How Interconnects Work: Anatomy of Crosstalk

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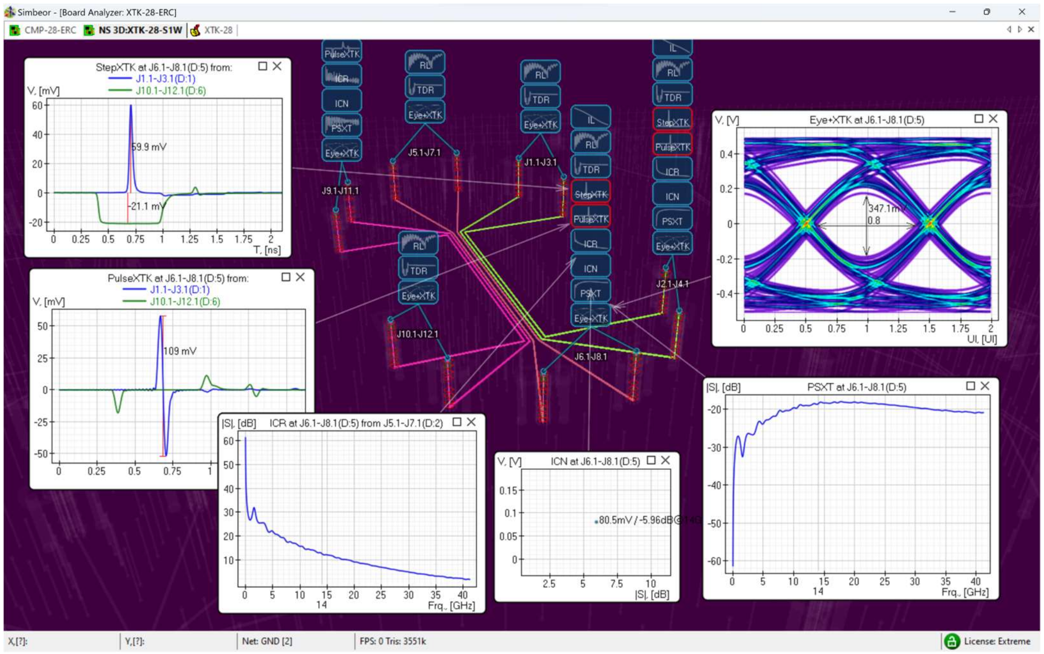


Crosstalk in PCB and packaging interconnects is arguably one of the most complicated phenomena that may cause signal degradation. It is caused by unwanted coupling between signal links and between signal links and power distribution system. The effect is deterministic, but very difficult to predict in many cases – too many variables and uncertainties. Crosstalk effects can be treated statistically as a deterministic jitter with a bounded distribution, but the distribution is usually not known and just guessed. A direct analysis of a worst-case crosstalk scenario may lead to a system overdesign. Neglecting it in design may cause a system failure that is difficult to find and fix later in a design process. On top of that, distortions caused by crosstalk cannot be corrected by signal conditioning techniques at a receiver side. **Thus, it is very important to understand the sources of crosstalk, how to quantify it and how to mitigate it efficiently.** This is the first part of the paper with an overview of crosstalk sources and terminology – just a slice through the complicated phenomenon. The second part will describe and compare different ways to quantify, compute, and measure crosstalk. This paper continues the "How Interconnects Work" series [1]-[4].

Crosstalk In the Balance of Power

The best way to describe "what happens to a signal on the way to the receiver" is to use the balance of power that can be written for a passive interconnect as follows [4]:

P_out = P_in - P_absorbed - P_reflected - P_leaked + P_coupled

This is applicable to both the time domain and the frequency domain over the bandwidth of a signal as defined in [1]. **P_in** is the power delivered by a transmitter to the interconnect (useful signal) and **P_out** is the power delivered to a receiver (degraded useful signal + noise). All other terms in the balance of power equation describe the signal distortion. The absorption and reflection terms (**P_absorbed** and

P_reflected) were discussed in the previous papers of the "How Interconnects Work" series [2]-[4]. This paper is about **P_leaked** and **P_coupled** or the crosstalk parameters:

P_leaked is a power leaked into other coupled interconnects, into the common mode and, possibly, into power distribution network (PDN is just another type of interconnects) and into free space (radiated) – that leak causes signal distortion and is a possible source of crosstalk in addition to being source of EMC/EMI;

P_coupled is a power gained from the other coupled interconnects, common mode, PDN and free space – this is the crosstalk;

The crosstalk in general is just unwanted noise from the coupled structures (**P_coupled**) caused by unwanted signal leaks (**P_leaked**) that degrade the useful signal and may reduce the data transmission rate and even cause complete link failure.

Crosstalk Types

Unwanted coupling in PCB and packaging interconnects can be separated into **a local and a distant coupling**:

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

The local couplings can be accurately simulated in general and taken into account during pre- and post-layout signal integrity analysis. Simbeor SDK provides multiple kits for the pre-layout solution space exploration that takes into account crosstalk. Simbeor SI Compliance Analyzer provides unique capabilities for the fast and easy post-layout analysis of crosstalk with possibility of the analysis and compliance verification automation.

Coupling in parallel traces can cause not only crosstalk and interference (unwanted noise), but also additional losses due to signal energy leaks to adjacent links (suck outs) – this is **P_leaked** term in the balance of power. The leak losses may be significant in traces routed on the surface of PCB (microstrips). Though, they are usually negligible for traces routed between parallel planes (striplines).

The distant coupling may cause a system-level interference and requires complete PCB or package analysis. The distant system-level coupling is very difficult to model and predict with sufficient accuracy. Though it can be either avoided (no rouging over split planes) or easily reduced by enforcing the localization for each structure that may potentially be coupled to parallel planes (transmission planes), surface dielectric layers, or to enclosure. Such coupling occurs at locations of changes in reference conductors. Un-localized viaholes is the major source of leaks and crosstalk that can be easily avoided with use of more stitching vias closer to signal vias for instance. **The system-level interference must be avoided by use of only the structures that are predictably localized up to the target frequency (conditional localization).** Simbeor SI Compliance Analyzer has unique capability to verify the localization and find sources of potential system-level coupling during the reference integrity analysis.

Crosstalk Origin

PCB and packaging interconnects, such as striplines, microstrip lines and coplanar waveguides, are open waveguiding structures. **It means that the signal energy propagates along the PCB and packaging traces mostly in dielectrics around the signal conductors**. It can be illustrated with the peak power flow density (vector product of electric and magnetic field) for a typical PCB stripline interconnect shown in Fig. 1. This is the peak power flow density (PDF) of a signal with 0.5 V magnitude normalized to maximal value and expressed in dB as 10*log|P|.

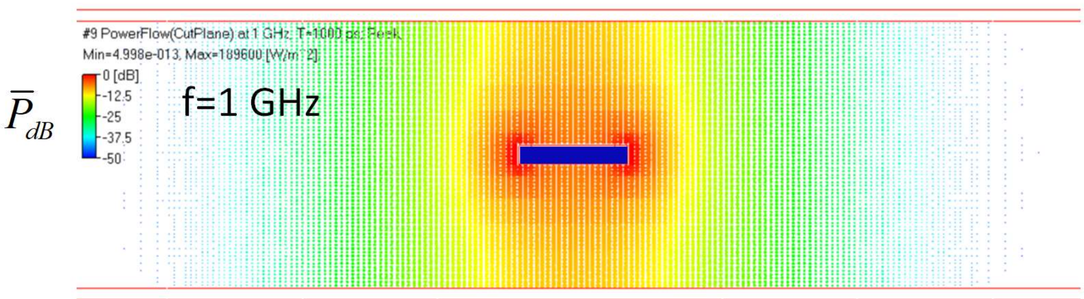


Fig. 1. Power flow density in stripline 1.2 mil thick, 7 mil wide, DK=3.76, LT = 0.006 @ 1 GHz, planes 0.77 mil thick, 17.2 mil apart, color scale is used to plot peak power flow density in W/m^2, computed with Simbeor THz.

As you can see, there is no exact localization of the signal energy. The power flow density or signal energy concentrates near the strip edges and between the strip and planes. Red and yellow area is where the most of the signal energy propagates. But it is also non-negligible in in the green area of 2-3 strip widths in this case. **Everything that gets into zone with green PFD (-25 to -30 dB level, 2-3 width of strip on both sides) become coupled and that coupling may cause interference or crosstalk and signal leaks.** In addition, the coupling changes the strip impedance. The picture in Fig.1 is for the dominant strip line mode that has equipotential reference planes.

The signal energy spreading around a signal conductor is even worse in traces routed on a surface of PCB (microstrips) as we can see from the Fig. 2.

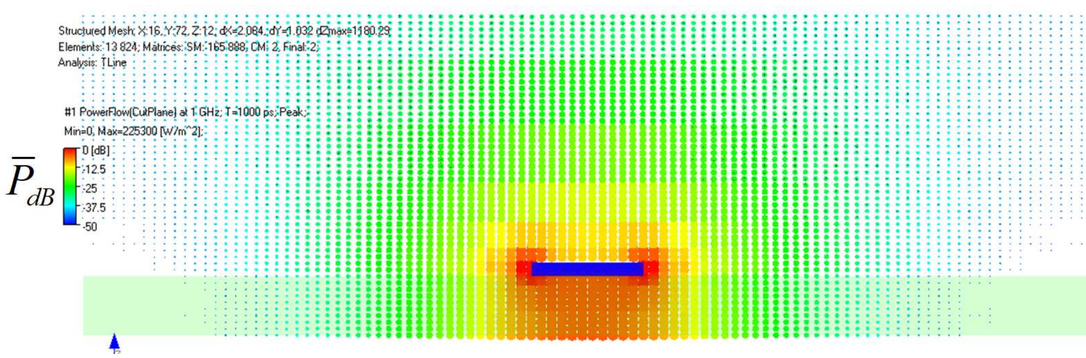


Fig. 2. Peak values of power flow density at 1 GHz in microstrip line with 8.256 mil wide trace on 4.5 mil FR4 substrate.

As we can see, the area of potential coupling is larger in the air as well as in the substrate – the trace can be effectively coupled to nearby traces as well as to any external objects that are in the area with green PFD (-25 to -30 dB).

Similar or even worse extension of the coupling area can be observed in unsymmetrical strip lines as illustrated in Fig. 3.

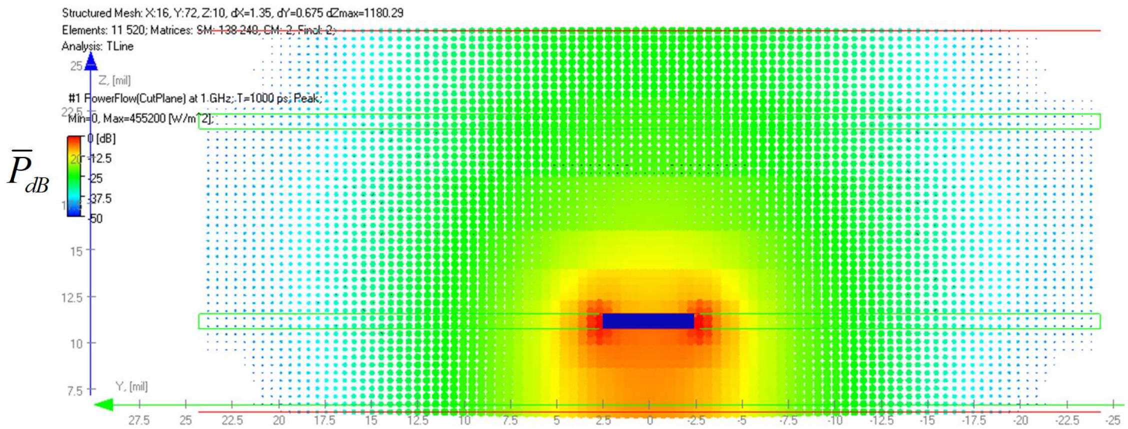


Fig. 3. Peak values of power flow density at 1 GHz in unsymmetric strip line with *5.4 mil trace, distance from strip to top plane 9.77 mil, distance to bottom plane 4.5 mil, DK = 4.2 (~50 Ohm).*

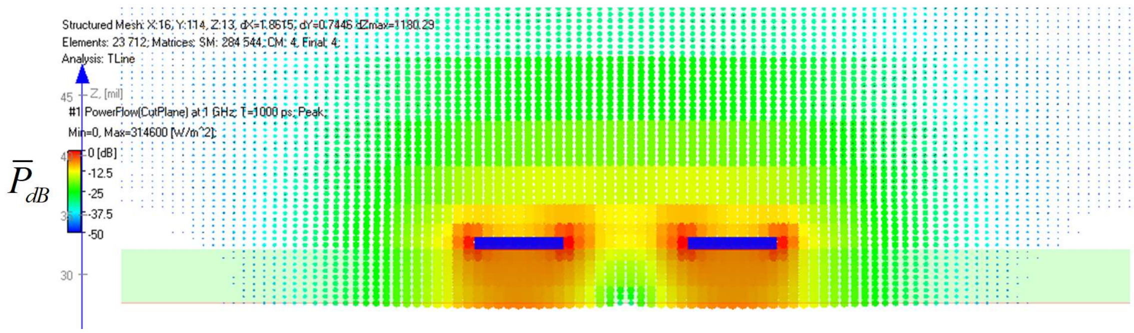


Fig. 4. Peak values of power flow density at 1 GHz of differential mode in differential microstrip line with 7.446 mil traces on *4.5 mil on substrate DK=4.1 (~100 Ohm).*

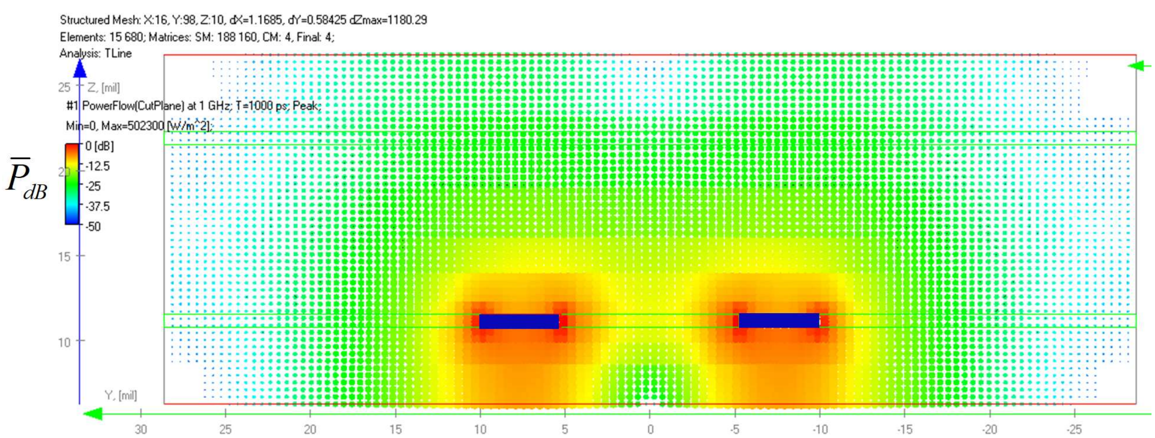


Fig. 5. Peak values of power flow density at 1 GHz of differential mode in unsymmetrical stripline with 4.674 mil wide strips, distance from strip to the top plane 9.77 mil, to the bottom plane 4.5 mil, DK = 4.2 (~100 Ohm).

Differential mode in loosely coupled differential traces has signal energy spread similar to the single-ended case as illustrated by the peak power flow density for a typical differential microstrip in Fig. 4. Peak PFD for a differential mode in a differential stripline with unsymmetrical reference planes is shown in Fig. 5. As we can see, the distant reference planes may substantially increase the area where strips can be coupled even in the differential cases. The PFDs in Fig. 4 and 5 are for the differential modes with excitation +0.5/-0.5 V. The power of the differential mode flows around each trace in the same direction mostly along the traces.

If you still think in terms of currents, please give up 😊. The currents at microwave frequencies are not “flowing” and not “returning” anywhere – they are just a part of the wave propagation process and conductor energy absorption. The power flow density is the best way to visualize the physics of a signal propagation. Though, the surface currents in the reference conductors can be also used to evaluate possible coupling areas, but it is not so intuitive and obvious as with the PFD. In general, it is important to understand that a **single trace is a 2-conductor transmission line or waveguiding structure – the second conductor is always the reference plane (microstrip) or two planes (stripline)**. Differential traces are a 3-conductor transmission line and the currents in the reference conductor are also spreading beyond the traces similar to the single-ended cases. Reference conductors or planes in the signal energy propagation area are as important, as the traces themselves. In a case of two reference planes, the equipotentiality of the planes along the signal propagation must be ensured with more stitching vias and even via fences, to avoid coupling to the dominant mode of the parallel plane structures - such coupling may occur at discontinuities such as viaholes and at dielectric inhomogeneities. Enforcement of the reference equipotentiality for coplanar transmission lines is even more complicated.

The bottom line is that the PCB and packaging interconnects are the open waveguiding structures with the signals propagating in space around the signal traces. Getting a signal with spectrum in microwave and millimeter-wave bandwidth from one component to another through an open waveguiding structure and without interaction with other signals is always a challenge. **See more on different signal couplings cases in “How Interconnects Work™ …” demo-videos at ether Simberian web site** https://www.simberian.com/ScreenCasts.php?view=table**or @simbeor YouTube channel**

- #2019_11: **How Interconnects Work™**: Crosstalk in microstrip lines and how to reduce it (use of tabbed lines), 12 min YouTube
- #2019_10: **How Interconnects Work™**: Where crosstalk may come from - case of coupling between differential striplines and vias, 8 min YouTube
- #2019_09: **How Interconnects Work™**: Where crosstalk may come from - case of stripline coupling through antipads in BGA breakout areas, 12 min YouTube
- #2019_05: **How Interconnects Work™**: Crosstalk in adjacent striplines and how to reduce it - visualization with power flow density, 14 min YouTube
- #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
- #2019_02: **How Interconnects Work™**: Visualization of mode conversion or skew in differential traces with power flow density, 16 min YouTube
- #2017_08: **How Interconnects Work™**: Crosstalk in microstrip traces crossing split planes, 10 min YouTube

- #2016_13: **How Interconnects Work™**: Crosstalk power flow in differential vias, 10 min
- #2016_12: **How Interconnects Work™**: Crosstalk power flow in single-ended vias, 11min
- #2016_11: **How Interconnects Work™**: Crosstalk power flow in microstrip lines, 12min

Useful Crosstalk Terminology

Before proceeding with crosstalk modeling and quantification, let's define some common terms.

- **Stripline** – model for traces routed on the internal layers of PCB/PKG with two reference planes and mostly homogeneous dielectric;
- **Microstrip line** – model for traces routed on surface of PCB/PKG with one reference plane and inhomogeneous dielectric;
- **Coplanar line** – model of traces with additional reference conductors in the same layer with the signal traces;
- **Aggressor** is a transmitter (Tx) or a link with a signal that may cause interference in other links;
- **Victim** is a receiver (Rx) or a link that may pick up unwanted interference caused by an aggressor link;
- **NEXT** – near end crosstalk is interference observed at the link side that is closer to the aggressor link transmitter or signal source;
- **FEXT** – far end crosstalk is interference observed at the link side opposite to the aggressor link transmitter or signal source;

There are many more terms for the crosstalk characterization. PSXT, MDXT, ICN, ICR and some other will be introduced and explained in the next part on the crosstalk quantification.

Crosstalk In Parallel Traces

To illustrate interference of signals in parallel segments of microstrip line (surface trace), we will use 2 single-ended links routed at a distance equal to one trace width over 1 inch distance. The characteristic impedance of each single trace is close to 50 Ohm. In the first case a signal propagates in the top trace from port p1 to port p3 as illustrated Fig. 6.

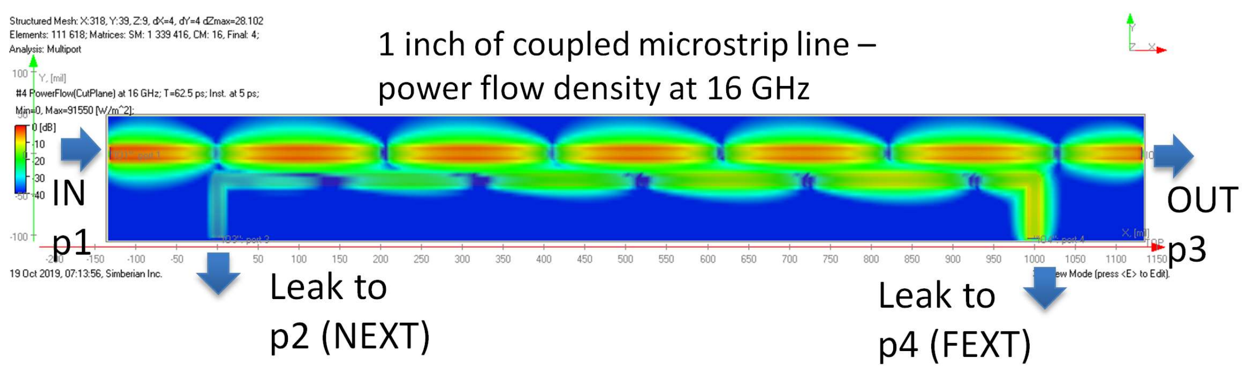


Fig. 6. Instantaneous values of power flow density in dB with 16 GHz 0.5 V signal propagating from p1 to p2. 16 mil traces on 8 mil substrate with Dk=3.9 coupled over 1 inch with 16 mil separation.

As we can see, the bottom link with ports p2 and p4 is literally in the "signal space" of the top trace. As the consequence, the bottom trace is coupled and "sucks out" some signal energy. Some useful signal power leaks into port p2 near the signal source (NEXT) and into port p4 on the side opposite to the signal source (FEXT). The leak is the result of the wave energy redistribution over the length of the coupled segment. How much energy is leaked and what are the consequences of that leak if there is a useful signal propagating in both links? The most fundamental and convenient way to describe this energy re-distribution phenomenon is with scattering parameters or S-parameters [5]. In case if you consider [5] as too long to read, there is shorter introduction into the subject [6]. Even shorter one-paragraph introduction in provided in Appendix I. It is very important to get familiar with the S-parameters. **In fact, practically all other interconnect quality metrics including crosstalk quantification parameters are simply derived from the S-parameters.** Magnitudes of the S-parameters describing process of the energy re-distribution for the structure from Fig. 6 are shown in Fig. 7.

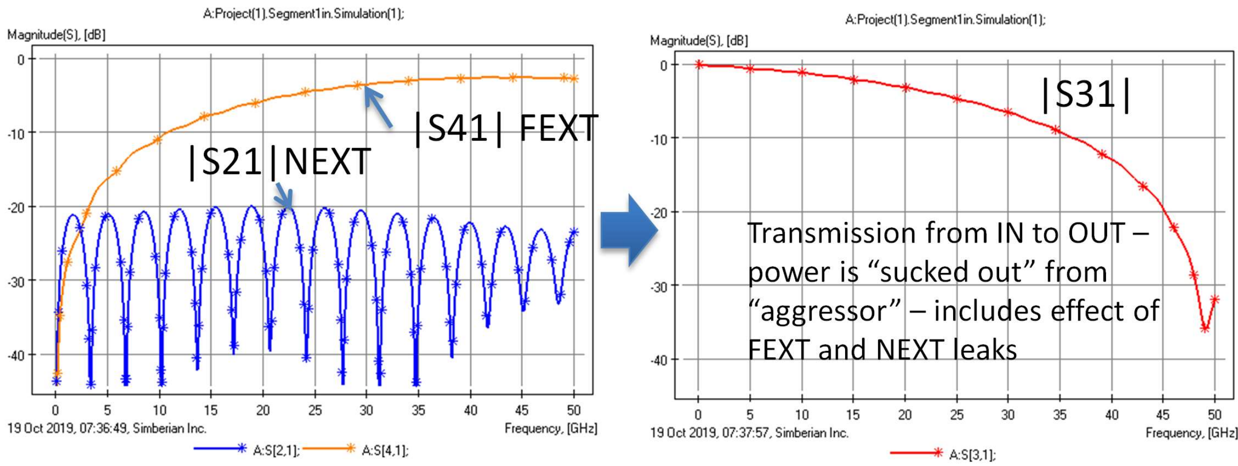


Fig. 7. Magnitudes of coupling S-parameters (|S41|and |S21|on the left graph) and transmission parameter (|S31| on the right graph) for microstrip traces with 1 inch coupled segment from Fig. 6.

Parameter S31 on the right graph of Fig. 7 is the transmission of the useful signal from port p1 to port p3. Parameters S41 and S21 are the leaks and potential crosstalk parameters. As we can see, the transmission|S31| is decreasing with the frequency, but with the slope that is much steeper than expected due to the material absorption losses [2]. The reflection or return losses are very small in this case. The reason for this is a leak into port p4 that is described by S-parameter S41 on the left graph in Fig. 7. The leak into port p2 is much smaller in this case. With a useful signal at port p4, parameter S41 becomes Far End Crosstalk or FEXT. It is frequency-dependent and the maxima and minima are defined by difference of propagation velocities of the even and odd modes in the coupled segment and by the segment length. There is no FEXT if this difference is zero or close to zero. FEXT can be observed in any lossy multi-conductor transmission line in general, but it becomes substantial only in cases of transmission lines with inhomogeneous dielectrics – microstrip lines for instance. Striplines with dielectric layers with different properties also have observable and not-negligible FEXT. FEXT increases with the length of the coupled traces up to some level and then decreases, reaches minimum. Then minima and maxima repeat periodically with the frequency. In general, more energy from an aggressor link may leak into a victim link on links with longer coupling sections. The leak can interfere with the victim signal and degrade it. On the other hand, the leak also degrades the aggressor signal - it causes additional losses that can be comparable or even larger than the reflection and the material absorption losses. In fact, there might be conditions when almost all energy of the aggressor signal becomes FEXT. In the case above almost complete suck out happens around 50 GHz. For the same traces coupled over longer 5-inch segment the complete suck out would happen around 10 GHz as illustrated in Fig. 8 – this is Nyquist frequency for 20 Gbps signal. **Extensive simulations must be used to detect and avoid such conditions early in design process.**

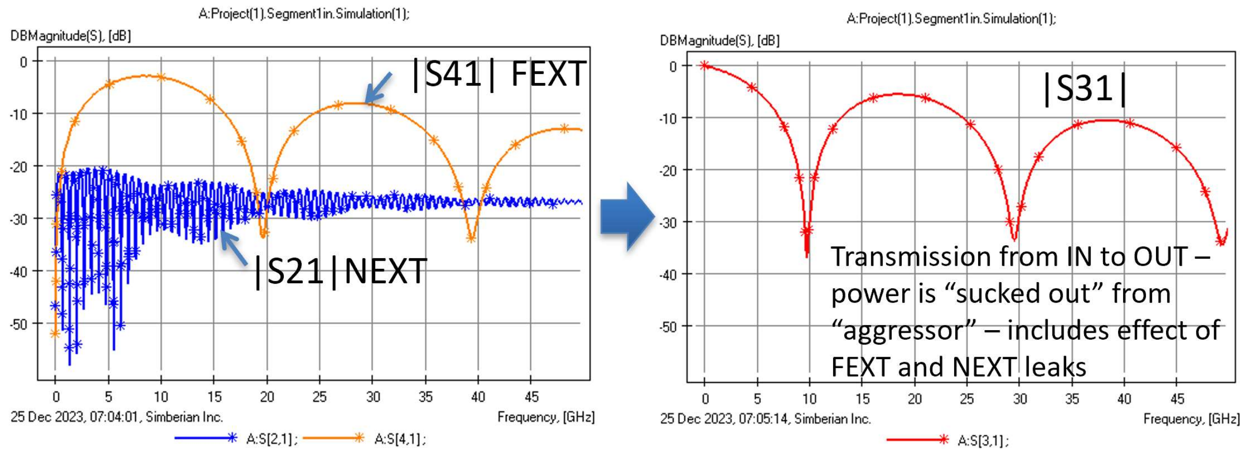


Fig. 8. Magnitudes of coupling S-parameters (|S41|and |S21|on the left graph) and transmission parameter (|S31| on the right graph) for microstrip traces with 5 inch coupled segment.

Note that S-parameters of passive interconnects are reciprocal. It means that transmission from port i to port j is always equal to transmission in the opposite direction from port j to port i. This is not very intuitive property. Applying it to crosstalk, we can state that S41=S14, S21=S12 – if the aggressor and the victim are switched, the crosstalk does not change. If the bottom link in Fig. 6 has useful signal, then that link becomes aggressor for the top link as illustrated in Fig. 9. The top link is the aggressor for the bottom link and the bottom link is the aggressor for the top link. The result is the superposition of signals in both links and both links have FEXT.

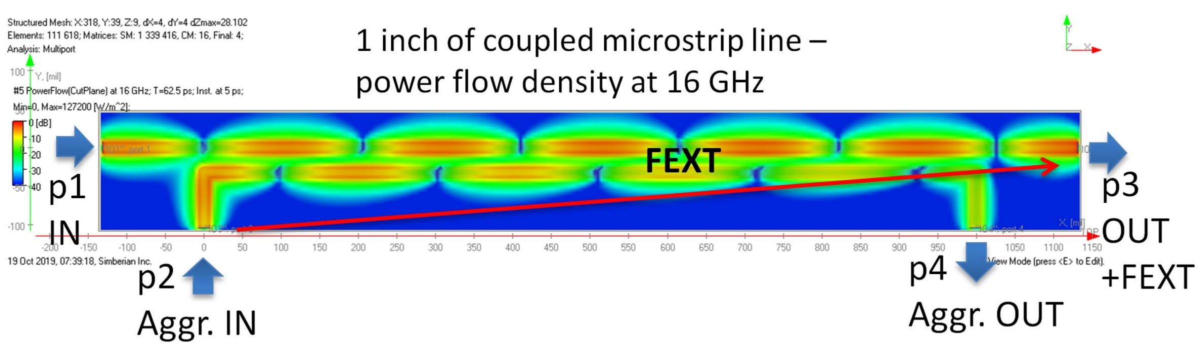


Fig. 9. Crosstalk power flow density in coupled microstrip line segment from Fig. 6 with two 16 GHz signals propagating in the same direction from p1 to p3 and from p2 to p4.

Up to this point we have investigated the coupling in the frequency domain or for the time-harmonic signals. Though, digital signals are usually transmitted by pulses [1]. Let’s take a look at the crosstalk in time domain. To do that a pulse response can be computed from the S-parameters of the 4-port structure, as illustrated in Fig. 10.

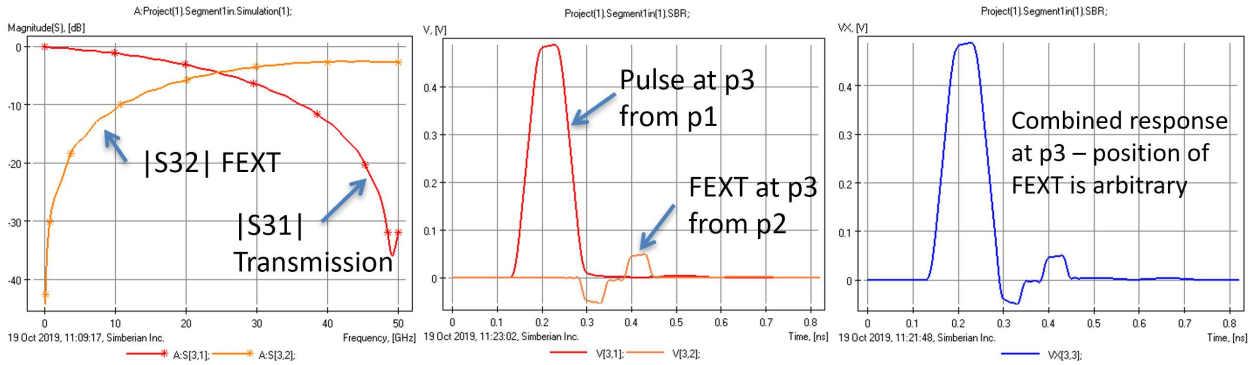


Fig. 10. Frequency-domain FEXT and transmission parameters (left graph) with corresponding pulse responses (middle graph) and superposition of the pulse responses (right graph).

Transmission from port p1 to port p3 is characterized by S-parameter S31 on the left graph and corresponding pulse response shown in the middle graph in Fig. 10. The bottom link is a potential aggressor in this case – it has leak to port p3 as illustrated with S-parameter S32 in Fig. 10. If the bottom link has similar pulse propagating from port p2 to p4, some part of the pulse energy is going to be leaked into port p3 of the top link. Corresponding pulse response for the crosstalk parameter is also shown in the middle graph in Fig. 10. If both links have signals propagating in the same direction over the coupled segment, the signals at port p3 will be a superposition of the link pulse response and the crosstalk pulse response as illustrated on the right graph in Fig. 10. Because of timing of the pulses at ports p1 and p2 is not synchronized, the crosstalk position is arbitrary with respect to the link pulse. That what makes the crosstalk difficult to quantify. We can find the worst-case relative timing, but it may never happen.

Let's return to the near-end crosstalk. S-parameters S21, S12, S34 and S43 describe the near-end crosstalk or NEXT in cases when the signals in the top and bottom links are propagating in the opposite directions as illustrated in Fig 11. The NEXT is frequency-dependent and has nulls at frequencies where the coupled segment is about a multiple of a half of wavelength as illustrated on the left graph in Fig. 12. The maxima in NEXT are approximately at frequencies with the wavelength equal to odd multiple of a quarter of wavelength. The near end coupling depends on the length of the coupling section. Though the frequency-domain pattern of the NEXT is considerably different from the FEXT.

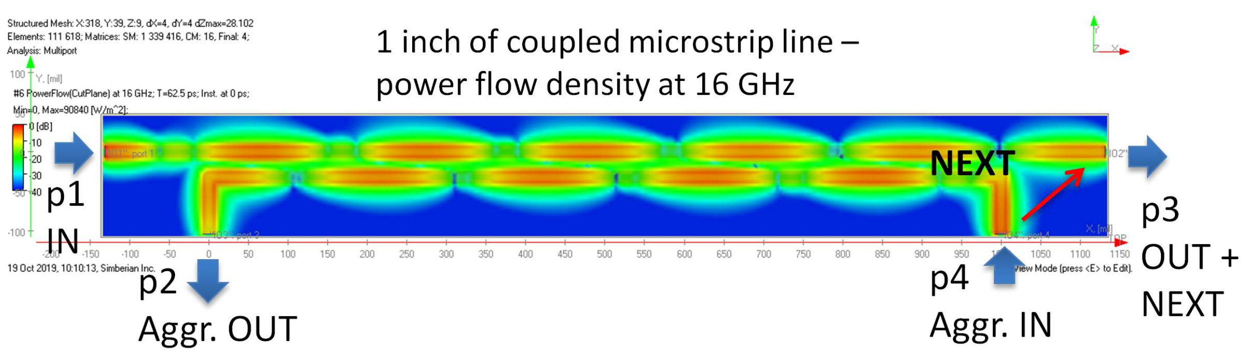


Fig. 11. Crosstalk power flow density in coupled microstrip line segment from Fig. 6 with two 16 GHz signals propagating in the opposite directions from p1 to p3 and from p4 to p2.

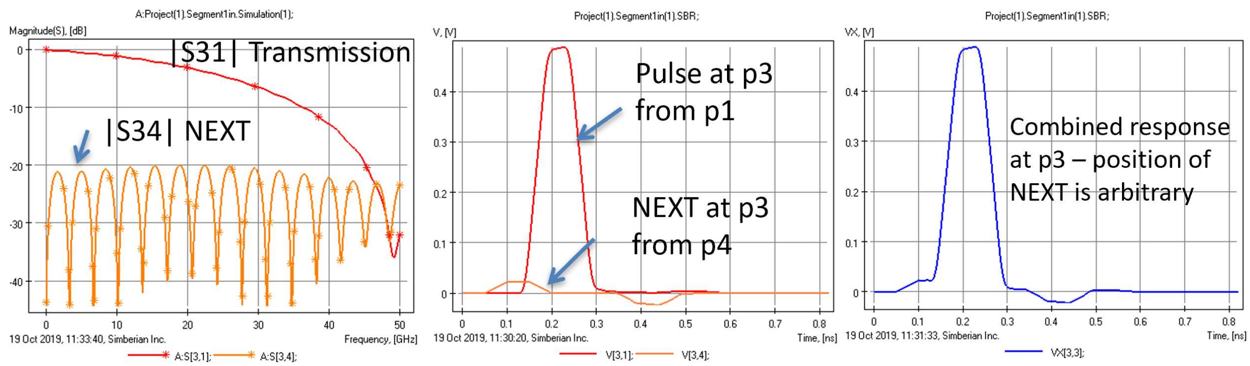


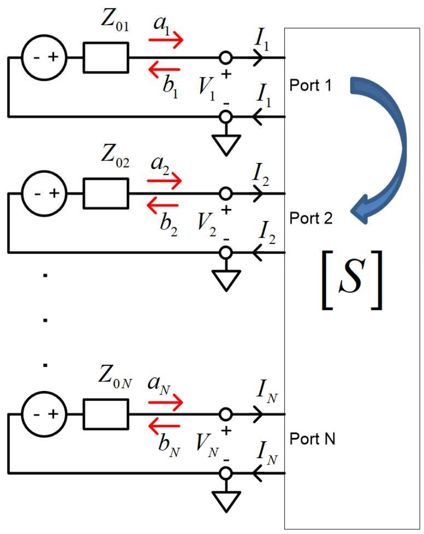
Fig. 12. Frequency-domain NEXT and transmission parameters (left graph) with corresponding pulse responses (middle graph) and superposition of the pulse responses (right graph).

As we can see the level of maxima of NEXT in this case is about -20 dB. That gives relatively small crosstalk pulse response (smaller than FEXT in this case) as shown on the middle graph in Fig. 12 together with the pulse transmitted from port p1 to port p3. It comes in two parts – the first appear at the victim port p3 almost immediately, and the second part appear with a delay. There might be no second part for a longer links with substantial attenuation. A superposition of the useful signal and crosstalk at the port p3 is shown on the right graph in Fig. 12. As in the case of FEXT, timing of the NEXT with respect to the pulse of the useful signal is also arbitrary. **Note that the signal in the victim link may be substantially attenuated at the received end – that means that much smaller values of NEXT can cause failure of the victim receiver.** All coupled nets have NEXT. Though, it is important how close to the victim receiver it appears.

Finally, here is how crosstalk in coupled transmission lines can be "explained". A signal in multi-conductor interconnects propagates as a superposition of multi-conductor line modes – this is a model presentation, but it allows to understand what happens there and what to expect. For instance, differential signal on the NEXT or FEXT aggressor side becomes a superposition of four waves in coupled 2-conductor transmission line segment. Those are the even and odd modes propagation in both directions. The signal cannot remain on just one trace due to the coupling. The waves propagating toward the aggressor side are observed as the NEXT. The waves propagating forward from the aggressor are observed as the useful signal in the aggressor link and also as the FEXT at the victim port. If all modes have identical propagation velocity (never happen in real lossy lines), there will be no FEXT. Though, if the modes are not ideally terminated at both ends, the waves forming NEXT are reflected and may appear at the other end as the FEXT. Waves that compose FEXT can be also reflected and become NEXT. To understand different coupling outcomes, a lattice diagram with the transmission line modes superposition can be used**. Though, the advanced modeling is easier way to account for all those reflections.**

Appendix I: S-parameters At a Glance

All signal and power interconnects can be described as multiport or black-box structures with just ports sticking out as illustrated in Fig. A1. Each port has 2 terminals – a signal terminal and a local reference terminal with currents and voltages defined as also shown in Fig. A1.



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

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

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

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

Scattering parameters:

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

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

Fig. A1. S-parameters definition.

With the currents and voltages, we can define Y-parameters (short-circuit parameters) and Z-parameters (open-circuit parameters). Though, instead of voltages and currents, S-parameters use “waves” defined at each port terminated with the normalization impedance Zo (that is usually just a resistance equal to 50 Ohm). The waves are variables that combines both current and voltage at each port in such way that the square of wave magnitude is the power of the wave. The waves may be real as in transmission line or, more often, are defined formally for modeling or measurement purpose. Similar to the currents and voltages, waves have directions as shown in Fig. A1. Incident waves (a=0.5*(I+Zo*V)/sqrt(Zo)) are directed into the multiport and reflected or outgoing waves (b=0.5*(I-Zo*V)/sqrt(Zo)) are directed from the multiport. S-parameters relate the incident and reflected/outgoing waves for each combination of ports terminated with the normalization impedance Zo. It is N by N matrix for the multiport shown above with the frequency-dependent elements. The matrix element Sij is the wave outgoing from port i (directed from multiport) with the unit incident wave at the port j only (no incident waves at all other ports). For passive interconnects it means that the magnitudes of S-parameters are always bounded by 1 – the power leaving the multiport should be always equal or smaller than power delivered to multiport. In cases of transmission line ports with the normalization impedance equal (or about equal) to transmission line characteristic impedance, Sij can be interpreted as a voltage at port i created by incident voltage 1 V at port j. Magnitudes on the S-parameter graphs are usually plotted in dB versus frequency (SdB=20*log(|S|)). dB value can be converted back into S-parameter magnitude by simple conversion formula 10^(|SdB|/20). 0 dB corresponds to unit value – for transmission parameter it means ideal transmission. For reflection parameters or return loss 0 dB or unit value means complete reflection or failure. -3dB corresponds to magnitude about 0.708 (or a half of power), -6dB to 0.5 (or a quarter of power), -20dB to 0.1, -40dB to 0.01, …. So, this is practically all you need to know about S-parameters, if you are not planning to write your own electromagnetic or signal integrity software 😊

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