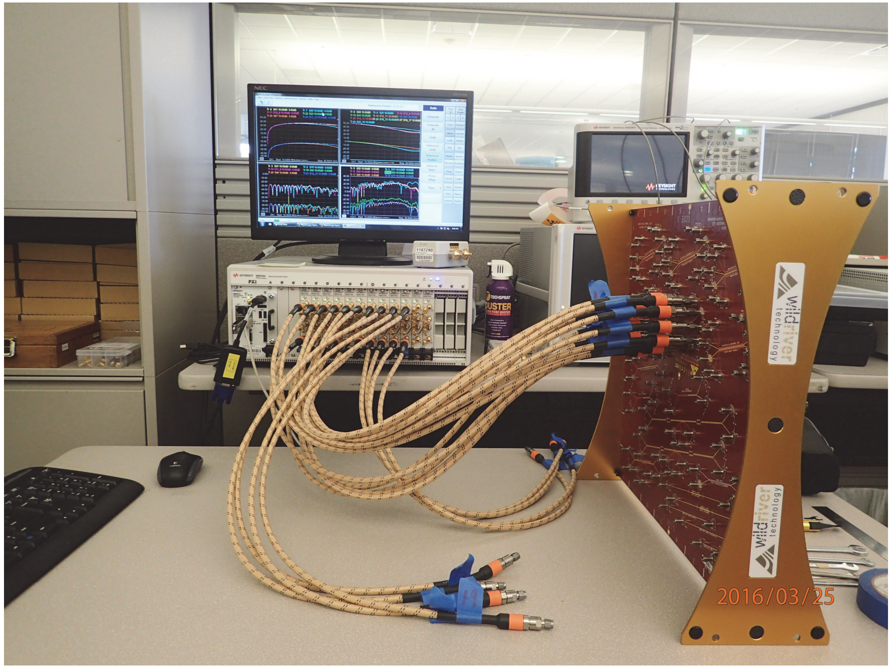
BT – bit time for NRZ
ST – symbol time for PAM4
RT – rise/fall time (0-100%)
Fn – Nyquist frequency
BW – bandwidth estimate is based on -25 dB spectrum drop-off (-20dB for 224 and higher)

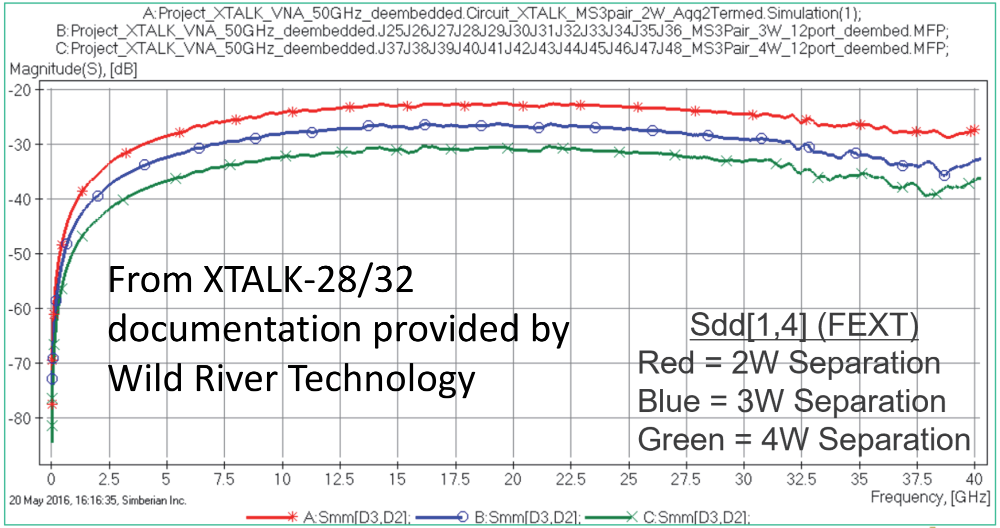
Bandwidth for the analysis may be further reduced by attenuation of high-frequency harmonics

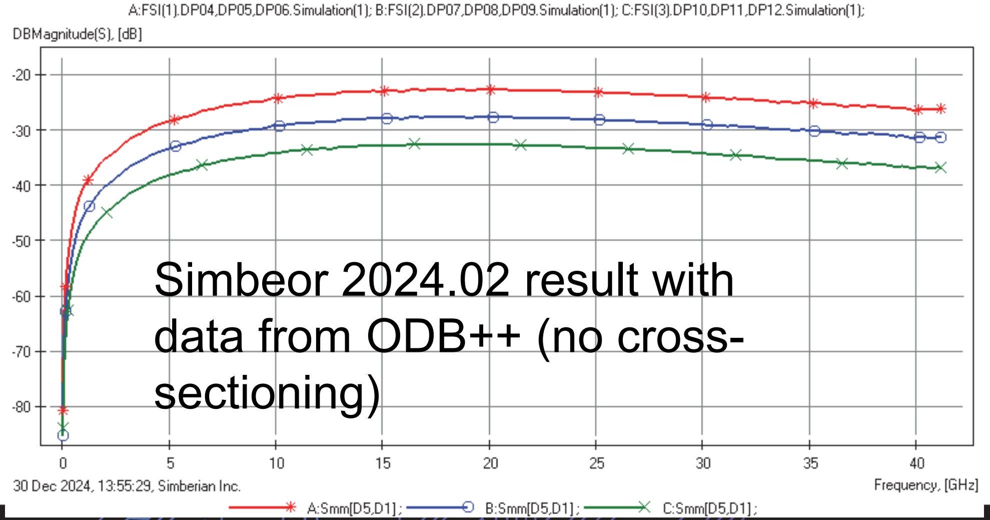
It may be not applicable to near-end crosstalk…

XTALK-28/32 – FEXT

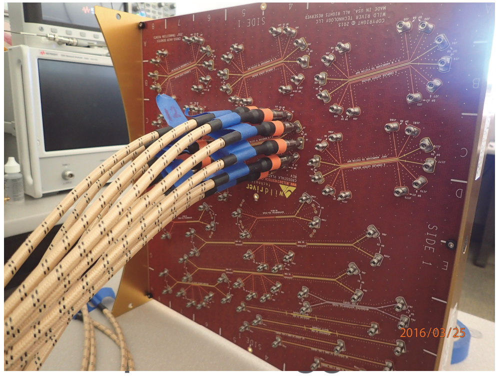
S-parameters is the foundation for all other xtalk metrics

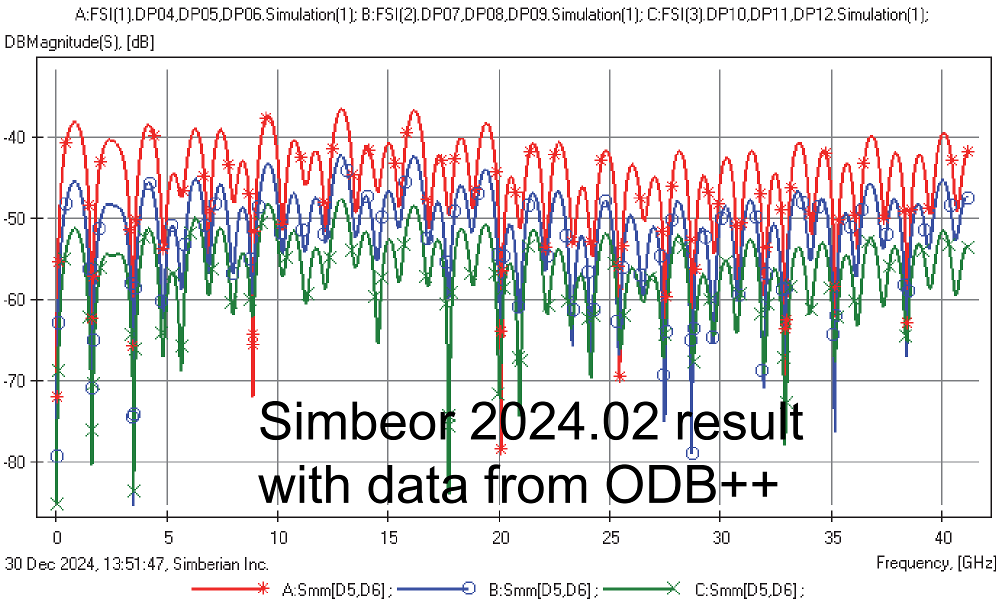


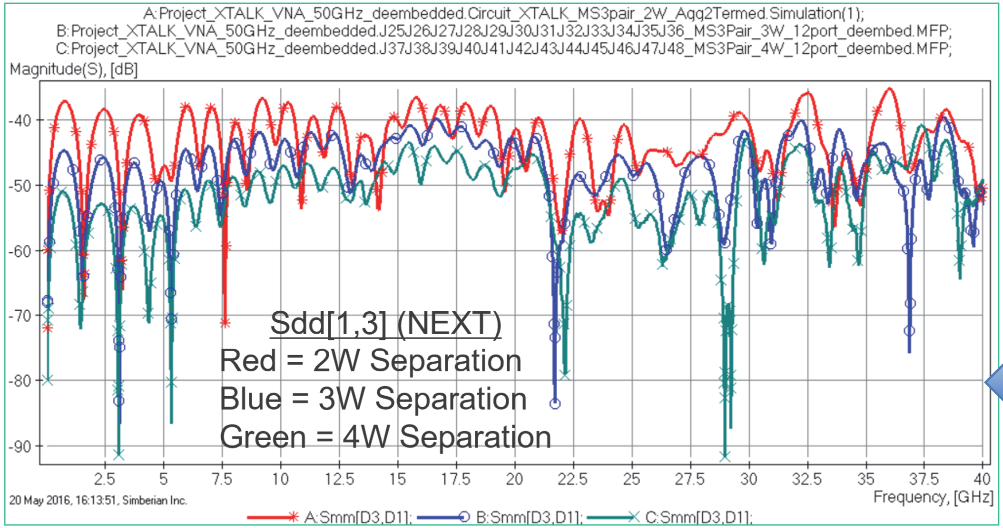




XTALK-28/32 – NEXT (no cross-sectioning)







From XTALK-28/32 documentation provided by Wild River Technology

Power Sum XTalk (PSXT)

- S-parameters of coupled links can be directly used to simulate the effect of coupling in time domain or evaluate the probability density function of crosstalk
- Preliminary evaluation of xtalk can be done with **Power Sum Crosstalk (PSXT)** (OIF-CEI and IEEE 802.3 standards):

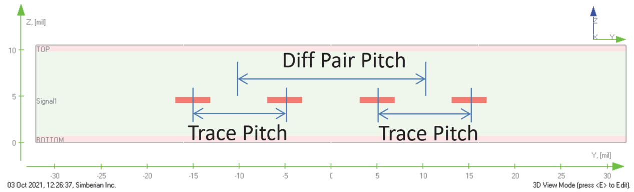
$$PSXT_i = 10 \cdot log\left(\sum_{j \in \Omega_{XT}} \left|S_{i,j}\right|^2\right) \text{ [dB]}$$ - total PSXT (MDXT)

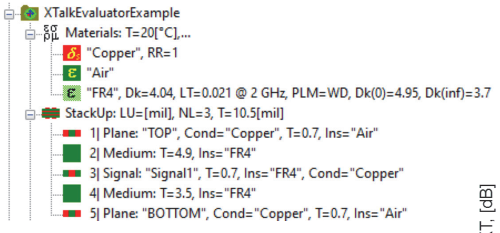
$$PSNEXT_i = 10 \cdot log\left(\sum_{j \in \Omega_{NEXT}} \left|S_{i,j}\right|^2\right) \text{ [dB]}$$ - Near-End PSXT (MDNEXT)

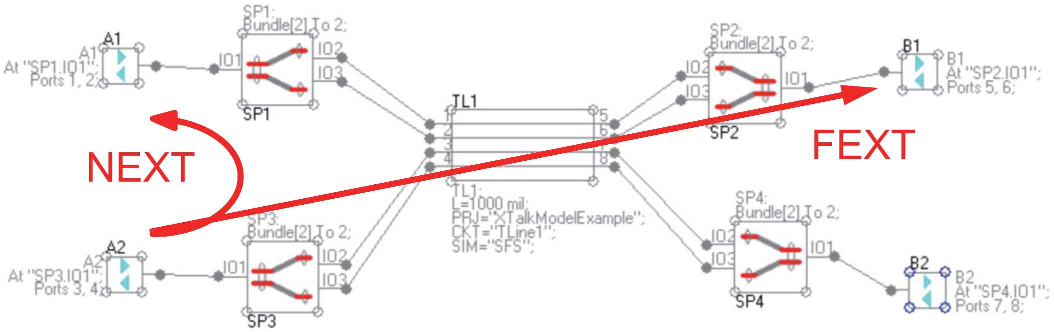
$$PSFEXT_i = 10 \cdot log\left(\sum_{j \in \Omega_{FEXT}} \left|S_{i,j}\right|^2\right) \text{ [dB]}$$ - Far-End PSXT (MDFEXT)

- **PSXT is a sum of squares of S-matrix elements from all possible aggressors at a victim receiver port expressed in dB**

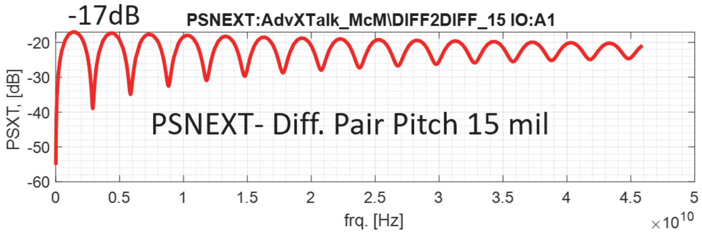
PSXT of Stripline Segment

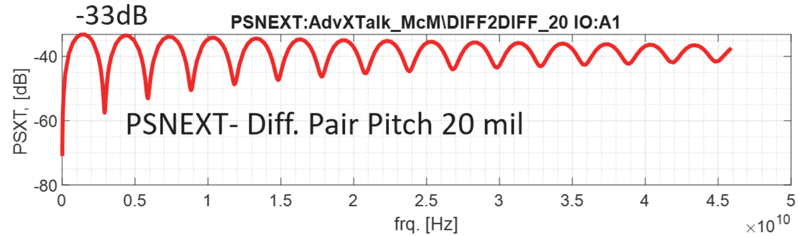




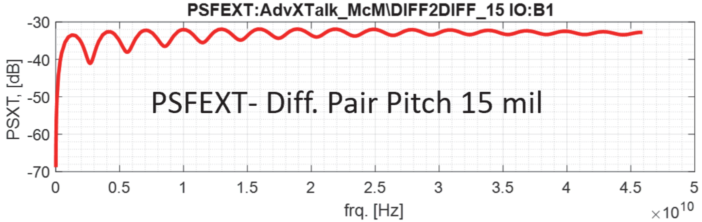


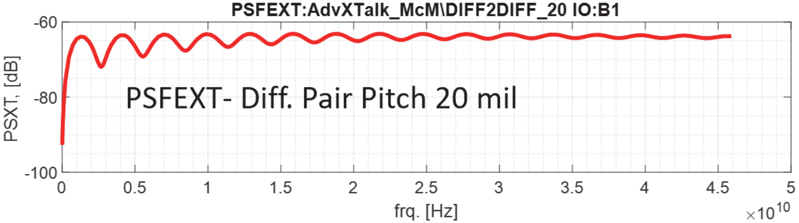
PSXT in 1 inch segment of coupled differential striplines (same as corresponding S-parameters)





Examples from Simbeor SDK AdvXTalkKit





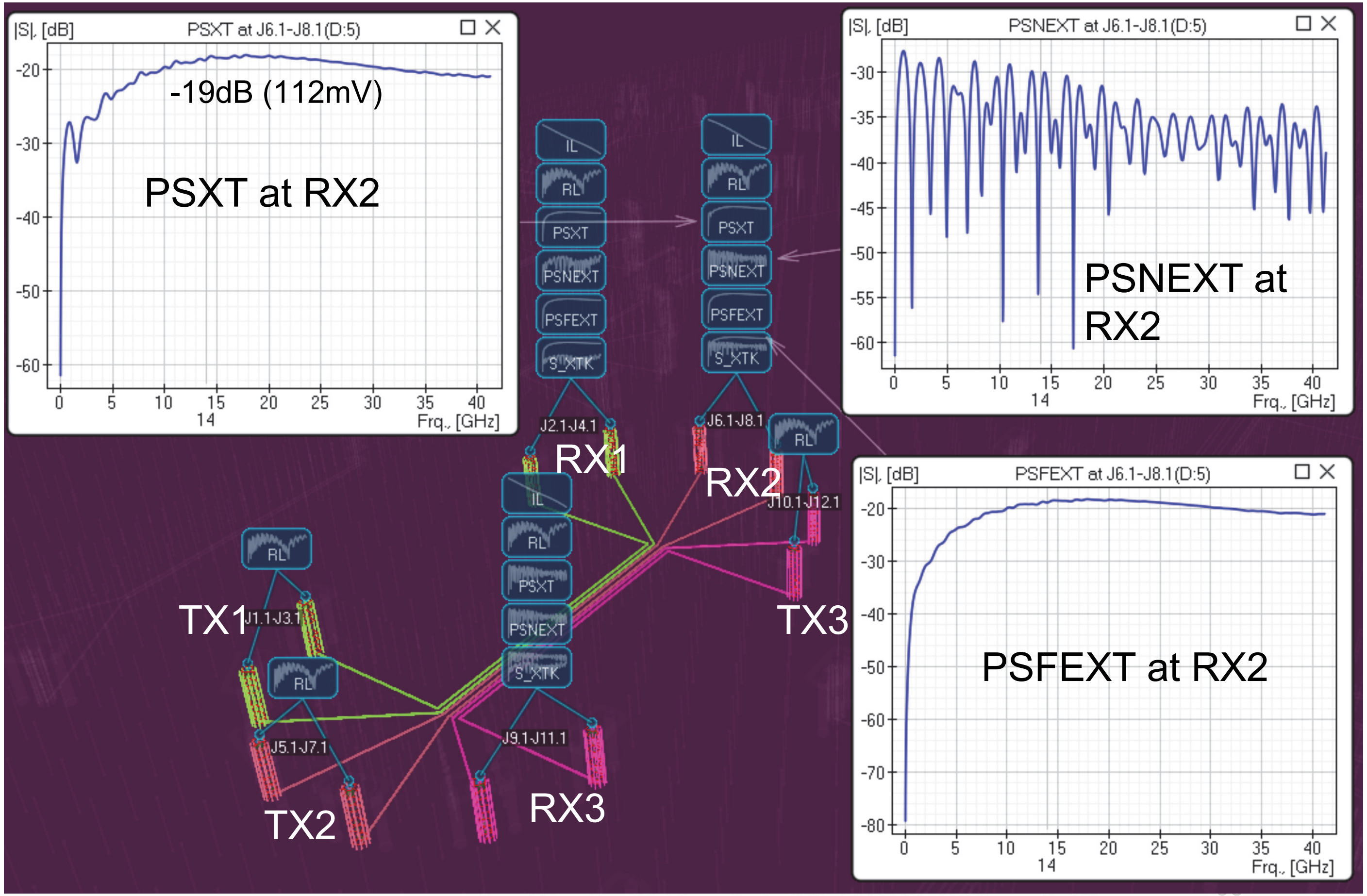
XTALK-28/32: PSXT S=1w

PSXT is a superposition of the aggressor's signals that does not account for the phases of signal harmonics

PSFEXT dominates here

PSNEXT contributes too

Trace width is 13.5mil and differential trace pitch is 37mil , diff. pair separation 1w, 2-inch coupling length



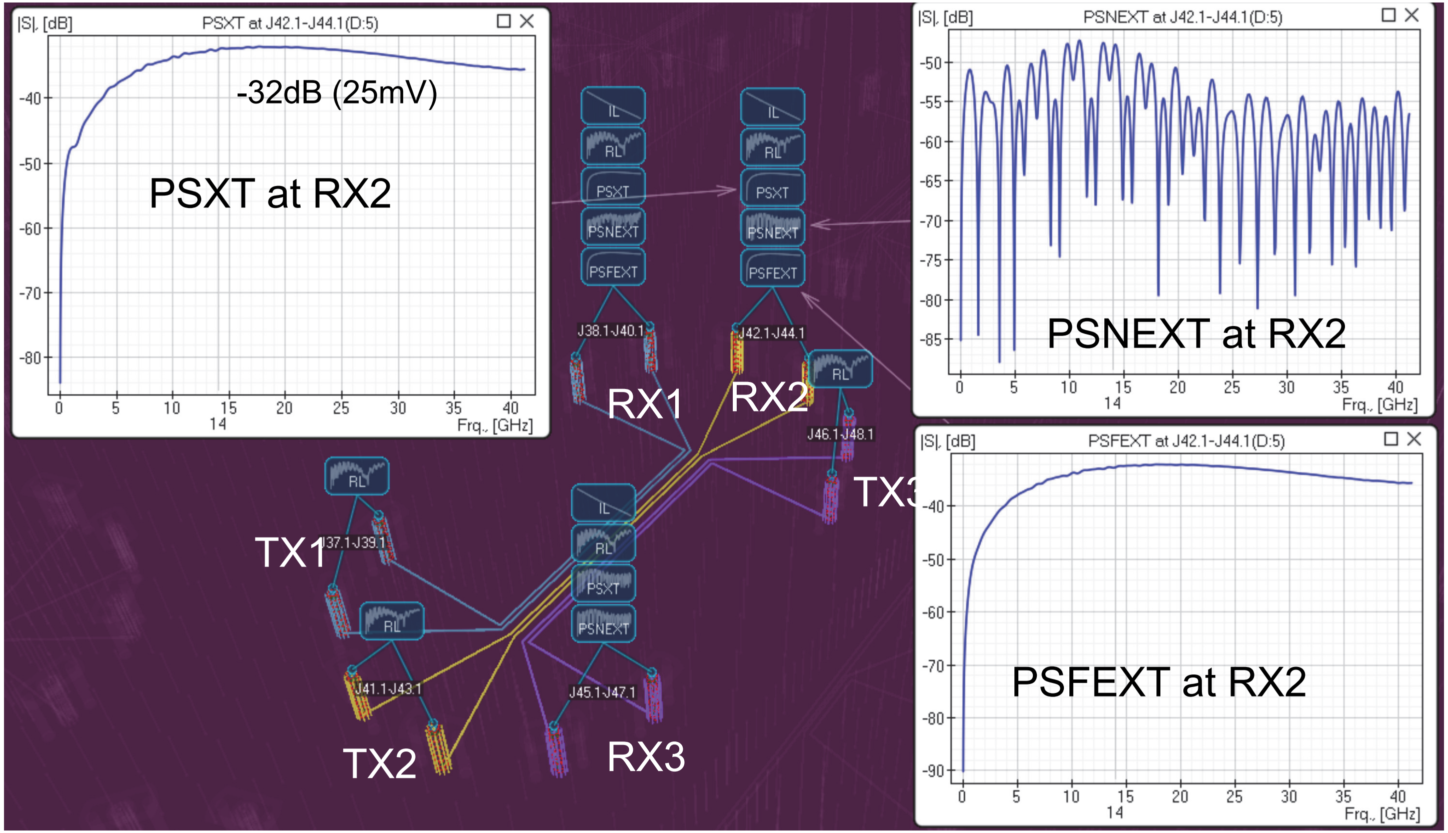
XTALK-28/32: PSXT S=4w

Separation reduces XTalk

PSNEXT becomes negligible

PSFEXT persists and dominates

Trace width is 13.5mil and differential trace pitch is 37mil, diff. pair separation 4w, 2-inch coupling length



Insertion Loss to Crosstalk Ratio (ICR)

- Same level of PSXT may be acceptable for a link with small losses, but cause failure in a link with large losses
- **Insertion loss to crosstalk ratio or ICR metric** can be used to evaluate and quantify the impact of the crosstalk:

$ICR_{i,j} = IL_{i,j} - PSXT_i$ [dB] ICR at port *i*

$IL_{i,j} = 20 \cdot log\left(\left|S_{i,j}(f)\right|\right)$ [dB] insertion loss at port *i* (negative)

$PSXT_i = 10 \cdot log\left(\sum_{j\in\Omega_{XT}}\left|S_{i,j}\right|^2\right)$ [dB] power sum crosstalk at port *i* (negative)

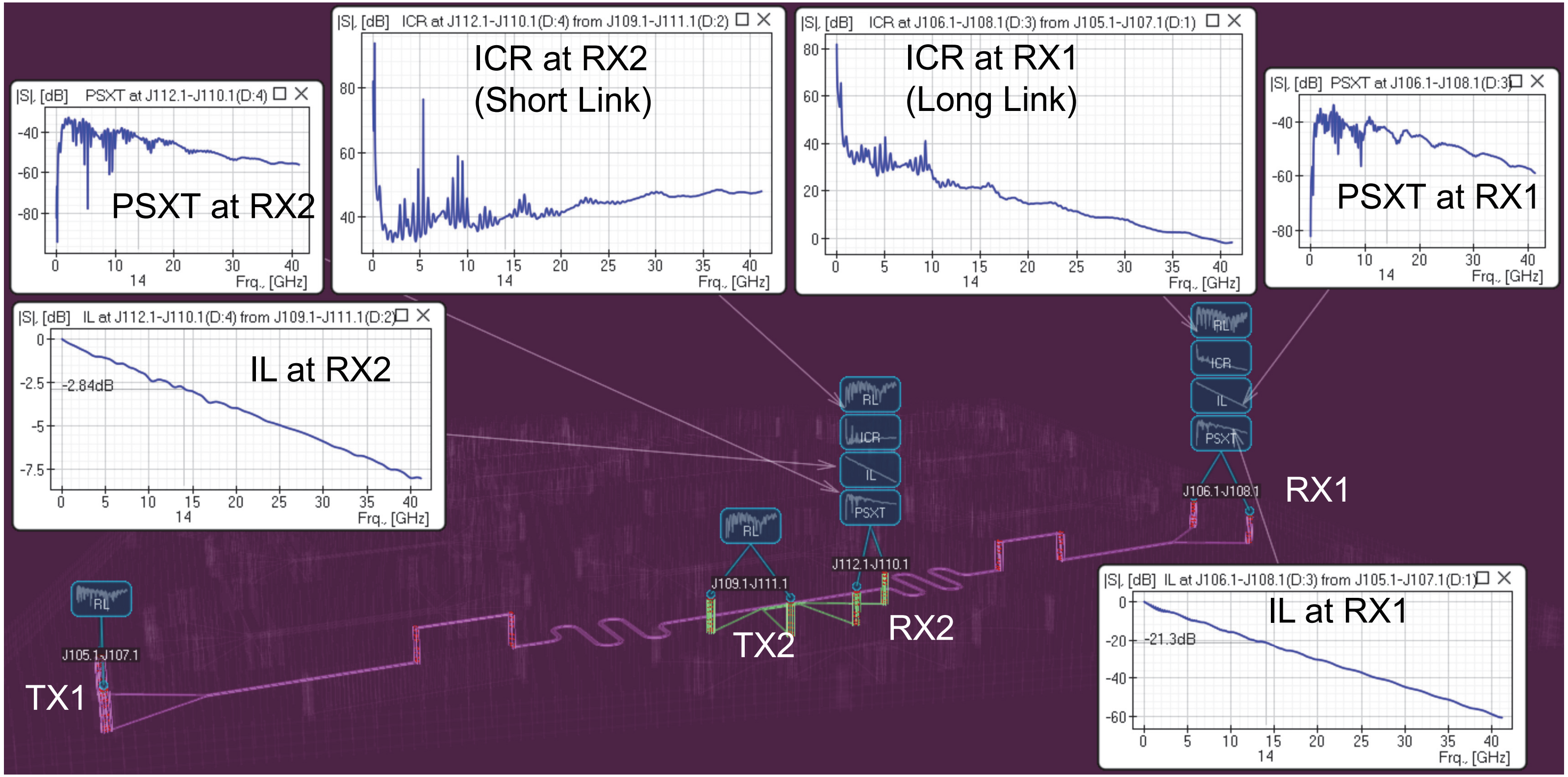
- ICR is a king of signal to noise ratio: IL is signal at a receiver and PSXT is noise
- **The larger values of ICR mean smaller impact of the crosstalk on the signal**

XTALK-28/32: ICRs for Long and Shor Links

Same PSXT in both links

ICR is larger (better) at high frequencies in short link

Trace width is 13.5mil and differential trace pitch is 37mil, diff. pair separation 1w , 1-inch coupling length



Integrated Crosstalk Noise (ICN)

- PSXT and ICR are frequency-dependent metrics – do not account for signal spectrum
- **Integrated crosstalk noise (ICN)** metric accounts for the signal spectrum and filtering properties of a transmitter and a receiver as follows

$$\sigma_{XTK} = \sqrt{\sigma_{NEXT}^2 + \sigma_{FEXT}^2}$$

$$\sigma_{NEXT} = \sqrt{2 \cdot \Delta f \cdot \sum_k W_{nt}(f_k) \cdot NXT(f_k)}\,, \quad NXT(f_k) = \sum_{j \in \Omega_{NEXT}} \left|S_{i,j}\right|^2$$

$$\sigma_{FEXT} = \sqrt{2 \cdot \Delta f \cdot \sum_k W_{ft}(f_k) \cdot FXT(f_k)}\,, \quad \mathrm{FXT}(f_k) = \sum_{j \in \Omega_{FEXT}} \left|S_{i,j}\right|^2$$

W_{NEXT} and W_{FEXT} are computed with the rise and fall time of the near and far end transmitters (aggressors), baud rate (bit or symbol rate), reference receiver and transmitter bandwidth and amplitudes of the near and far end aggressors - - *see definitions in in IEEE Std. 802.3 standard* and more in *M. Shimanouchi, H. Wu, M. P. Li, Evolution of Various Crosstalk Metrics and Evaluation Methods for High-Speed Serial Link and Their Complementary Characteristics, DesignCon 2019.*

$$W_{nt}(f_n) = (A_{nt}^2/f_b)\,\mathrm{sinc}^2(f_n/f_b)\left[\frac{1}{1+(f_n/f_{nt})^4}\right]\left[\frac{1}{1+(f_n/f_r)^8}\right]$$

$$W_{ft}(f_n) = (A_{ft}^2/f_b)\,\mathrm{sinc}^2(f_n/f_b)\left[\frac{1}{1+(f_n/f_{ft})^4}\right]\left[\frac{1}{1+(f_n/f_r)^8}\right]$$

XTALK-28/32: ICN vs. IL at Nyquist Frequency

Parameters defining weight W(f)

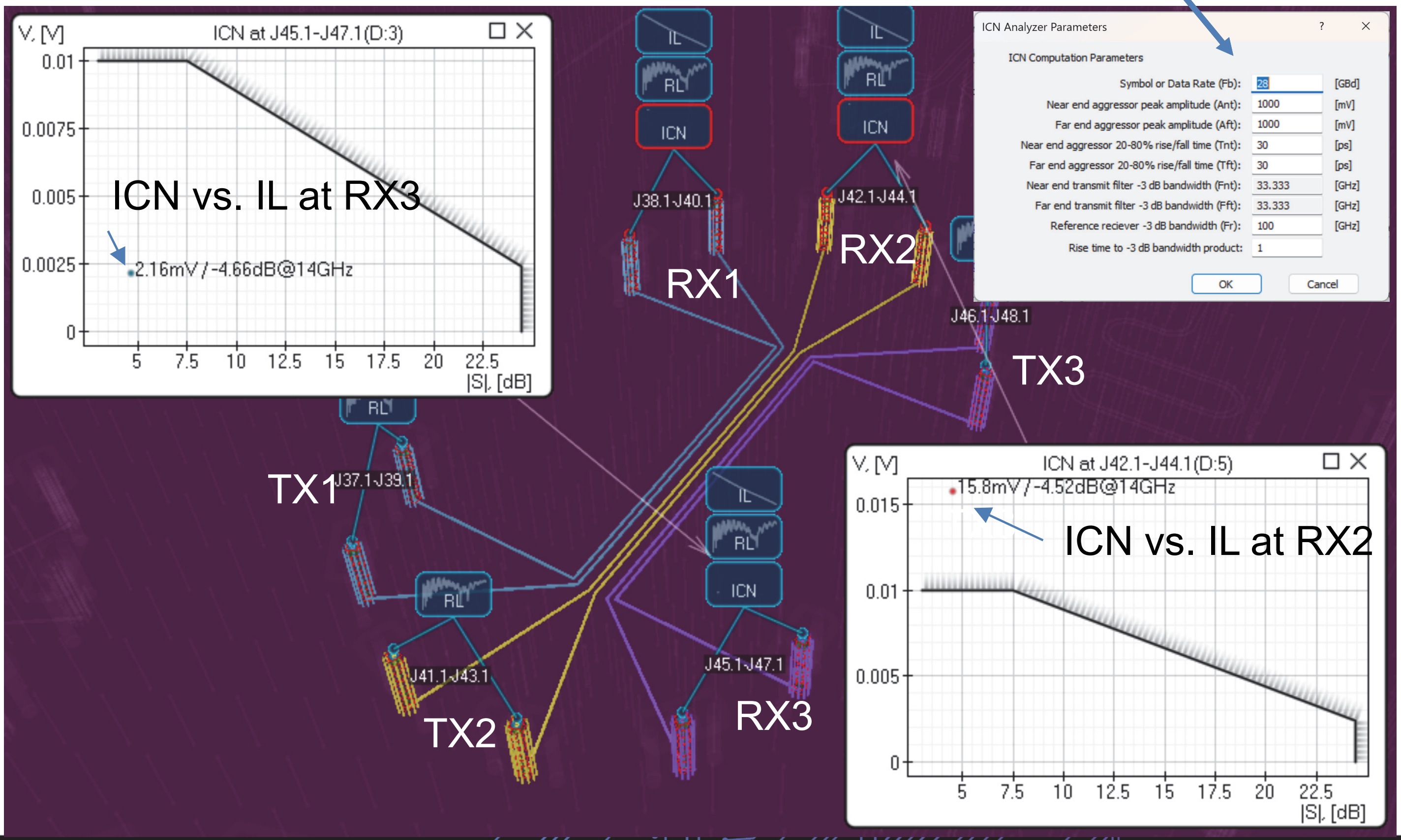
Masks from *IEEE Std. 802.3 standard:* allows larger crosstalk in links with smaller insertion losses

IL is the same in all links

RX3 has small NEXT - pass

RX1 and RX2 have large FEXT – do not pass

Trace width is 13.5mil and differential trace pitch is 37mil, diff. pair separation 4w, 2-inch coupling length

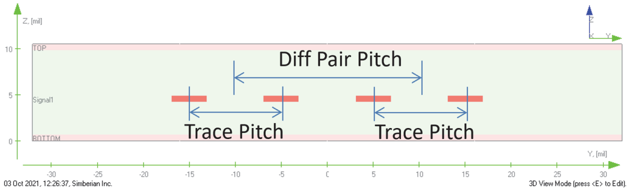


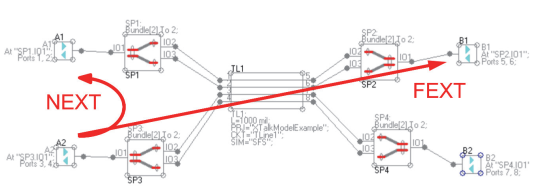
XTalk in Time Domain

- **Measure the crosstalk values directly in time domain as maximal peak to peak value of a voltage response** at a victim IO with a stimulus attached to the aggressor transmitter IO
- This type of analysis can be done with realistic models of transmitter and receiver and accounts for the reflections from non-ideal terminations
- Also, the analysis of a victim link in time domain with one or multiple aggressors is useful to understand the “evasive” nature of the cross-talk
- **Analysis with bitstream signals can be used to understand how crosstalk affect the eye diagram and bit error rate**
- Analysis in time domain is practically always done with S-parameters extracted in frequency domain – rational compact models of S-parameters are used here

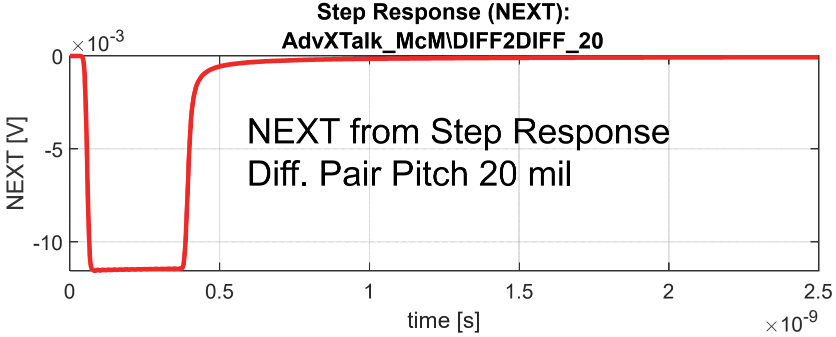
XTalk from Step and Pulse Response

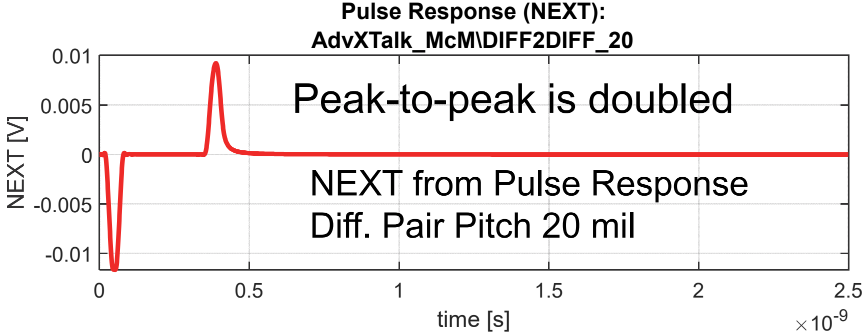
1 inch segment, trace pitch 10 mil, trace width is 3.8 mil, rise time 20ps

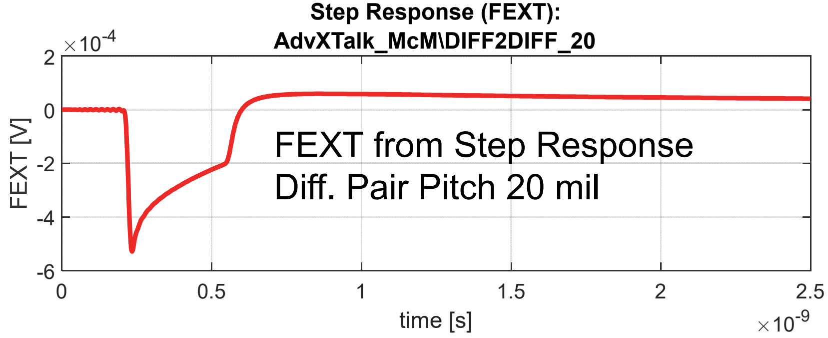


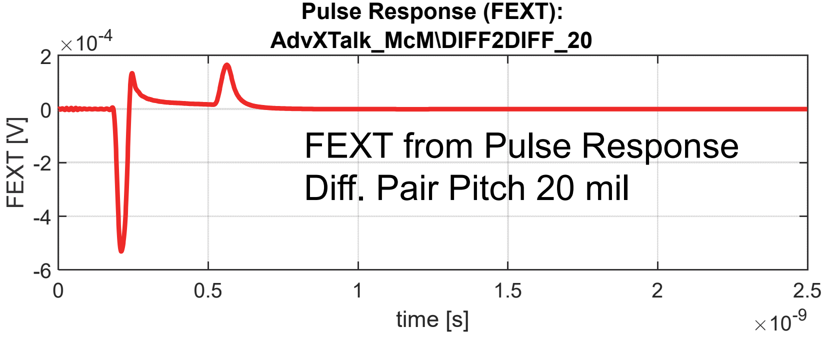


Example from Simbeor SDK AdvXTalkKit





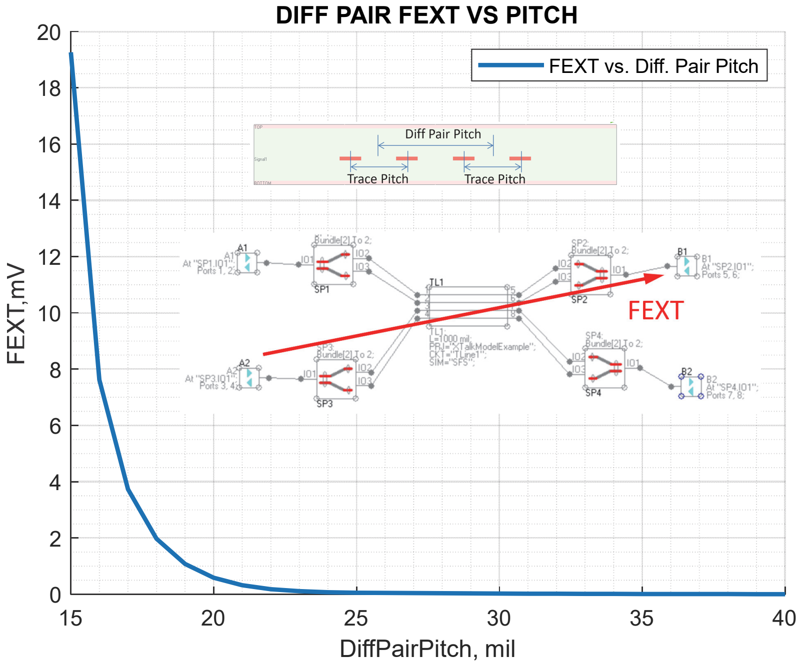


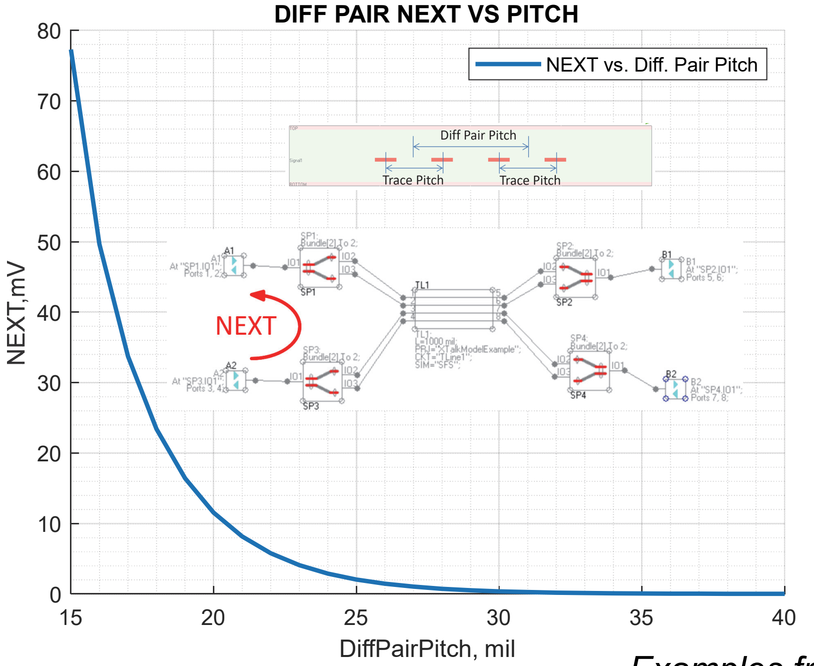


XTalk from Step Response

1 inch segment, trace pitch is 10 mil, trace width is 3.8 mil, xtalk from step response with 20ps rise time (stimulus 2V in series with 100 Ohm)

Includes reflections and losses – unlike analysis with coupling coefficients (compare)





Examples from Simbeor SDK AdvXTalkKit

XTALK-28/32: XTalk in TD, S=1w

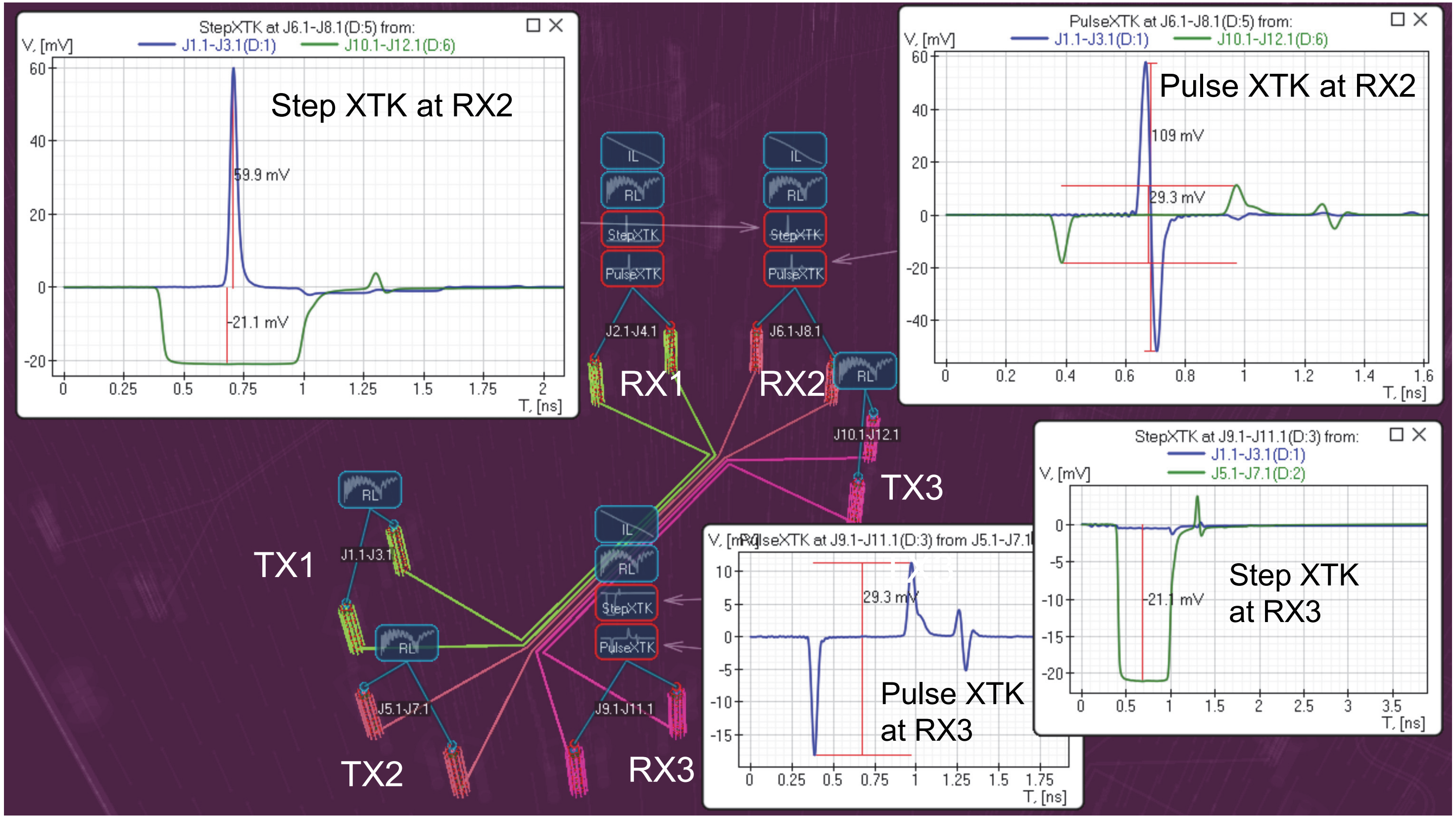
Comparison of step and pulse xtalk responses

Stimulus 2V, 100 Ohm (effective 1V), 25ps rise time (10-90%), Fast SI Analysis

Pulse response may double the observed xtalk values

NEXT + FEXT ~140mV

Trace width is 13.5mil and differential trace pitch is 37mil, diff. pair separation 1w , 2-inch coupling length

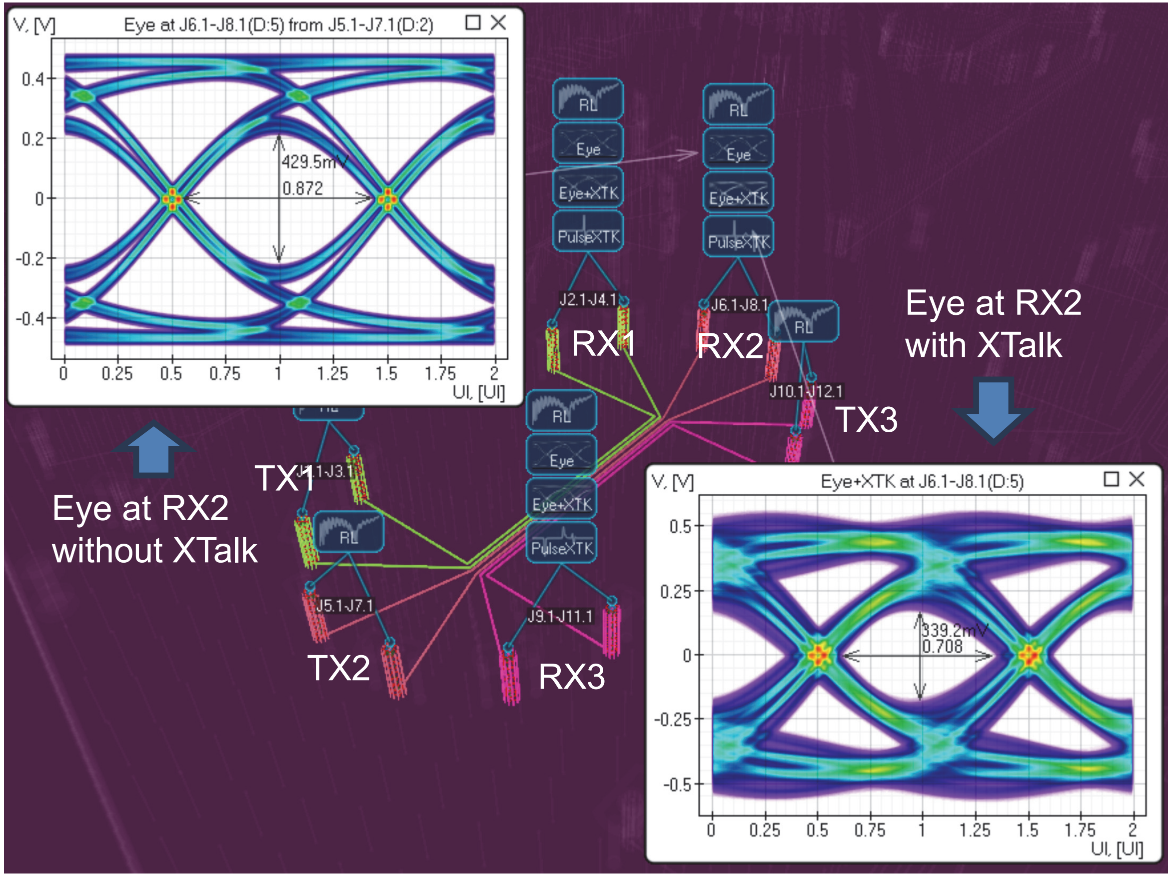


XTALK-28/32: XTalk in TD, S=1w

XTalk on Eye Diagram

Stimulus 28 Gbps PRBS32, 25ps rise time (10-90%), 1V, 100 Ohm, Fast SI Analysis

Trace width is 13.5mil and differential trace pitch is 37mil, diff. pair separation 1w , 2-inch coupling length



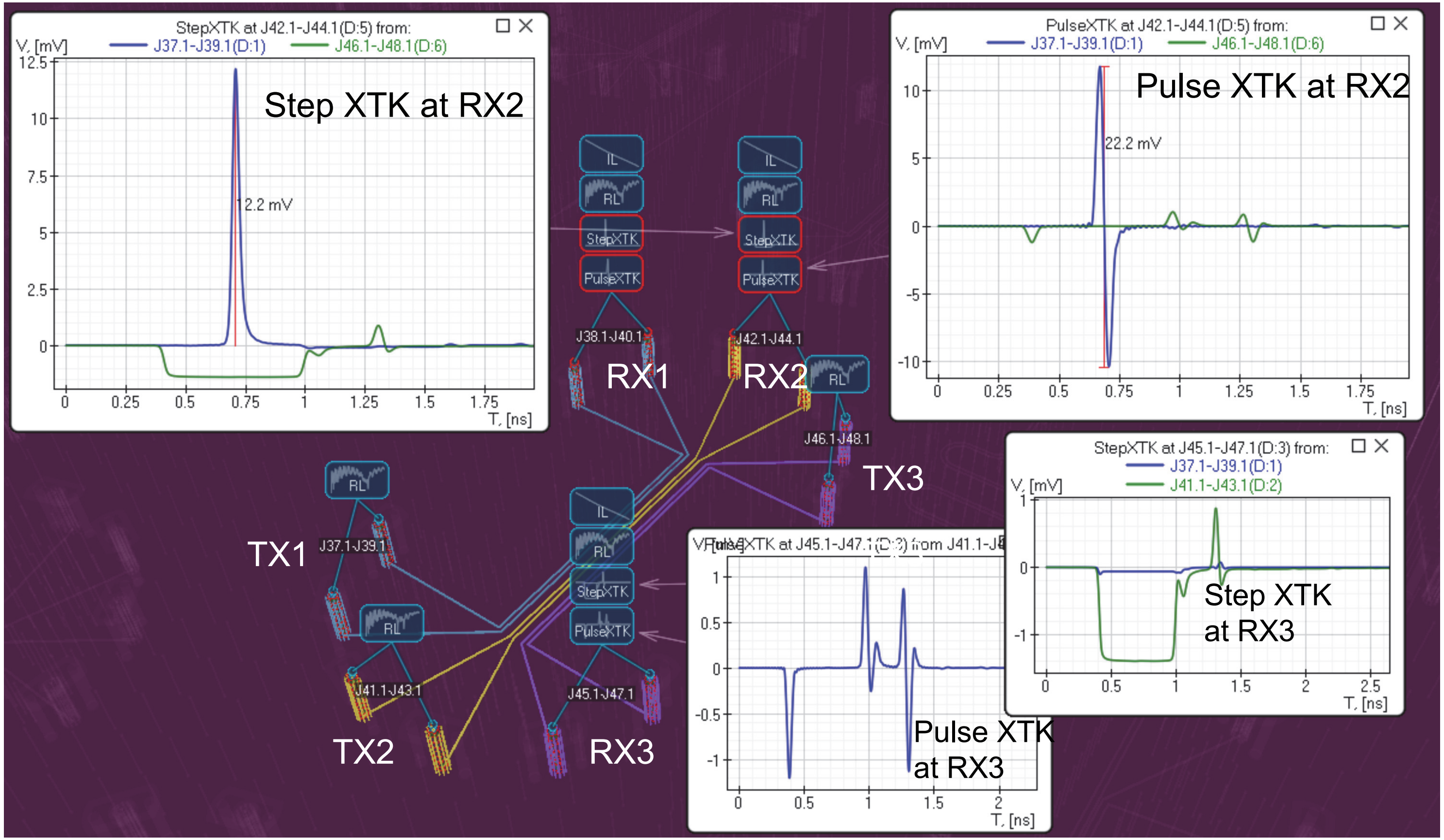
XTALK-28/32: XTalk in TD, S=4w

Comparison of step and pulse xtalk responses

Stimulus 2V, 100 Ohm (effective 1V), 25ps rise time (10-90%), Fast SI Analysis

Pulse response may double the observed xtalk values

Trace width is 13.5mil and differential trace pitch is 37mil, diff. pair separation 4w , 2-inch coupling length

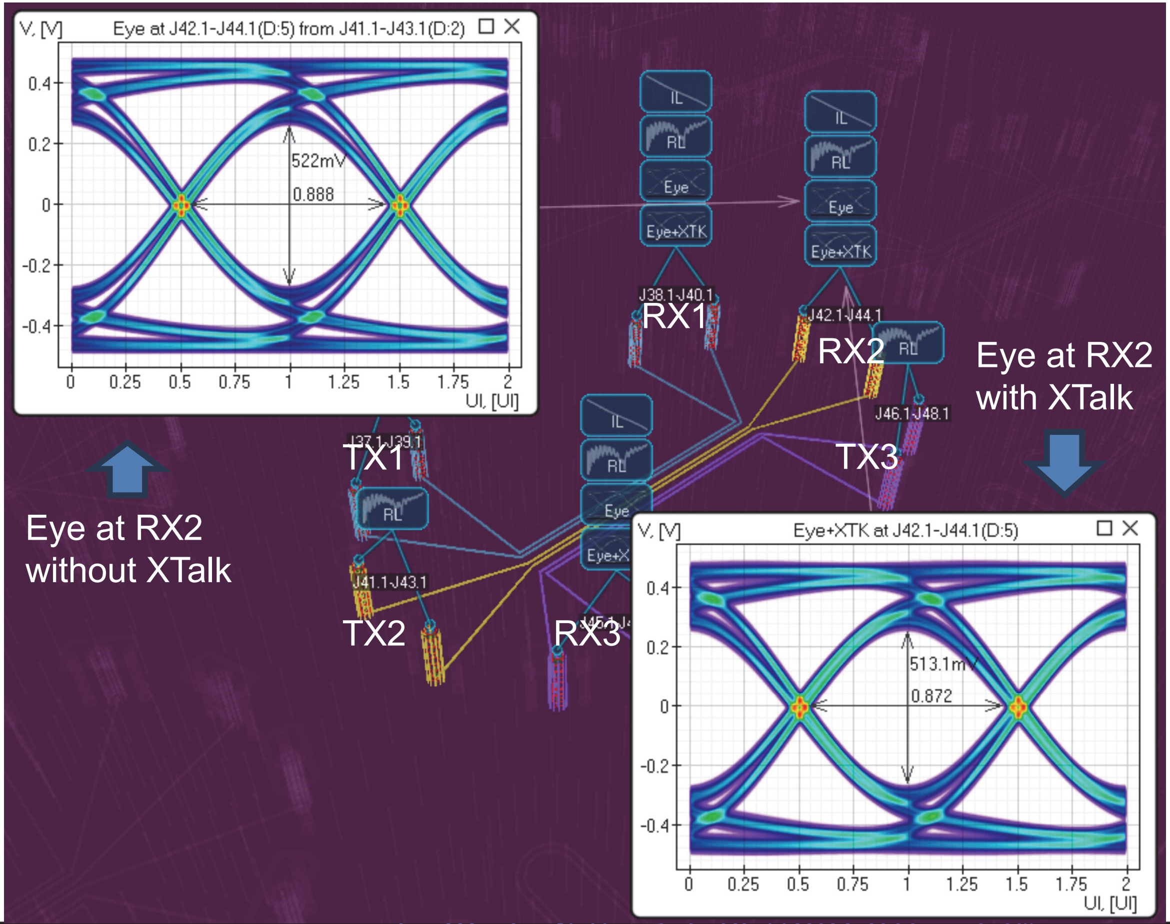


XTALK-28/32: XTalk in TD, S=4w

XTalk on Eye Diagram

Stimulus 28 Gbps PRBS32, 25ps rise time (10-90%), 1V, 100 Ohm, Fast SI Analysis

Trace width is 13.5mil and differential trace pitch is 37mil, diff. pair separation 4w , 2-inch coupling length



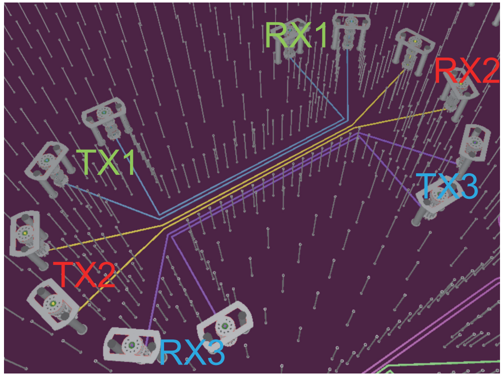
Signal and XTalk Superposition

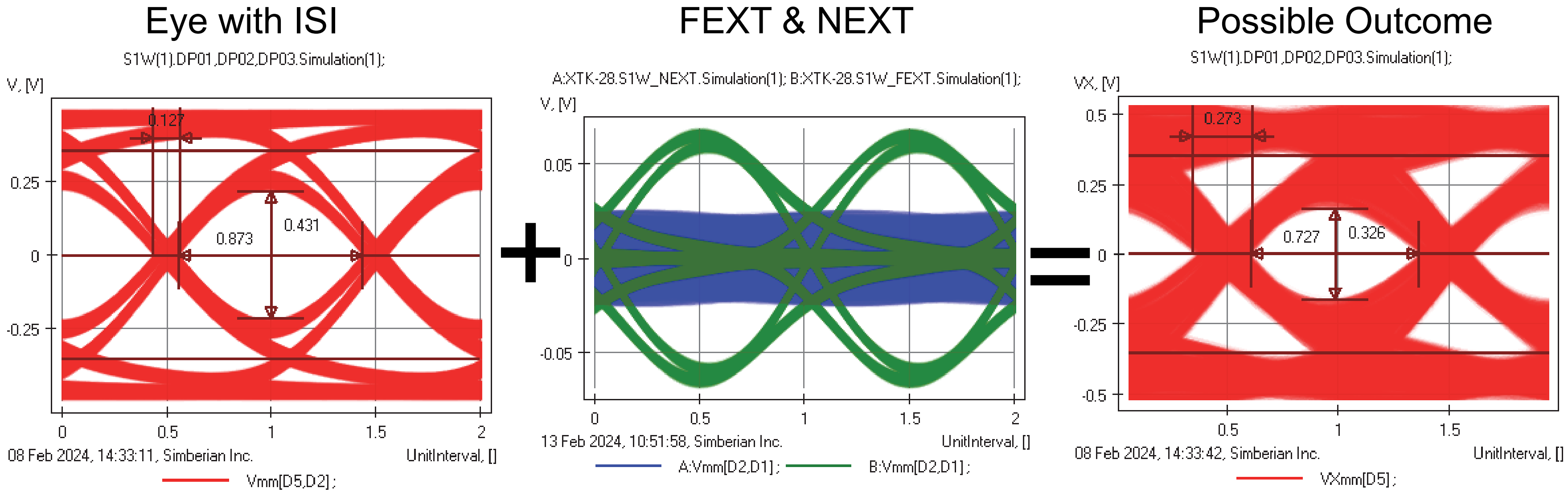
The ultimate metric for a link performance is bit error rate (BER) or eye diagram height at a specified BER

Statistical methods are usually used to evaluate BER or the eye diagram opening

Statistical approach to requires a statistical model for a crosstalk

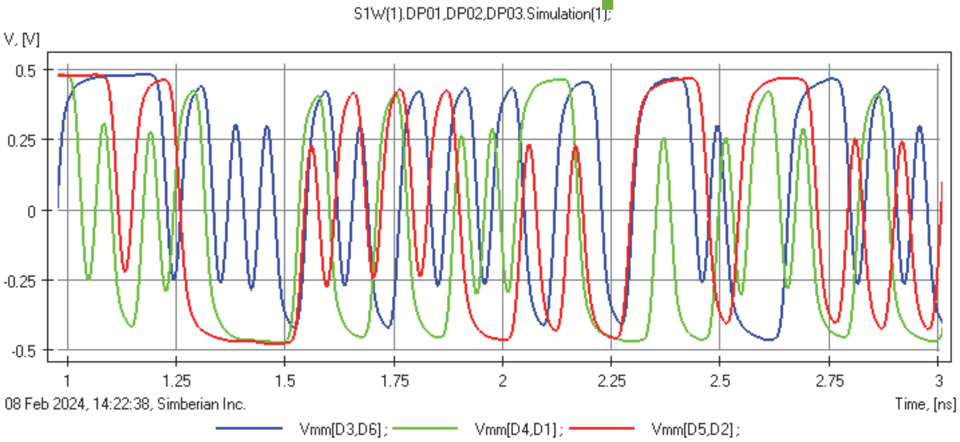
The crosstalk is not random and also bounded by our worst-case estimates

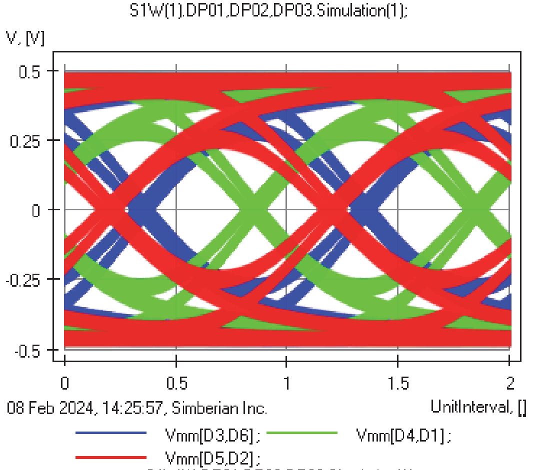


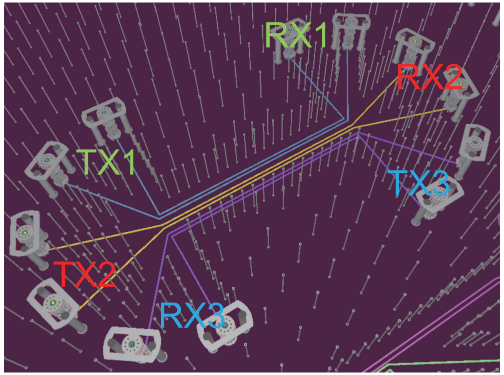


Superposition – Jitter or Amplitude Noise?

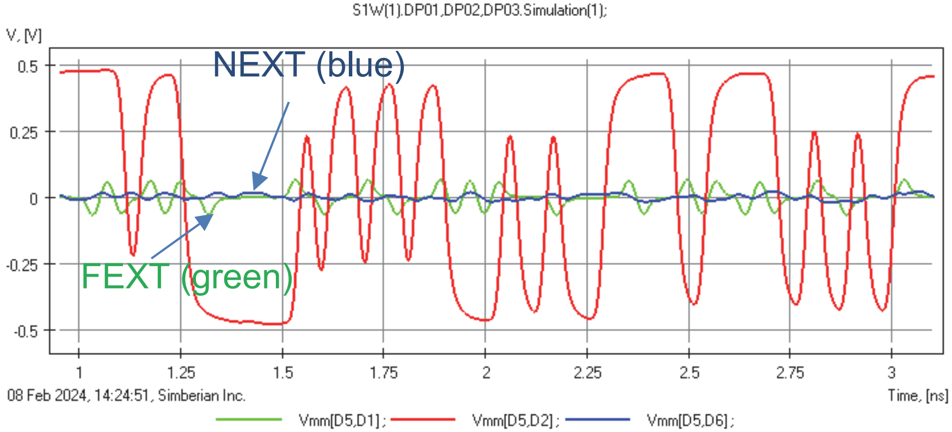
Bits in 3 adjacent coupled links at RX1 (green), RX2 (red) and RX3 (blue)

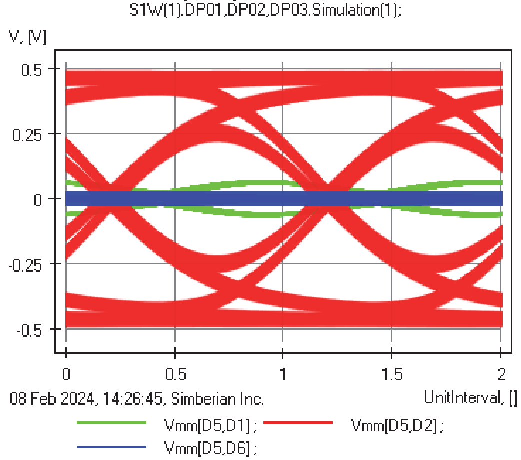






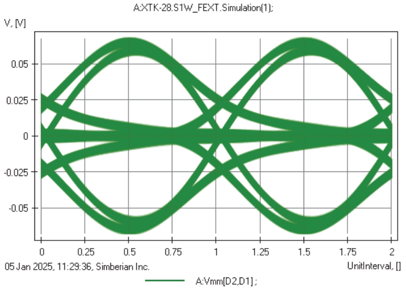
Bits at RX1 (red) and Xtalk from TX1 (green) and TX3 (blue)



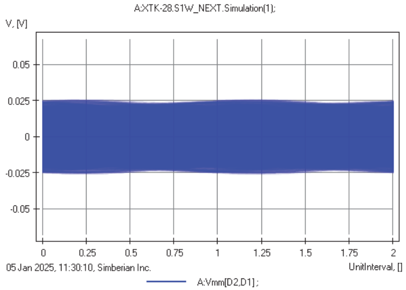


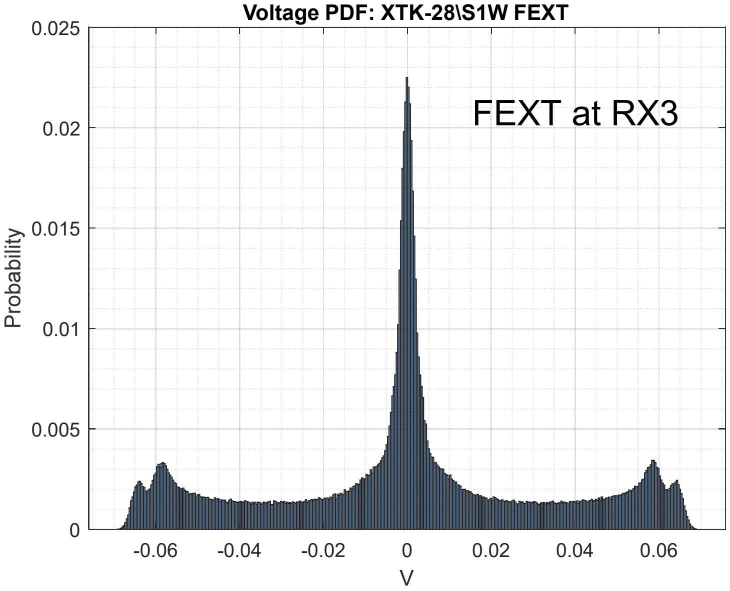
XTALK-28/32: w=13.5mil and, pitch 37mil, diff. pair separation 1w , 2-inch coupling length

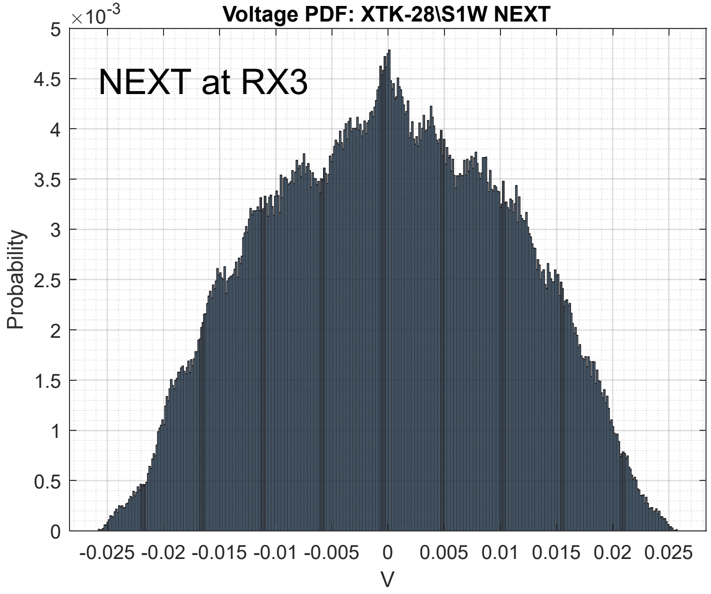
Xtalk Probability Density: Normal or Not?



28 Gbps PRBS32, rise time 10-90% - 20 ps, T=1us, dt=2ps XTALK-28/32: w=13.5mil and, pitch 37mil, diff. pair **separation 1w** , 2-inch coupling length

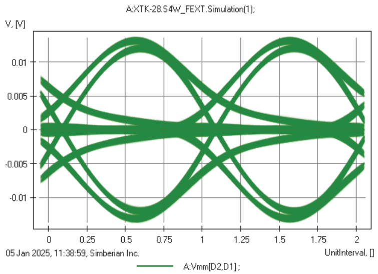




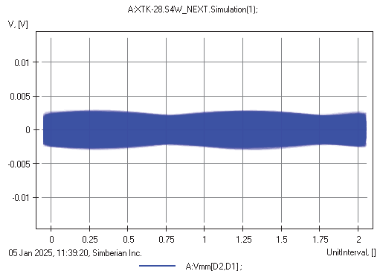


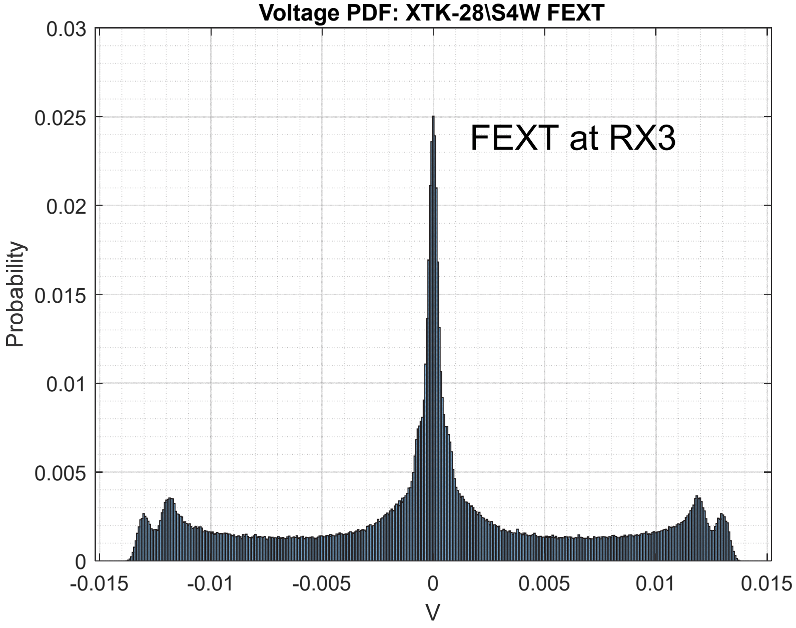
Computed with Simbeor SDK

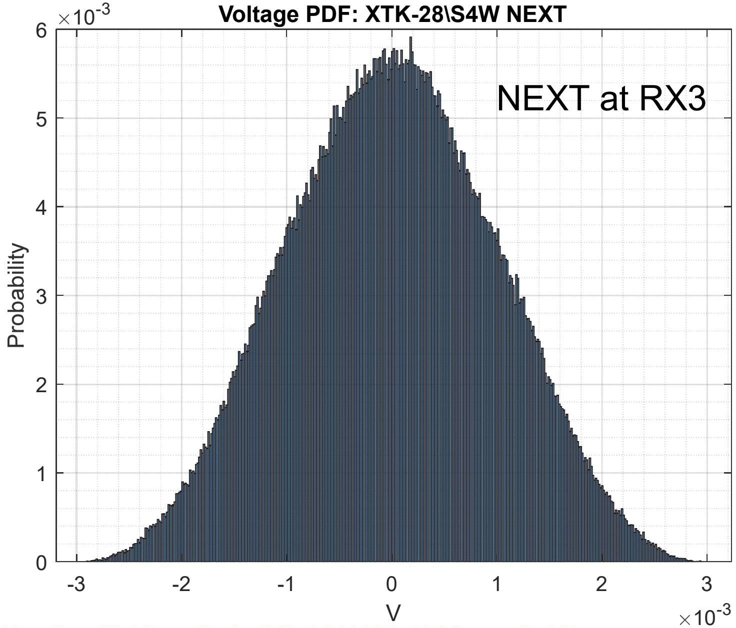
Xtalk Probability Density: Normal or Not?



28 Gbps PRBS32, rise time 10-90% - 20 ps, T=1us, dt=2ps XTALK-28/32: w=13.5mil and, pitch 37mil, diff. pair **separation 4w** , 2-inch coupling length

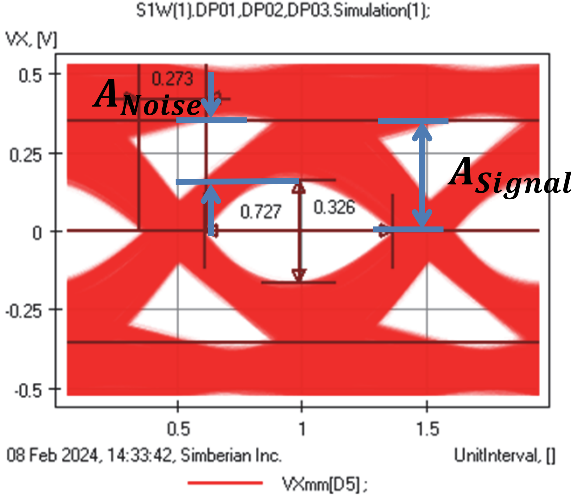






Computed with Simbeor SDK

Channel Operating Margin (COM)

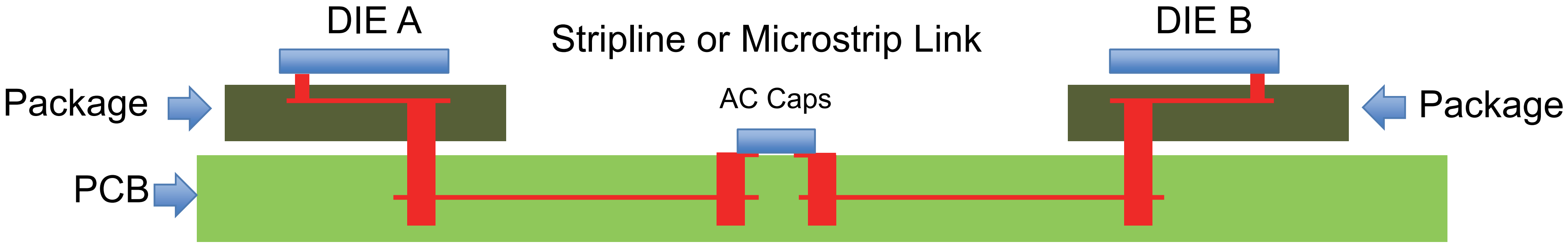


$$COM = 20 \cdot \log\left(\frac{A_{Signal}}{A_{Noise}}\right)$$

>3 dB for pass threshold

A_{Signal} *Peak Signal – includes all losses accounted in model, reference package and equalization*

A_{Noise} *Peak SER Noise – includes losses, ISI, **crosstalk** and random noise with some assumptions and equalization*



COM Computation

1) S-parameters computed or measured over sufficient bandwidth:
  a) S-parameters (s4p) of a link from TP0 to TP5
  b) **S-parameters of aggressor links (s4p)**
  c) Package models – from vendors (defined in Excel spreadsheet)

2) COM script – Matlab script from IEEE P802.3ck group and Excel spreadsheet with parameters;

R. M. Mellitz, A. Ran, M. P. Li, V. Ragavassamy, Channel Operating Margin (COM): Evolution of Channel Specifications for 25 Gbps and Beyond, DesinCon 2013

H. Wu, M. Shimanouchi, M. P. Li, COM & IBIS-AMI How They Relate & Where They Diverge, DesignCon 2019

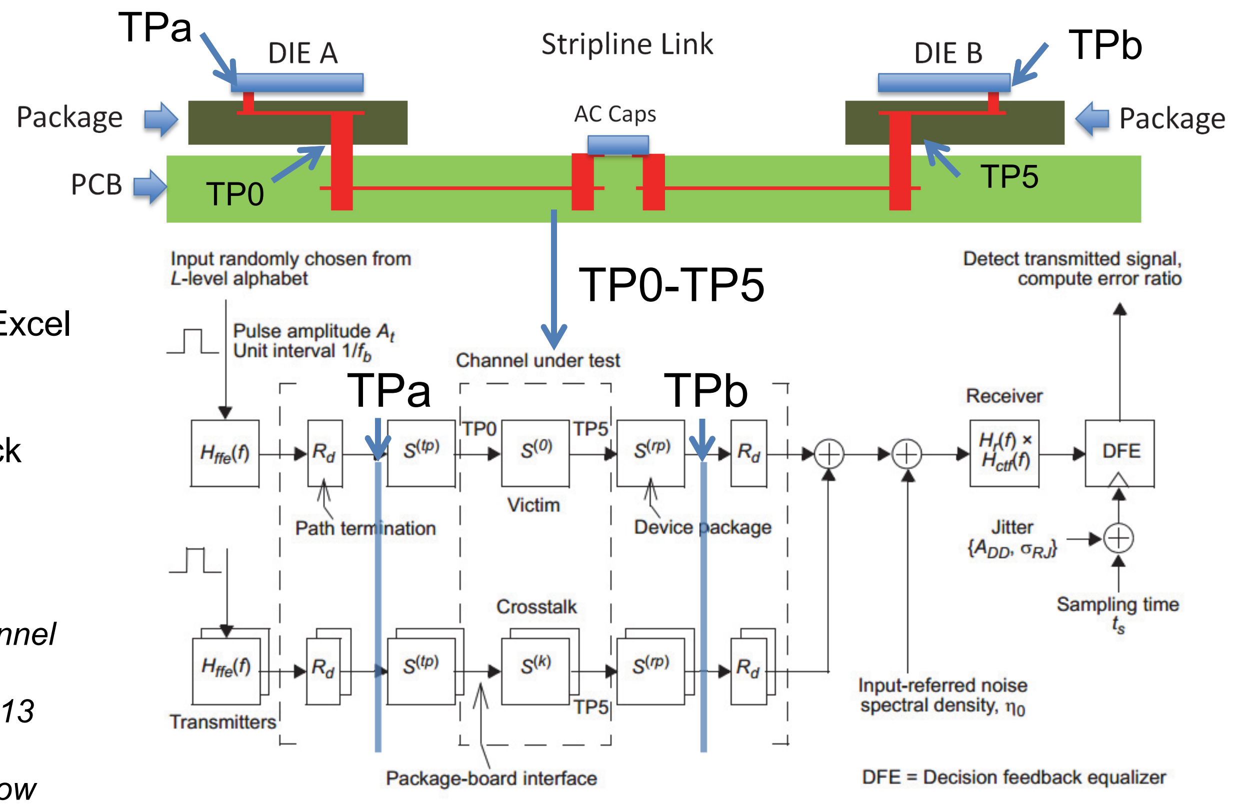


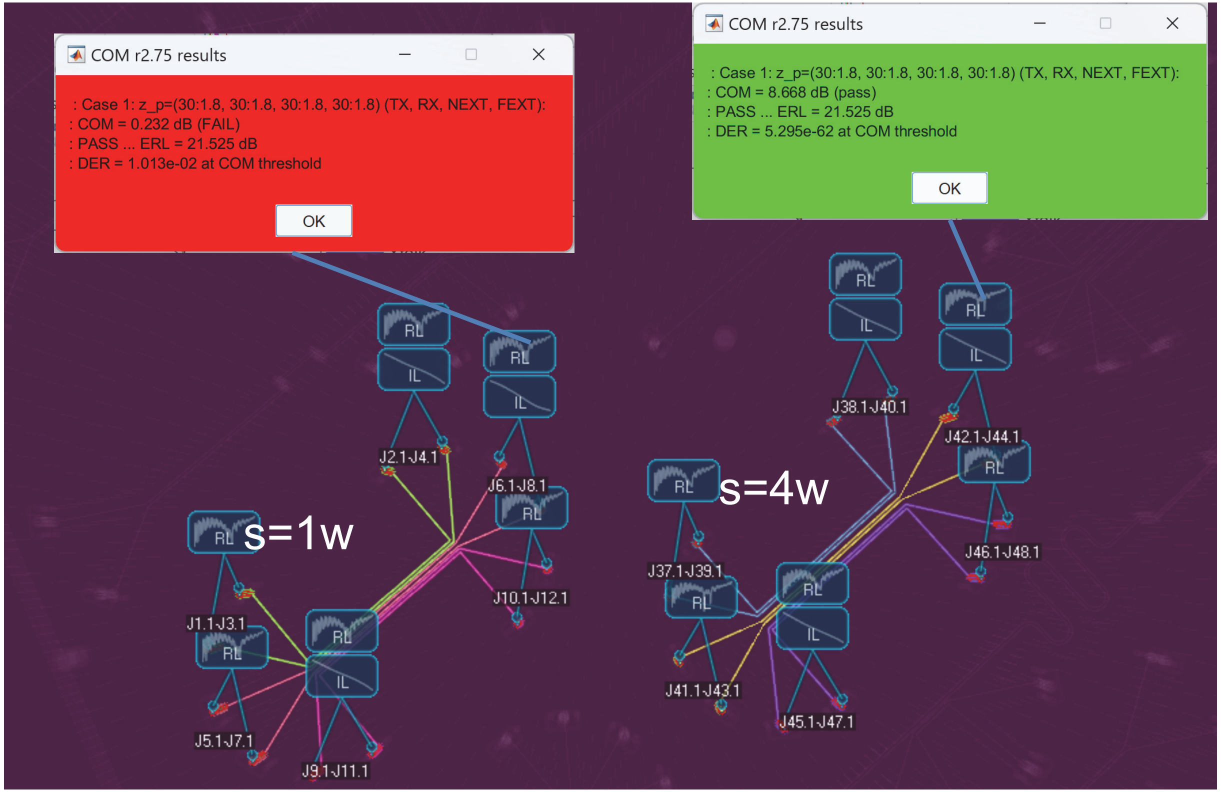
Figure 93A–1—COM reference model

XTALK-28/32: COM

For each crosstalk source COM computes probability density functions (PDFs) and convolve them with the bit PDF

The IEEE COM tool is used with default reference transmitter and receiver parameters

DER limit is 1e-4 and the pass value of COM is 3 dB in this case



Takeouts

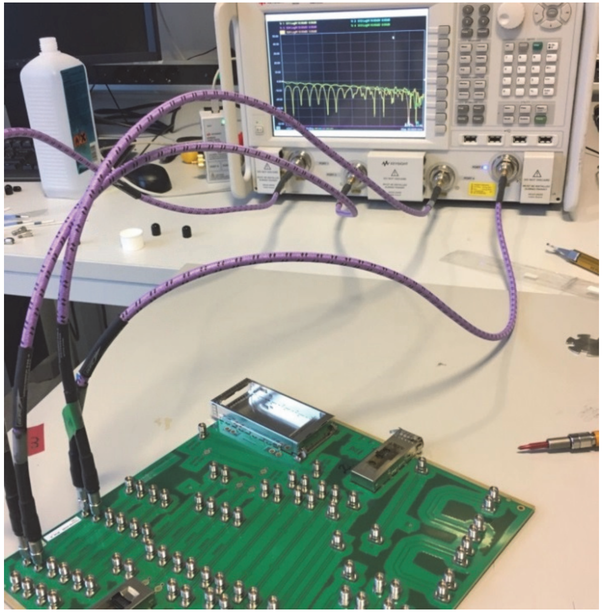
- There are multiple methods to evaluate and model crosstalk – each has its own metrics that can be used either for preliminary evaluation or for troubleshooting (post-mortem examination)
- Equation-based estimates have limited accuracy and applicability – electromagnetic analysis is required for accurate crosstalk modeling
- **Frequency domain, S-parameters and multiport theory** are the foundation for time and probabilistic domain models
- Time domain approach has timing uncertainties, depends on rise time,…
- Probabilistic approach is an alternative to the total localization…

OUTLINE

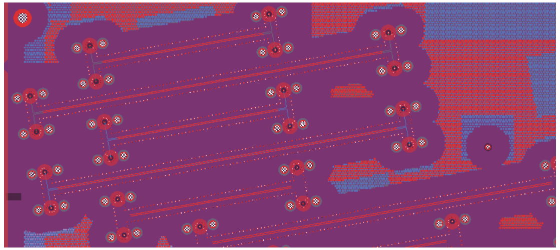
- Introduction
- Basics: Fields and S-parameters
- Crosstalk Anatomy - Qualitative Analysis
- Crosstalk Quantification
- Distant Crosstalk - Sources and Mitigation
- Conclusion

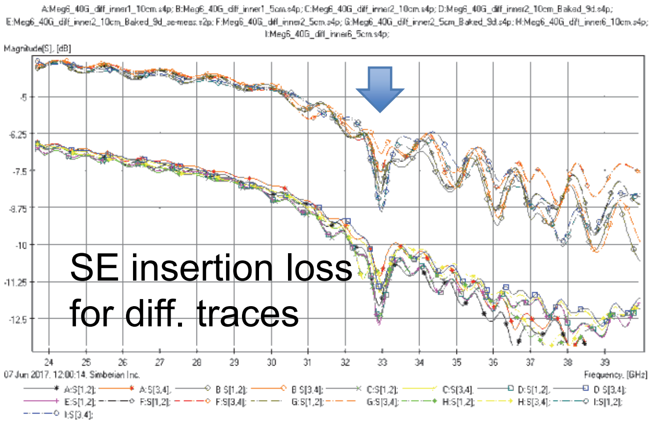
Localization and Multipath Propagation

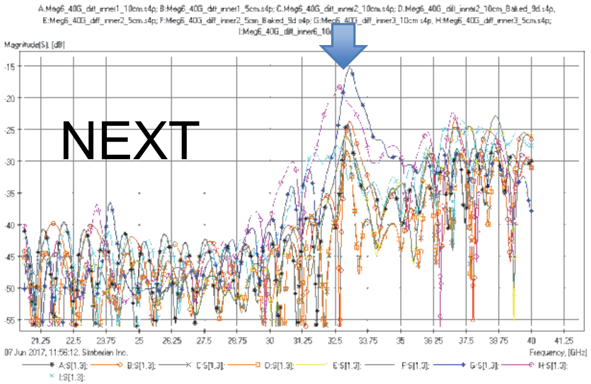
EvR-1 platform: *M. Marin, Y. Shlepnev, 40 GHz PCB Interconnect Validation: Expectations vs. Reality, DesignCon2018 – App. Note #2018_01, Webinar #8*

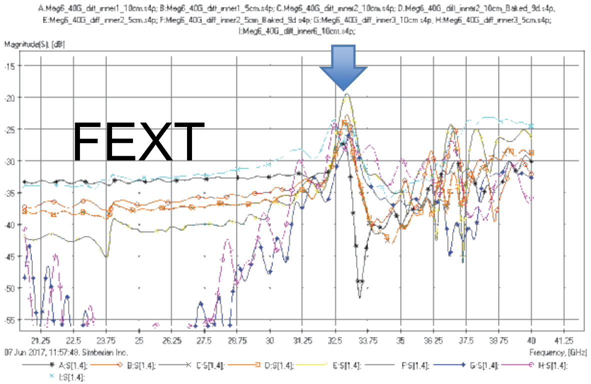


5 and 10 cm diff. traces in INNER1, INNER2, INNER6 and BOTTOM





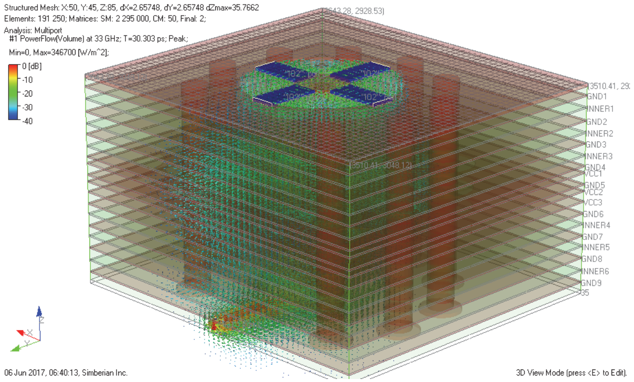


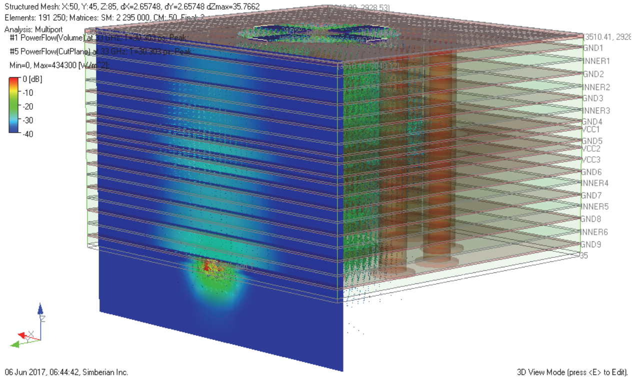


EvR-1 – Launch Localization Breakout

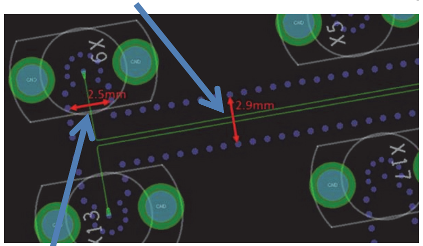
Microstrip launch peak power flow density at 33 GHz

Compression-mount connector launch from EvR-1 - demo video: [#2018_01: How Interconnects Work™: Localization of coaxial connector launch, 10 min](https://youtu.be/WMfLvJYWNDY) – YouTube
https://youtu.be/WMfLvJYWNDY



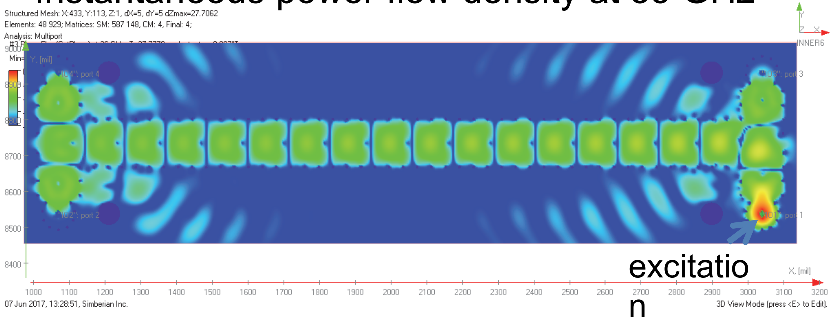


29 GHz cutoff frequency



32 GHz cutoff frequency

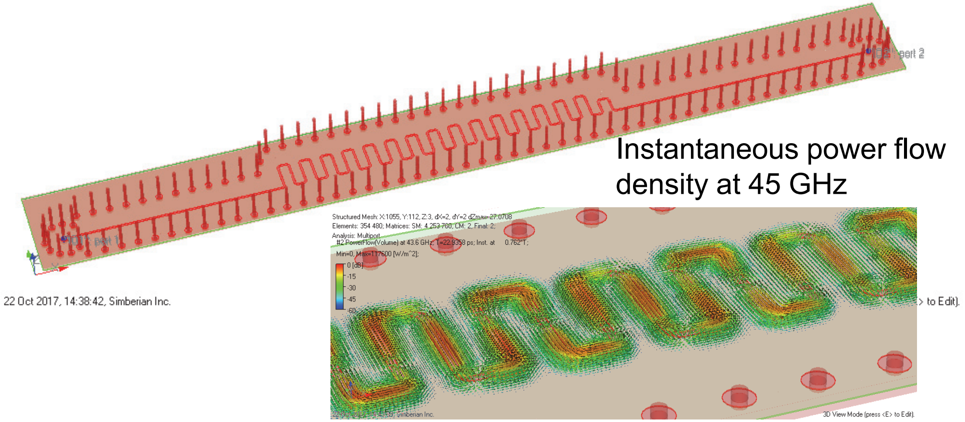
Instantaneous power flow density at 35 GHz



excitation

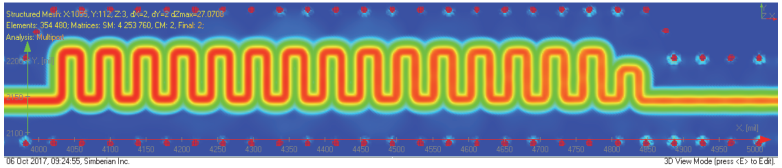
Energy leaked from the launches goes into Substrate Integrated Waveguide (SIW) and appear at the other ports

Multipath Propagation XTalk

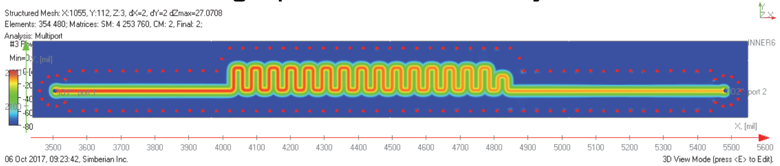


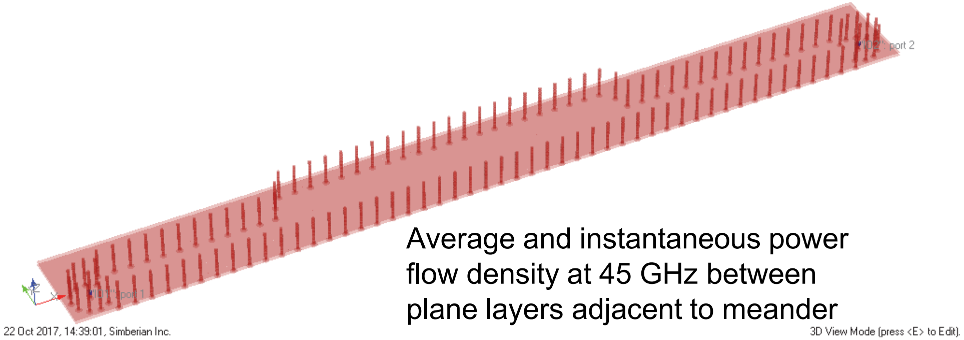
Instantaneous power flow density at 45 GHz

Max power flow density at 45 GHz

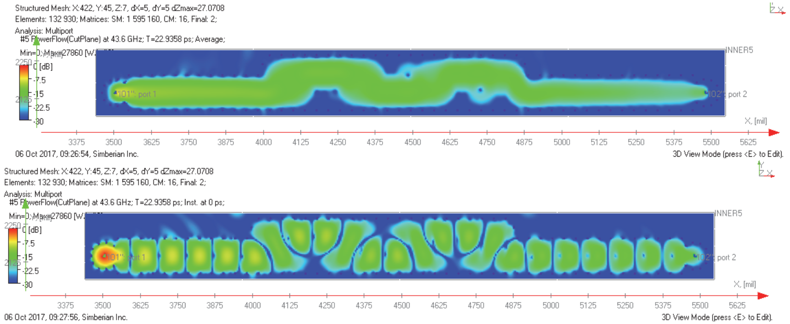


Average power flow density at 45





Average and instantaneous power flow density at 45 GHz between plane layers adjacent to meander



See more at M. Marin, Y. Shlepnev, 40 GHz PCB Interconnect Validation: Expectations vs. Reality, DesignCon2018

Whole-Board Analysis or Localization?

- Vias are coupled to PDNs and surface layers of PCB
- Vias are the source of distant crosstalk and multipath propagation
- Distant crosstalk may require the whole board analysis
- Is it possible to evaluate possible crosstalk with analysis in isolation?
- What include into such simulation?
- What boundary conditions to use?

Let's investigate distant coupling and try to derive a metric for the localization and possible coupling…

XTalk and Dissipation in Balance of Power

$$P_{out} = P_{in} - P_{reflected} - P_{dissipated} - P_{leaked} + P_{coupled}$$

$P_{dissipated}$ includes energy absorbed by materials (P_{absMat}) and by boundary conditions (P_{absBC})

$$P_{dissipated} = P_{absMat} + P_{absBC}$$

$$P_{in} = |a_1|^2 \, [Wt], a_2 = 0$$

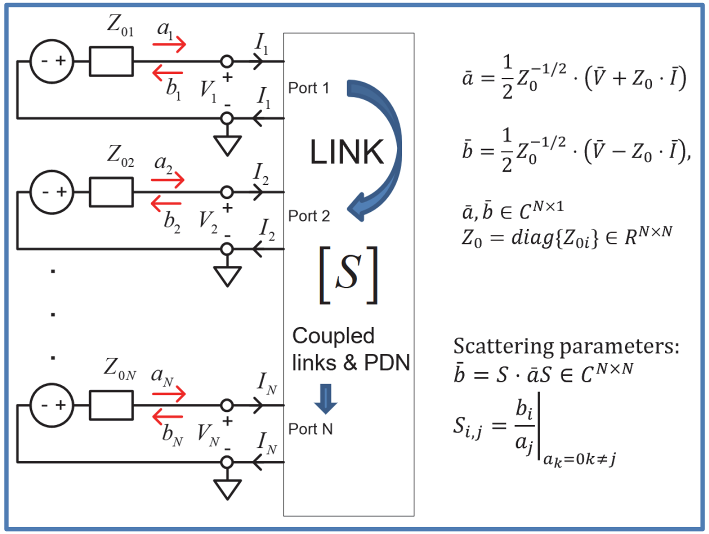
$$P_{out} = |S_{2,1}|^2 \, P_{in}$$

$$P_{reflected} = |S_{1,1}|^2 \, P_{in}$$

$$P_{dissipated} = \left(1 - \sum_k |S_{k,1}|^2\right) P_{in}$$

$$P_{leaked} = \left(\sum_{k \neq 1,2} |S_{k,1}|^2\right) P_{in}$$

$$P_{coupled} = \sum_{k \neq 1,2} |S_{2,k}|^2 \, P_{ink}$$



1. Can $P_{coupled}$ be evaluated in isolation with PML boundary conditions (BC)?
2. Can $P_{dissipated}$ evaluated for via isolated with PML BC be used as a metric of via localization and possible coupling or xtalk?

*(*)Y. Shlepnev, How Interconnects Work: **Reflections from Discontinuities** , Simberian App Note #2022_01, January 10, 2022.*

Single-Ended Vias Coupling and Localization

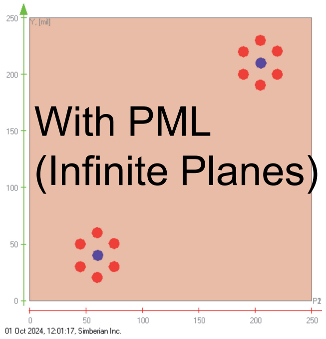
Two 0.77mil copper planes, separated by dielectric with Dk=3, LT=0.001

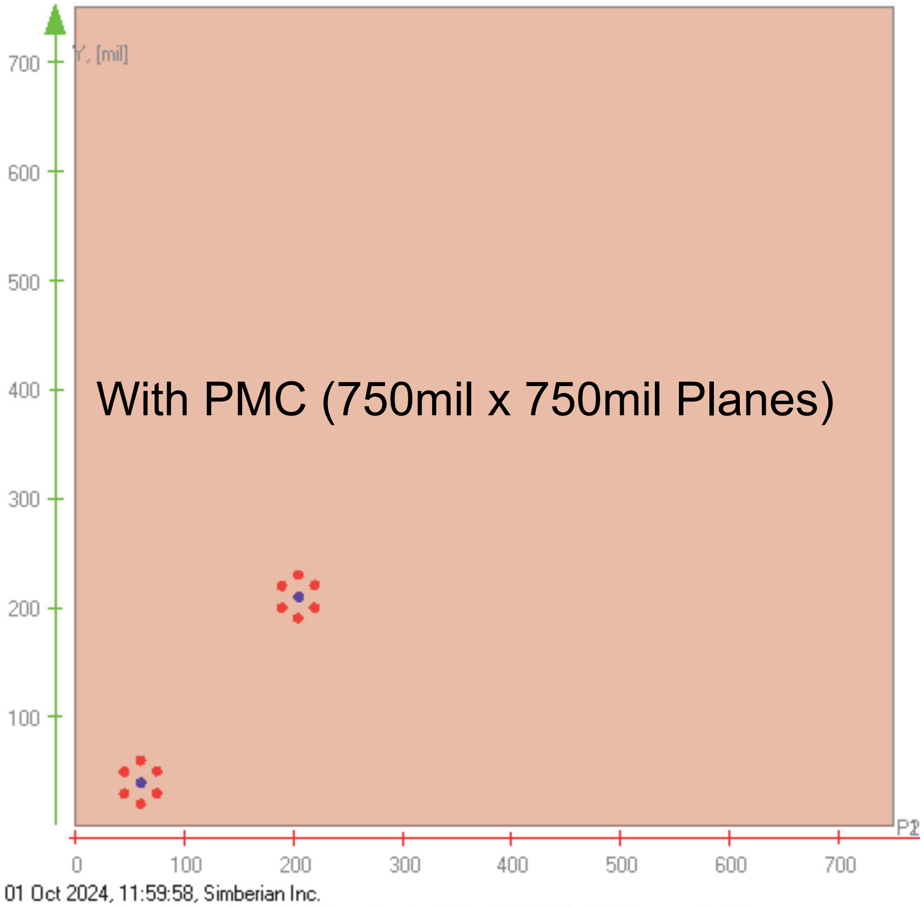
Two signal vias at 220mil (10mil diameter)
Number of stitching vias (Nstv) from 0 to 6 at about 20mil distance from signal

4-port structure with 50Ohm terminations

Physics-based model with 2D analysis in Simbeor 3DTF solver

Solution:
Coupling_PPW_SE





SE Vias Coupling - H=9mil

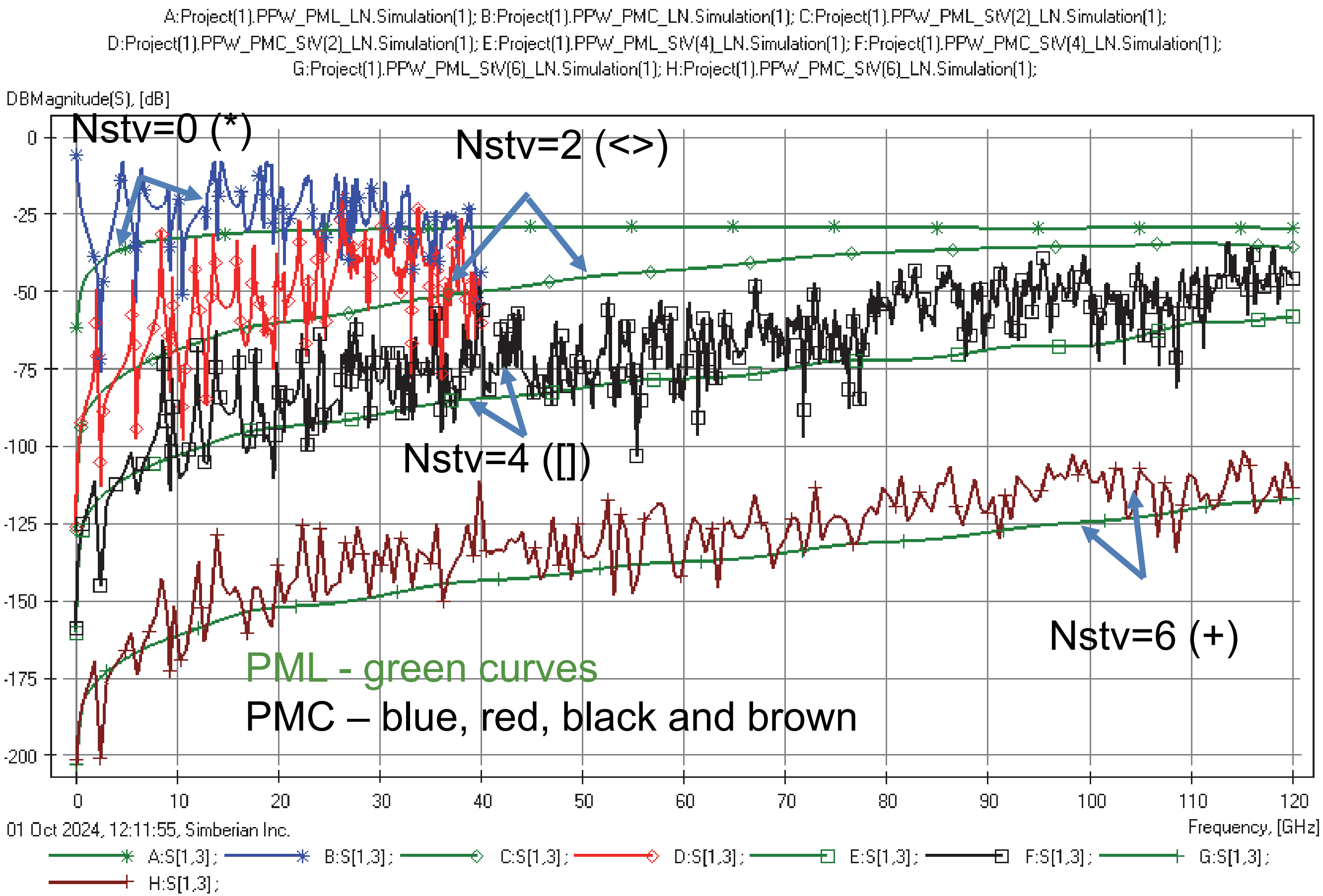
Coupling is reduced by increase of Nstv

Resonances in PPW affect coupling

Coupling saturates as frequency increases (remarkable)

Two stitching vias is not enough

Analysis complexity for real-life problems is enormous

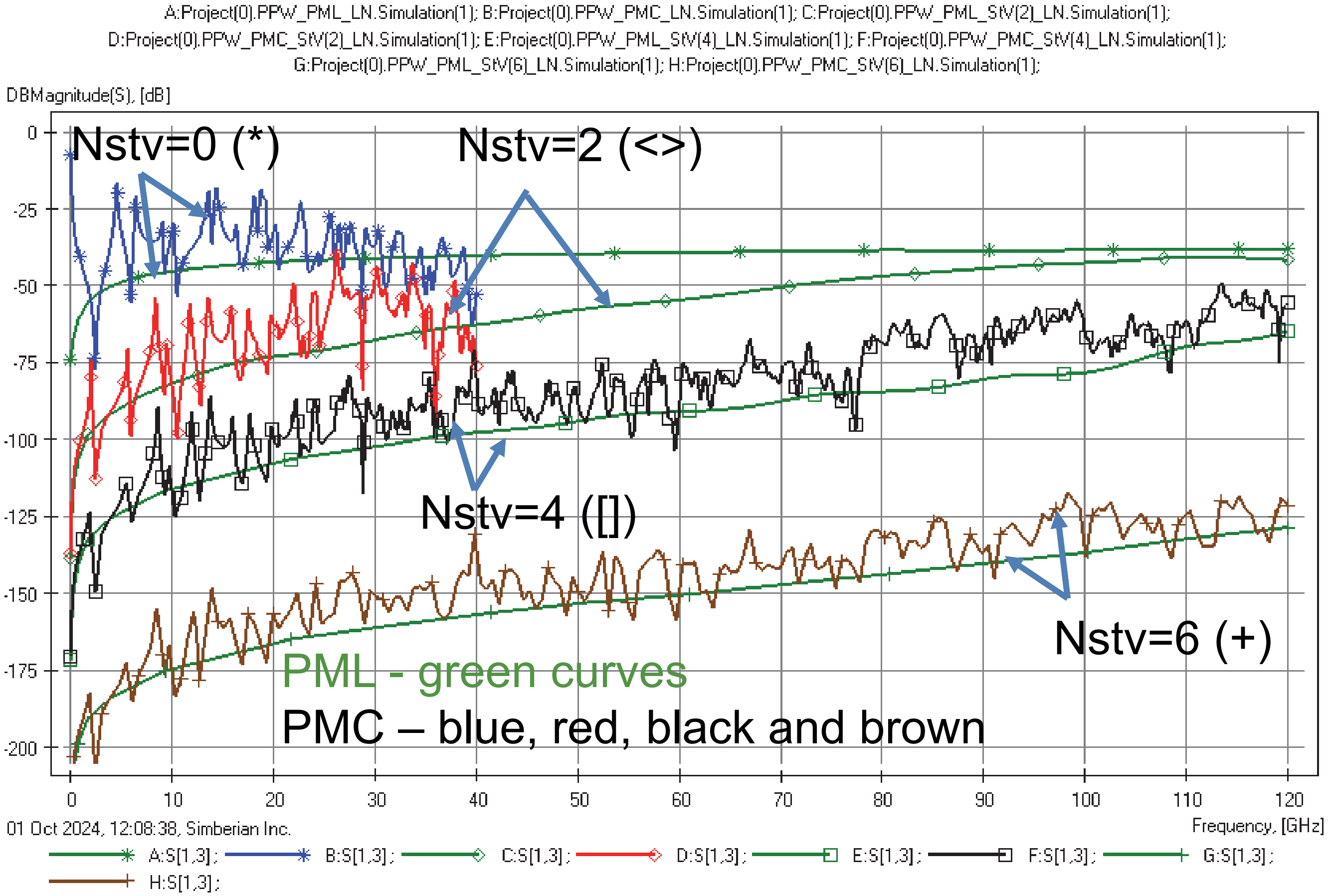


SE Vias Coupling - H=2mil

Coupling decreases with reduction of plane separation

Coupling is reduced by increase of Nstv

2 or 4 stitching vias may be not enough

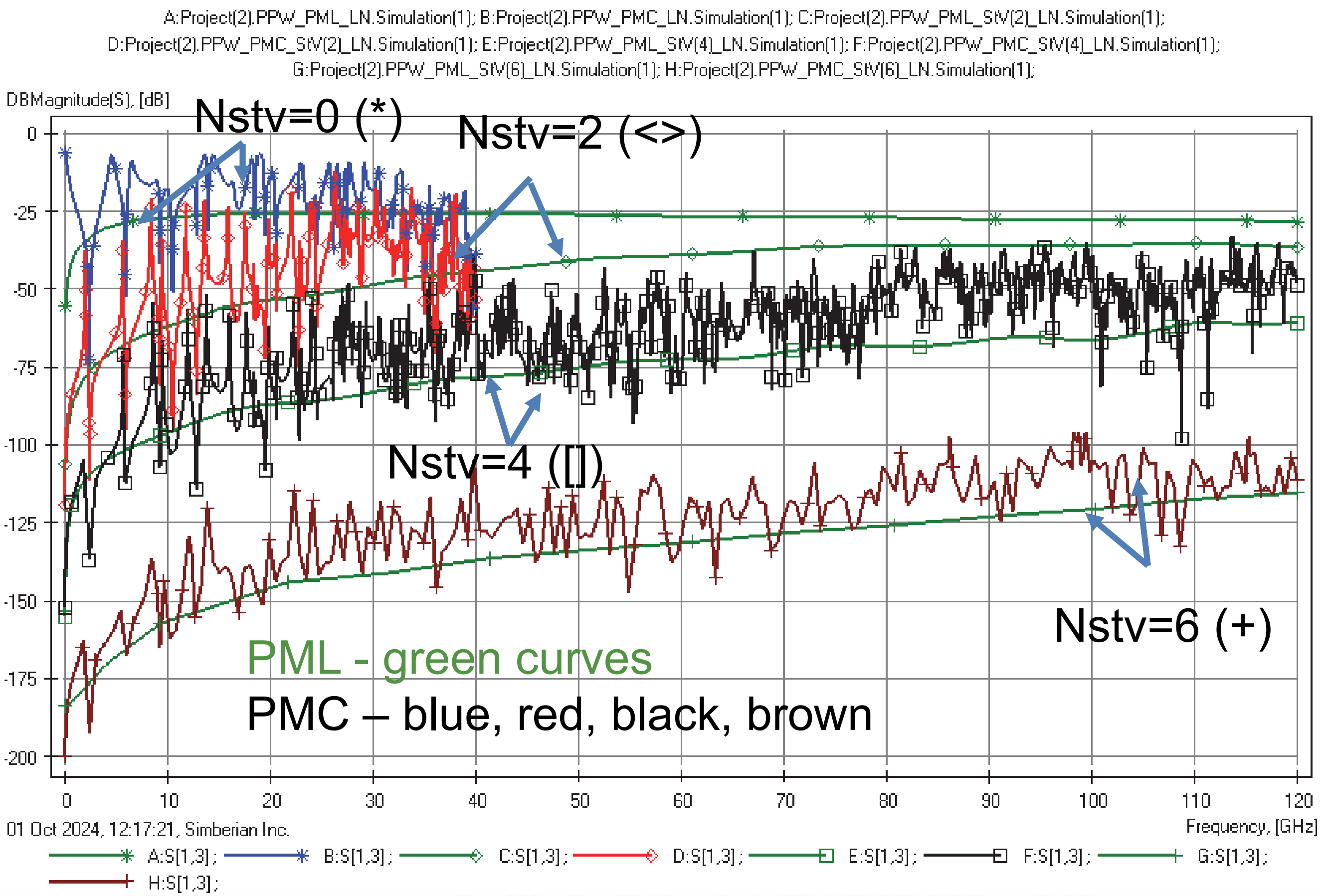


SE Vias Coupling - H=20mil

Coupling increases with increase of plane separation

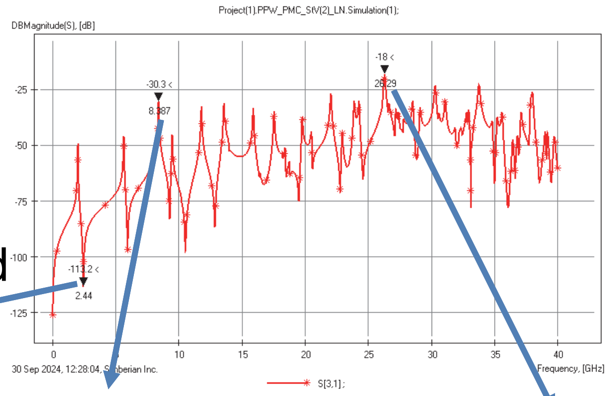
Coupling is reduced by increase of Nstv

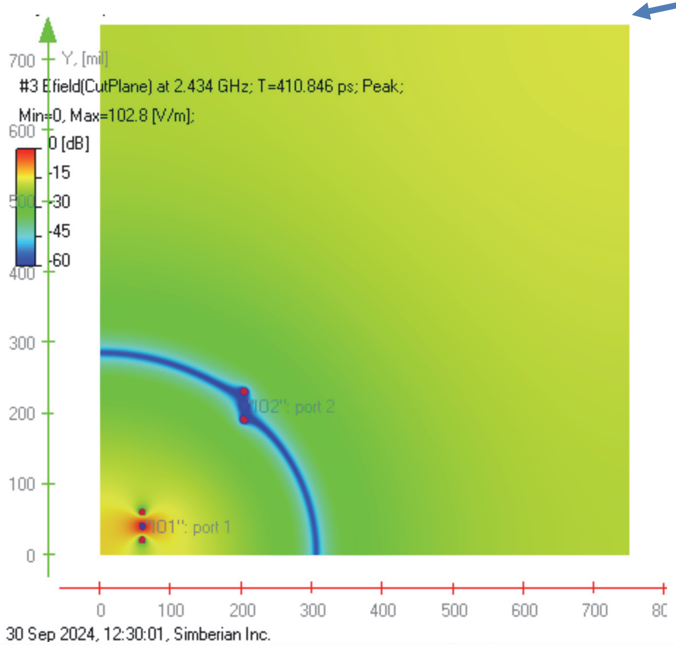
2 or 4 stitching vias is not enough

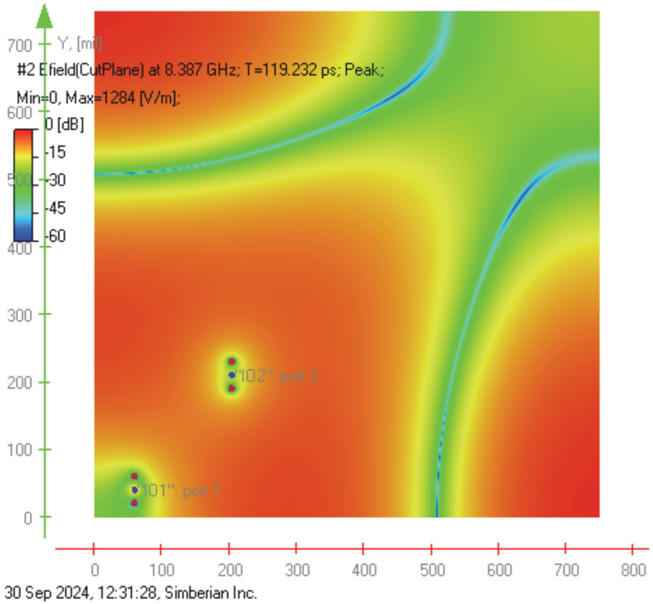


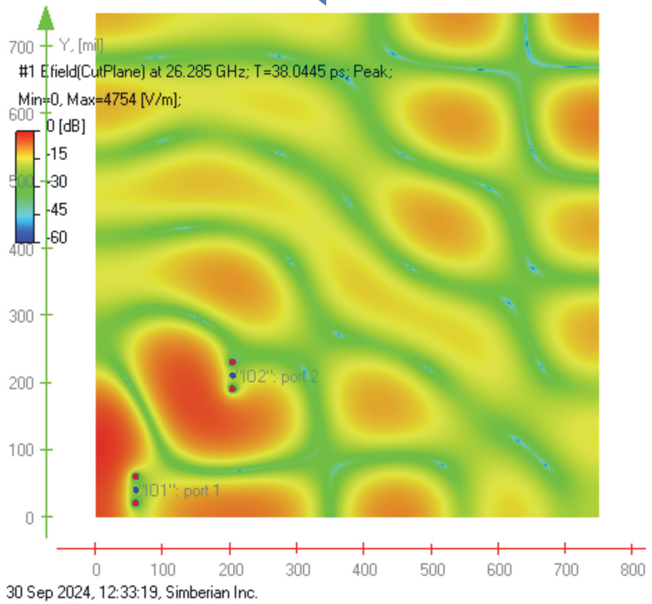
PPW Resonances: H=9mil

750mil x 750mil with PMC, Nstv=2
Electric field with 0.5V excitation at corner via
Minima and maxima depends on PDN geometry and all terminations – difficult to account for





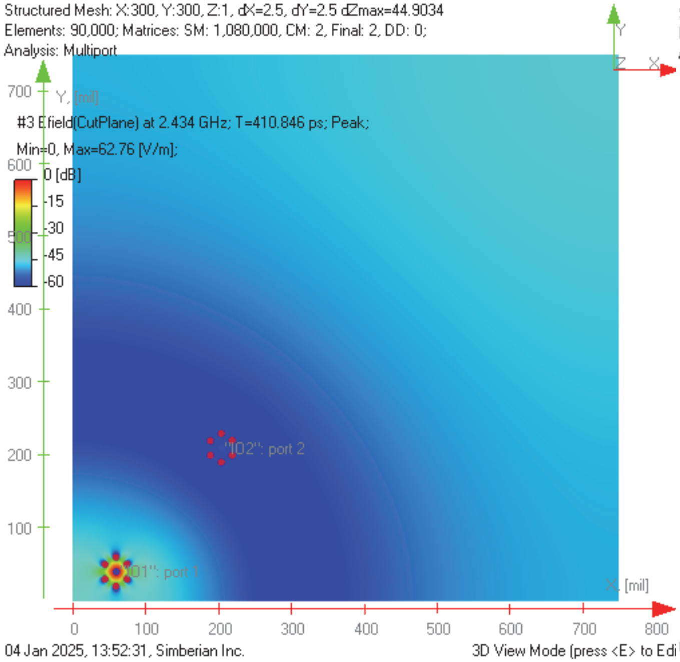


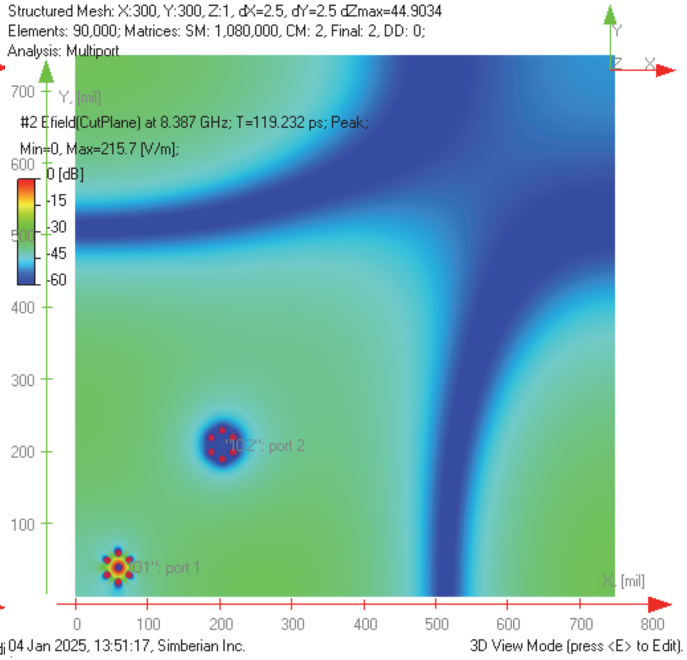


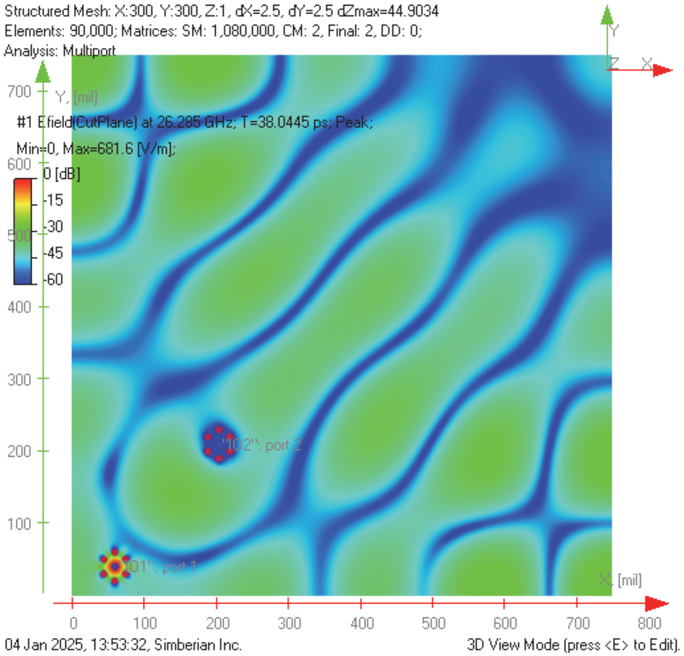
Coupling Reduction: H=9mil

750mil x 750mil with PMC, Nstv=6
Electric field with 0.5V excitation at corner via
No need to simulate the whole board

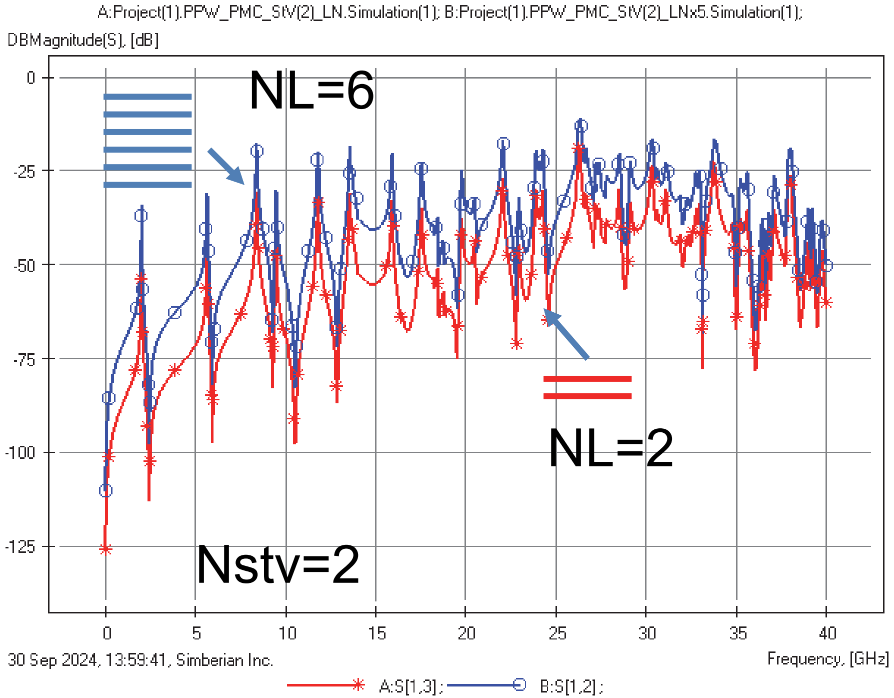
Coupling is not sensitive to distance – defined mostly by maxima and minima or PP resonances

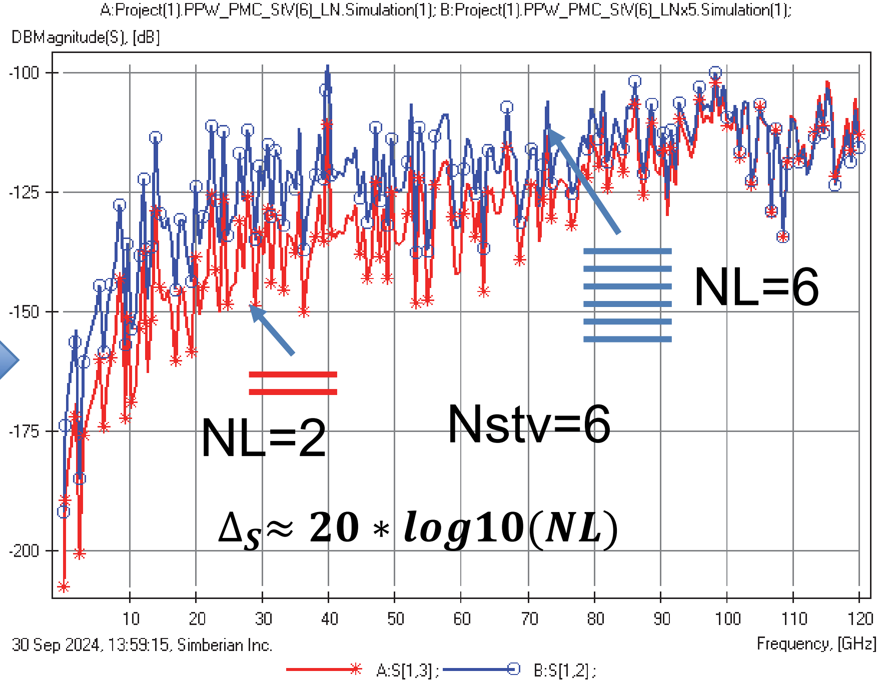






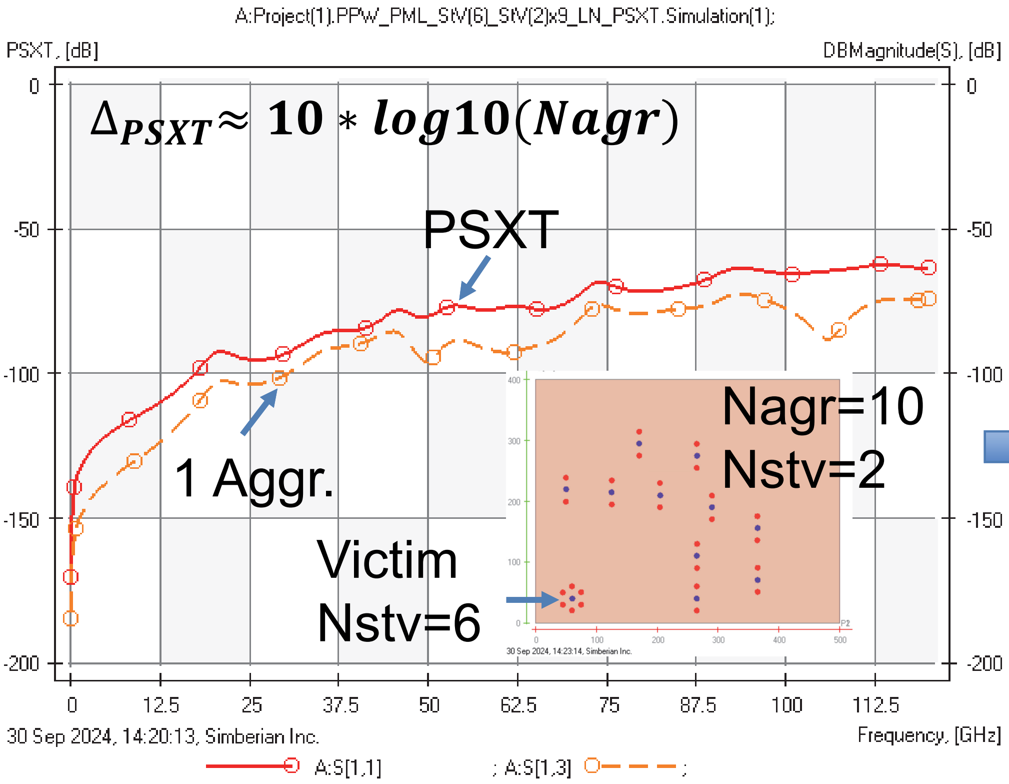
Increase of Plane Count from 2 to 6 (H=9mil)

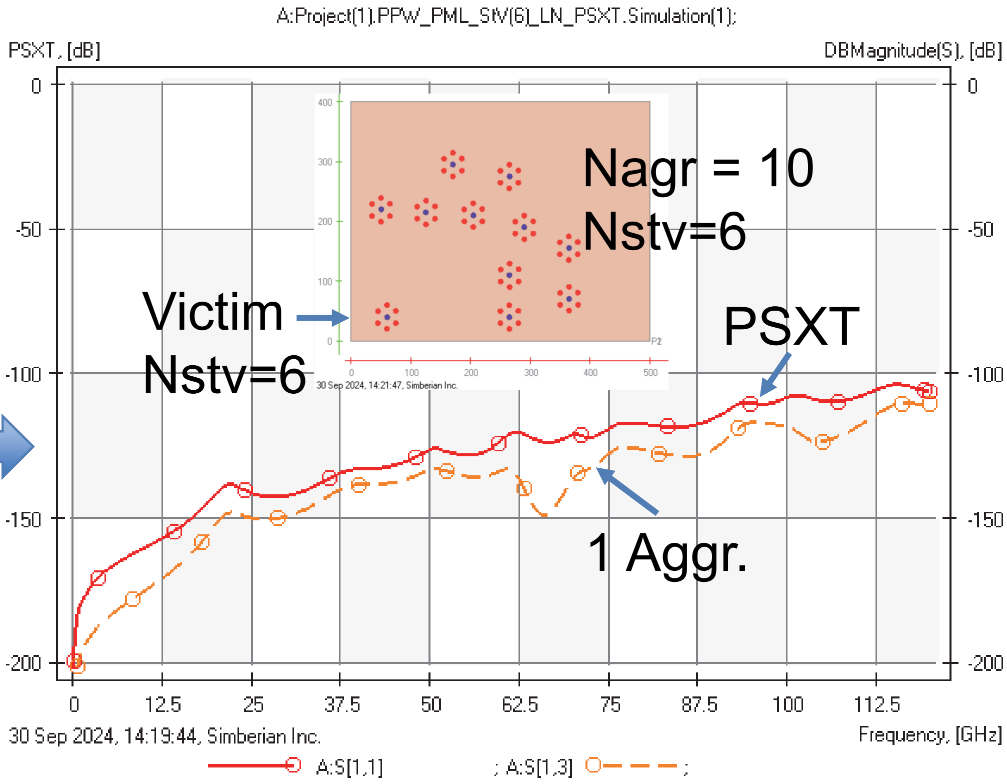




Multiple planes increase xtak, but stitching vias reduce it…

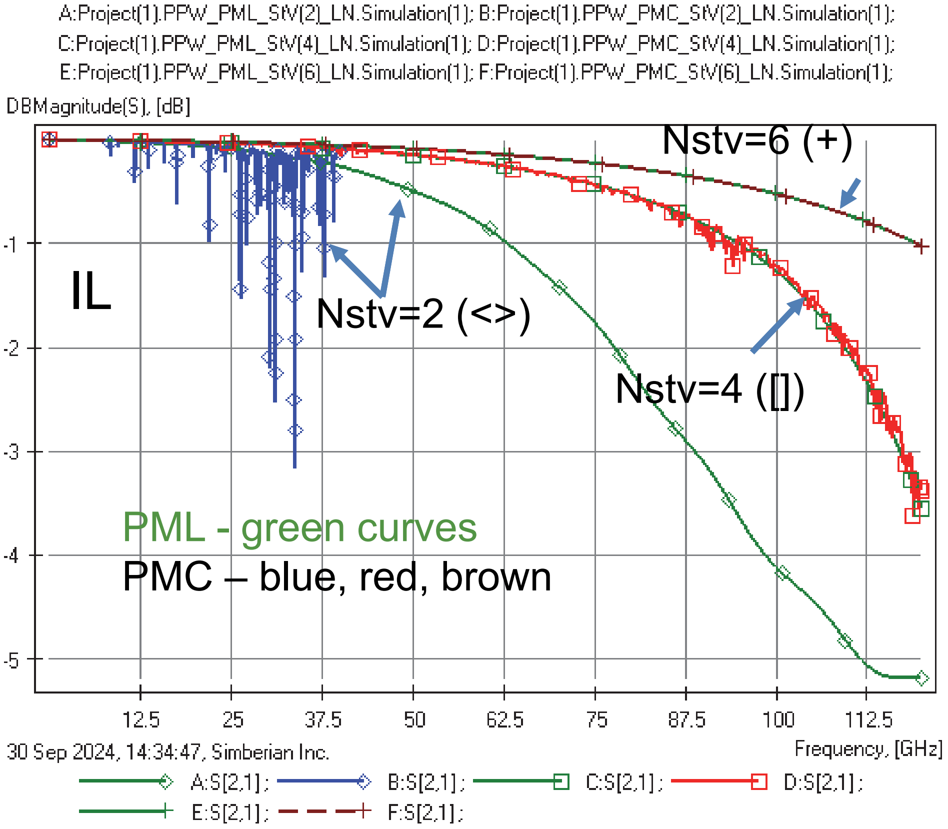
Increase of Aggressors Count (H=9mil)

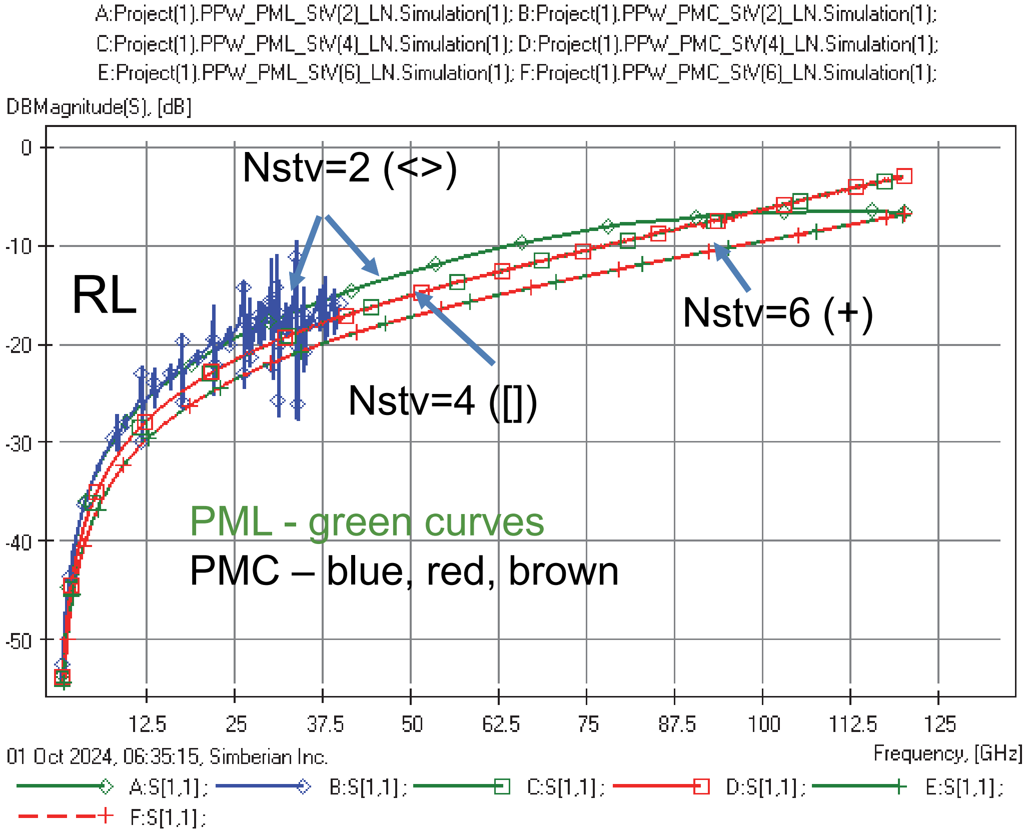




Multiple disturbers increase xtalk, but stitching vias reduce it…

Stitching Vias and IL & RL (H=9mil)





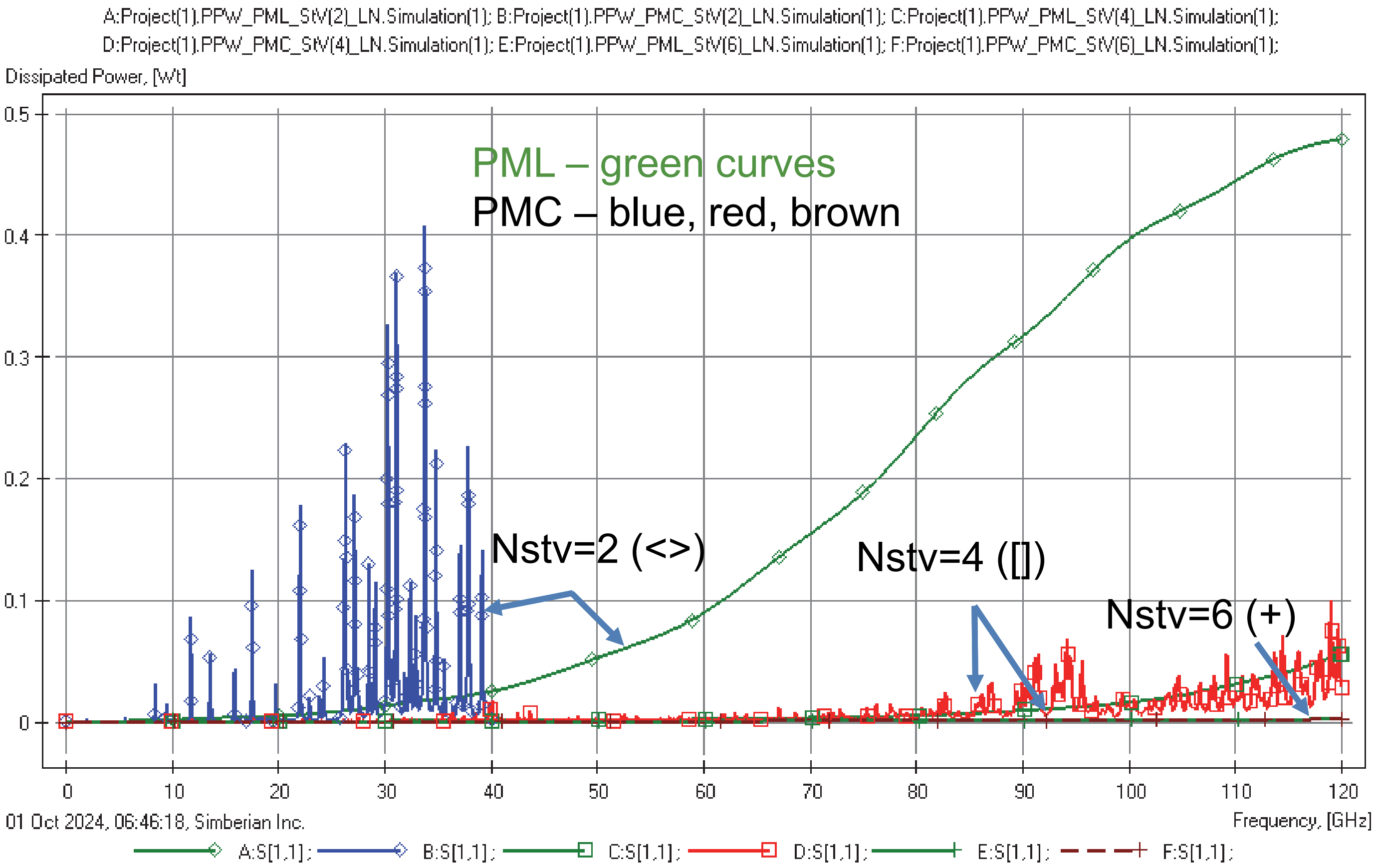
Coupling to cavities causes resonances in IL and RL, but stitching vias reduce it…

SE Via Dissipated Power and Stitching (H=9mil)

$$P_{dissipated} = \left(1 - \sum_k \left|S_{k,1}\right|^2\right) P_{in}$$

Stitching vias reduce power dissipation (leaks)

Power dissipation for sufficiently localized structures can be evaluated with PML boundaries (infinite planes)



SE Via Dissipated Power – 3D EM Model

$$P_{dissipated} = \left(1 - \sum_k |S_{k,1}|^2\right) P_{in}$$

3D EM model for estimation of dissipated power and comparative analysis

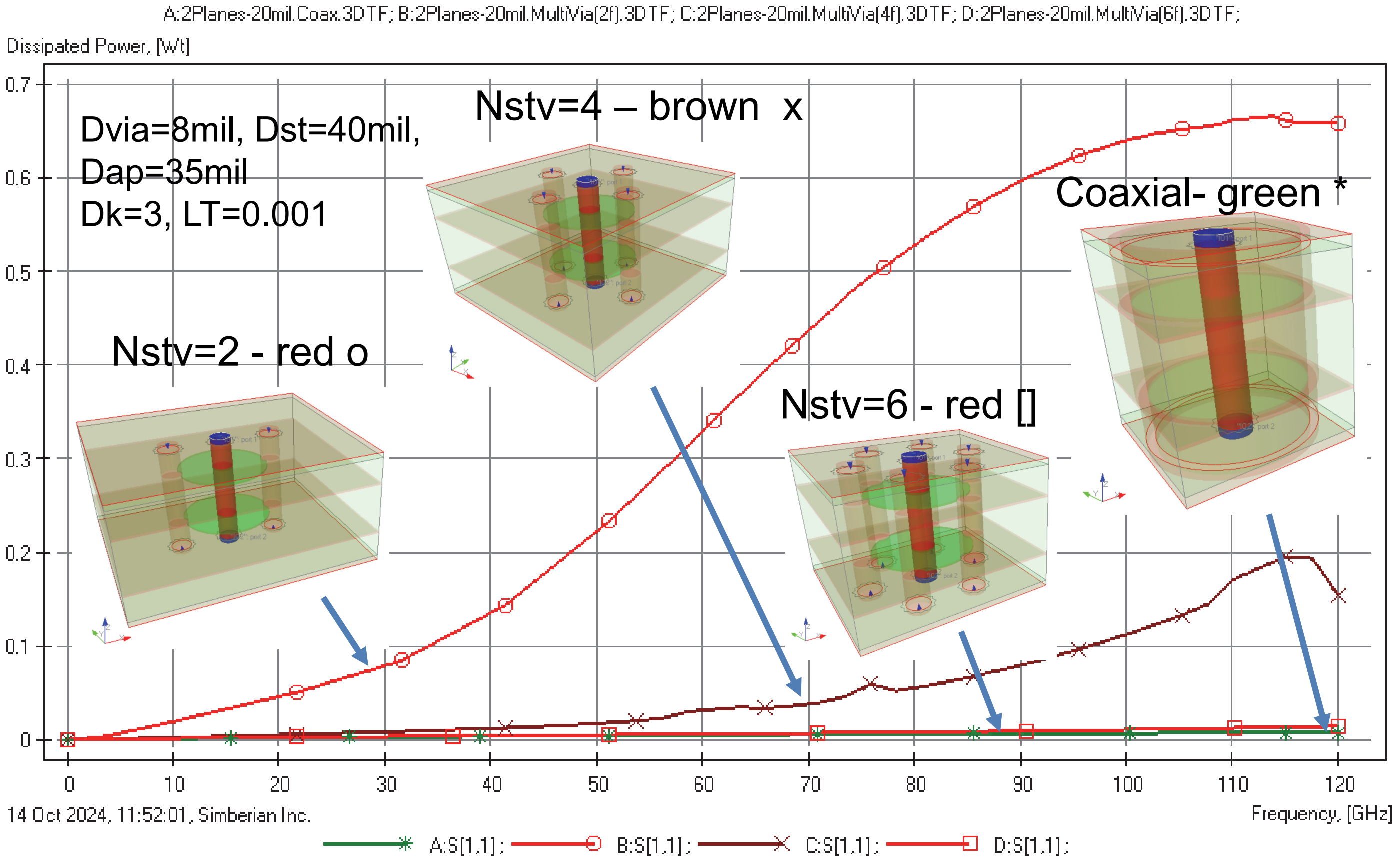
2 Planes, H=20mil

Excitation: 1 Wt

Stitching vias reduce dissipated power and potential coupling

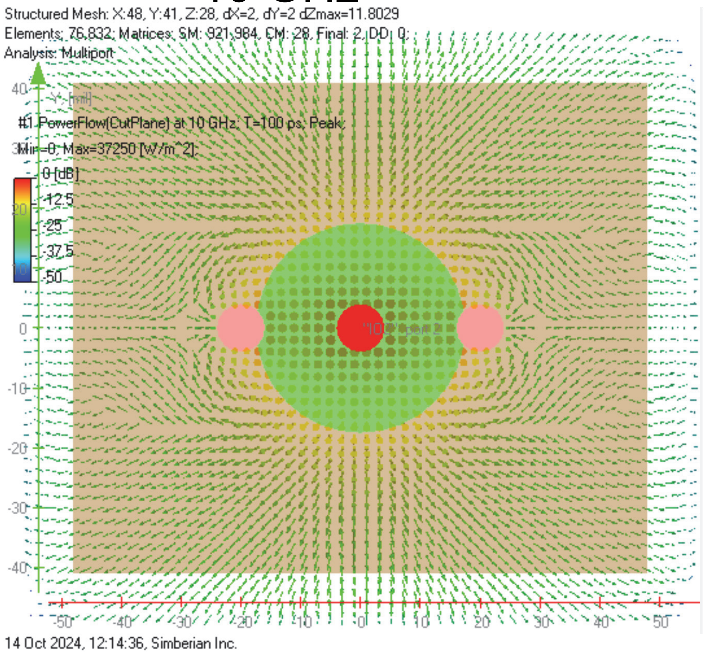
Simbeor 3DTF, PML

Solution: ViaBasics_SE_SICW



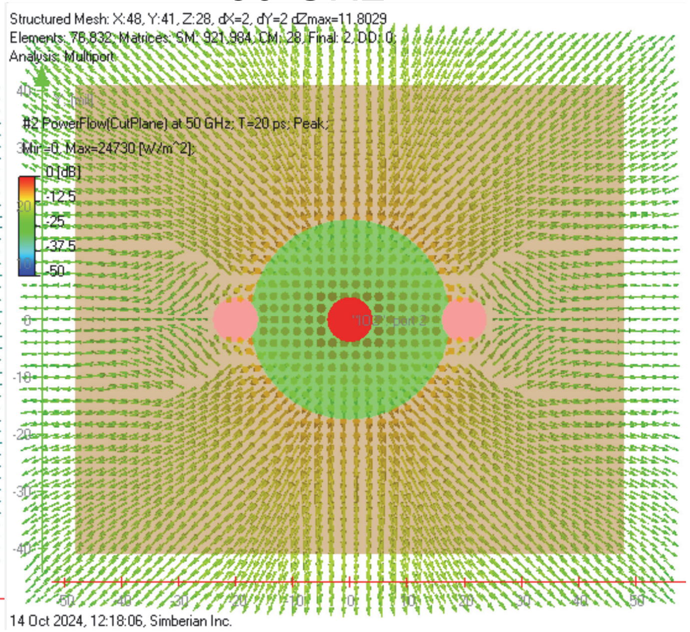
SE Via Power Flow Density, Nstv=2

10 GHz



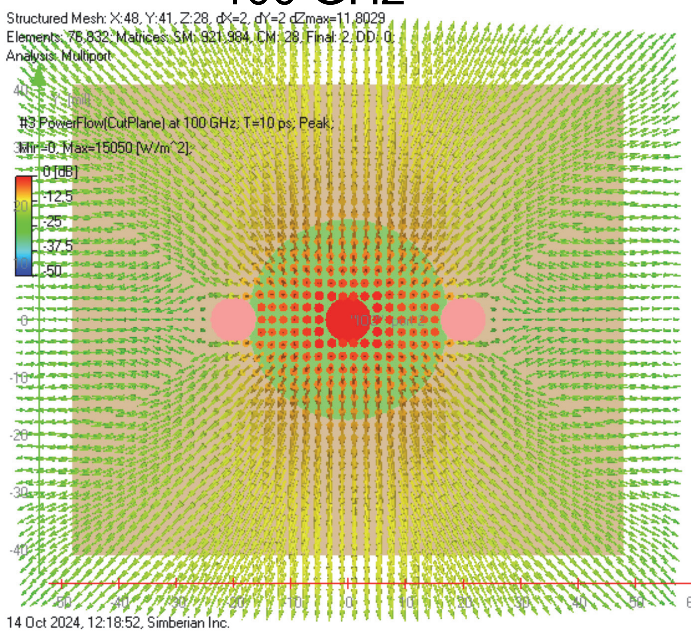
DP=2%

50 GHz



DP=22%

100 GHz

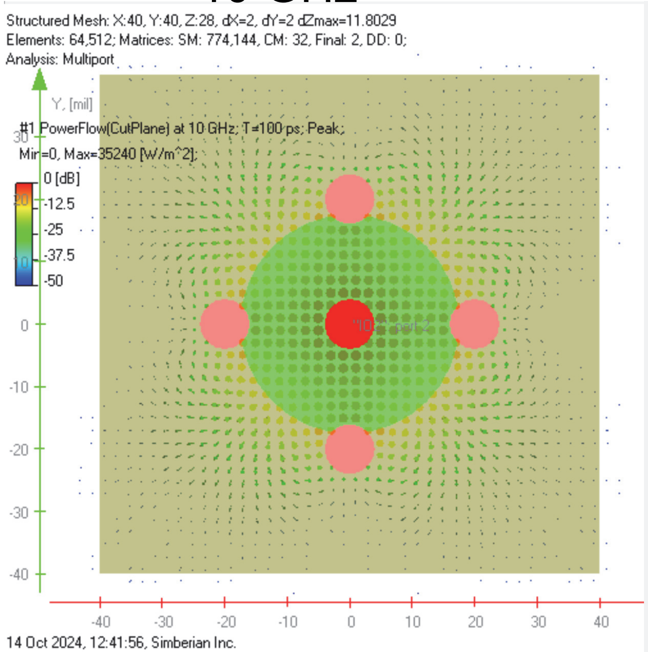


DP=64%

Peak PFD, Simbeor 3DTF, PML BC, Dvia=8mil, Dst=40mil, Dap=35mil, Dk=3, LT=0.001

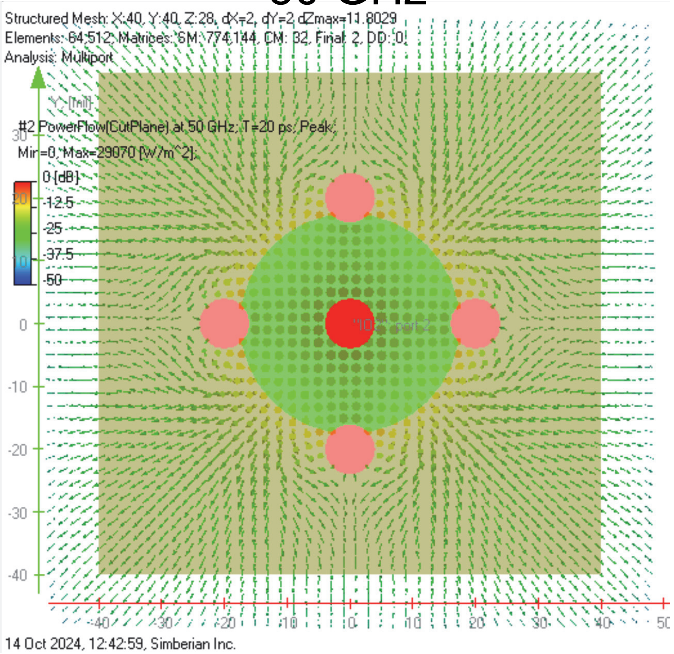
SE Via Power Flow Density, Nstv=4

10 GHz



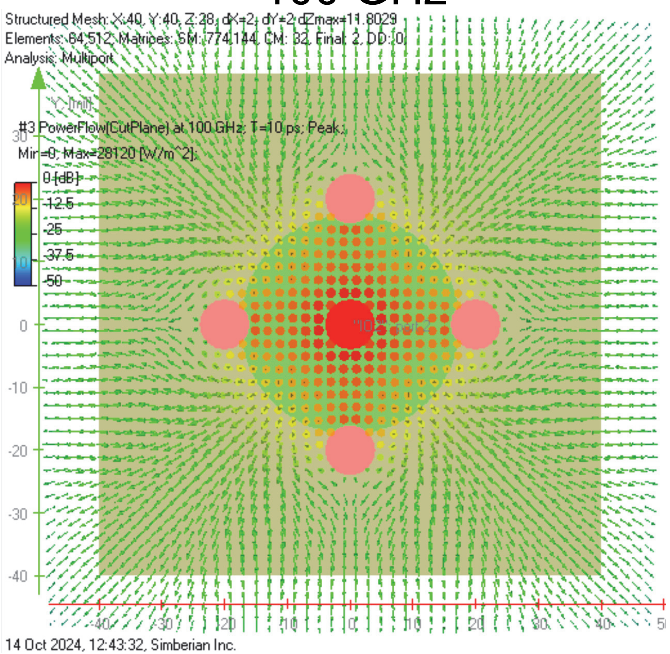
DP=2.4%

50 GHz



DP=1.7%

100 GHz

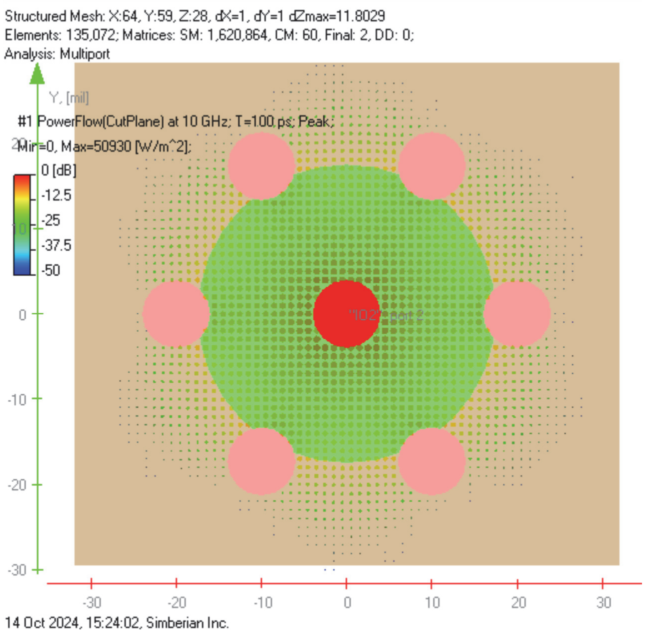


DP=11.4%

Peak PFD, Simbeor 3DTF, PML BC, Dvia=8mil, Dst=40mil, Dap=35mil, Dk=3, LT=0.001

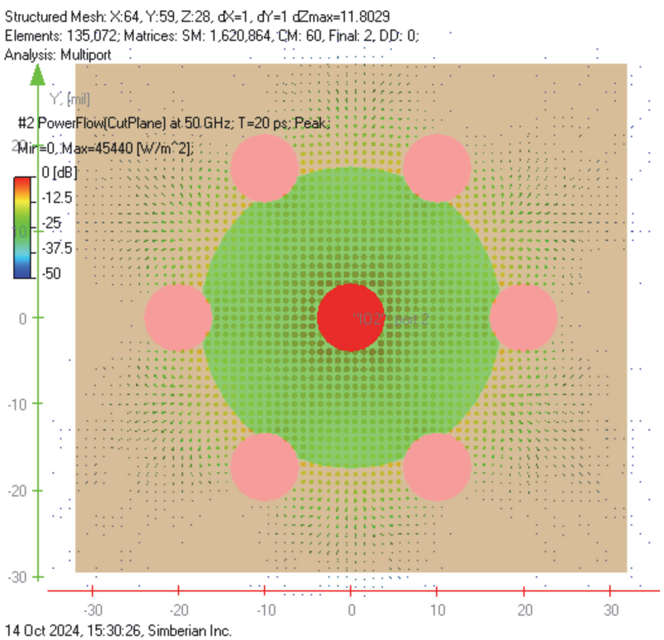
SE Via Power Flow Density, Nstv=6

10 GHz



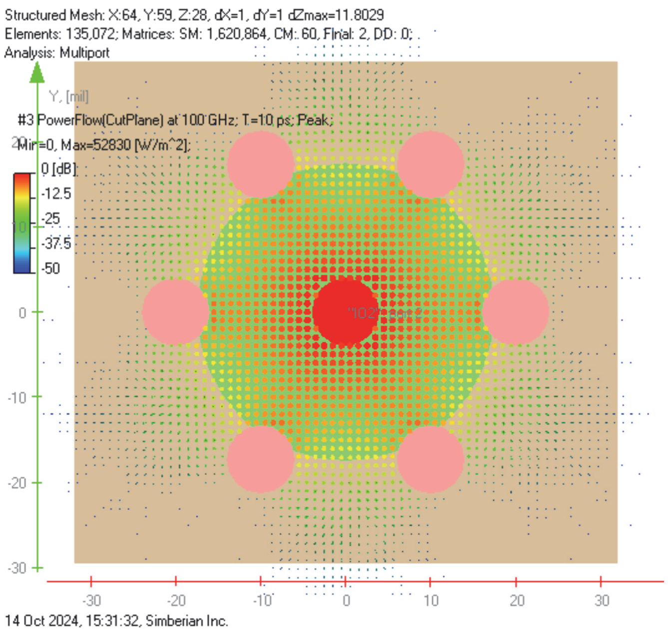
DP=0.15%

50 GHz



DP=0.46%

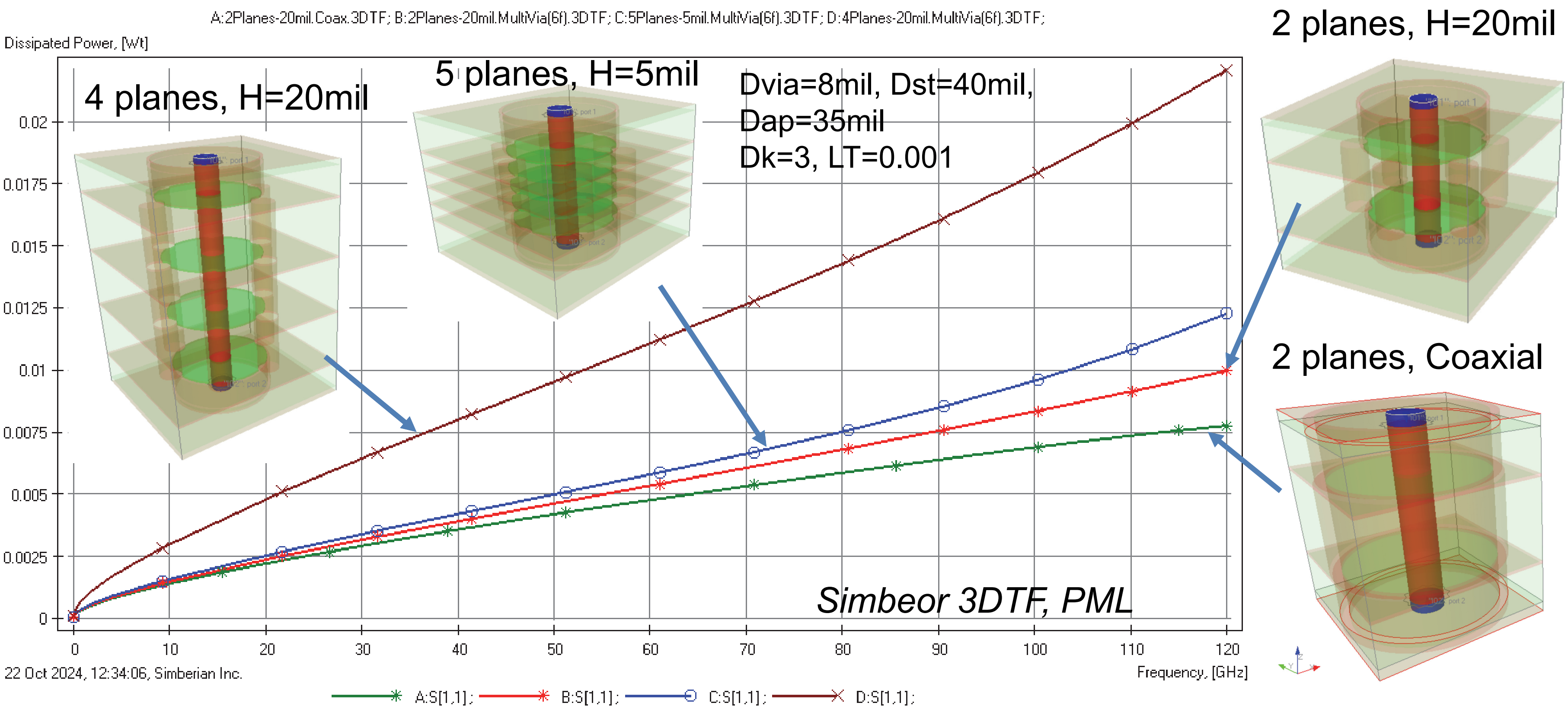
100 GHz



DP=0.8%

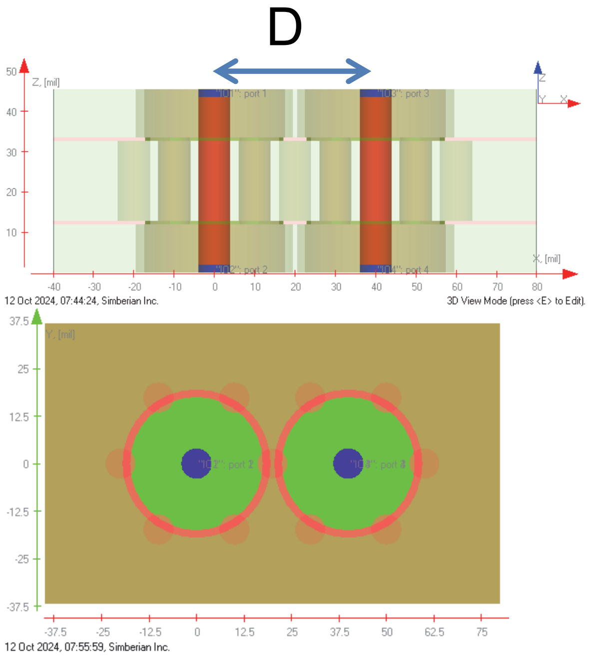
Peak PFD, Simbeor 3DTF, PML BC, Dvia=8mil, Dst=40mil, Dap=35mil, Dk=3, LT=0.001

SE Via Dissipated Power, Nstv=6

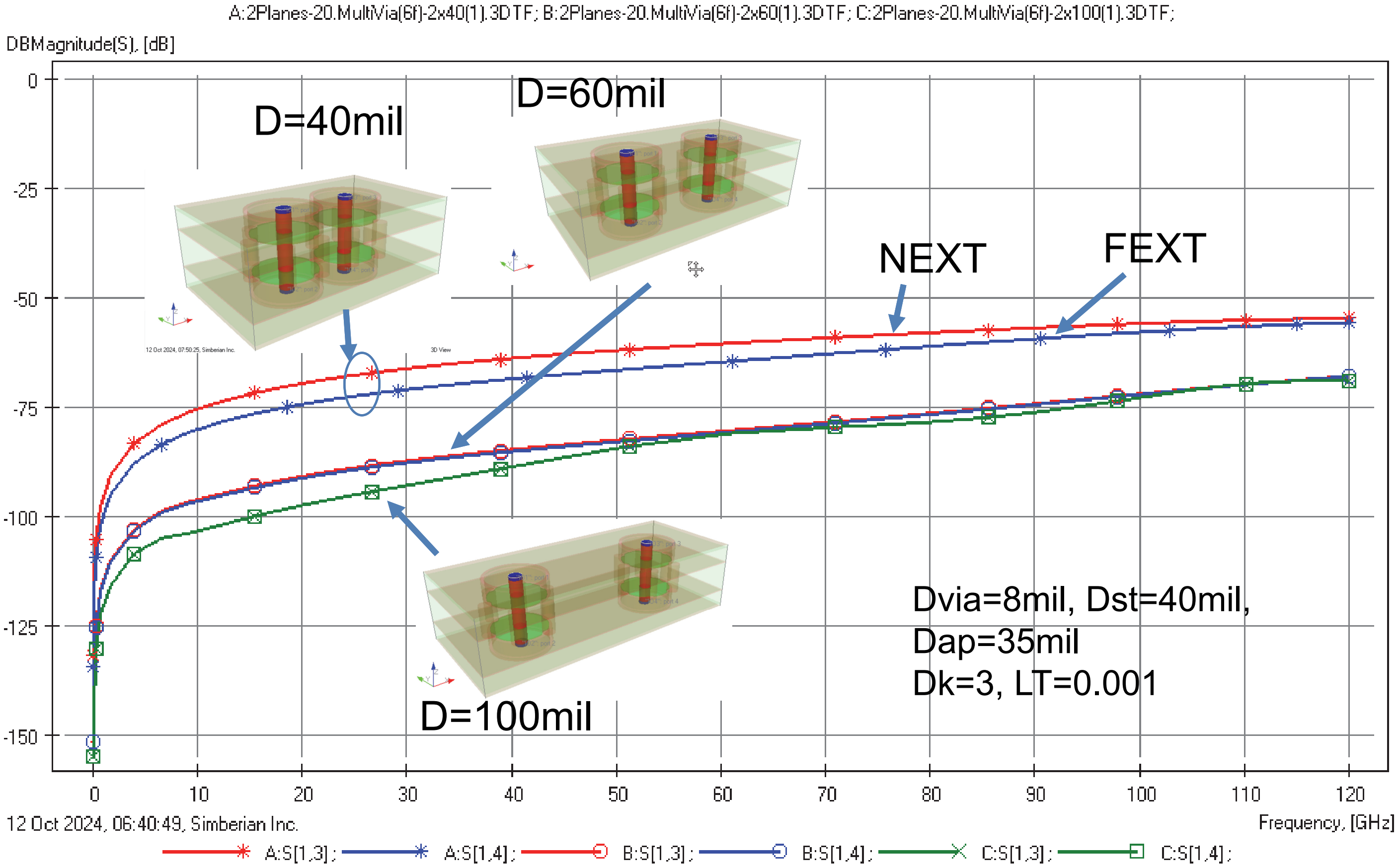


DP increases with number of layers NL - proportional to NL with the same separation H

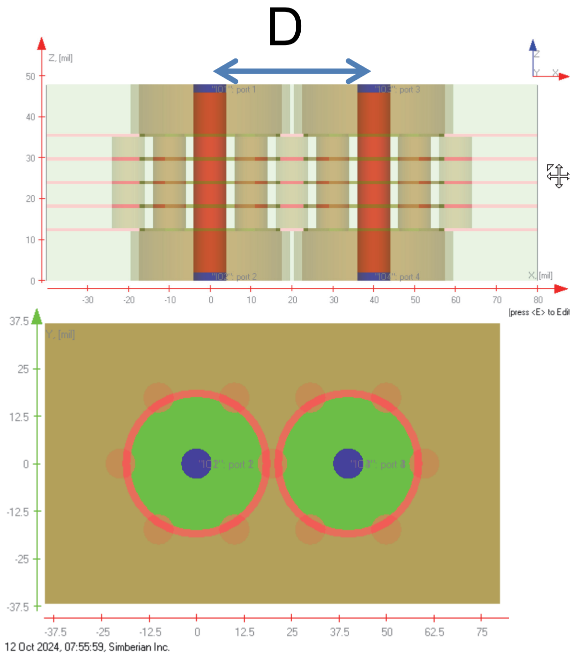
SE Vias Local Coupling – 2 planes, H=20 mil



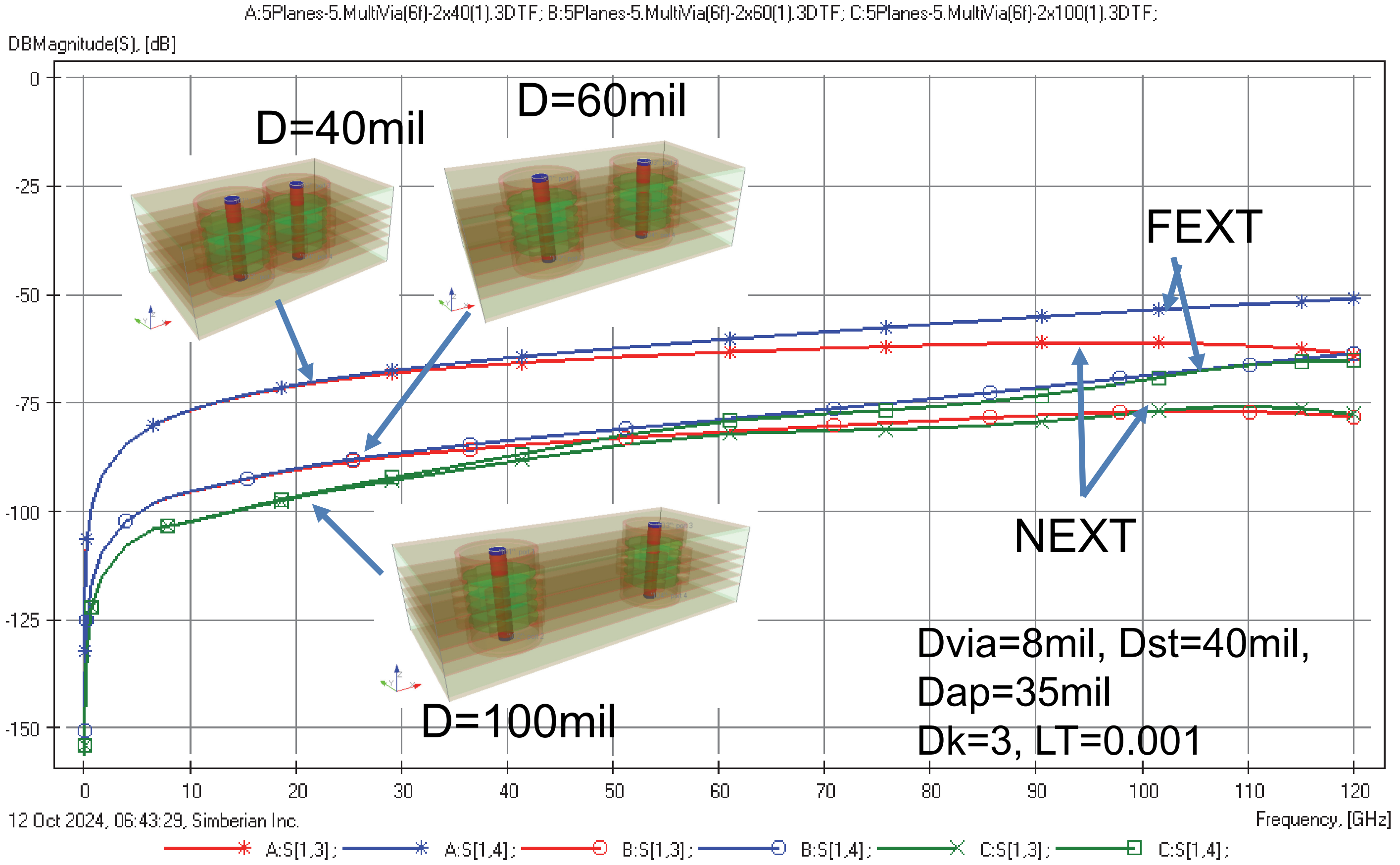
Dvia=8mil, Nstv=6
Dk=3, LT=0.001
Dst=40, Dap=35
Simbeor 3DTF, PML



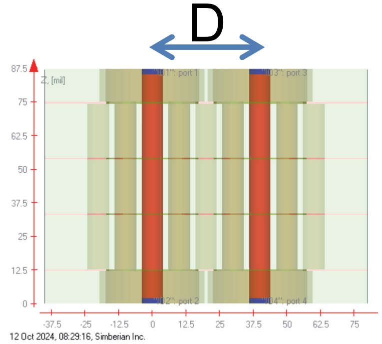
SE Vias Local Coupling – 5 planes, H=5mil

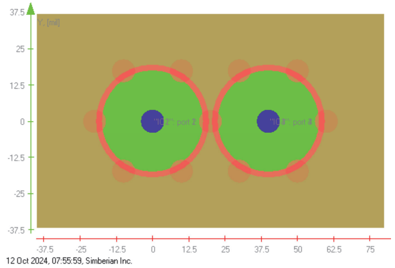


Simbeor 3DTF, PML
Dvia=8mil, Nstv=6
Dk=3, LT=0.001
Dst=40, Dap=35

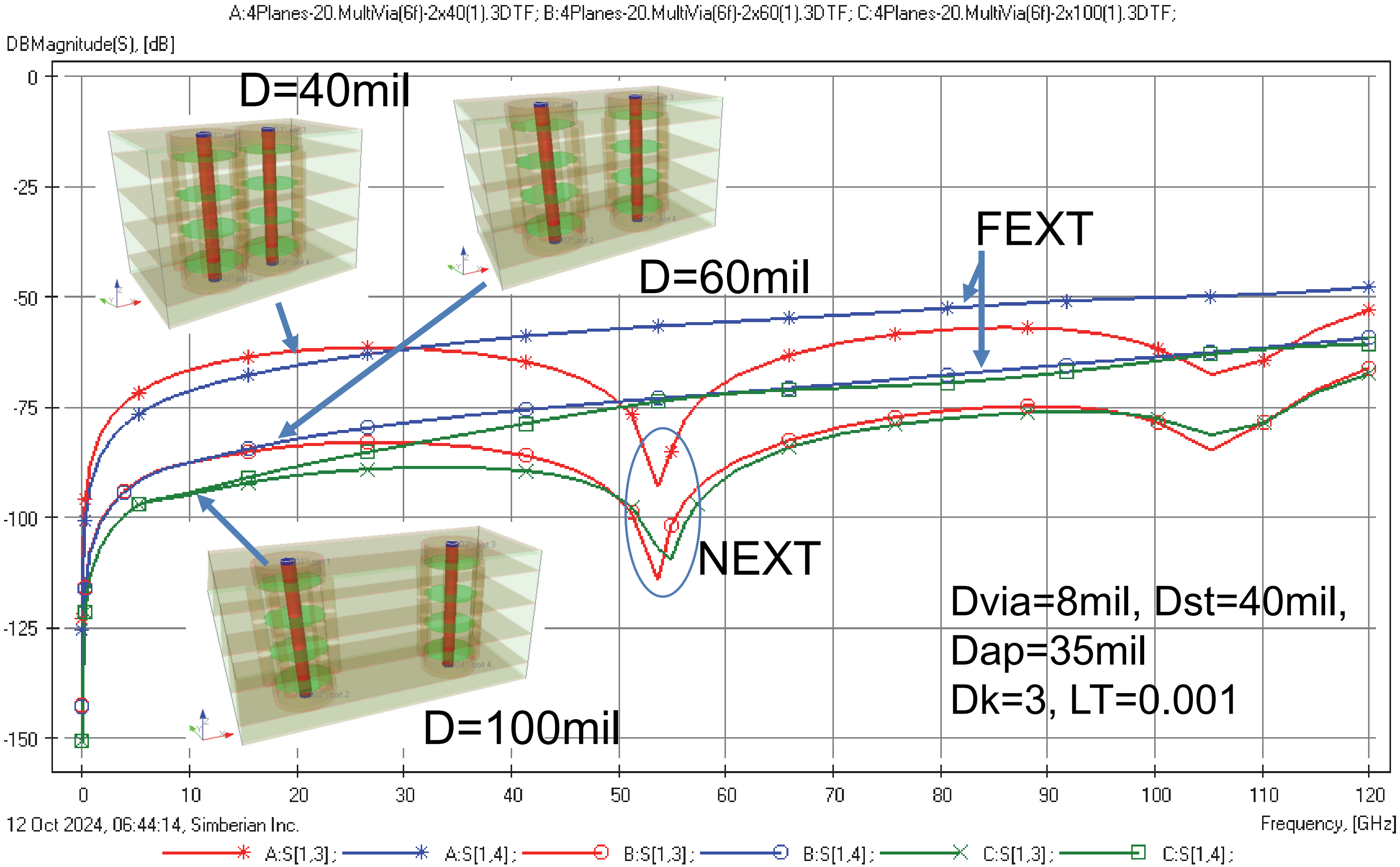


SE Vias Local Coupling – 4 planes, H=20mil





Simbeor 3DTF, PML
Dvia=8mil, Nstv=6
Dk=3, LT=0.001
Dst=40, Dap=35

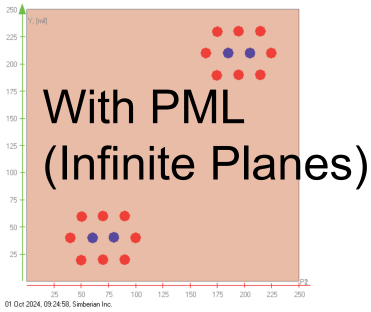


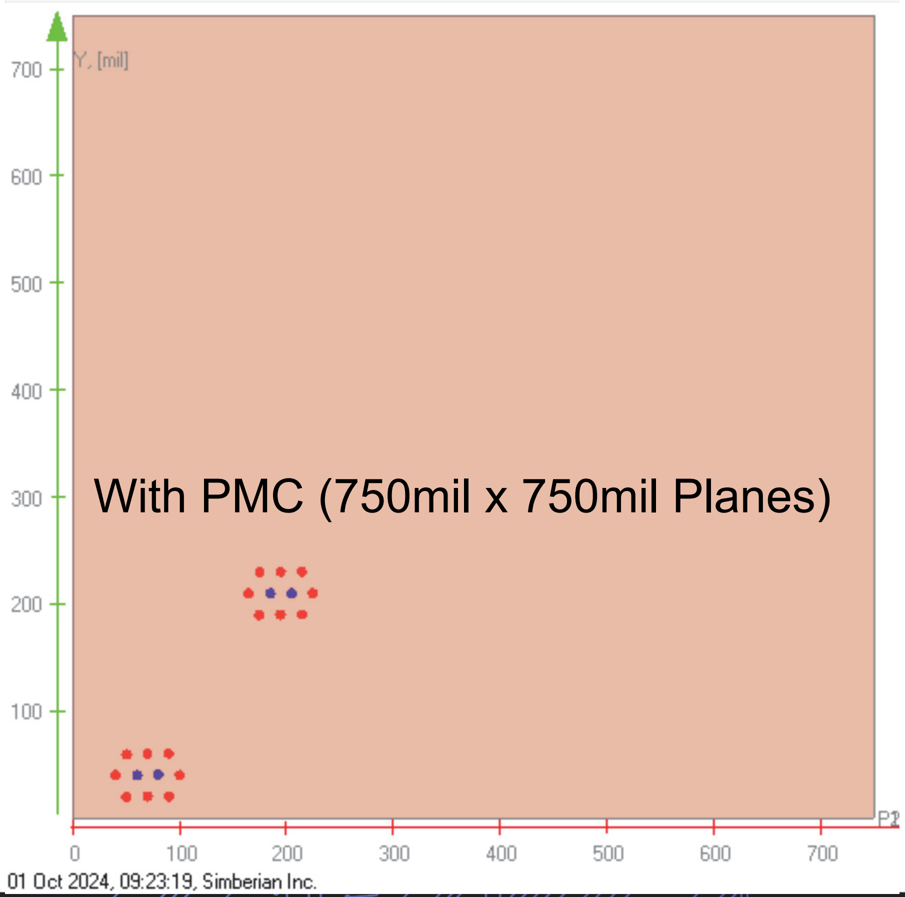
Diff. Vias Localization and Distant Coupling

Two 0.77mil copper planes, separated by 9mil dielectric with Dk=3, LT=0.001
Differential signal vias 20mil distance (10mil diameter), two pairs at ~220mil
Number of stitching vias (Nstv) from 0 to 8 at about 20mil distance from signal
8-port structure with 50Ohm terminations

Physics-based model with 2D analysis in Simbeor 3DTF solver

Solution:
Coupling_PPW_DIFF





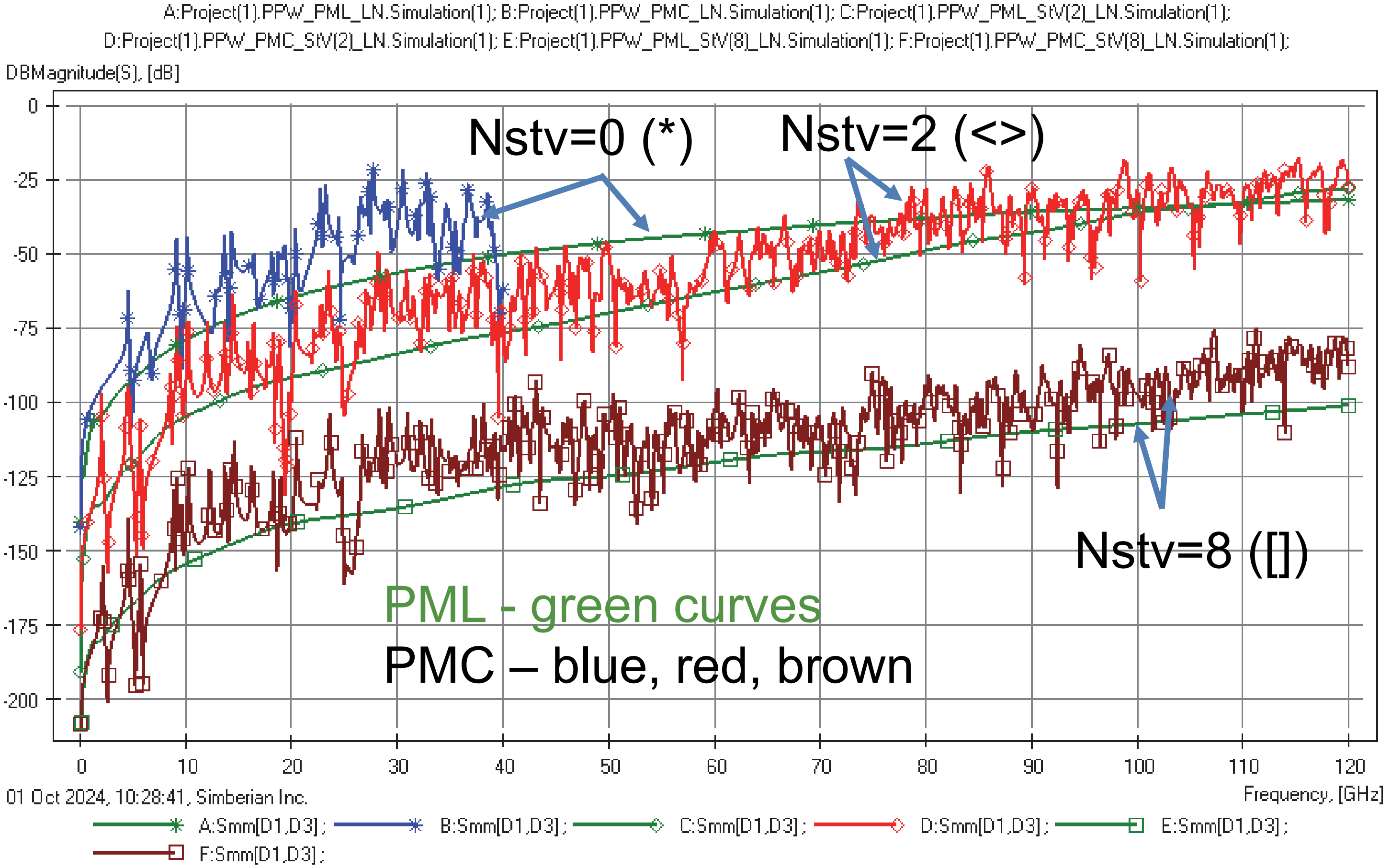
Differential Mode Coupling

DM coupling is reduced by increase of Nstv

Two stitching vias may be not enough for extended bandwidth

Common mode coupling is much worse (similar to SE)
In addition, there is substantial common to differential mode conversion

Analysis complexity for real-life problems is enormous



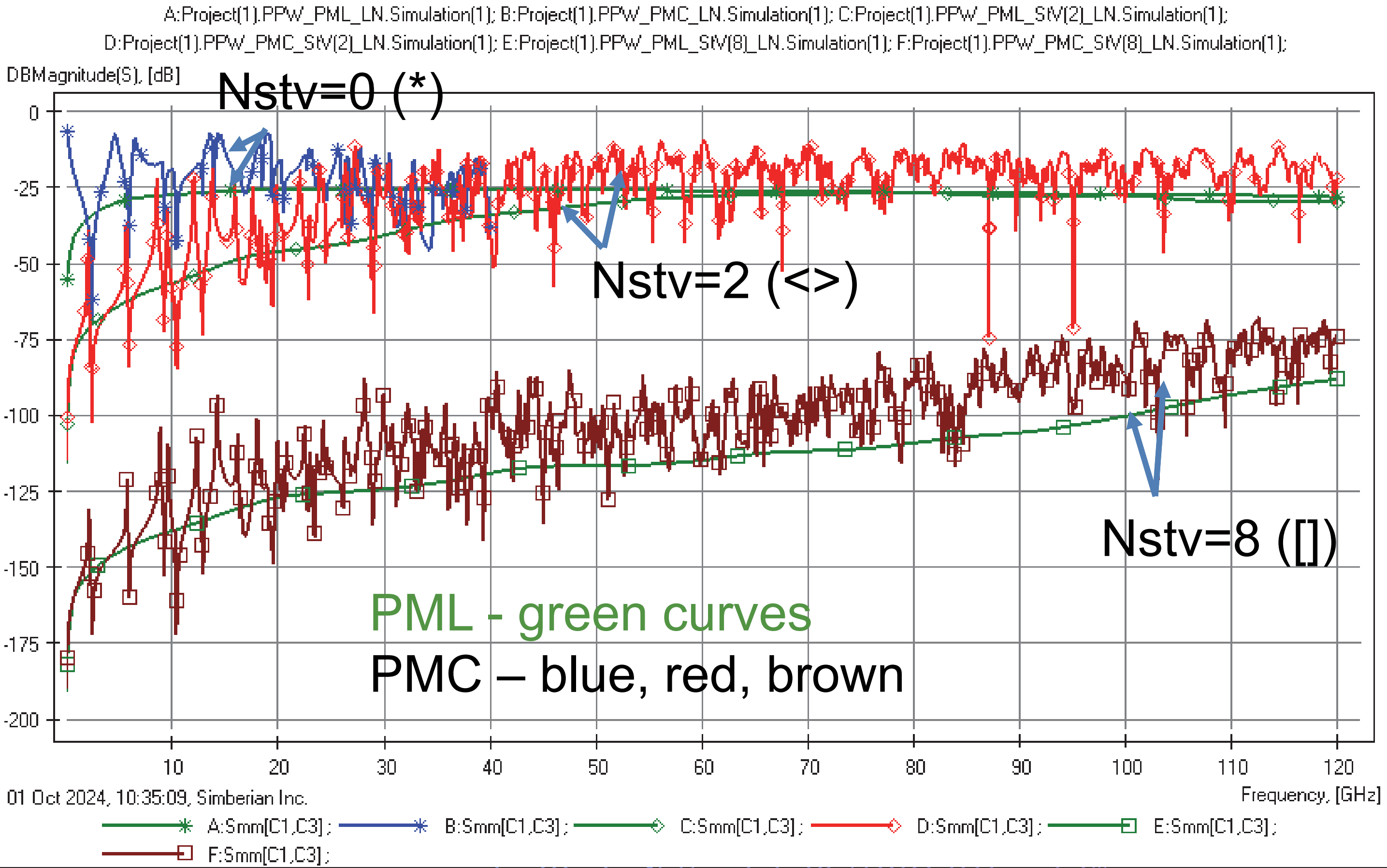
Common Mode Coupling

CM is much larger than DM

CM coupling is reduced by increase of Nstv (similar to SE)

Two stitching vias is not enough

If CM isolation is needed – more stitching vias is required

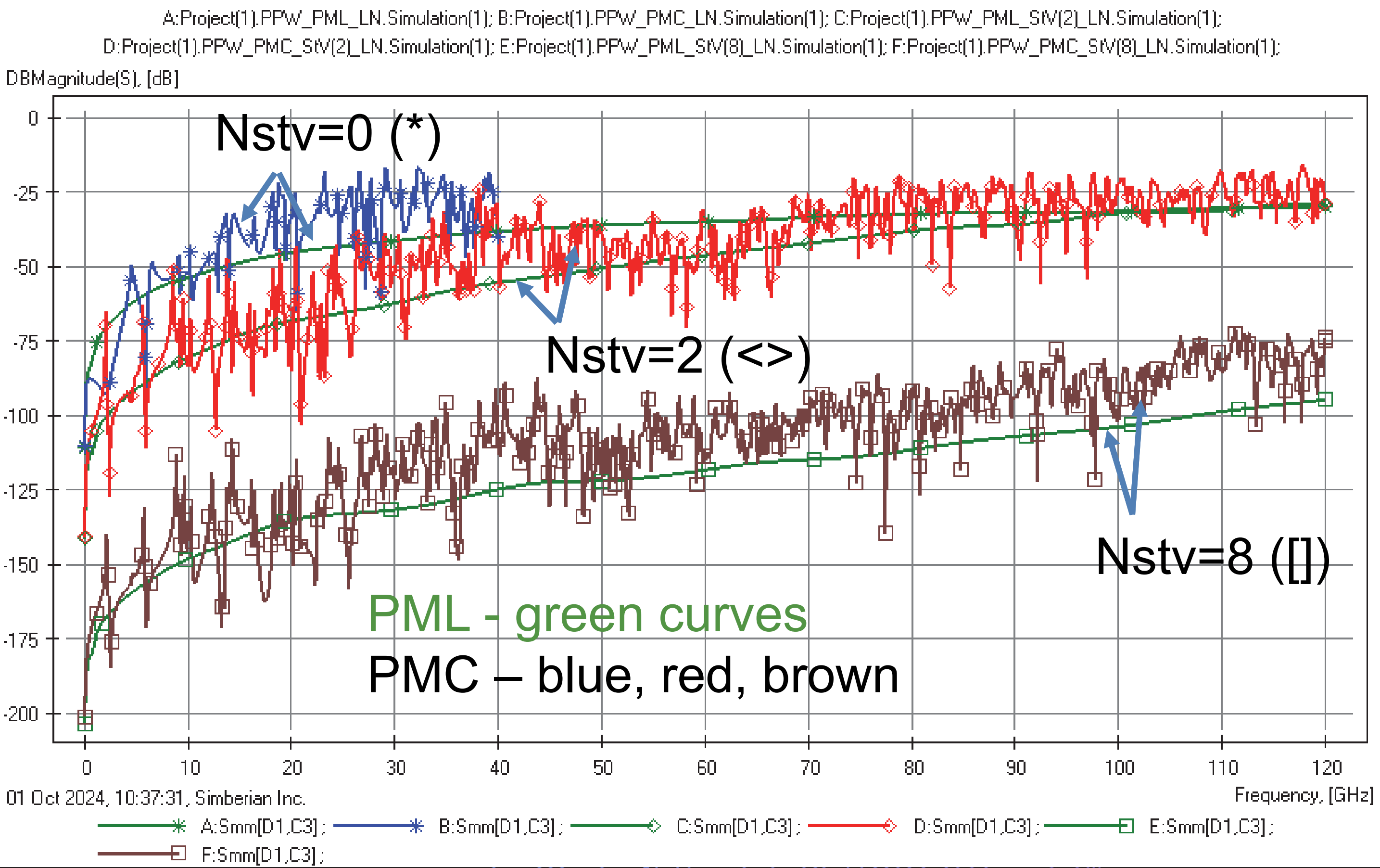


Common to Differential Mode Coupling

Modal coupling through PDN even in symmetrical vias

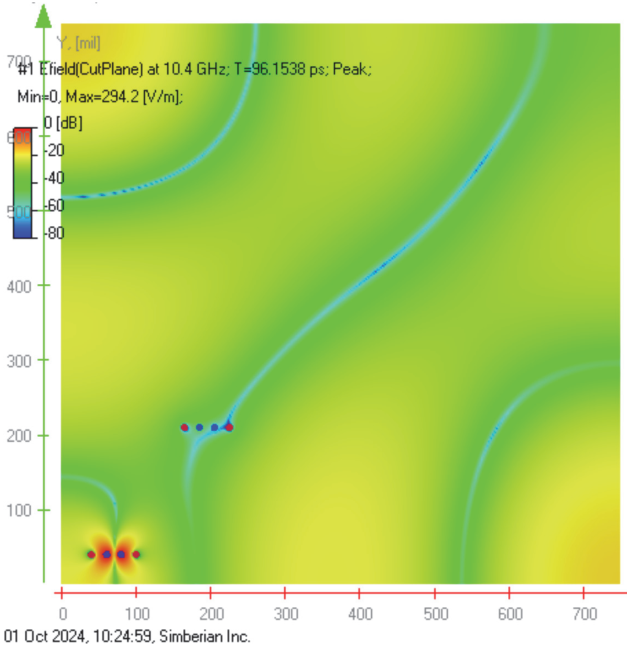
Depends on PDN geometry and resonances

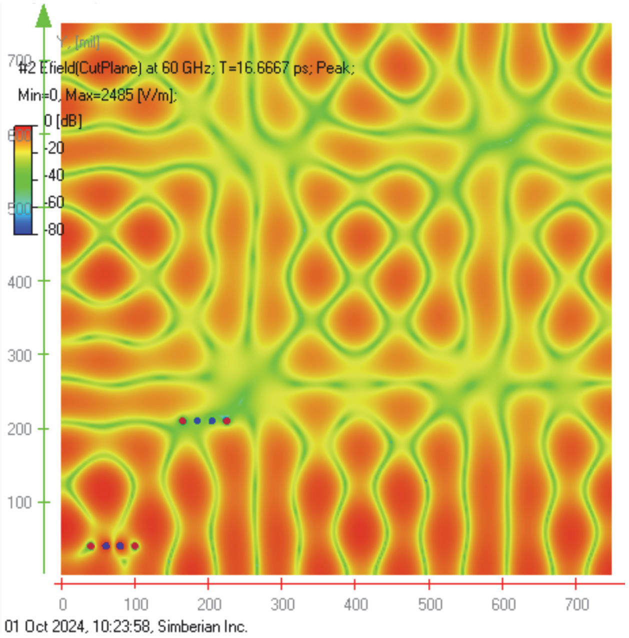
It is reduced by increase of Nstv

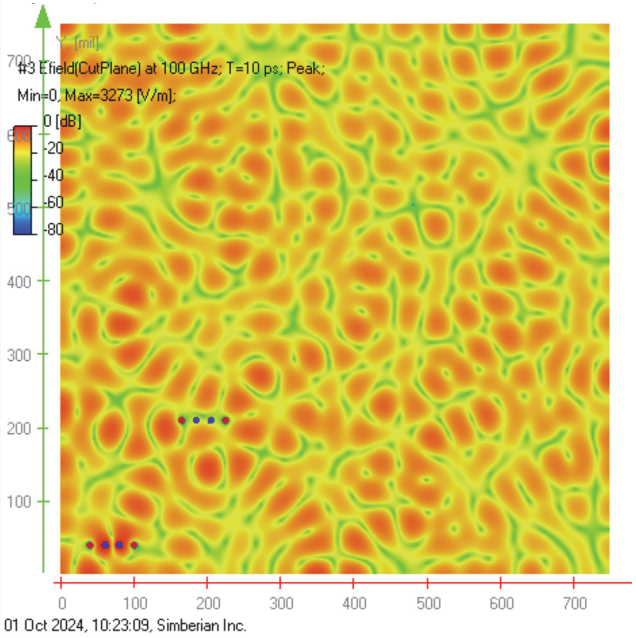


Coupling and Mode Conversion – Fields

750mil x 750mil with PMC
Electric field for 1V differential source at corner vias
Difficult to account in 3D EM analysis on real boards





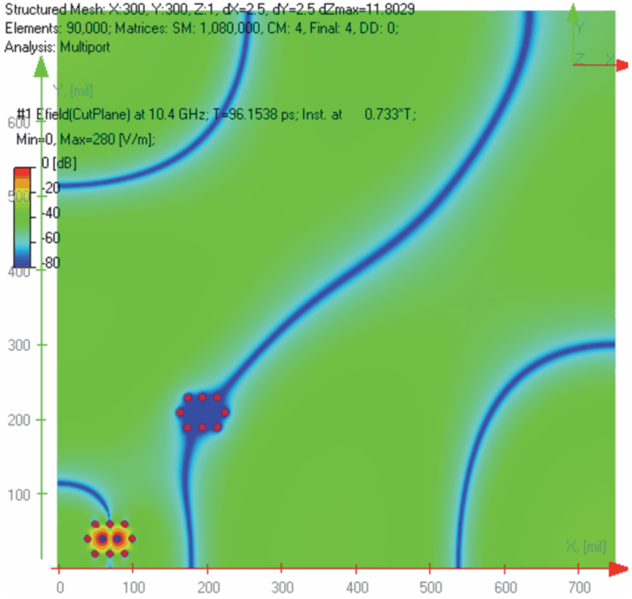


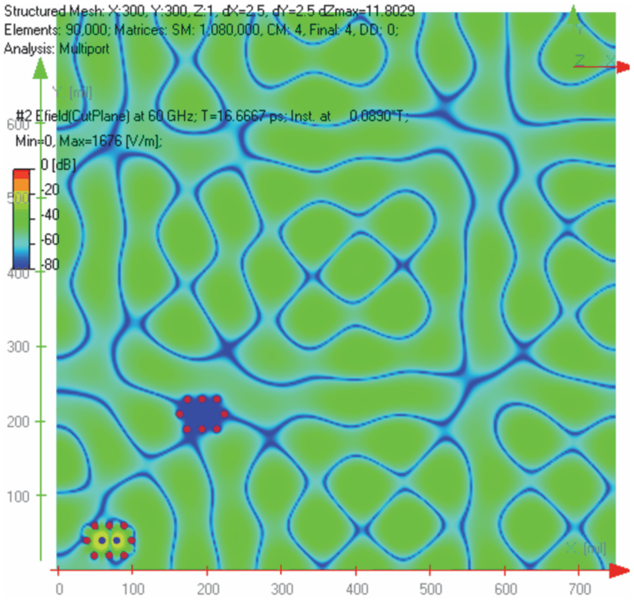
Coupling Reduction with Stitching Vias

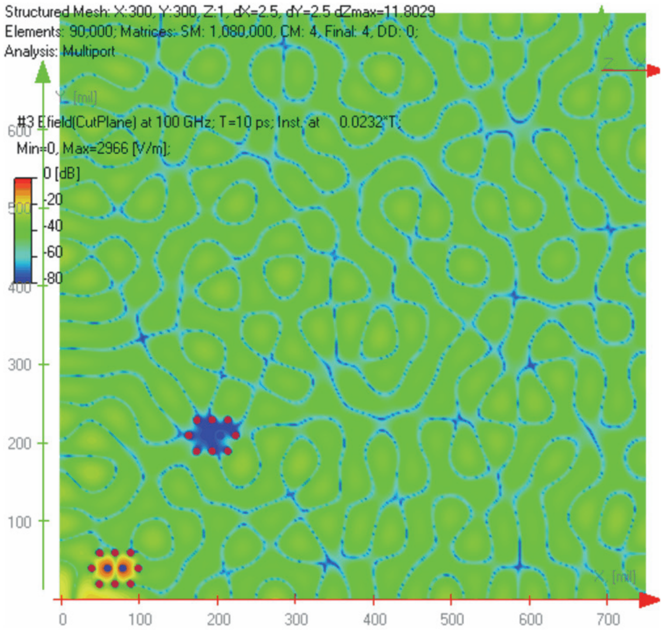
750mil x 750mil with PMC, Nstv=8

Electric field for 1V differential source at corner vias

Does not need 3D EM analysis of whole boards







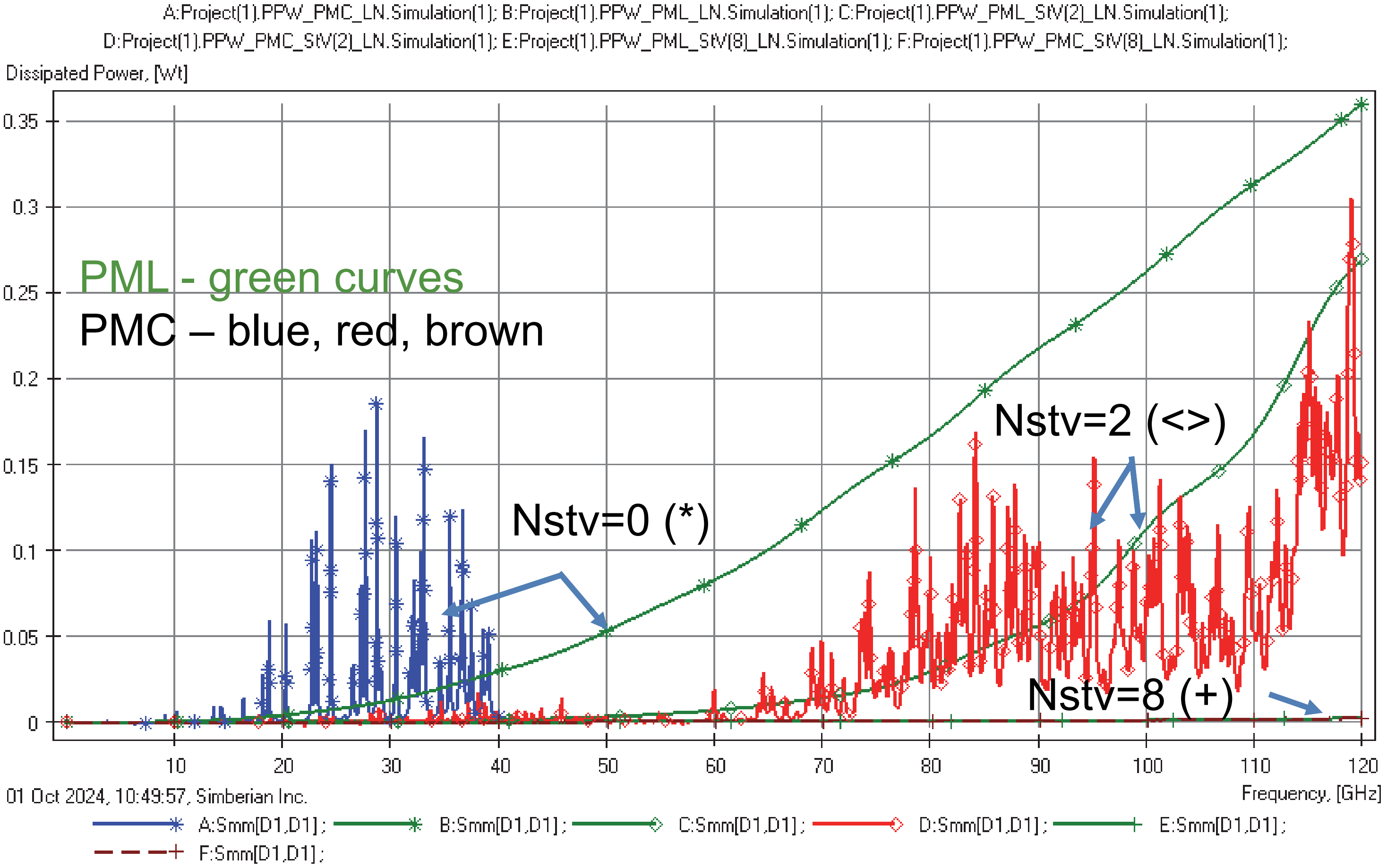
(animated)

Differential Mode Dissipated Power

$$P_{dissipated} = \left(1 - \sum_k |S_{k,1}|^2\right) P_{in}$$

Stitching vias reduce power dissipation (leaks)

Power dissipation for sufficiently localized structures can be evaluated with PML boundaries (infinite planes)



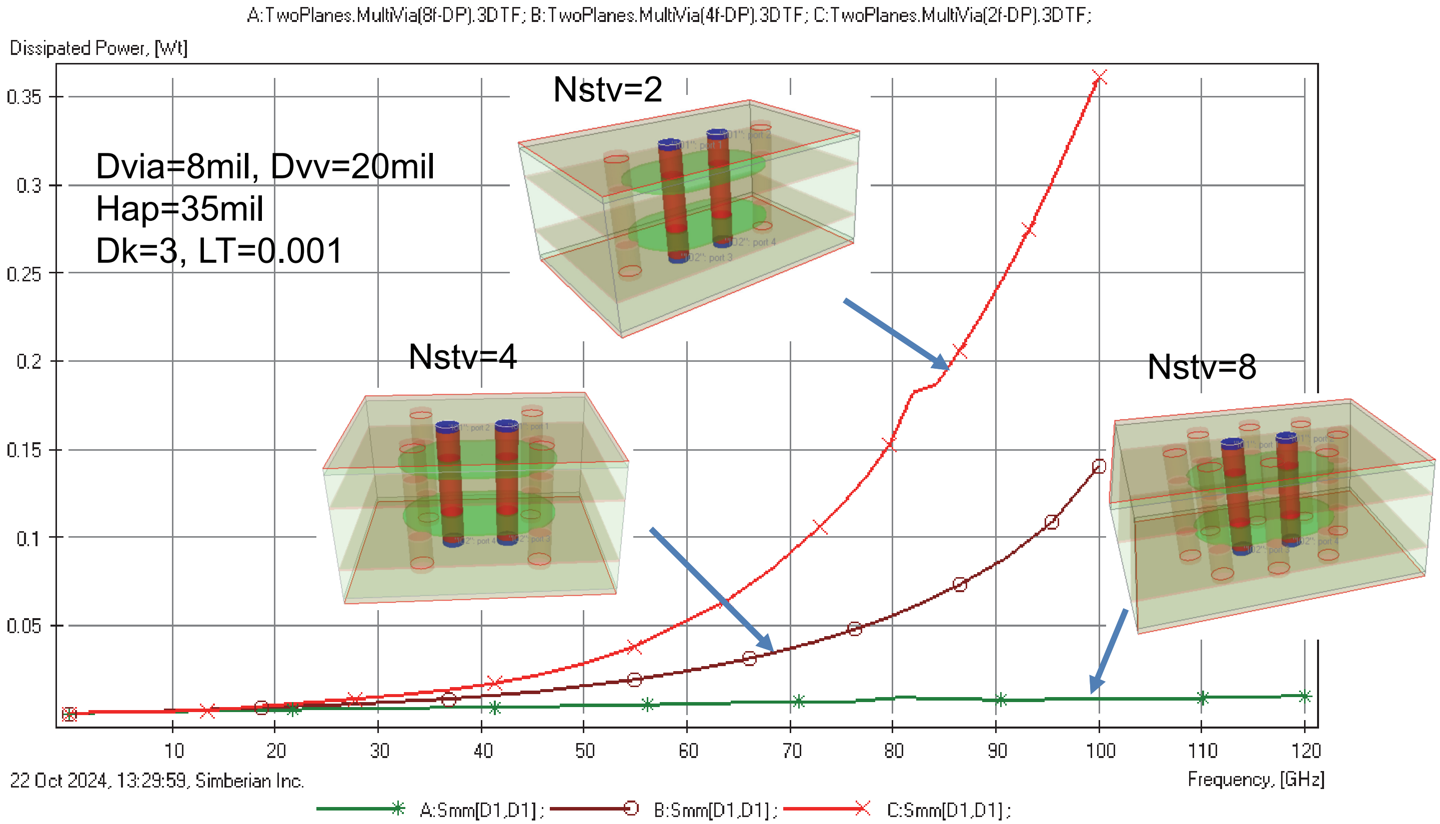
Differential Vias – Dissipated Power, 3D EM

$$P_{dissipated} = \left(1 - \sum_k \left|S_{k,1}\right|^2\right) P_{in}$$

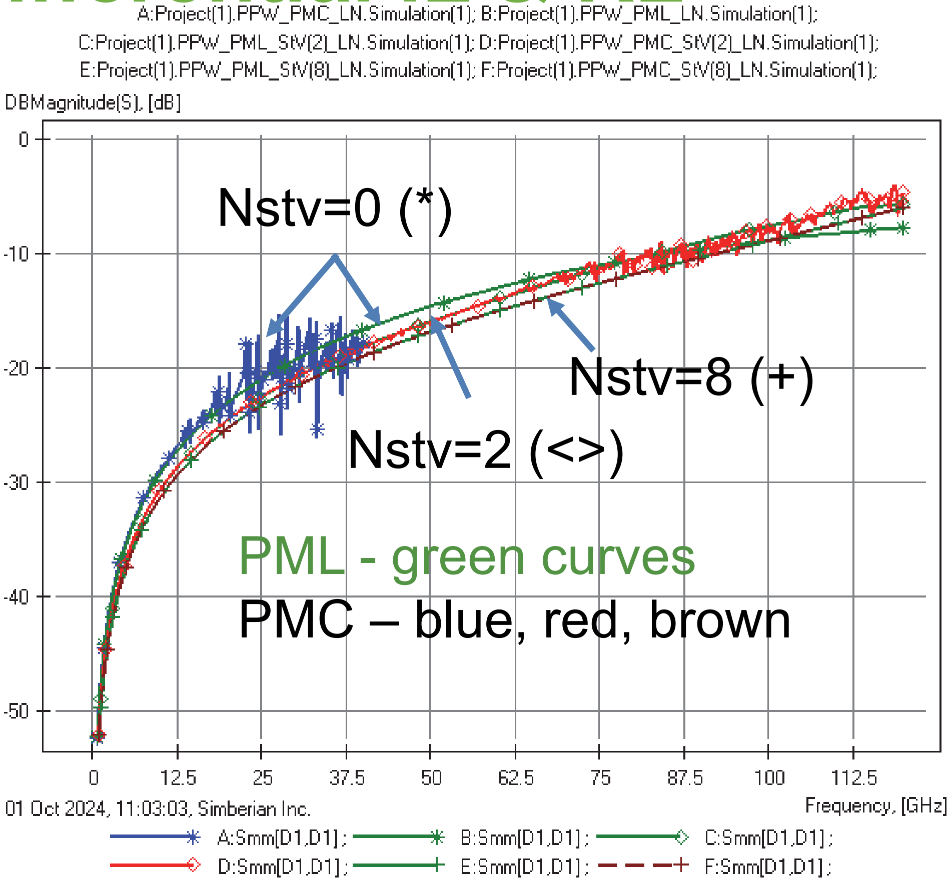
Diff. Excitation 1 Wt

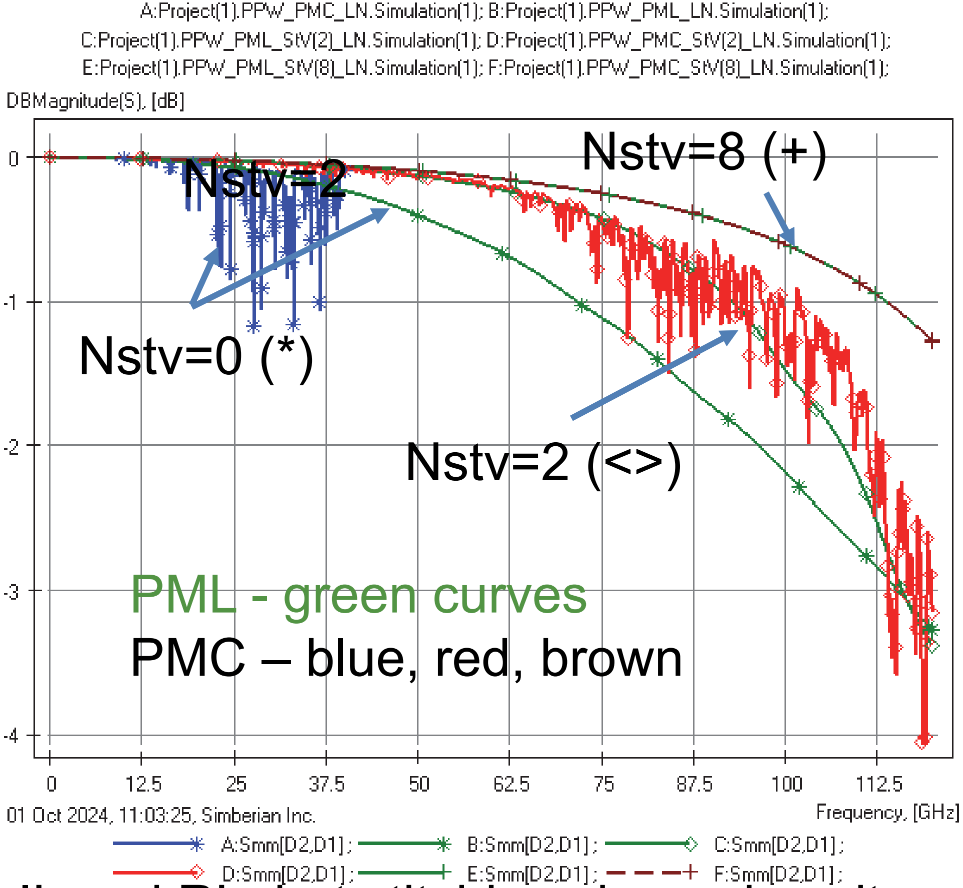
Stitching vias reduce dissipated power and possible leaks and coupling

Simbeor 3DTF, PML



Differential IL & RL



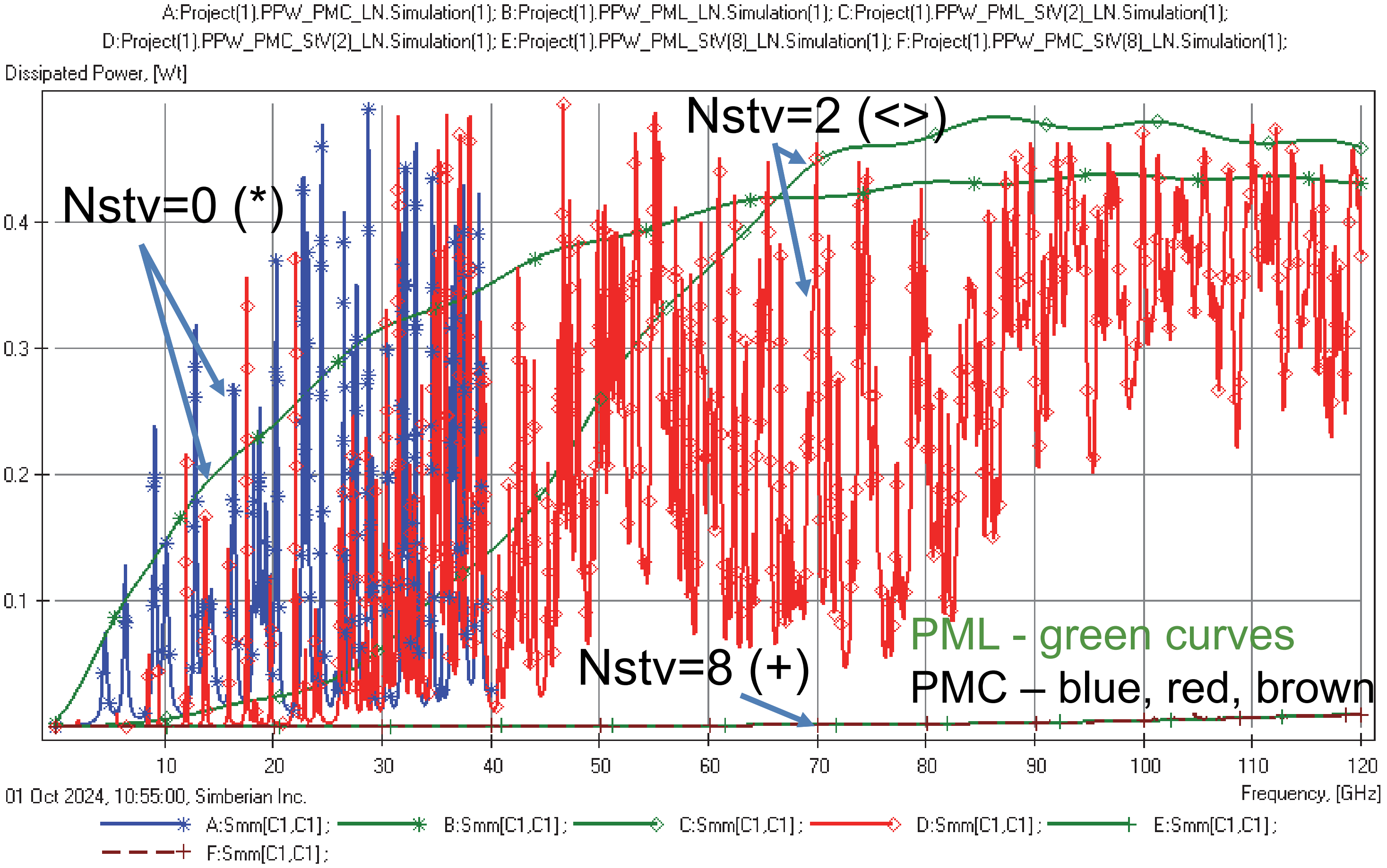


Coupling to cavities causes resonances in IL and RL, but stitching vias reduce it…

Common Mode Dissipated Power

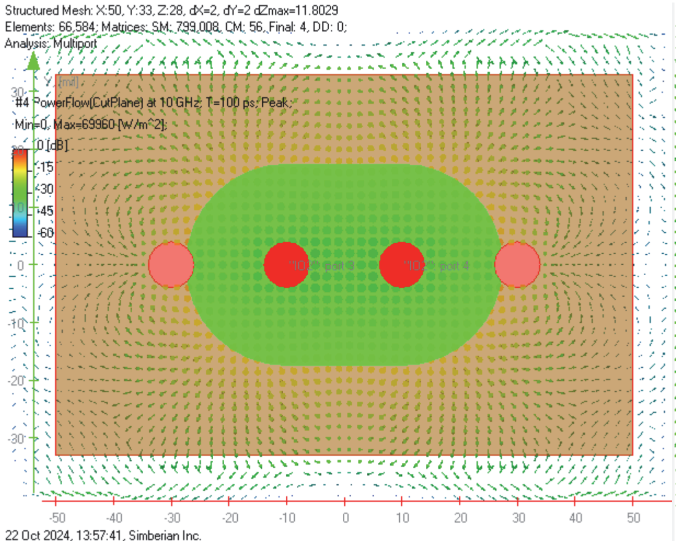
Stitching vias reduce leaks of common mode

Require more stitching vias for ideal de-coupling from PDN



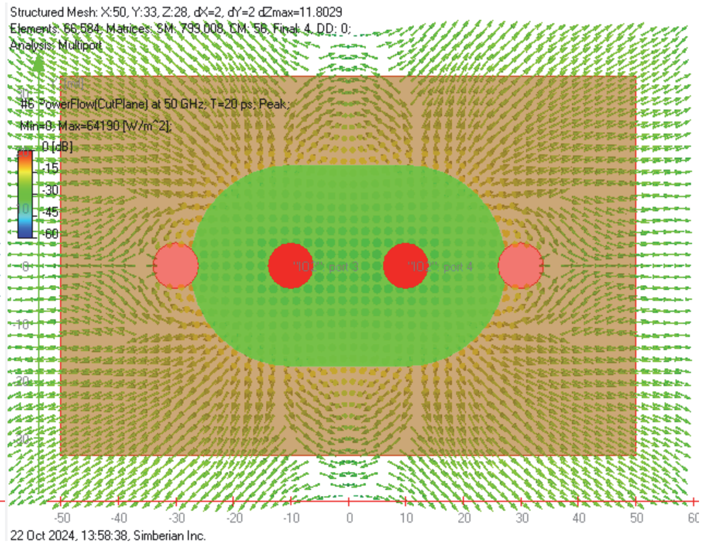
Power Flow Density: Diff. Mode, Nstv=2

10 GHz



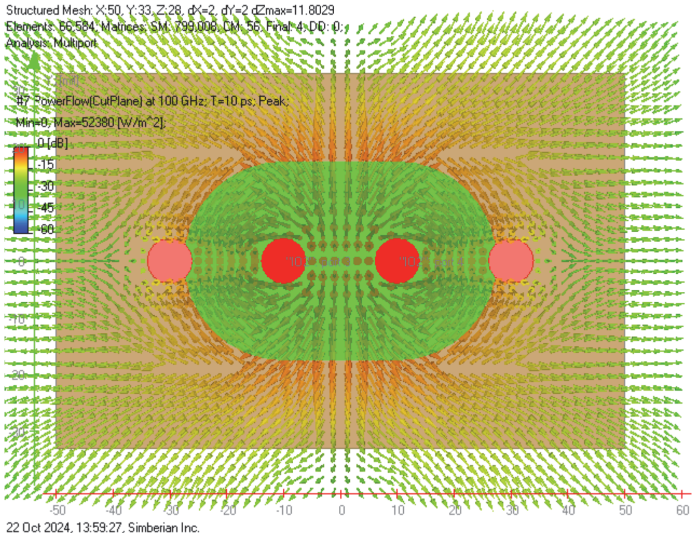
DP=0.12%

50 GHz



DP=2.9%

100 GHz

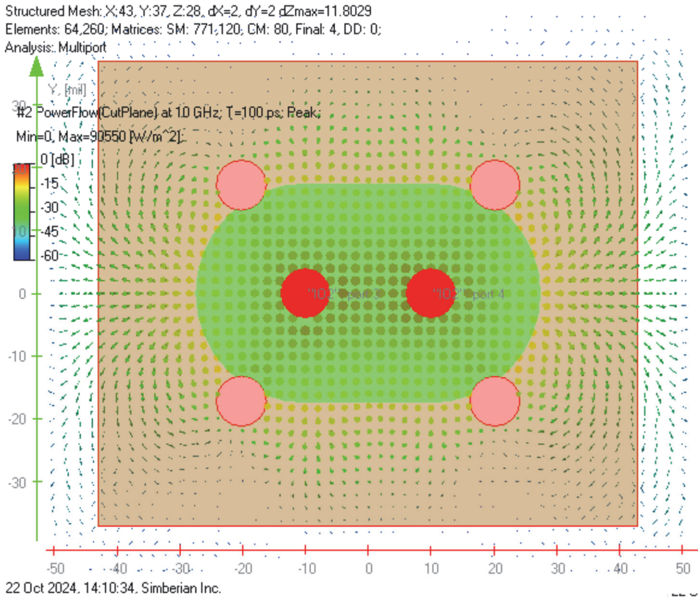


DP=36%

Simbeor 3DTF, PML BC, Dvia=8mil, Dvv=20mil, Hap=35mil, Dk=3, LT=0.001

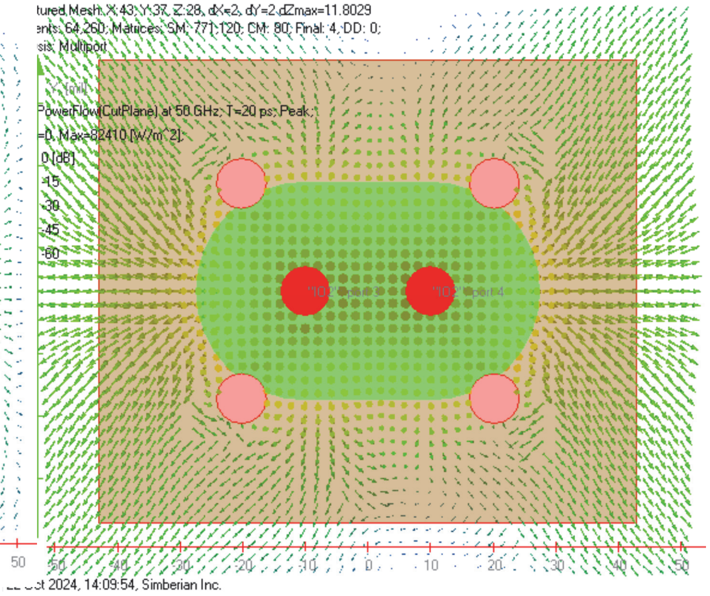
Power Flow Density: Diff. Mode, Nstv=4

10 GHz



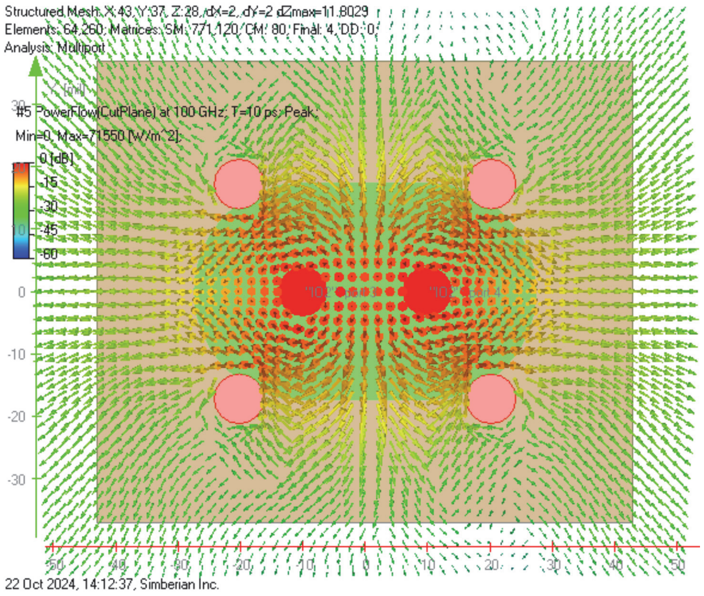
DP=0.14%

50 GHz



DP=1.6%

100 GHz

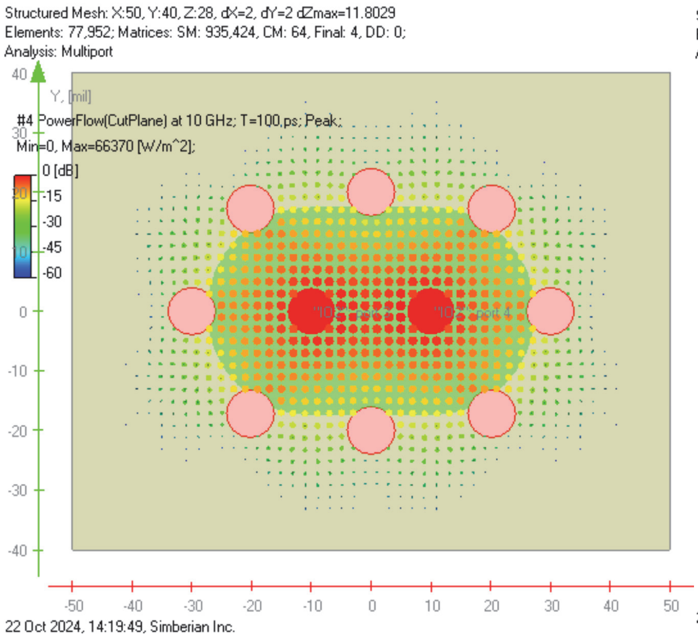


DP=14%

Simbeor 3DTF, PML BC, Dvia=8mil, Dvv=20mil, Hap=35mil, Dk=3, LT=0.001

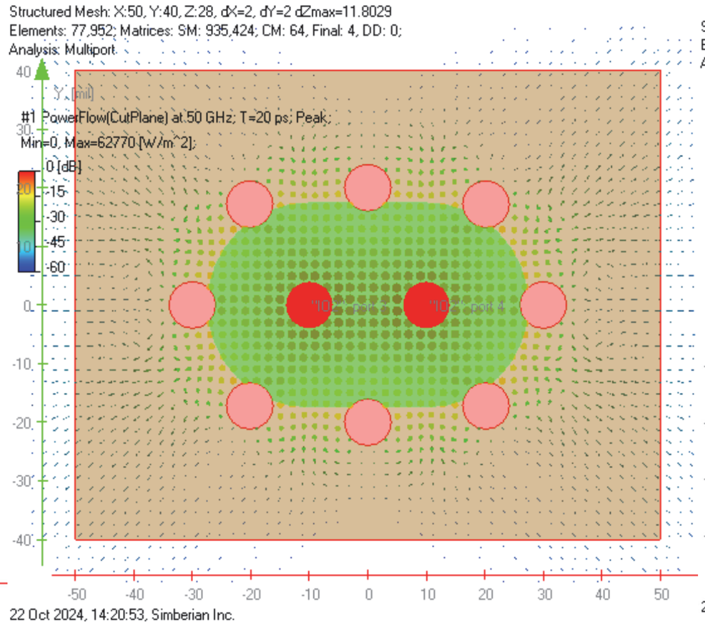
Power Flow Density: Diff. Mode, Nstv=8

10 GHz



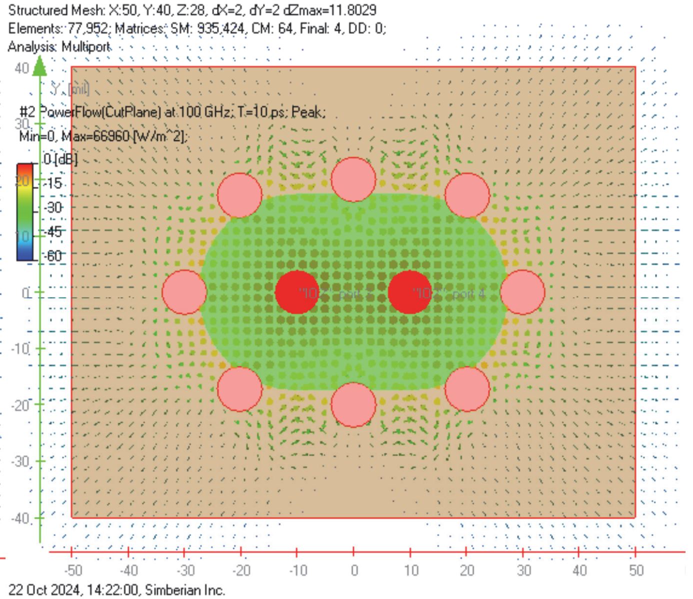
DP=0.13%

50 GHz



DP=0.45%

100 GHz



DP=0.84%

Simbeor 3DTF, PML BC, Dvia=8mil, Dvv=20mil, Hstv=40mil, Hap=35mil, Dk=3, LT=0.001

Localized vs. Non-Localized

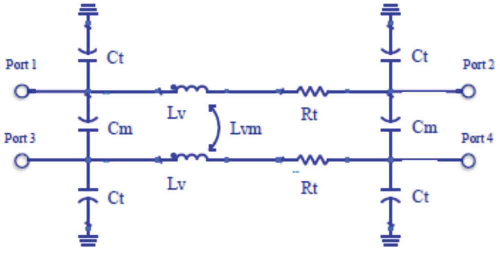


Fig. 2. Model 1 - RLC π-type circuit model

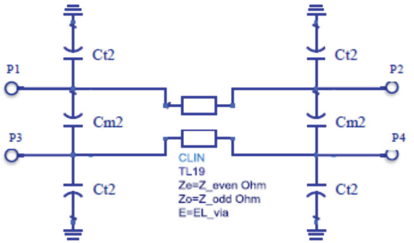
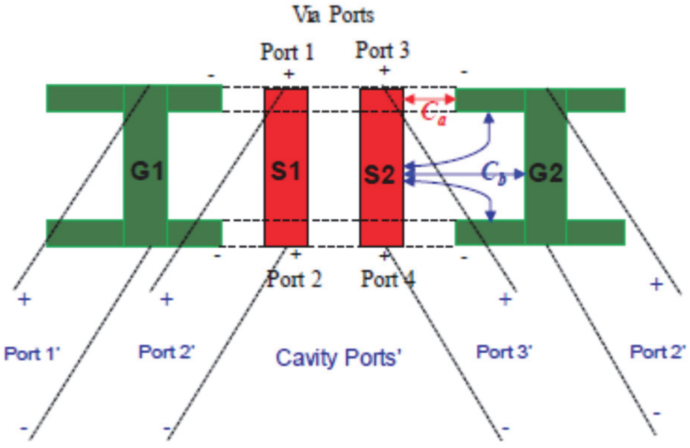
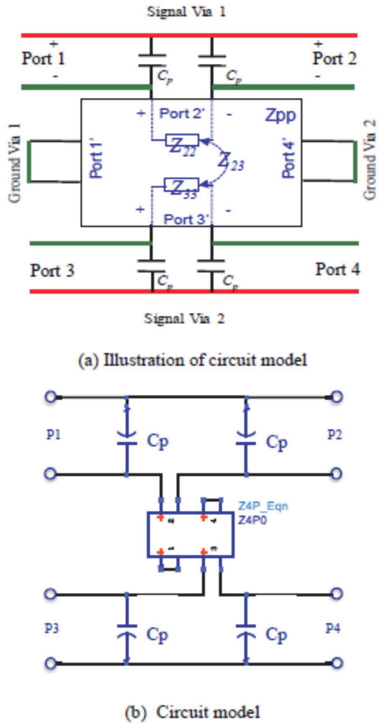


Fig. 3. Model 2 - Transmission line model with via-plate capacitance



(a)Illustration of four ports between two parallel plates



(a) Illustration of circuit model

(b) Circuit model

Fig. 5. Model 4 - Parallel plates Impedance *Zpp* model

J. Xu et al., "A survey on modeling strategies for high-speed differential Via between two parallel plates," 2017 IEEE International Symposium on Electromagnetic Compatibility & Signal/Power Integrity (EMCSI), Washington, DC, USA, 2017, pp. 527-531.

Takeouts – Localize or Face Uncertainties…

Localization Frequency

Frequency

Localized	Non-Localized
Predictable with analysis in isolation	Requires analysis with PDNs
Any boundary conditions can be used for analysis in isolation	Low-impedance ABC are required for simulation in isolation
Local coupling can be included	Hybrid 2D+3D analysis is required
EASY	DIFFICULT

Localization of any via breaks as frequency grow…

Use Dissipated Power as the localization metric

More on waveguiding approach to via design at *A. Manukovsky, Y. Shlepnev, J, Nutzati, A. Kuntsevych, I. Peleg, S. Mordooch, Via Design for 112 Gbps & Beyond: Theory & Reality, DesignCon 2025, Thursday, January 30, 8:00 AM - 8:45 AM Pacific Time (US & Canada), Ballroom B.*

The Other Ways to Mitigate Crosstalk

- Modal transmission
  - *F. Broydé and E. Clavelier, "An overview of modal transmission schemes," 2013 17th IEEE Workshop on Signal and Power Integrity, Paris, France, 2013.*
- Ensemble None-Return to Zero (ENRZ)
  - *S. S. Chen, Z. Xu, A. Tajalli and B. Holden, "Crosstalk Performance Analysis: ENRZ, NRZ, PAM3, and PAM4," in IEEE Transactions on Signal and Power Integrity, vol. 2, pp. 53-63, 2023.*
- Synchronous transmission and equalization
  - *J. F. Buckwalter and A. Hajimiri, "Cancellation of crosstalk-induced jitter," in IEEE Journal of Solid-State Circuits, vol. 41, no. 3, pp. 621-632.*
  - *K.-J. Sham, "Crosstalk mitigation techniques in high-speed serial links," Ph.D dissertation, Univ. Minnesota, Minneapolis, MN, USA, 2009.*

- **What else?...**

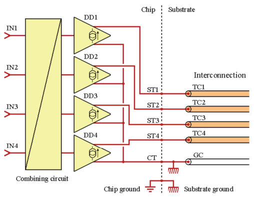


Fig. 2. A pseudo-differential TX-circuit for the ZXtalk method or another modal signaling scheme.

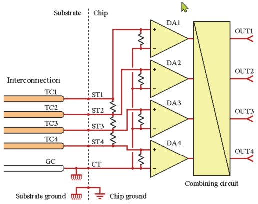


Fig. 3. A pseudo-differential RX-circuit and an on-chip M-type termination circuit for the ZXtalk method.

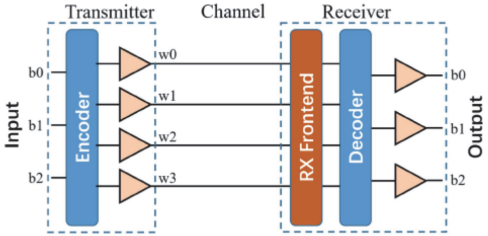


Fig. 1. ENRZ system diagram. b0–b2: bits stream. w0–w3: ENRZ lanes.

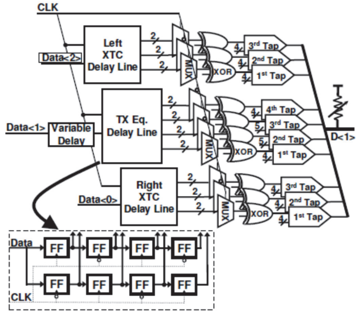


Figure 6.5: Design architecture of the center transmitter with pre-emphasis and XTC

Conclusion

- Crosstalk is a complex and persistent phenomenon that is here to stay
- The primary objective of crosstalk evaluation is to ensure that the design performs at the specified bit error rate
- There is no single universal and accurate method to determine the impact of crosstalk
- This tutorial covers some fundamental techniques for evaluating crosstalk
- Approaches to managing crosstalk include either total localization (overdesign) or system-level analysis (which can be overly complex)

Some References on Crosstalk

- H. Johnson, M. Graham, High-Speed Digital Design – A Handbook of Black Magic, 1993
- J.A.B. Faria, Multiconductor Transmission-Line Structures: Modal Analysis Techniques, 1993
- C.R. Paul, Analysis of Multiconductor Transmission Lines, 1994
- F. Olyslager, Electromagnetic Waves and Transmission Lines, 1999
- B. Young, Digital Signal Integrity – Modeling and Simulation with Interconnects and Packages, 2000
- S.H., Hall, G.W. Hall, J.A. McCall, High-Speed Digital System Design – A Handbook of Interconnect Theory and Design Practices, 2000
- E. Bogatin, Signal and Power Integrity – Simplified, 2004, 2010
- S.H. Hall, Advanced Signal Integrity for High-Speed Digital Designs, 2009
- S.C. Thierauf, Understanding Signal Integrity, 2011
- F. Broyde & E. Clavelier, Tutorial on Echo and Crosstalk in Printed Circuit Boards and Multi-Chip Modules – Lecture Slides, 2012

Thank you!

QUESTIONS?

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