

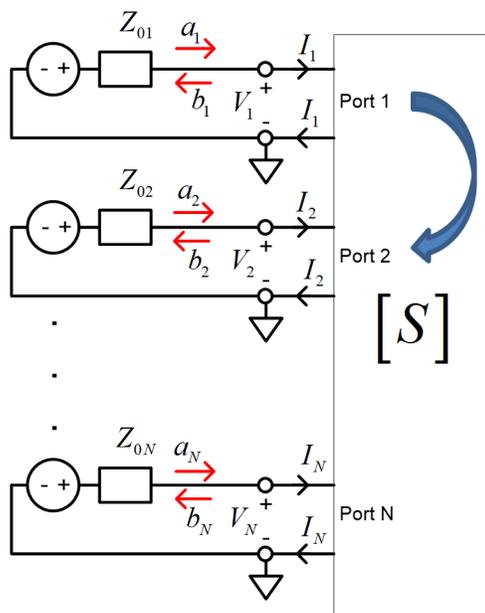
## How Interconnects Work: Reflections from Discontinuities

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An ideal digital interconnect is a lossless transmission line with characteristic impedance and phase delay flat over the signal bandwidth and termination resistors equal to the characteristic impedance. In such interconnect bits generated by transmitter would flow seamlessly into the receiver with no limits on the bit rate. Such utopian transmission line exists only in our imagination and in textbooks as well. The physics of our world prohibits that. One way to describe “what happens to the signal on the way to a receiver” is to use the **balance of power** that can be written for the passive interconnect as follows:

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

This is in frequency domain over the bandwidth of the signal as defined in [1].  $P_{out}$  is the power delivered to the receiver and  $P_{in}$  is the power delivered by transmitter to the interconnect. All other terms in the balance of power equation describe the signal distortion. The formula above expresses all we need to know about the interconnects (it should be “casted in granite” ☺). As they say, “a formula is worth a thousand words”, almost literally in this case. To understand it, just imagine the interconnect system as a multiport with the transmitter at port 1, receiver at port 2 and multiple other ports for links coupled to the link connecting port 1 and 2 and terminations to real impedance (not necessary identical at all ports). Something like shown below together with the definition of waves and scattering parameters (or S-parameters):



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

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

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

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

Scattering parameters:

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

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

The Harrington’s concept of “loaded scatterers” allows us to turn any interconnect system (even malfunctioning as an antenna) into a multiport and describe it with the waves and scattering parameters as it is done on the picture above. Also choice of real  $Z_0$  for all ports makes those waves the “power waves”. Now we can define the powers in the balance of power equation through the multiport waves  $a$  and  $b$  as follows:

$P_{in}$  is power delivered into the interconnect by transmitter:  $P_{in} = |a_1|^2 [Wt]$ ,  $|a_1|$  is magnitude of the incident wave at the transmitter end or port 1;

**P<sub>out</sub>** is power delivered to the receiver:  $P_{out} = |b_2|^2 [Wt]$ ,  $|b_2|$  is magnitude of the transmitted wave at the receiver end port 2 assuming that  $|a_1| \neq 0$  and no incident waves on the other ports  $|a_k| = 0$  for all  $k \neq 1$ ;

**P<sub>absorbed</sub>** is power absorbed or dissipated by dielectrics and conductors and possibly by absorbing boundary conditions in interconnect models (more on that is below);

**P<sub>reflected</sub>** is power reflected back to the transmitter:  $P_{reflected} = |b_1|^2 [Wt]$ ,  $|b_1|$  is magnitude of reflected wave at the transmitter end, assuming that no incident waves on the other ports  $|a_k| = 0$  for all  $k \neq 1$ ;

**P<sub>leaked</sub>** is power leaked into the other coupled interconnects, modes and, possibly, into power distribution network (PDN – another type of interconnects) and into free space (radiated):  $P_{leaked} = \sum(|b_k|^2 [Wt], k \neq 1,2; |b_k|$  are waves transmitted to the ports of the coupled interconnects,  $|a_k| = 0$  for all  $k \neq 1$ ;

**P<sub>coupled</sub>** is power gained from the other coupled interconnects, modes, PDN and free space:  $P_{coupled} = |b_1|^2 + |b_2|^2 [Wt]$ ,  $a_1 = 0, a_2 = 0, a_k \neq 0$  for all other  $k$ ;

Here is a summary for definition of elements in the interconnect balance of power:

$$P_{in} = |a_1|^2, a_2 = 0$$

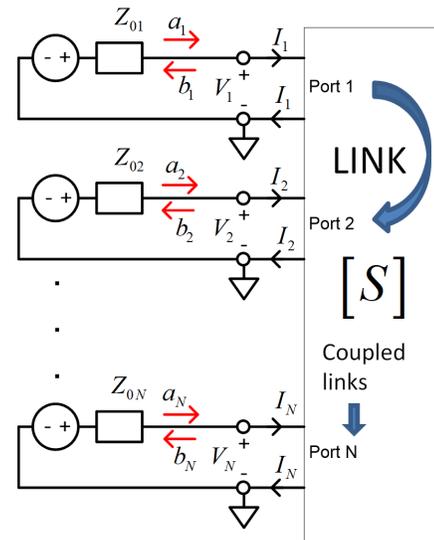
$$P_{out} = |b_2|^2, a_1 \neq 0; a_k = 0, k \neq 1$$

$$P_{reflected} = |b_1|^2, a_1 \neq 0; a_k = 0, k \neq 1$$

$$P_{leaked} = \sum_{k \neq 1,2} |b_k|^2, a_1 \neq 0; a_k = 0, k \neq 1$$

$$P_{coupled} = \sum_{k=1,2} |b_k|^2, a_{1,2} = 0; a_k \neq 0, k \neq 1,2$$

$$P_{absorbed} = |a_1|^2 - \sum_k |b_k|^2, a_1 \neq 0; a_k = 0, k \neq 1$$



Let's examine those terms further. First, let's assume that the link is localized somehow – no coupling or  $P_{leaked} = P_{coupled} = 0$  (the coupling will be a subject for another article). It is hard to do it for the real life “open waveguiding systems” such as PCB or packaging interconnects, but is close to reality for properly designed interconnect systems. The remaining balance of power is as follows:

$$P_{out} = P_{in} - P_{absorbed} - P_{reflected}$$

What can we learn from this simple expression? **P<sub>out</sub> characterizes the transmitted signal and includes signal distortions from absorption or dissipation [2] as well as from all kinds of reflections.** This is important to know. If more power is reflected or absorbed then less power is delivered to the receiver – that means more signal degradation or distortion. It is a little more complicated with the complex termination impedances. So let's stay with the real termination impedances – it is not a simplification or approximation because of all reactive parts of the termination impedance can be easily embedded into S-parameters – the “black box” can take everything in.

**P<sub>out</sub> can be expressed as Insertion Loss (IL, positive number) in dB as follows:**

$$IL = 10 \cdot \log(P_{in}/P_{out}) [dB]$$

This is by definition. We can rewrite it with the magnitudes of the multiport waves as follows:

$IL = 10 \cdot \log(|a_1|^2/|b_2|^2) = 20 \cdot \log(|a_1|/|b_2|) = -20 \cdot \log(|b_2|/|a_1|) = -20 \cdot \log(|S_{21}|)$   
 S21 is the transmission parameter – transmission from port 1 to port 2. **The insertion loss is just negative magnitude of the transmission parameter S21 in dB.**

**P\_absorbed characterizes all power losses to heat the materials** (inevitable contribution to the increase of entropy) or dissipated by absorbing boundary conditions in interconnect models. It can be expressed as follows (no coupling):

$$P_{\text{absorbed}} = P_{\text{in}} - P_{\text{out}} - P_{\text{reflected}} = |a_1|^2 - |b_2|^2 - |b_1|^2 \text{ [Wt]}$$

This is useful formula in case if you want to evaluate the absorbed or leaked (if leaks are modeled with the absorbing boundary conditions for instance).

**P\_reflected characterizes just the reflections** and can be expressed as the Return Loss (RL, positive number) in dB as follows:

$$RL = 10 \cdot \log(P_{\text{in}}/P_{\text{reflected}}) \text{ [dB]}$$

Again, this is by definition. We can also rewrite it with the magnitudes of the multiport waves as follows:

$$RL = 10 \cdot \log(|a_1|^2/|b_1|^2) = 20 \cdot \log(|a_1|/|b_1|) = -20 \cdot \log(|b_1|/|a_1|) - 20 \cdot \log(|S_{11}|)$$

Here S11 is the reflection parameter – reflection at port 1. **The return loss is just negative magnitude of the reflection parameter S11 in dB.**

Here are the expressions for insertion and reflection loss through power:

$$\frac{P_{\text{out}}}{P_{\text{in}}} = |S_{2,1}|^2 \Rightarrow IL = 10 \cdot \log\left(\frac{P_{\text{in}}}{P_{\text{out}}}\right) = -20 \cdot \log|S_{2,1}|$$

$$\frac{P_{\text{reflected}}}{P_{\text{in}}} = |S_{1,1}|^2 \Rightarrow RL = 10 \cdot \log\left(\frac{P_{\text{in}}}{P_{\text{reflected}}}\right) = -20 \cdot \log|S_{1,1}|$$

The absorption or dissipation losses in dielectrics and conductors were recently discussed in [2]. Such losses are inevitable, but can be effectively mitigated at the stackup planning stage – selection of dielectric and conductor materials and stackup geometry defines the maximal possible communication distance for a particular data rate and power required to transmit the signal.

Considering the reflections, they can be further separated into the following 3 categories:

1. Reflections from mismatch of transmission line impedance and terminations [3];
2. Reflections from single discontinuities – transitions, vias, AC caps, length compensation structures, gaps in reference plane,...;
3. Reflections from periodic discontinuities – fencing vias and cut outs, fiber-weave effect,...;

Item 1 was discussed in [3]. The most important thing for the reflection understanding from that is the formula for reflection parameter from a segment of transmission line with complex propagation constant Gamma, characteristic impedance Zc and length l terminated with Zo:

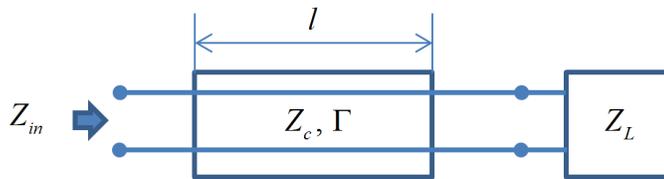
$$S_{1,1} = (Z_c^2 - Z_0^2) / (Z_c^2 + Z_0^2 + 2 \cdot Z_c \cdot Z_0 \cdot \text{cth}(\Gamma \cdot l))$$

The reflection reaches the maximal values at frequencies where the transmission line length is the quarter of wavelength in line (Lambda) or Lambda/4+n\*Lambda/2 in line (see definitions in [3]):

$$\max(S_{1,1}) = (Z_c^2 - Z_0^2) / (Z_c^2 + Z_0^2 + 2 \cdot Z_c \cdot Z_0 \cdot \text{th}(\alpha \cdot l))$$

$$\max(S_{1,1}) = (Z_c^2 - Z_0^2) / (Z_c^2 + Z_0^2), \text{ if } \alpha = 0$$

Here  $\alpha$  is the attenuation constant that increases with the frequency [2]. This is another remarkable formula – the reflections are smaller for lines with more losses (greater  $\alpha$ ) and at higher frequencies [3]. The last part is valid only if there are no discontinuities. **It makes everything perfectly clear with the choice of characteristic impedance – the closer  $Z_c$  to the termination impedance  $Z_0$ , the lower the reflections.** A link with  $Z_c=Z_0$  would be the best link, if it would be possible to design and build it like that. What prevents us from doing it are all kinds of transitions. In reality just some parts of interconnects look like and can be modeled as the transmission lines. There are also transitions between the transmission lines – packages, connectors, viaholes, length compensation structures, AC coupling capacitors, transitions between the transmission lines of different types or with different cross-sections. **Accurate analysis of a complete link with the discontinuities usually requires validated software with 3D electromagnetic analysis capabilities, such as Simbeor electromagnetic signal integrity software (available at [www.simberian.com](http://www.simberian.com)).** Before providing examples of such numerical analysis, let's write a couple of practical formulas for qualitative and quantitative analysis of the reflections from the discontinuities. The first formula is for the input impedance of a transmission line segment loaded with  $Z_L$  impedance as follows:



$$Z_{in} = Z_c \frac{Z_L + Z_c \cdot th(\Gamma \cdot l)}{Z_c + Z_L \cdot th(\Gamma \cdot l)}$$

It can be easily derived starting with the exponential expansion [3] by setting the boundary conditions on the right side. **The formula for the impedance can be used recursively starting from the receiver by including discontinuities in a transmission line that have expressions for the impedance – capacitances, inductances or combinations, stubs and so on.** All you need to know is how to connect the complex impedances in parallel and in series. As soon as we know the input impedance of the link at the transmitter side, the reflection parameter can be computed as follows:

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

Note that this is exactly the same value as the reflection coefficient that is usually designated with Gamma [4]. There are no differences in this context. The formula can be also inverted, to compute  $Z_{in}$  from the reflection parameter.

Now, let's examine a few cases. First, if  $Z_L = Z_c$  then  $Z_{in} = Z_c$  and the reflection parameter is  $S_{11}=(Z_c-Z_0)/(Z_c+Z_0)$  – this is reflection from the ideally terminated transmission line segment or from the infinite transmission line. It correlates well with the  $S_{11}$  provided earlier for the transmission line segment with the infinite length ( $cth$  is the unit in this case). Another limit case is  $Z_L=infinity$  – this is open-circuited segment of transmission line. In this case  $Z_{in}$  is particularly simple:

$$Z_{in} = Z_c \cdot cth(\Gamma \cdot l)$$

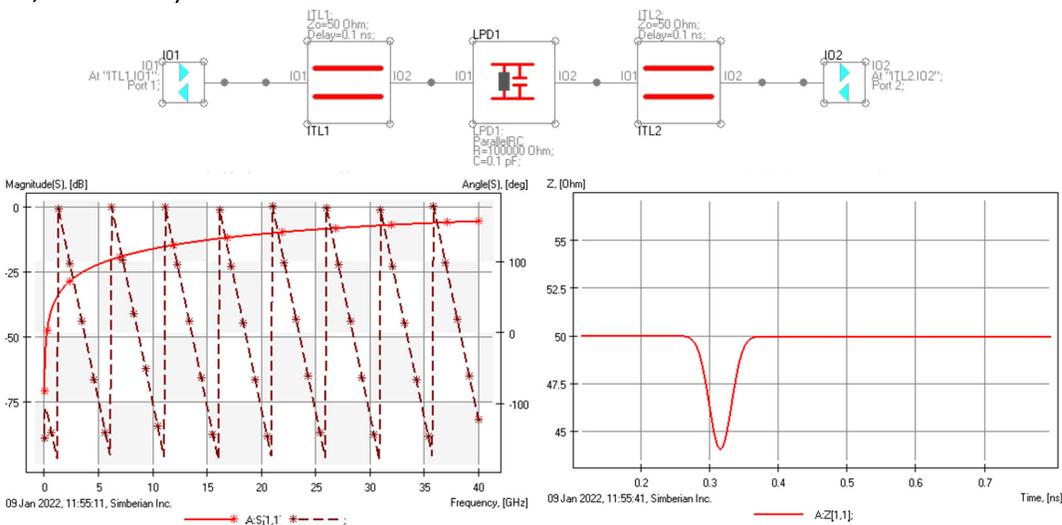
What can we learn from that in the interconnects context? It can be used to model and explain the effect of stubs. The first order approximation for a via-hole discontinuity is a transmission line model. If via is going through all stackup layers, but connects only one surface layer with an interior layer for instance, the remaining part of the via is a stub connected in parallel to the link. If stub length is a

quarter of wavelength in corresponding equivalent transmission line model we have input impedance of the stub as follows:

$$Z_{in} = Z_c \cdot th(\alpha \cdot l)$$

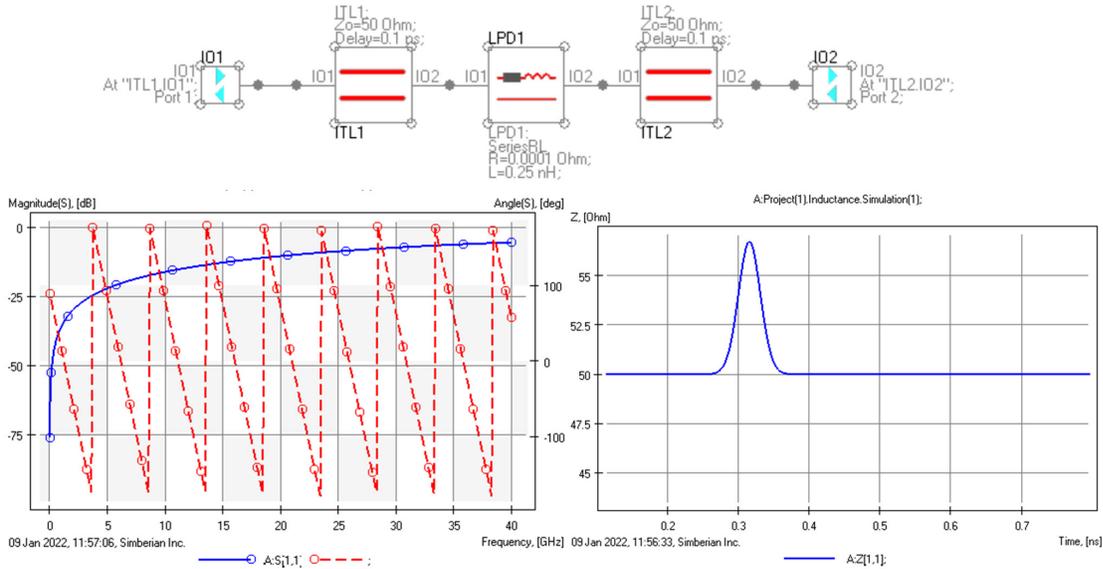
It is zero if alpha or attenuation is zero – than means it short-circuits the link. With zero  $Z_{in}$ , the reflection is  $S_{11}=-1$  – it does not matter what are the other parts of the link look like. Never let it happen at the frequencies close the Nyquist frequency! So, it is all the transmission line trigonometry – though, hyperbolic in this case, literally.

After having some fun with the formulas and the limit cases, let’s analyze a couple of ideal discontinuities first. The analysis can be done with the formulas as well (a good exercise), but it is much easier to do it with the software – Simbeor in this case. A 0.1 pF capacitor connected in parallel into segment of ideal 50 Ohm transmission line (yes, ideal transmission lines exist in the software as well) will have magnitude of reflection (left graph, left axis, red line) and phase of reflection (left graph, right axis, brown line) as shown below:



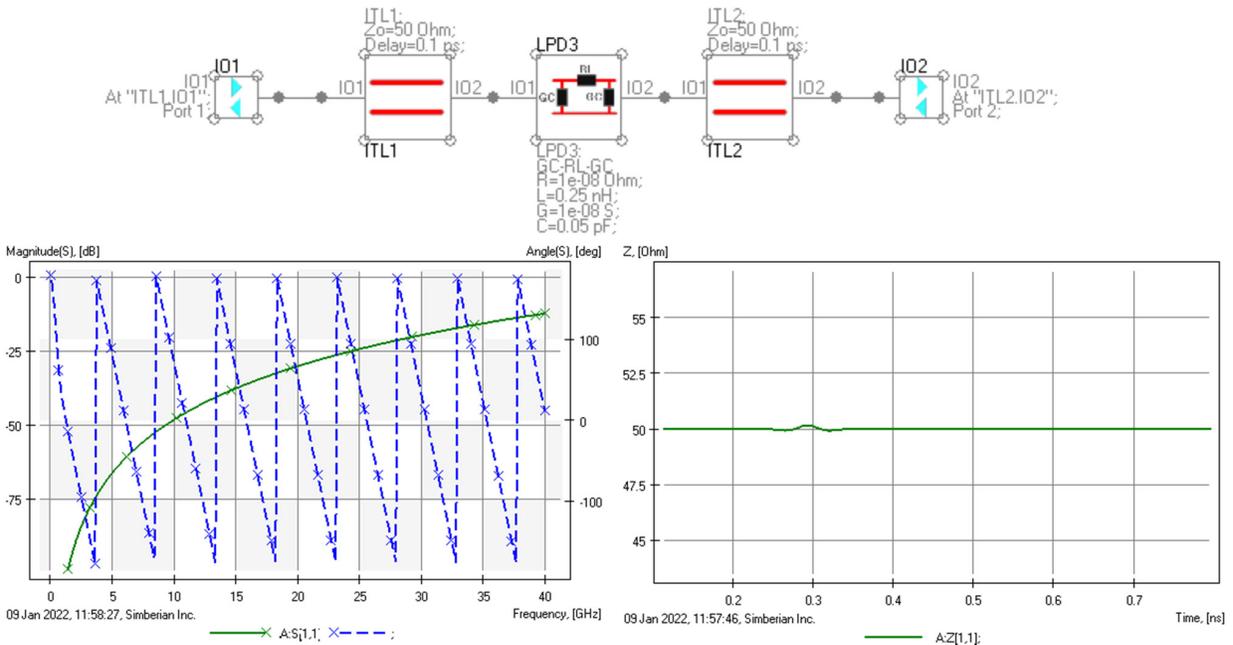
Time-domain reflectometry (TDR) is also shown on the right plot. It is much easier to recognize the capacitance on the TDR plot as the “capacitive” dip. The TDR is computed here directly from the frequency-domain response with 40 ps Gaussian step function.

Next, let’s take a look at 0.25 nH inductance connected in series into the same ideal 50 Ohm transmission line segment shown below:



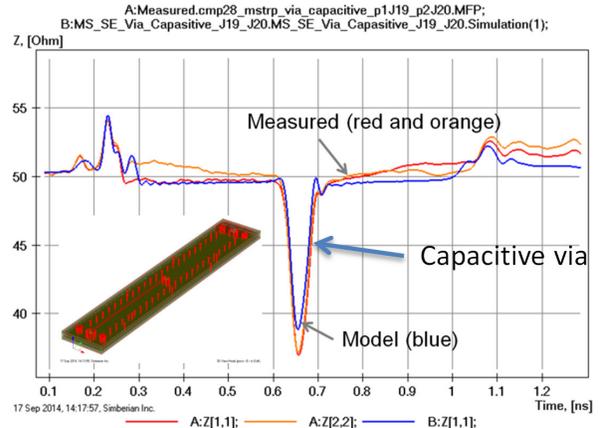
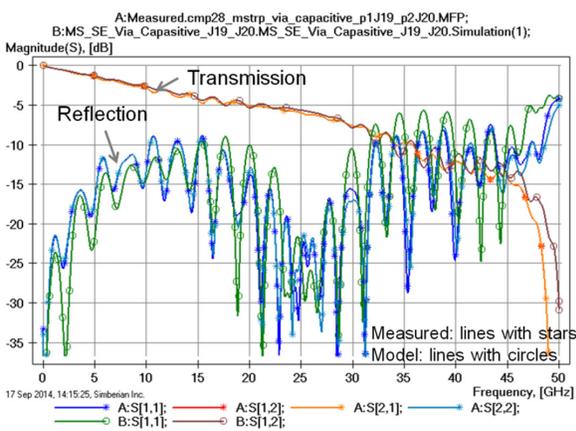
The frequency-domain reflection magnitude shown in blue on the left graph is exactly the same as the reflection from the capacitance (the values are intentionally selected to have it like that). However, the phase is different (though difficult to conclude something from it) and the TDR shows clear “inductive” bump.

What if we combine 0.1 pF capacitance and 0.25 nH inductance as C/2-L-C/2 circuit – the reflection will go down as shown below:



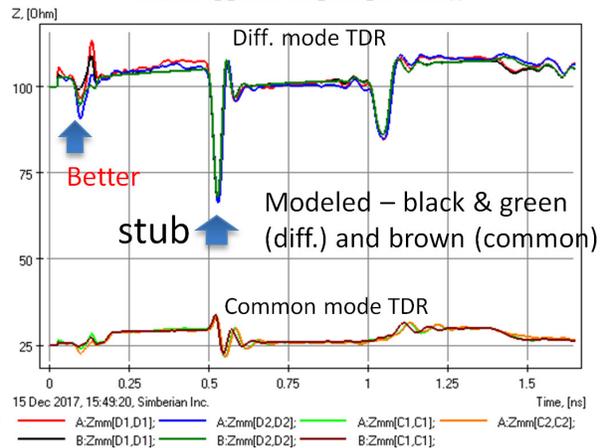
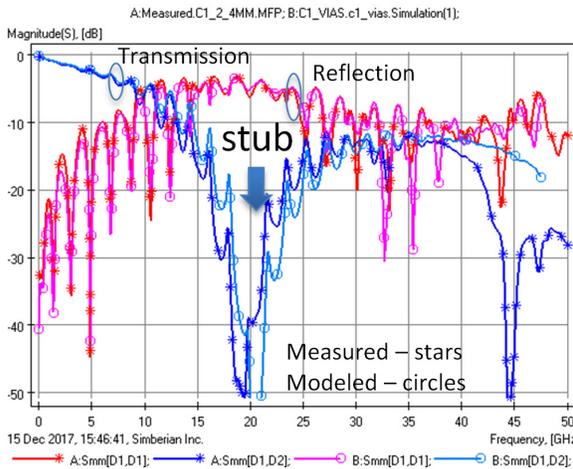
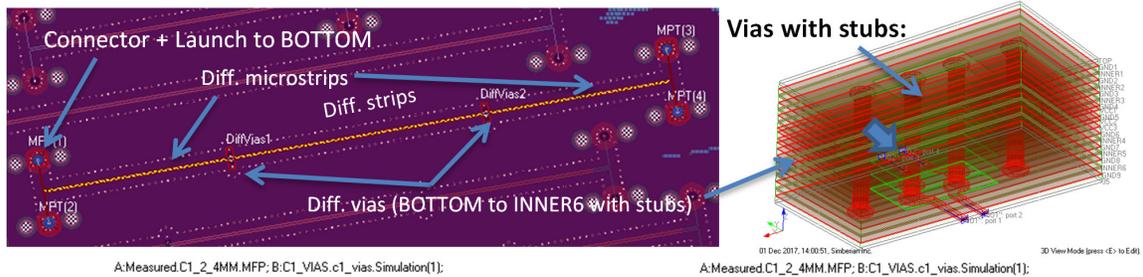
That actually explains how the transmission line works – t-line has both capacitance and inductance with the ratio equal to the characteristic impedance -  $Z_0 = \sqrt{L/C} = 50 \text{ Ohm}$ , as in this case. Though, the reflections in the frequency domain are small, but not zero. This is because of the model uses lumped L and C in the middle of continuous transmission line segment and it behaves as the t-line only at lower frequencies. Though, TDR with 40 ps rise time is almost flat and does not resolve such lumped discontinuity in this case.

Finally, a few practical examples of the electromagnetic discontinuity analysis with validation used in our DesignCon 2020 tutorial [5]. The first example is the capacitive via from CMP-28 validation platform designed by Wild River Technology.



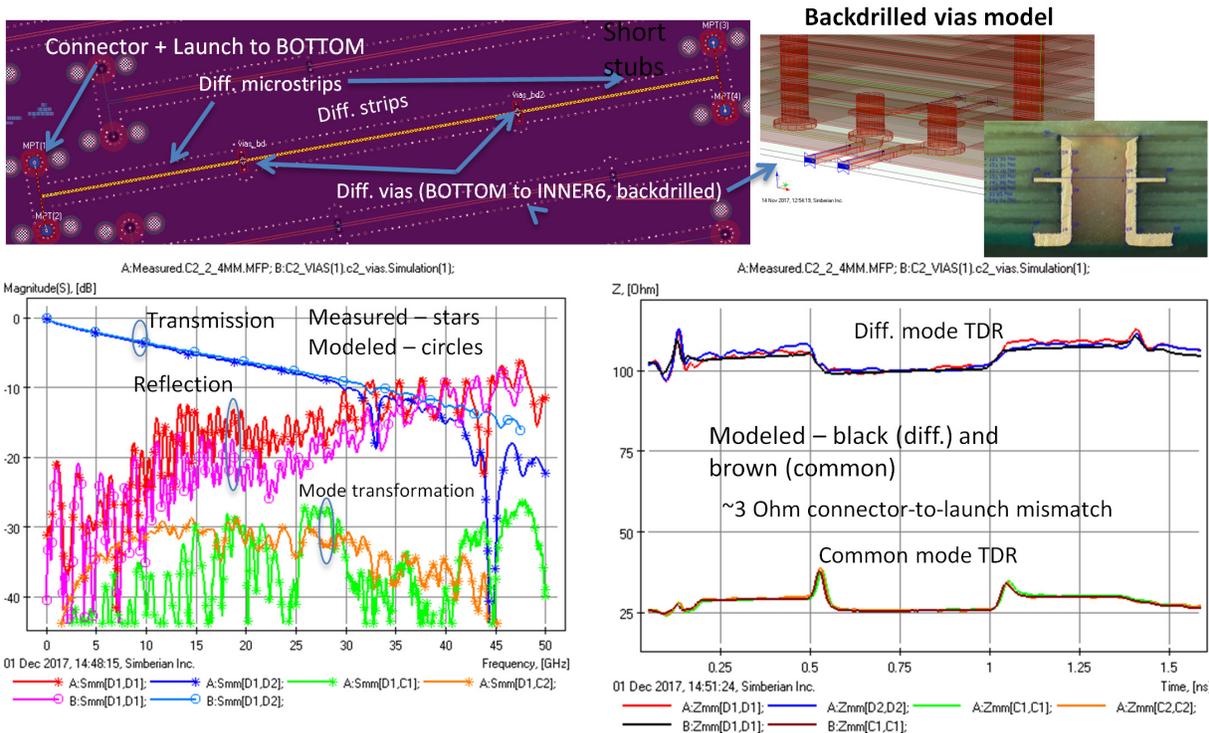
Modeled and measured magnitudes of the transmission and reflection parameters are shown on the left graph and corresponding TDRs is on the right. The link includes coaxial connectors and launches (inductive bumps at the beginning and end). The reflections in this case are not show stopper – the link is short and the added capacitance is not so big. It was intentionally capacitive – see more analysis to measurement correlation examples from CMP-28 platform at [6].

Another two practical examples with the analysis to measurement correlation are from EvR-1 project [7] (see also app notes #2018\_01, 2018\_07 and webinar #8 at [www.simberian.com](http://www.simberian.com)). Below are the modeled and measured magnitudes of the differential reflection and transmission parameters and differential and common-mode TDRs for a differential link with 2 vias from BOTTOM to INNER6 layers with stubs (structure EvR1-C1):



The via stubs almost short-circuited the link at about 20 GHz – the reflection parameter (reddish lines with circles) went up and the transmission parameters (bluish lines with stars) went down around the stub resonance frequency. With the transmission below -30 dB, the signal harmonics at those frequencies will not go through. Corresponding differential TDR shows big capacitive dip at the via location – though, it is difficult to conclude from TDR at which frequency the via would destroy the signal. **Notice that the correlation between the model and measurements is as good as it usually gets – this is because of the “sink or swim” approach used in [7].**

What can be done to avoid such harmful reflections? Via back-drilling and optimization – below is structure EvR1-C2 with the back-drilled and optimized viaholes for the transition between the same layers:



The reflection is below -10 dB up to 30 GHz and no bumps at the location of the vias – the link is as good as it can be for all practical purpose. Note that the correlation of smaller reflections is more difficult to achieve because of the manufacturing variations that was also investigated for this particular board in [7].

**So the bottom line is the reflections even from a single discontinuity can destroy your signal.**

Though, the effect of the major discontinuities can be predicted with validated models before the board designed (pre-layout analysis) and must be predicted after the board is designed and before it goes into manufacturing (post-layout analysis). Examples of discontinuity analysis and optimization for typical PCB links with Simbeor electromagnetic signal integrity software are provided at the demo-video section of Simberian web site <https://www.simberian.com/ScreenCasts.php?view=list>

**REFERENCES:**

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6. [Guide to CMP-28/32 Simbeor Kit, CMP-28 Rev. 4, Sept. 2014.](#)
7. [M. Marin, Y. Shlepnev, 40 GHz PCB Interconnect Validation: Expectation vs. Reality](#), DesignCon2018, January 31, 2018, Santa Clara, CA.

**Appendix:** Interconnect balance of power expressed through S-parameters:

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

$$P_{in} = |a_1|^2, a_2 = 0$$

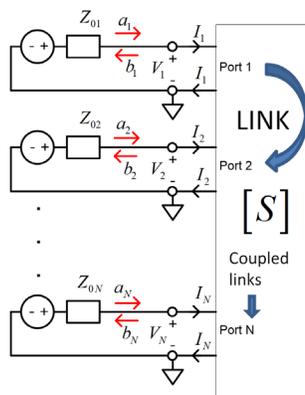
$$P_{out} = |b_2|^2, a_1 \neq 0; a_k = 0, k \neq 1$$

$$P_{reflected} = |b_1|^2, a_1 \neq 0; a_k = 0, k \neq 1$$

$$P_{leaked} = \sum_{k \neq 1,2} |b_k|^2, a_1 \neq 0; a_k = 0, k \neq 1$$

$$P_{coupled} = \sum_{k=1,2} |b_k|^2, a_{1,2} = 0; a_k \neq 0, k \neq 1,2$$

$$P_{absorbed} = |a_1|^2 - \sum_k |b_k|^2, a_1 \neq 0; a_k = 0, k \neq 1$$



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

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

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

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

Scattering parameters:

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

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

**Insertion Loss, Return Loss, Coupling and Leaks:**

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

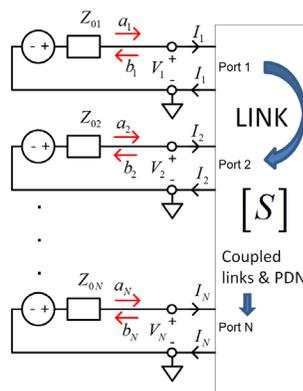
$$\frac{P_{out}}{P_{in}} = |S_{2,1}|^2 \Rightarrow IL = 10 \cdot \log\left(\frac{P_{in}}{P_{out}}\right) = -20 \cdot \log|S_{2,1}|$$

$$\frac{P_{reflected}}{P_{in}} = |S_{1,1}|^2 \Rightarrow RL = 10 \cdot \log\left(\frac{P_{in}}{P_{reflected}}\right) = -20 \cdot \log|S_{1,1}|$$

$$\frac{P_{coupled}}{P_{in}} = \sum_{k \neq 1,2} |S_{1,k}|^2 + \sum_{k \neq 1,2} |S_{2,k}|^2 \quad \frac{P_{leaked}}{P_{in}} = \sum_{k \neq 1,2} |S_{k,1}|^2$$

$$PSXT_2 \leftarrow -10 \cdot \log\left(\frac{P_{coupled}}{P_{in}}\right) = -10 \cdot \log\left(\sum_{k \neq 1,2} |S_{2,k}|^2\right)$$

$$P_{absorbed} = 1 - \sum_k |S_{k,1}|^2$$



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

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

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

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

Scattering parameters:

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

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