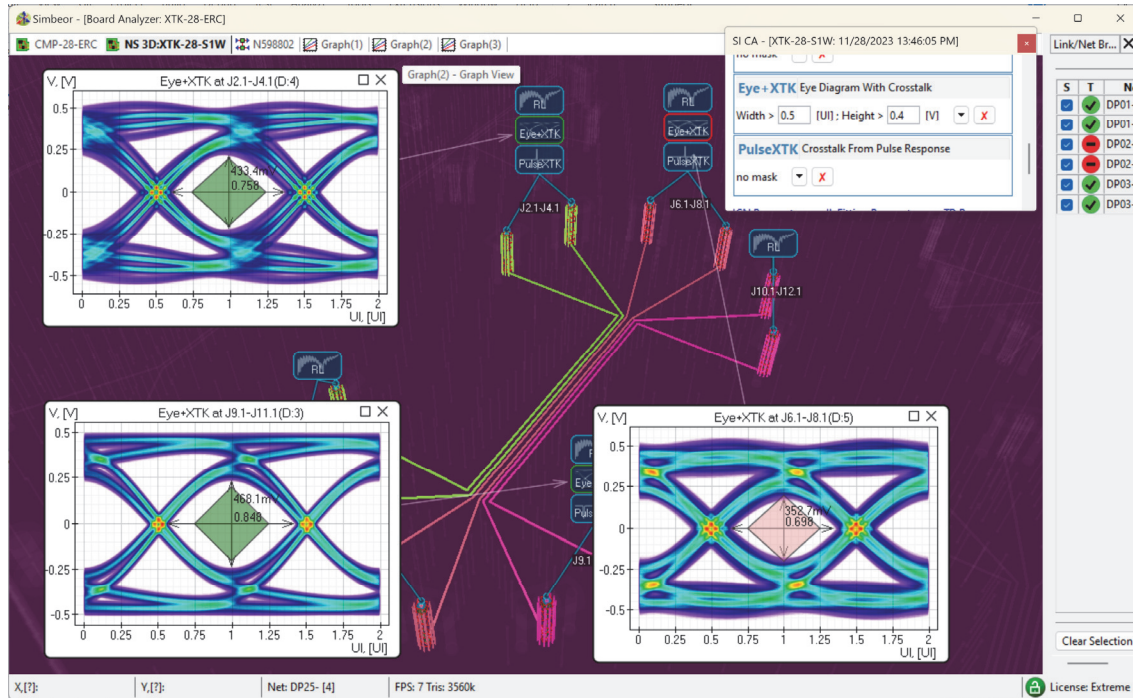


## How Interconnects Work: Crosstalk Quantification

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**Crosstalk is unwanted noise from structures coupled to a signal link that degrade the useful signal and may reduce the data transmission rate and even cause complete link failure.** All possible signal degradation effects, including the crosstalk, can be expressed with the balance of power as follows:

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

The formula is valid in the time-domain and in the frequency-domain over the bandwidth of the signal [1]. Here  $P_{out}$  is the power of signal at a receiver. The power from a driver ( $P_{in}$ ) is absorbed in dielectrics and conductors ( $P_{absorbed}$ ) [2], reflected back to the driver ( $P_{reflected}$ ) [3] - [4] and, possibly, leaked ( $P_{leaked}$ ) to coupled structures.  $P_{leaked}$  is a type of loss that includes local leaks to other signal links discussed in the first part of this article [5] and leaks to power distribution systems and radiation that happen mostly at the un-localized via holes [6]. The leaked power can be predicted or prevented relatively well – it requires analysis of a link with potentially coupled links (local coupling) and analysis of via-holes with absorbing boundary conditions. Only small leaks on via-holes can be accurately predicted with the analysis in isolation from the rest of the board and the large leaks must be prevented (that will be covered in the third part of this article). Overall, the leakage effects on a signal can be predicted relatively well in links with vias localized up to a target frequency. This is unlike the last term in the balance of power – the power gained from the coupled structures or  $P_{coupled}$ .

**There are a lot of uncertainties related to  $P_{coupled}$  and, as the result, there are multiple ways to characterize it.** This is because of a signal from the coupled links or aggressors are involved in the analysis (see definitions of terms in Appendix I). It can be one or multiple aggressors, signal similar to the victim signal as in parallel buses, or signals with different data rates or/and rise time in cases of accidental or distant coupling. Peaks of crosstalk are defined by the aggressor signal – the timing of the peaks and victim signals are not synchronized, but they are not completely random as well. The impact

of the coupling also depends on the strength of the signal in the victim link and location of the coupling. Very small coupling at the victim receiver where the useful signal is already degraded by the absorption and reflection losses may have much greater impact on the signal comparing to larger coupling at the driver end of the victim link. **As a consequence of that uncertainty, there are multiple ways to quantify the crosstalk phenomenon and this article discusses most of them.** Overall, the crosstalk simulation and quantification can be separated into three categories:

- 1. Coupling coefficients:** analysis of transmission line cross-sections at one frequency point and use of approximate equations for backward and forward coupling ( $K_b$  and  $K_f$ );
- 2. Frequency domain:** extraction of S-parameters with coupling in frequency domain and use of crosstalk metrics PSXT, ICR, ICN;
- 3. Time domain:** simulation in time domain with step, pulse or PRBS excitation signals (peak voltages or eye distortion).
- 4. Probabilistic:** statistical evaluation of crosstalk impact on bit error rate (BER) and channel operating margin (COM).

Approach #1 is useful only for the evaluation of local coupling in parallel or nearly parallel traces and can be effectively used for quick pre-layout investigations or to find the locations of crosstalk in post-layout analysis. Approach #2 is the most universal and is the foundation for the time-domain and for the statistical approach. It can be used for both local and distant coupling (see definitions in [5]). The time-domain analysis #3 is also universal and, technically, is the most accurate evaluation of the actual crosstalk values. The time-domain response is usually computed from the frequency-domain S-parameters, to account the frequency-dependent dispersion. But the results are useful on its own, especially for understanding the phenomenon. The time-domain analysis is useful for evaluation of the crosstalk from a step or pulse (single bit or symbol) excitation. It can be used to simulate a crosstalk from pseudo-random bit streams (PRBS), but the bits in a victim and the bits in possible aggressors are not correlated and the timing of the rise and fall edges in aggressors and victim are not synchronized – different bit sequences and timing produces different crosstalk impact. To handle these uncertainties, the crosstalk can be treated as a noise with a bounded probability density function identified from the time-domain analysis. This is the most modern probabilistic approach #4 for the crosstalk treatment in the IEEE 802.3 and OIF-CEI signaling standards. It may be also the most pessimistic crosstalk model.

Simbeor software is used to generate all examples for this article. All pre-layout analysis examples are from AdvXTalkKit provided with Simbeor SDK. Crosstalk validation platform XTALK-28/32 from Wild River Technology is used to illustrate the crosstalk quantification for the post-layout examples. Some analysis tips are provided in Appendix II.

## 1. Crosstalk Quantification with Coupling Coefficients

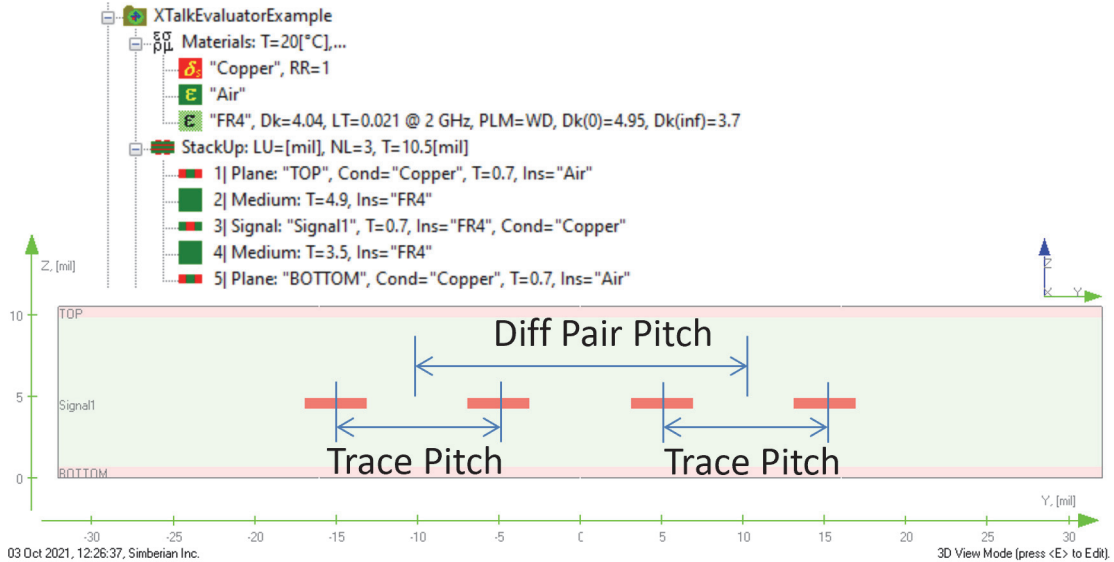
The fastest and the simplest way to quantify crosstalk is to simulate a cross-section of coupled traces with a field solver at one frequency point and use approximate equations for evaluation of forward and backward coupling. With this approach, capacitance [C] and inductance matrices per unit length are computed for a cross-section with the coupled traces first. Then an equation is used, to evaluate possible backward ( $K_b$ ) and forward ( $K_f$ ) coupling coefficients for a transmission line segment with length  $l$  for a step signal with unit amplitude and rise time  $t_r$ . Jarvis derived such formulas in [7] for single-ended symmetric traces. The Jarvis's formulas were further improved for small segments and for non-symmetric coupling cases by Bracken in [8]:

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

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

Here  $C_{21}$  and  $L_{21}$  are mutual capacitance and inductance,  $T_1$  and  $T_2$  are flight times and  $\zeta_1$  and  $\zeta_2$  are impedances of the coupled traces or impedances of differential modes for differential traces. The coefficients are the voltage step responses at the near end ( $K_b$ ) and the far end ( $K_f$ ) of the coupled transmission line segment, assuming 1 V step excitation in the aggressor. If only one coupled segment is involved,  $K_b$  is the voltage of near-end crosstalk (NEXT) and  $K_f$  is the voltage of far-end crosstalk (FEXT). The equations are relatively accurate for lossless or low-loss cases. Though, they do not work well with high losses or long segments (it may overestimate the forward crosstalk that is attenuated by the losses). **The formula can be used as an estimate of the maximal possible step cross-talk.** The formulas also assume ideal transmission line segment termination and do not account for possible reflections (yes, the crosstalk is reflected too). Note that the reflected NEXT can be observed as FEXT and the formulas provided above can be further improved as it is done in [9]. Though, the frequency or time-domain simulation of crosstalk is always preferable, to account for the reflections. Also, it should be noted that **the step excitation may under-estimate the actual peak-to-peak crosstalk for short links by up to 2 times or by up to +6dB** – that will be explained later.

The coupling coefficients is a convenient tool for both pre-layout and post-layout crosstalk investigation. Different kinds of sweeps can be used to define design rules in the pre-layout process for instance. A static or quasi-static field solver is needed to begin such investigation. Such as Simbeor SFS available in Simbeor software. It can be done by scripting in MATLAB or Python and use of Simbeor SFS field solver through Simbeor SDK as it is done for this article. As the first example of pre-layout analysis of coupling coefficients, let's investigate a cross-section shown in Fig. 1. The geometry is similar to the case investigated on page 5 in [10].



**Fig. 1. Stackup and cross-section for differential-to-differential pair crosstalk investigation in striplines, trace width is set to have 80 Ohm differential impedance.**

The backward coupling coefficient  $K_b$  is plotted as a function of separation between two differential pairs (Diff Pair Pitch) in Fig. 2. Trace Pitch is 10 mil, trace width is adjusted to have 80 Ohm differential impedance.  $K_b$  is a potential crosstalk value in Volts, assuming that the aggressor signal is 1 Volt. A coupling coefficient can be converted into % by multiplying it by 100. It can be also converted into dB as  $20 \cdot \log(|K|)$ . As we can see from Fig. 2, the crosstalk is quickly reduced by increase of separation between the differential pairs. Though, the result will depend on a particular stackup and material parameters. This is one of the parameterized examples of crosstalk analysis from AdvXTalkKit provided with Simbeor SDK – see details in [11]. It can be quickly adjusted for a different geometry and sweeps. For instance, what if the Trace Pitch parameter is increased to 14 mil – how it compares to the case with 10 mil. Fig. 3 shows the results of such experiment and this time  $K_b$  is plotted in dB. We can see that the loosely coupled differential traces have substantially larger coupling – it may require larger spacing to reach the transmitter-to-receiver isolation -72 dB in this case (dashed bottom line in Fig. 3 as suggested in [10]).

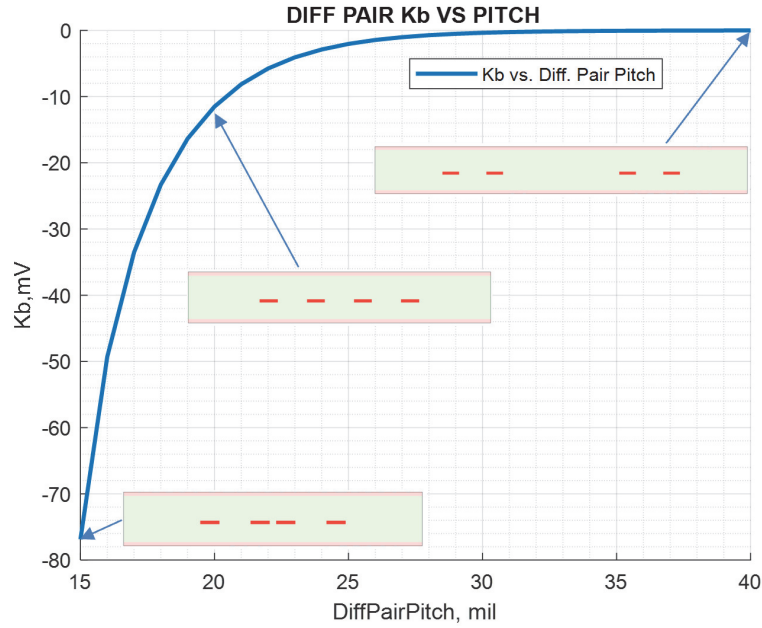


Fig. 2. Backward coupling coefficient  $K_b$  vs distance between two differential pairs for 1 inch segment of coupled differential striplines with Trace Pitch 10 mil and 20 ps rise time.

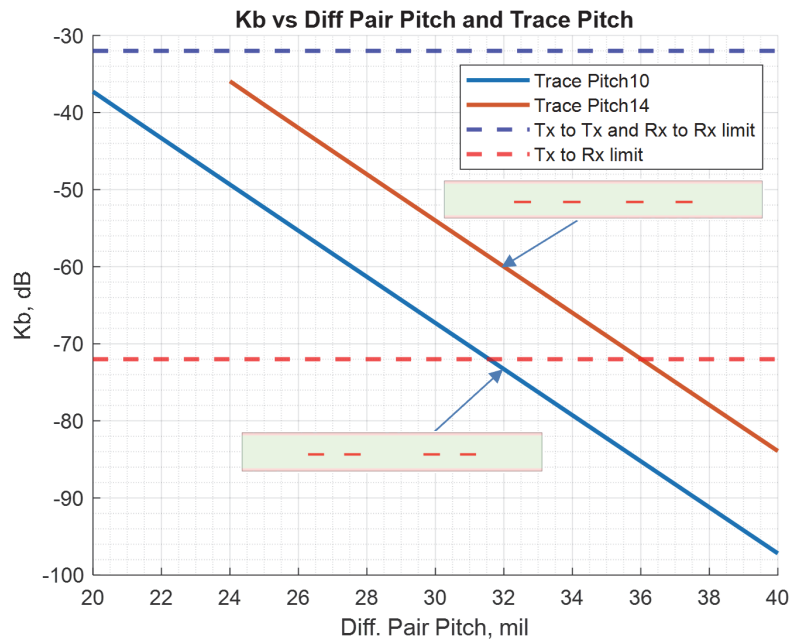


Fig. 3. Backward coupling coefficient  $K_b$  in dB vs distance between two differential pairs for 1 inch segment of coupled differential striplines with Trace Pitch 10 mil (blue line) and 14 mil (brown line) and 20 ps rise time.

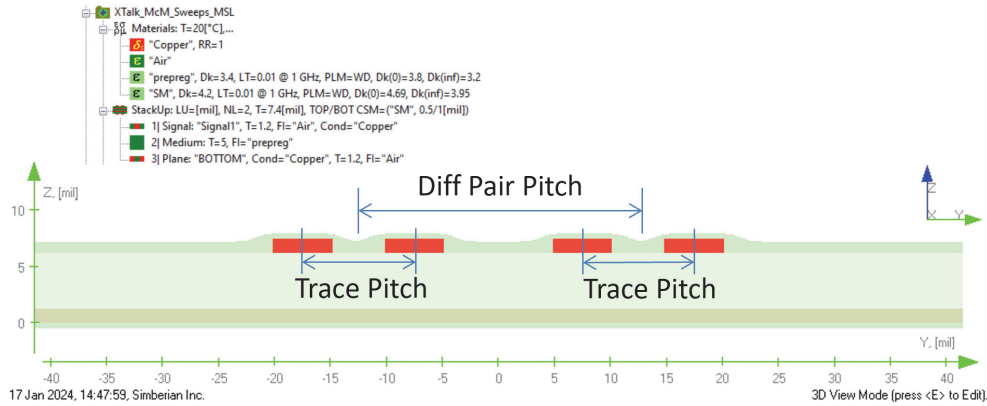


Fig. 4. Stackup and cross-section for differential-to-differential pair crosstalk investigation in microstrip line, Trace Pitch is 10 mil, trace width is adjusted for 100 Ohm differential impedance.

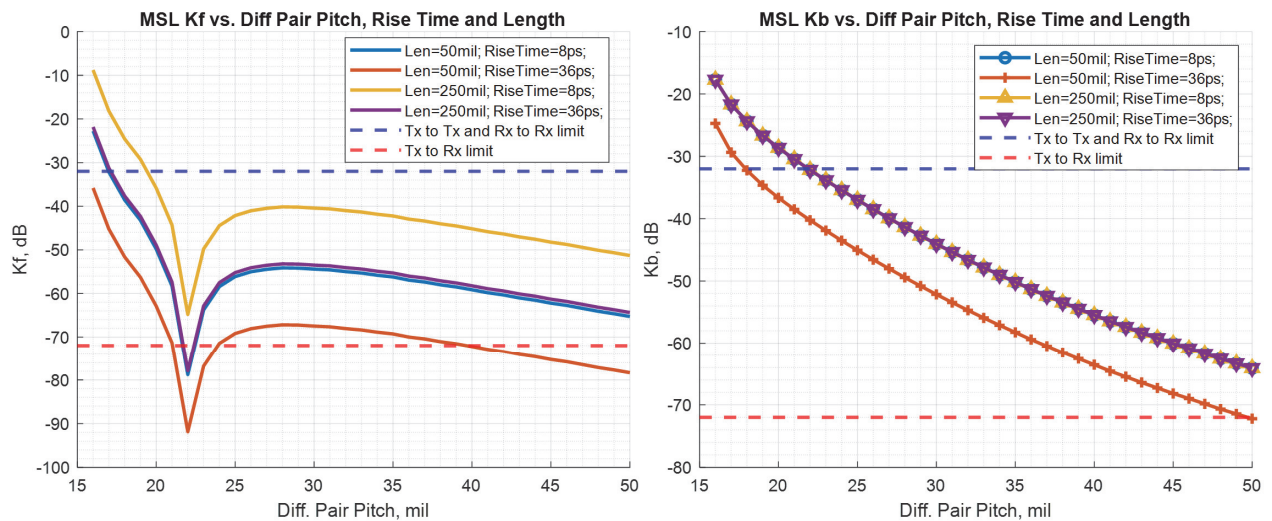


Fig. 5. Forward coupling coefficient  $K_f$  in dB (left plot) and backward coupling coefficient ( $K_b$ ) vs distance between two differential pairs (Diff Pair Pitch) for 50 mil and 250 mil segments of coupled microstrip line and 8 ps and 36 ps rise times.

The striplines usually have mostly backward coupling and forward coupling is negligible due to homogeneous dielectric filling. This is unlike the microstrip structures, where the forward coupling dominates. Example of pre-layout differential microstrip coupling investigation is shown in Fig. 4 and Fig. 5. The example is similar to example from page 14 and 15 of [10]. The traces are covered by conformal solder mask. The trace pitch is fixed and trace width is adjusted to have 100 Ohm differential impedance. As we can see from the left plot in Fig. 5, the forward coupling depends on the length and rise time. More coupling observed for longer traces and for shorter rise time. The forward coupling has a distinct minimum, but overall, it would be difficult to make something useful out of it due to high sensitivity to geometry variations and to material parameters. The backward coupling in microstrip differential traces is smaller for the short segment and longer rise time, but it practically identical for longer segment and shorter rise times as shown on the right plot in Fig. 5.

The coupling coefficients can be also used for preliminary investigation of existing PCB layout, to find location where the coupling coefficients exceed some threshold. It can be done literally with one

button click with advanced simulation-based Electrical Rule Checking (ERC) mode in Simbeor SI Compliance Analyzer tool (only signal rise time is required for this type of analysis). The crosstalk validation platform XTALK-28/32 from Wild River Technology is a perfect tool to validate and illustrate different ways to analyze the crosstalk (this is in addition to the investigation of interconnects predictability with the analysis to measurement correlation). The results of the crosstalk analysis for four 2-inch coupled differential segments and long link coupled to short link are shown in Fig. 6. The insert shows details of the crosstalk evaluation for 2-inch structure with 1w separation between the pairs – the forward coupling  $K_f = -69.69\text{mV}$  dominates the backward coupling  $K_b = 21.37\text{mV}$  for the bottom differential pair.

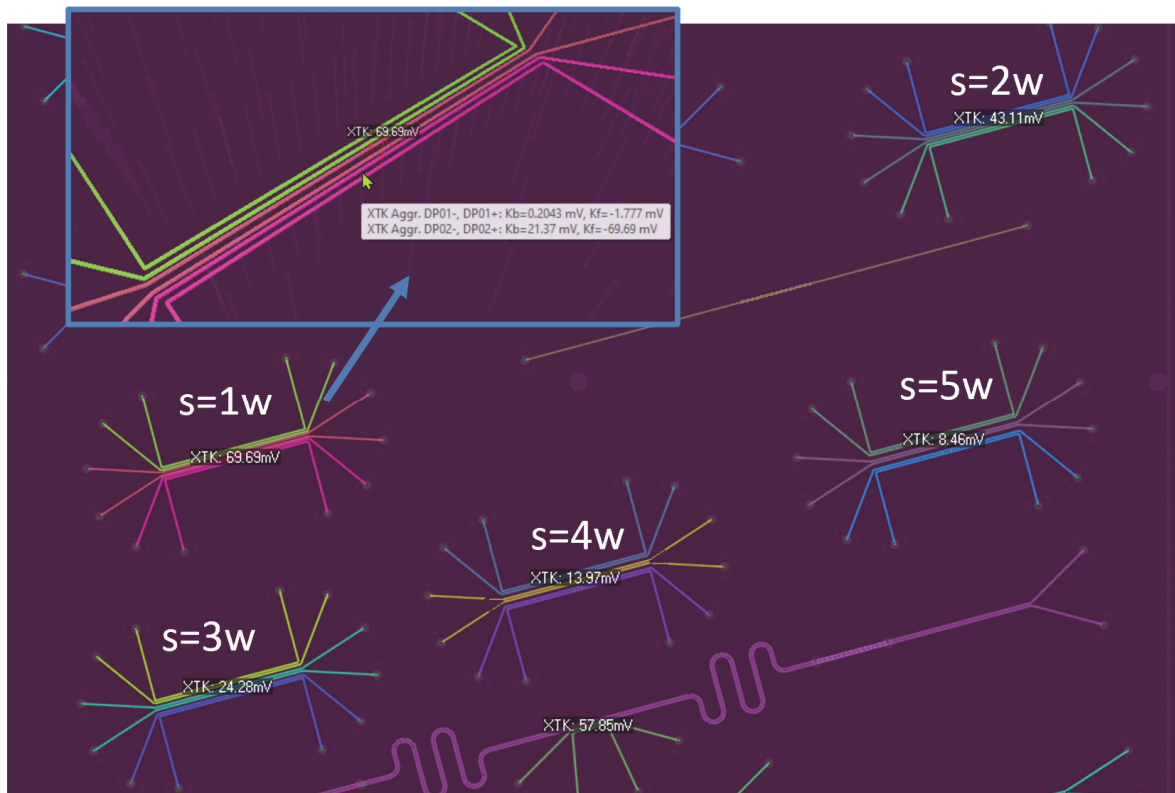


Fig. 6. Example of crosstalk evaluation on XTALK-28/32 platform in Simbeor SI Compliance Analyzer for five 2-inch differential microstrip structures with the edge-to-edge separation from 1 to 5 trace widths and for long to short link coupling structure (bottom), 25 ps rise time (10-90%).

Stackup for XTALK-28/32 board is practically the same as for CMP-28 validation platform featured in [12]. Trace widths are 13.5 mil (wide traces are used to reduce the effect of manufacturing variations). As we can see from Fig. 6, the larger separation gives smaller coupling. See some tips on the post-layout analysis of crosstalk in Appendix II and view of XTALK-28/32 platform in Appendix III.

## 2. Crosstalk Quantification in Frequency Domain

**More accurate way to quantify the crosstalk is to simulate a segment of multi-conductor transmission line in frequency domain over a signal spectrum bandwidth.** S-parameters of a coupled line segment can be either extracted separately for an analysis in isolation (to generate rules in pre-layout process) or used as an element of a model for coupled links that contains segments of coupled traces. Note that the S-parameters with coupling can be directly used to quantify the crosstalk. For instance, a transmission parameter between two ports from different links is a coupling parameter that describes the crosstalk. 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 [13], [14]. Though, a metric called **Power Sum Crosstalk (PSXT)** may be useful for preliminary evaluation of overall crosstalk in a link with multiple aggressors (see [13] or OIF-CEI and IEEE 802.3 standards):

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

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

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

PSXT is the total power sum crosstalk. PSNEXT and PSFEXT are PSXTs from the near and far end aggressors. PSXTs are functions of frequency and are computed from the S-parameters at a set of frequency points. As we can see, the **PSXT is just a sum of squares of S-matrix elements from all possible aggressors at a victim receiver port expressed in dB.** If there is just one aggressor, PSXT sum contains one S-parameter element. In this case, PSXT is equal to corresponding S-parameter magnitude expressed in dB. PSXT is different from S-parameters only if there are multiple disturbers or aggressors. In such cases PSXT may be called Multiple Disturber PSXT (MDXT, MDFEXT, MDNEXT).

As the first example, let's simulate 1 inch segment of coupled differential striplines with the cross-section shown in Fig. 1. The example is similar to the pre-layout example for the coupling coefficients. Trace Pitch in this case is 10 mil, trace width is fixed to 3.8 mil. This is another example from Simbeor SDK AdvXTalkKit [11]. The results of the analysis for two differential-to-differential pair separations are shown in Fig. 7. The left graphs are computed with 15 mil separation and the right graphs with 20 mil separation. The top plots are for the near end PSXT and the bottom plots are for the far end PSXT. This is stripline structure and, as we can see, the NEXT is considerably larger. Though, the FEXT is not negligible in case of the closely-spaced differential pairs. Because of just one potential aggressor, the PSXTs in this case are just magnitudes of the differential transmission from port A2 to port A1 (PSNEXT) and from port A2 to port B1 (PSFEXT). The maximal value of PSXT can be used directly as a rough estimate of the upper limit of the crosstalk values (transform PSXT into voltage with  $10^{(PSXT/20)}$ ). In this case, the PSNEXT for 15 mil case is about -17dB, that corresponds to 140mV and for 20 mil separation the value is about -33dB, that corresponds to 22mV. Notice that that values are larger than the values computed with the coupling coefficients as shown in Fig. 2. **The peak-to-peak crosstalk in time domain may be even larger than the upper limit estimated with PSTX.** It depends on how actual signal harmonics are superimposed at the victim port. Though, most likely, the actual crosstalk values are going to be smaller than the upper limit [13].



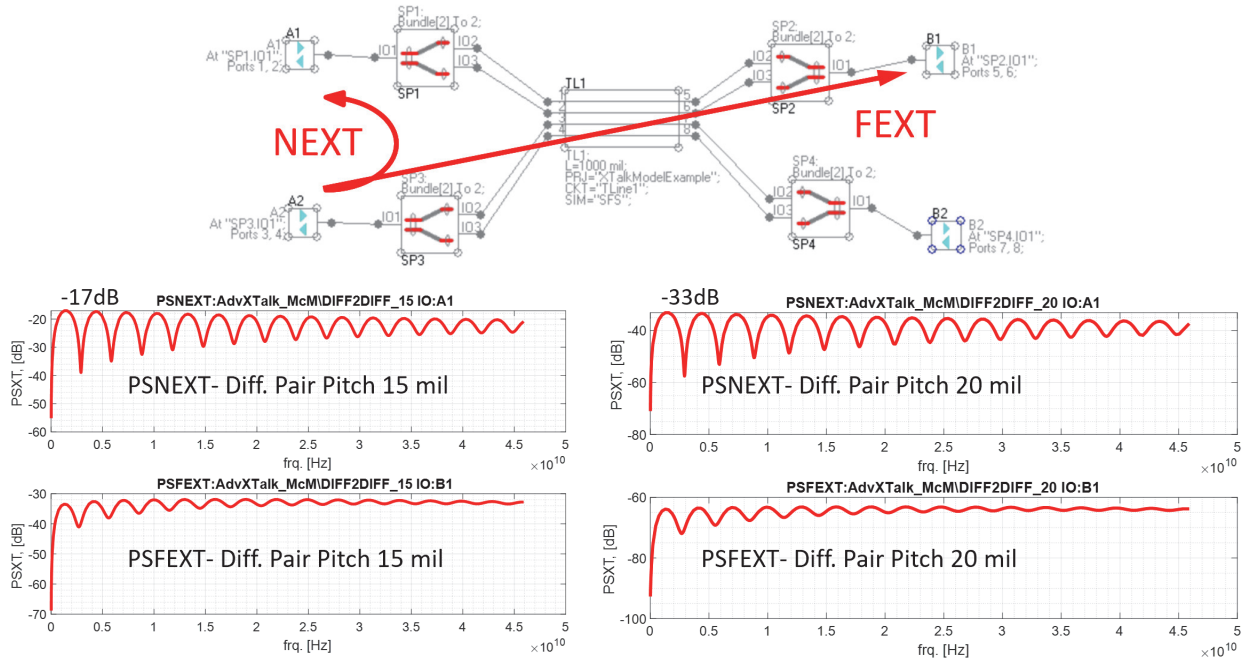


Fig. 7. Example of pre-layout analysis of PSXT in 1 inch segment of coupled differential striplines with cross-section shown in Fig. 1 for Diff. Pair Pitch 15 mil (left plots) and 20 mil (right plots).

As an example of the post-layout crosstalk analysis, let's compute PSXTs for some differential coupled links from XTALK-28/32 platform. Very similar to ERC, it can be done with one button click in Simbeor SI Compliance Analyzer tool (either Fast SI or 3D EM analysis can be used for the crosstalk modeling). Though, it requires additional setup for the transmitters and receivers, to properly define victims and aggressors. The bit rate and rise time have to be also defined – the frequency sweep is auto-defined from these data or can be manually re-defined if necessary. Simbeor uses de-compositional electromagnetic analysis [15], that accounts for the coupling between transmission lines.

The results of PSXT analyses for two structures with three differential links coupled over 2-inch parallel segments are shown in Fig. 8. Those are two structures from Fig. 6 marked as  $s=1w$  and  $s=4w$ . Differential trace width is 13.5 mil and differential trace pitch is 37 mil. Those are loosely-coupled microstrip differential pairs and they are very susceptible to the interference, as we can see from these examples. In each case, two links have transmitters on one side (TX1 and TX2) and one link has transmitter on the opposite side (TX3). Each receiver has multiple disturbers in this case. Crosstalk on the victim receiver RX2 is shown in Fig. 8. RX2 has one near end aggressor TX3 and one far end aggressor TX1. As in the case of pre-layout example, the corresponding PSNEXT and PSFEXT are exactly the magnitudes of corresponding transmission parameters. Though, the total PSXT is the sum of squares of magnitudes expressed in dB. **PSXT is a superposition of the aggressor's signals that does not account for the phases of signal harmonics.** The PSFEXT dominates in both cases. Though, PSNEXT is also substantial for the case with smaller separation between coupled differential pairs.

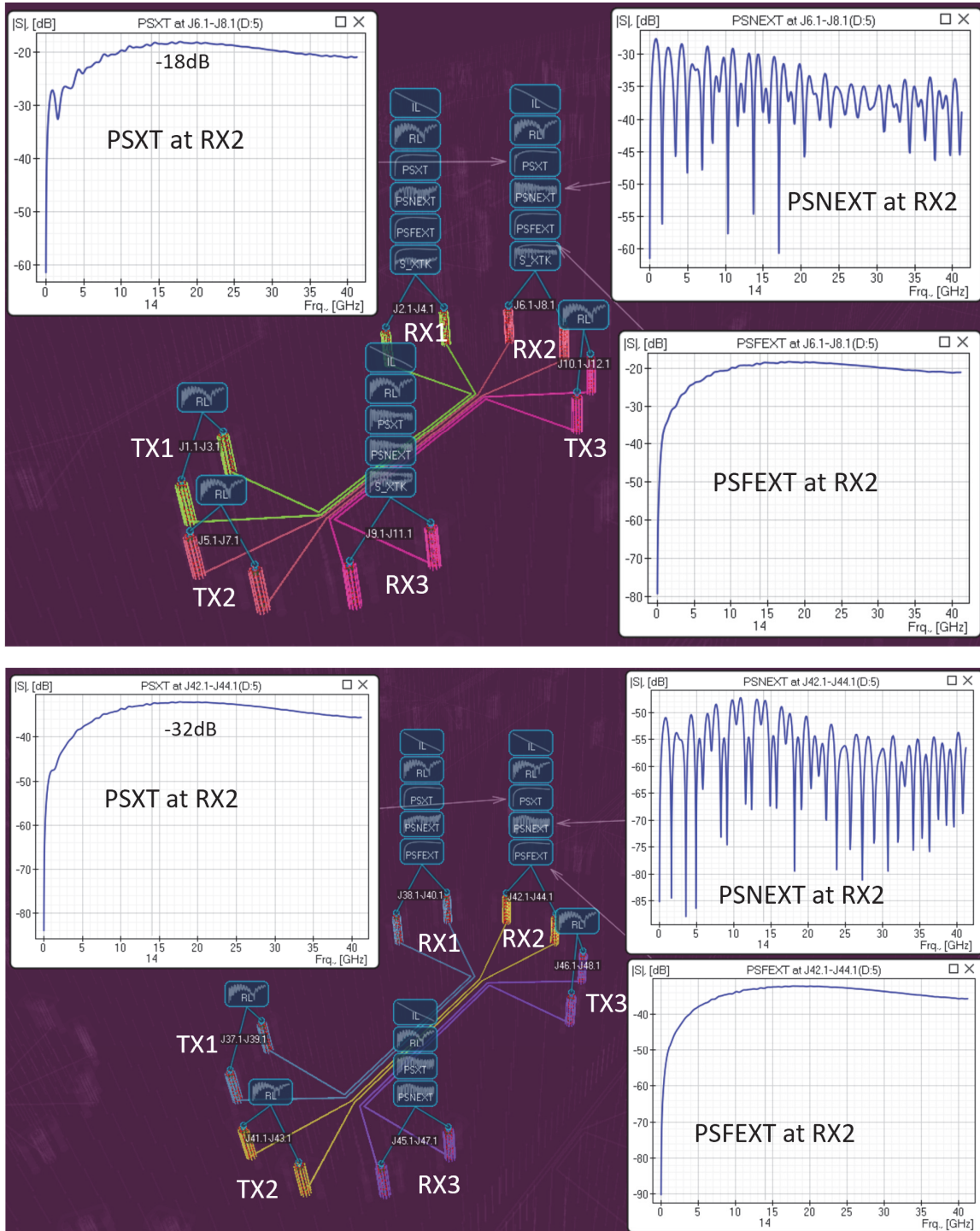


Fig. 8. Examples of post-layout PSXT analysis on XTALK-28/32 platform in Simbeor SI Compliance Analyzer for 2-inch differential microstrip structures with the edge-to-edge separation equal to 1 trace width (top screenshot) 4 trace widths (bottom screenshot). TX1-TX3 are the transmitter sides and RX1-RX3 are the receiver sides.

The  $s=1w$  case has maximal PSXT about -18dB, that is about 126mV. The 4w case has maximal PSXT about -32dB or 25mV. It is an estimate of the superposition of the crosstalk with two aggressors (near and far end). Fig. 6 shows only the largest values of the forward or backward coupling. Note that the ERC mode does not account for the actual signal propagation direction and assumes no reflections. The model used for PSXT computation does not have such limitations.

PSXT can be used to evaluate the crosstalk and set the limits or verify compliance for some signaling standards that provide compliance masks specifically for PSXT. Though, the 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 on a particular link:

$$ICR_{i,j} = IL_{i,j} - PSXT_i \text{ [dB]}$$

where  $IL_{i,j} = 20 \cdot \log(|S_{i,j}(f)|)$  is the insertion loss at port  $i$  and  $PSXT_i$  is the power sum crosstalk at the same port (both values are in dB).

ICR is a kind of signal to noise ratio [13] (IL is the signal at a receiver and PSXT is the noise). **The larger values of ICR mean smaller impact of the crosstalk on the signal.** To understand the ICR, let's use structure with coupled long and short links from XTALK-28/32 platform. The results of the analysis are shown in Fig. 9. ICR at RX1 is computed for longer link and ICR at RX2 is for the shorter link. We can see that the shorter link has much larger ICR, that means smaller impact of the crosstalk. This is because of much smaller insertion loss in the shorter link. PSXT and IL for both links are also shown in Fig. 9 for comparison.

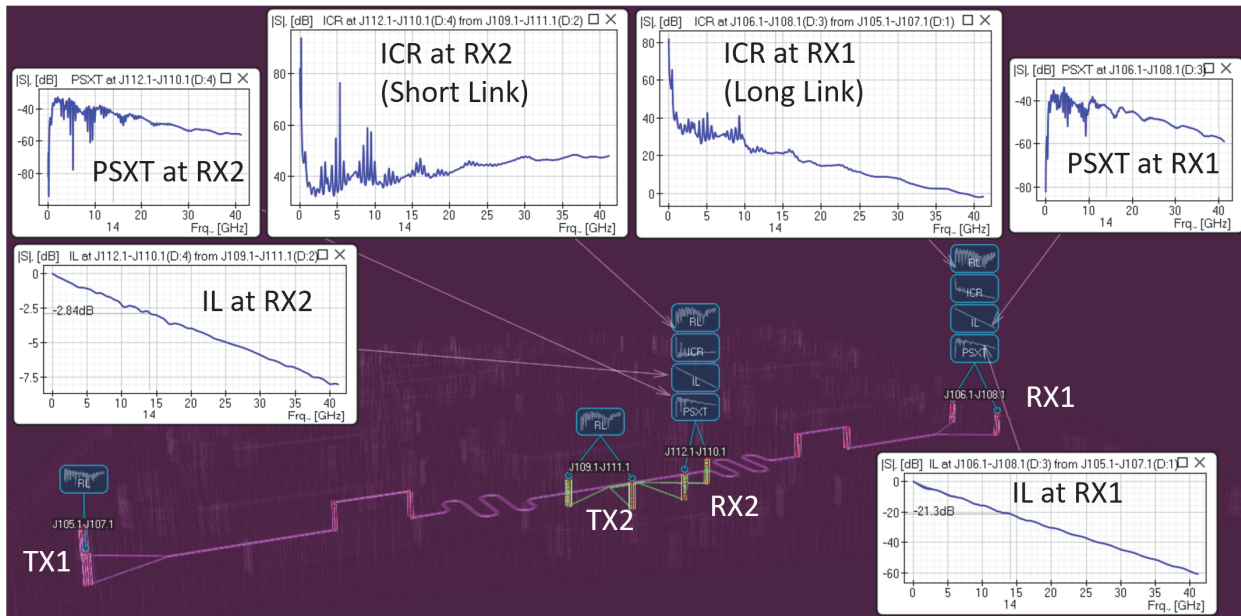


Fig. 9. Examples of post-layout ICR analysis on XTALK-28/32 platform in Simbeor SI Compliance Analyzer for long and short coupled links. TX1-RX1 is the long link and TX2-RX2 is the short link. ICRs for both receivers are shown at both receiver ports together with corresponding PSXT and IL.

The PSXT and ICR are useful metrics for a preliminary crosstalk evaluation. Though, those are pure frequency-domain metrics. The actual amount of crosstalk noise for a particular signal depends on the signal spectrum and may be also altered by filtering properties of a transmitter and a receiver packages. **Integrated crosstalk noise (ICN)** metric was introduced to account for the signal spectrum and filtering properties of a transmitter and a receiver [13]. ICN is just RMS of weighted PSFEXT and PSNEXT computed as follows:

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

$$\sigma_{NEXT} = \sqrt{2 \cdot \Delta f \cdot \sum_k W_{NEXT}(f_k) \cdot 10^{\frac{PSNEXT(f_k)}{10}}}$$

$$\sigma_{FEXT} = \sqrt{2 \cdot \Delta f \cdot \sum_k W_{FEXT}(f_k) \cdot 10^{\frac{PSFEXT(f_k)}{10}}}$$

The frequency-dependent weights  $W_{NEXT}$  and  $W_{FEXT}$  account for spectrum of random bit sequence. They 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 [13] or in IEEE Std. 802.3 standard). The result of the ICN computation is near end  $\sigma_{NEXT}$ , far end  $\sigma_{FEXT}$  and total crosstalk  $\sigma_{XTK}$  estimated in Volts (rms value). The total ICN is one number that is very convenient for qualitative assessments of the crosstalk impact.

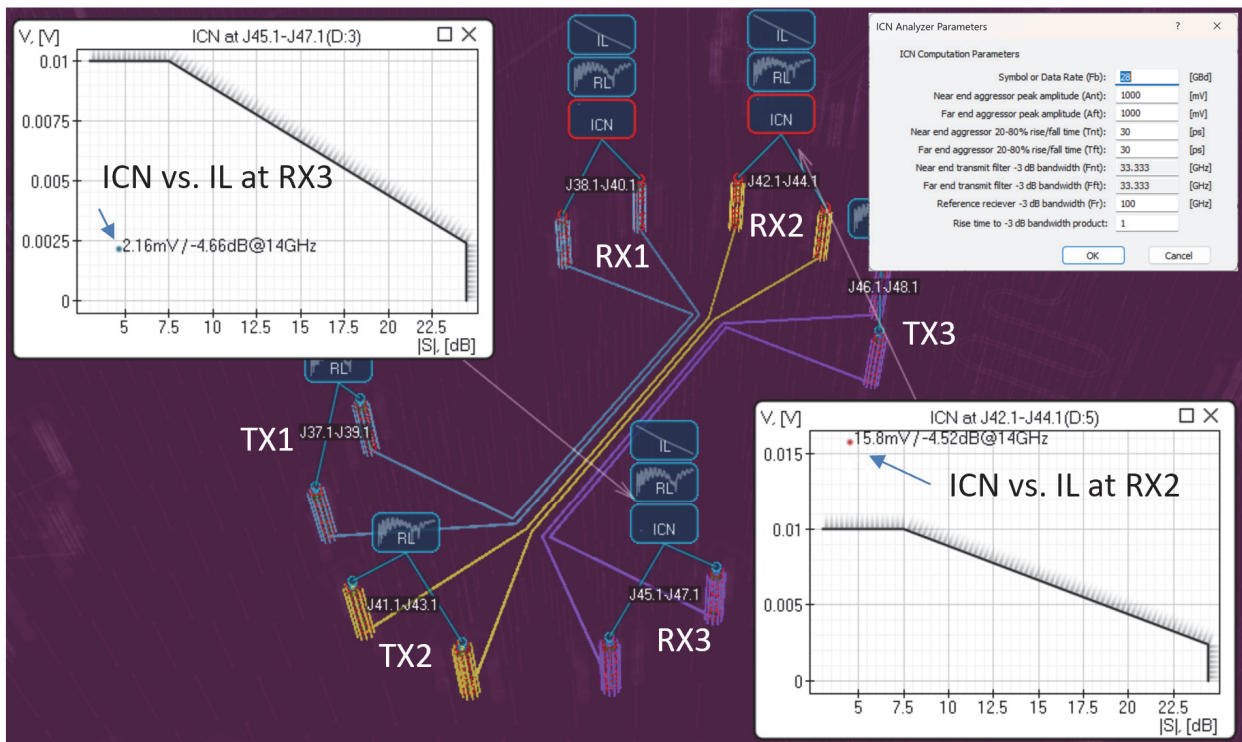


Fig. 10. Examples of total ICN analysis on XTALK-28/32 platform in Simbeor SI Compliance Analyzer for 28 Gpbs signal in 2-inch differential microstrip structure with the edge-to-edge separation between differential pairs equal to 4 trace widths - TX1-TX3 are transmitter sides and RX1-RX3 are receiver sides.

The limit on the ICN is usually set and plotted versus the insertion loss at the Nyquist frequency (see Appendix I for definition). An example of ICN computation and plotting for three differential links coupled over 2-inch length with 4 widths separation between the differential pairs is shown in Fig. 10. The ICN values are plotted together with a typical compliance mask. The mask allows larger crosstalk in links with smaller insertion losses. The aggressor signals are set to 1 V in this case and the other parameters for computation of ICN are shown in Fig. 10. As we can see, the receivers RX3 and RX2 have almost the same insertion losses at Nyquist frequency, but the crosstalk at RX2 is larger – corresponding dot is in the failure area of the compliance mask. RX3 in this case has only near-end crosstalk from two transmitters TX1 and TX2. Receiver RX2 has both far end crosstalk from TX1 (larger one) and near end from TX3. Note that the ICN sometime explained as an average value of expected crosstalk or even as a standard deviation for a crosstalk with the normal probability distribution [13]. Thus, it is usually the bottom estimate.

### 3. Crosstalk Quantification in Time Domain

**Yet another way to quantify crosstalk is to compute step or pulse response of a link with coupling and 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 more 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. If the aggressor signals are not synchronous with the victim signal, the crosstalk does not correlate in time with the single bit response [5]. Thus, it cannot be mitigated as the other types of signal degradation factors like reflection and losses. The time-domain analysis in Simbeor is done with the rational approximation of S-parameters computed for a segment or a complete link.

As the first example, let’s simulate 1 inch segment of coupled differential striplines with the cross-section shown in Fig. 1. The example is similar to the pre-layout example for the coupling coefficients and is the same as investigated in section 2 in the frequency-domain. Trace Pitch in this case is 10 mil, trace width is fixed to 3.8 mil. This is also example from Simbeor SDK AdvXTalkKit [11]. The results of the crosstalk analysis for 20 mil differential-to-differential pair separation are shown in Fig. 11. The left graphs are computed with the step excitation and the right graphs with the pulse excitation. The top plots are for the near end crosstalk and the bottom plots are for the far end crosstalk responses. This is stripline structure and, as we can see, the NEXT is considerably larger. Though, there is some FEXT and it is not negligible. **The peak-to-peak NEXT is almost 2 times larger than the value obtained from the step response because of the crosstalk reflections in a relatively short segment.** Examples of pre-layout investigation of FEXT and NEXT dependency from the differential pair pitch calculated from step response with 20 ps rise time are shown in Fig. 12 – this case is similar to the investigation with the coupling coefficients shown in Fig. 2. However, the direct analysis in time domain is more accurate – it includes reflections, losses and dispersion of the signal.

As an example of the post-layout crosstalk analysis, let’s simulate differential coupled links from XTALK-28/32 platform that was previously investigated with the coupling coefficients in section 1 and in the frequency domain in section 2. Again, it can be done with just one button click in Simbeor SI Compliance Analyzer tool. Though, in addition to setting up the transmitters and receivers, time-domain stimulus should be defined before the crosstalk simulation. The minimal setup requires the bit rate and rise time. Amplitudes of the sources and type of bit stream and possible jitter parameters may also need adjustments (see Appendix II on how to do it in Simbeor).

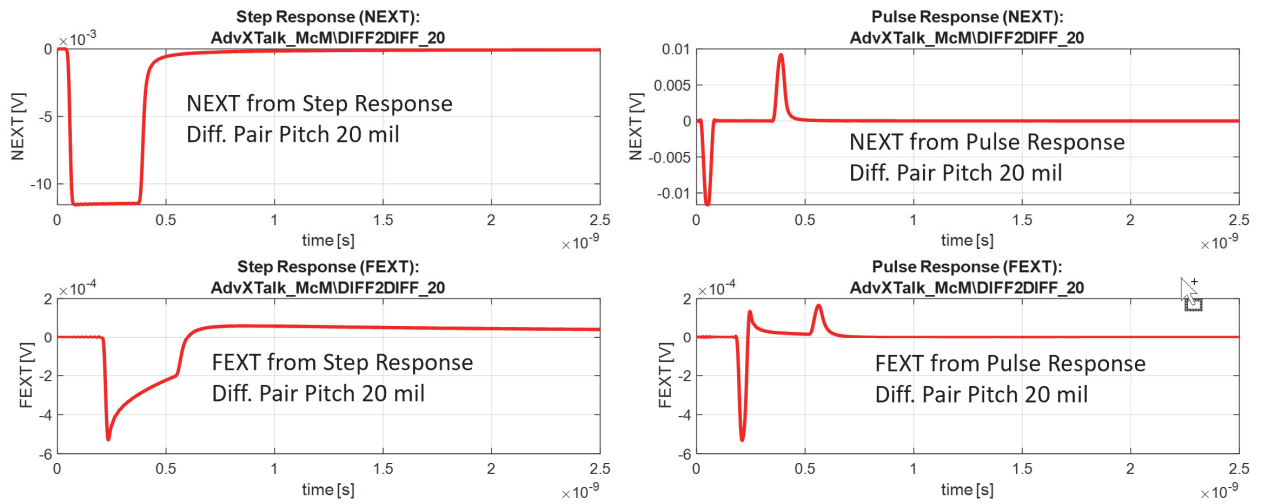


Fig. 11. Example of pre-layout simulation of crosstalk from step response (left graphs) and from pulse response (right graphs) in 1 inch segment of coupled differential striplines with cross-section shown in Fig. 1 for Diff. Pair Pitch 20 mil, rise time 20 ps (0-100%).

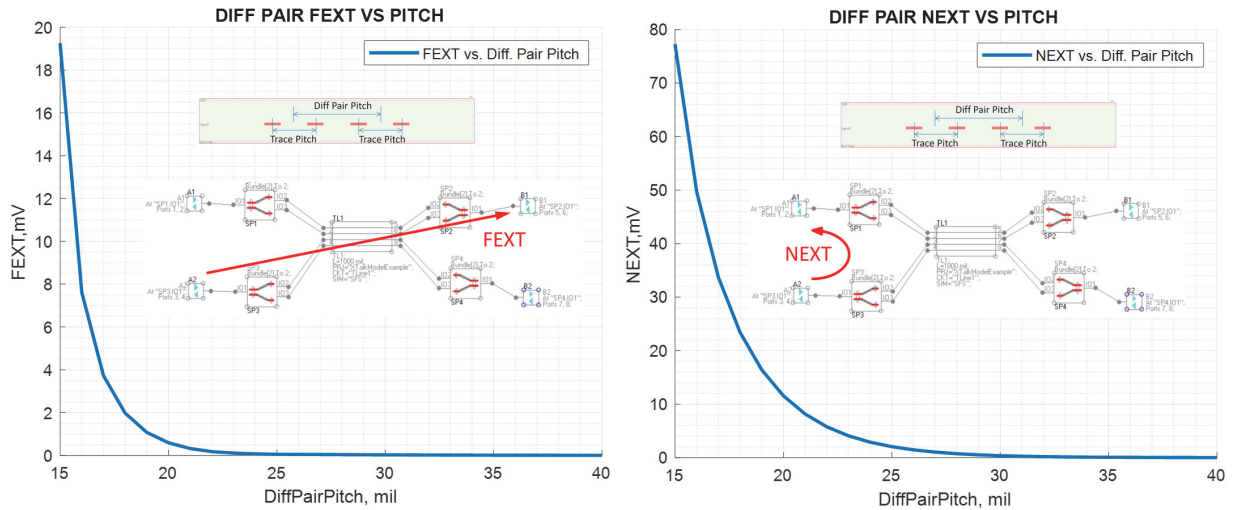


Fig. 12. Example of pre-layout simulation of crosstalk from step response in 1 inch segment of coupled differential striplines with cross-section shown in Fig. 1 versus Diff. Pair Pitch, rise time 20 ps (0-100%), similar to investigation with coupling coefficients shown in Fig. 2.

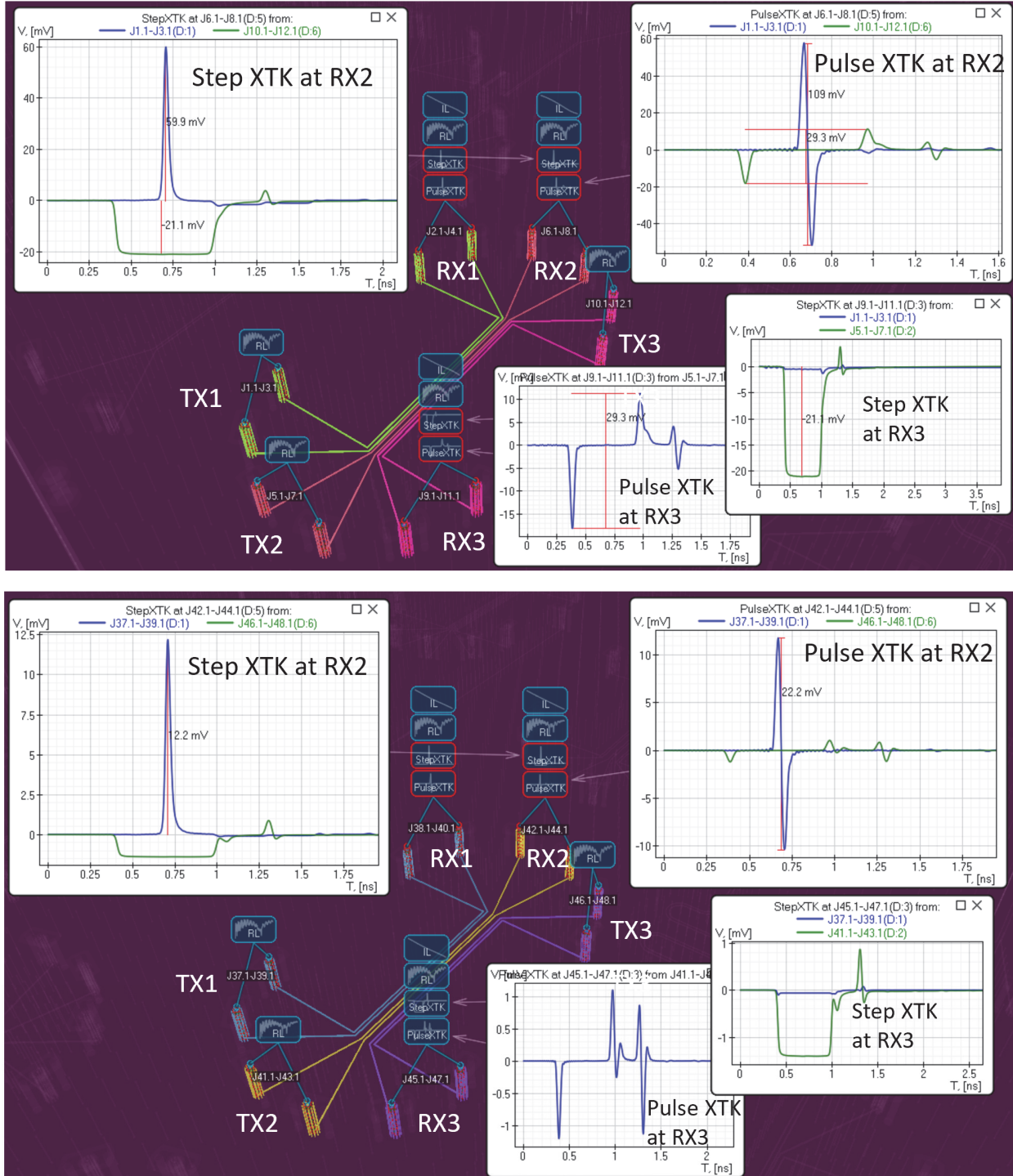


Fig. 13. Examples of post-layout time-domain crosstalk analysis with step and pulse responses (25 ps rise time) on XTALK-28/32 platform in Simbeor SI Compliance Analyzer for 2-inch differential microstrip structures with the edge-to-edge separation equal to 1 trace width (top screenshot) and 4 trace widths (bottom screenshot). TX1-TX3 are the transmitter sides and RX1-RX3 are the receiver sides.

The results of step and pulse crosstalk analyses for two structures with three differential links coupled over 2-inch parallel segments are shown in Fig. 13. Those are two structures from Fig. 6 marked as  $s=1w$  and  $s=4w$ . Differential trace width is 13.5 mil and differential trace pitch is 37 mil. In each case, two links have transmitters on one side (TX1 and TX2) and one link has transmitter on the opposite side (TX3). The pulse and step crosstalk are shown at receiver RX2 and RX3. RX2 has two aggressors – far end TX1 (blue lines on the plots) and near end TX3 (green lines on the plots). RX3 has two near-end aggressors TX2 and TX1. Overall crosstalk at RX3 is much smaller comparing to RX2. Also, the far end crosstalk at RX2 dominates, that is consistent with the investigation in frequency domain shown in Fig. 8. **Notice that the peak-to-peak far end crosstalk computed from the pulse response is almost 2 times larger than computed from the step response.** It peaks at the rising and falling edge of the pulse. Though, this is not always the case and each coupled link should be simulated, to find actual values of the crosstalk. Considering the crosstalk values, as we can see the FEXT peak-to-peak value is about 110 mV and the NEXT is about 30 mV for the structure with  $1w$  separation between the differential links. Both are below the “upper limit” 126 mV evaluated from PSXT. Though, a possible superposition of the FEXT and NEXT peaks gives us new upper boundary of the crosstalk 140 mV. The value is specific to the rise time and happens when the peaks of crosstalk from the NEXT and FEXT aggressors coincide in time – it is highly unlikely, but possible.

The step and pulse response may be useful for a preliminary quantification of the crosstalk and the upper bound evaluation. The actual signals in the links are sequences of bits for NRZ/PAM2 signal or symbols for PAM4 signals and eye diagrams are usually used to evaluate the signal distortion. Examples of eye diagrams without and with crosstalk computed at RX2 for two links with different separations between the links are shown in Fig. 14. The crosstalk impact on the eye is clearly visible in case of strong coupling (top picture in Fig. 14). The eye height is reduced by about 90 mV and width by about 0.16 of UI. This is below our upper bound estimate 140 mV – this particular sequence of bits did not include the worst-case condition that could produce ideal superposition of the FEXT and NEXT peaks at the middle of the victim eye. A smaller reduction of the eye size is observed for the links separated by  $4w$  – about 10 mV reduction in eye height and about 0.016 UI reduction in width. In both cases we observe crosstalk as additional jitter (eye width reduction) and amplitude noise (eye height reduction). That is not the worst case as well. In this example, the eyes were computed directly in time domain with PRBS32 bit stream signals in the victim as well as in the receivers. The sequences of bits in different links are not correlated and the phase or time offset of the bit rise times is defined randomly at the beginning of the analysis. The result of such analysis will depend on the particular phase and bit sequence as illustrated in Fig. 15. The top time domain plot shows a small subset of bits at the receiver ends in three coupled links. The bit sequences are not correlated and the offset between switching time is selected randomly at the beginning of the analysis. At RX2 we can observe the useful signal (red line in Fig. 15) and the crosstalk noise shown at the bottom plots in Fig. 15. The peaks in the noise are defined by the bit sequences and timing in the aggressors – those are not correlated with the victim signal. As the result, we can observe peak crosstalk at any time within the eye as illustrated by eye diagrams on the right. All graphs show the eye diagrams without the crosstalk. The eye diagram with the crosstalk for that case is shown at the top part of Fig. 14. Some bit sequences and timing offsets may degrade eye more and some less. In fact, the crosstalk may even improve the eye opening or reduce the jitter. This is possible, but highly unlikely. **Though, we are more interested in the probability of the worst case and statistical methods should be used to quantify this.**



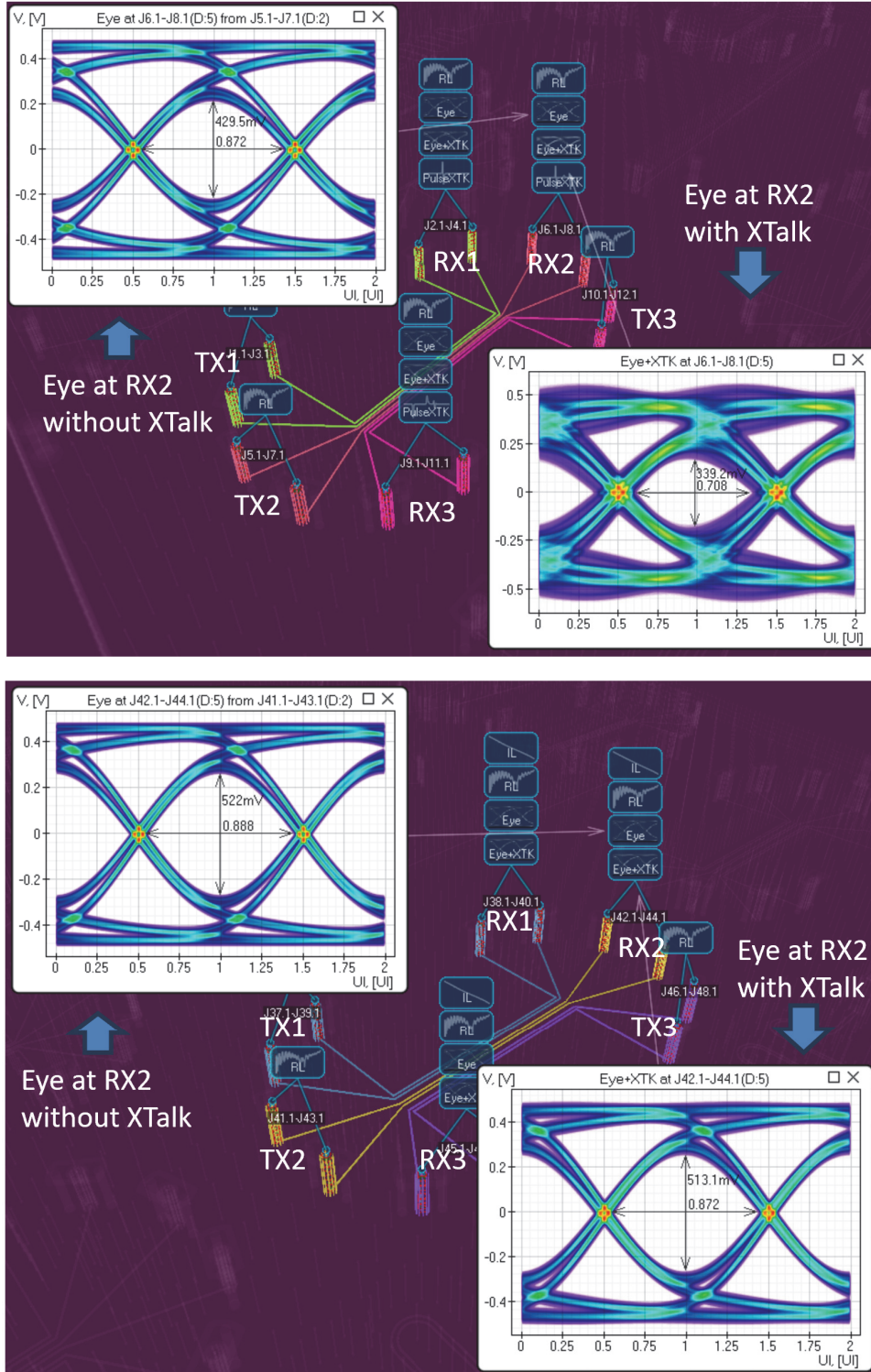


Fig. 14. Eye diagrams at RX2 without and with crosstalk computed in SI Compliance Analyzer for 2-inch differential microstrip structures from XTALK-28/32 platform with the edge-to-edge separation 1 trace width (top screenshot) and 4 trace widths (bottom screenshot), 28 Gbps, 20 ps rise/fall time (10-90%).

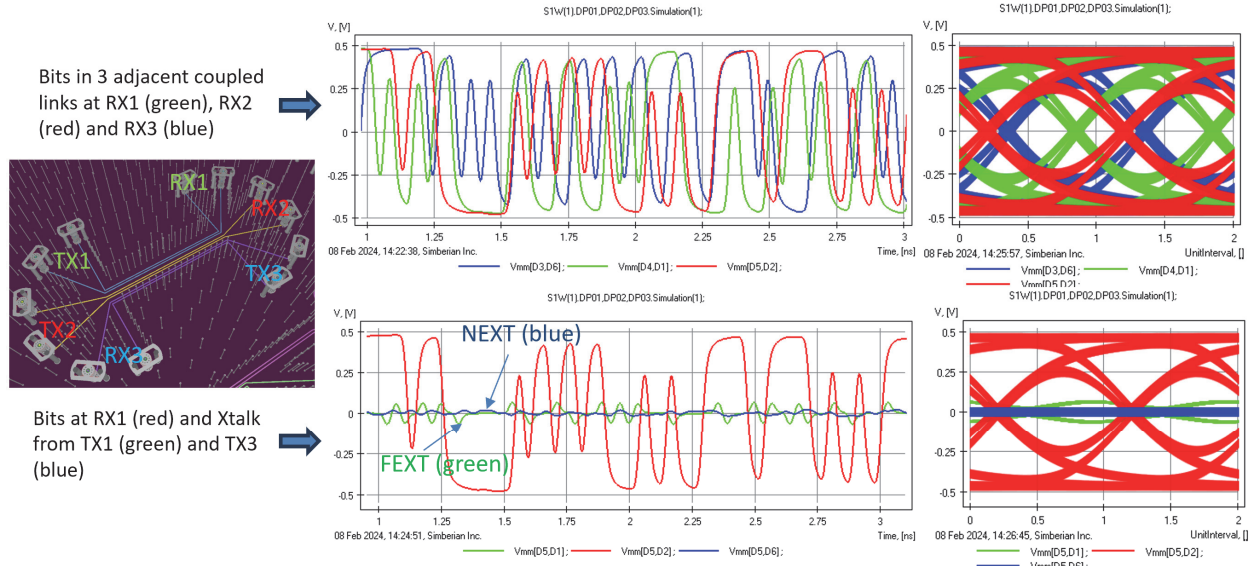


Fig. 15. Signal and crosstalk superposition example. Bit sequences in three links (top plots) and corresponding eye diagrams on the right for 28 Gbps PRBS32, 20 ps rise/fall time (10-90%). The bottom plot show bits at RX2 and the crosstalk noise from bits in the coupled links and corresponding eye diagram with location of the crosstalk noise.

#### 4. Statistical Crosstalk Quantification

The frequency and time domain analyses of the crosstalk are useful tools, but the ultimate metric for a link performance is the 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. The statistical approach to the BER evaluation requires a statistical model for a crosstalk. Though, the crosstalk is not random in general and also bounded by our worst-case estimates. A possible superposition of the victim signal with crosstalk from two aggressor links is illustrated in Fig. 16.

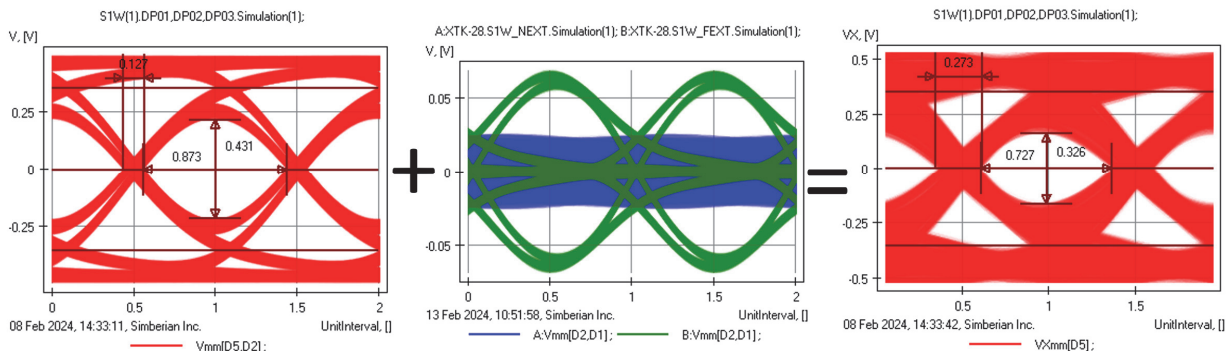


Fig. 16. Signal and crosstalk superposition example. Eye diagram without crosstalk (left picture), near and far end crosstalk components from the aggressor links (middle plot, green is FEXT and blue is NEXT) and possible uperposition of signal and crosstalk noise (right plot).

This is the same middle link case with two aggressors as at the top section in Fig. 14 and also in Fig 16. As we can see the near-end crosstalk looks like a noise, but the far end crosstalk does not look like a random signal. What is the probability to have the peak noise from FEXT and NEXT? Using the time-domain analysis we can evaluate the probability density function (PDF) for both cases as shown in Fig. 17. PDFs are computed for the tightly coupled links ( $s=1w$ , top plots) and for loosely coupled links ( $s=4w$ , bottom plots). As we can see, the NEXT distribution looks like normal – this is similar to observed in [13]. It is getting more “normal” for the loosely coupled links. Thought, the normality test is required to evaluate it. However, the FEXT distribution does not look as normal at all (PDFs on the left side in Fig. 17) and the probability to have maximal possible values is not negligible. Both distributions are bounded by the maximal possible values for a particular rise time. But the NEXT and FEXT are independent – **the total crosstalk PDF is a convolution of the two distributions – it is also bounded by the maximal possible value observed from the pulse crosstalk analysis.**

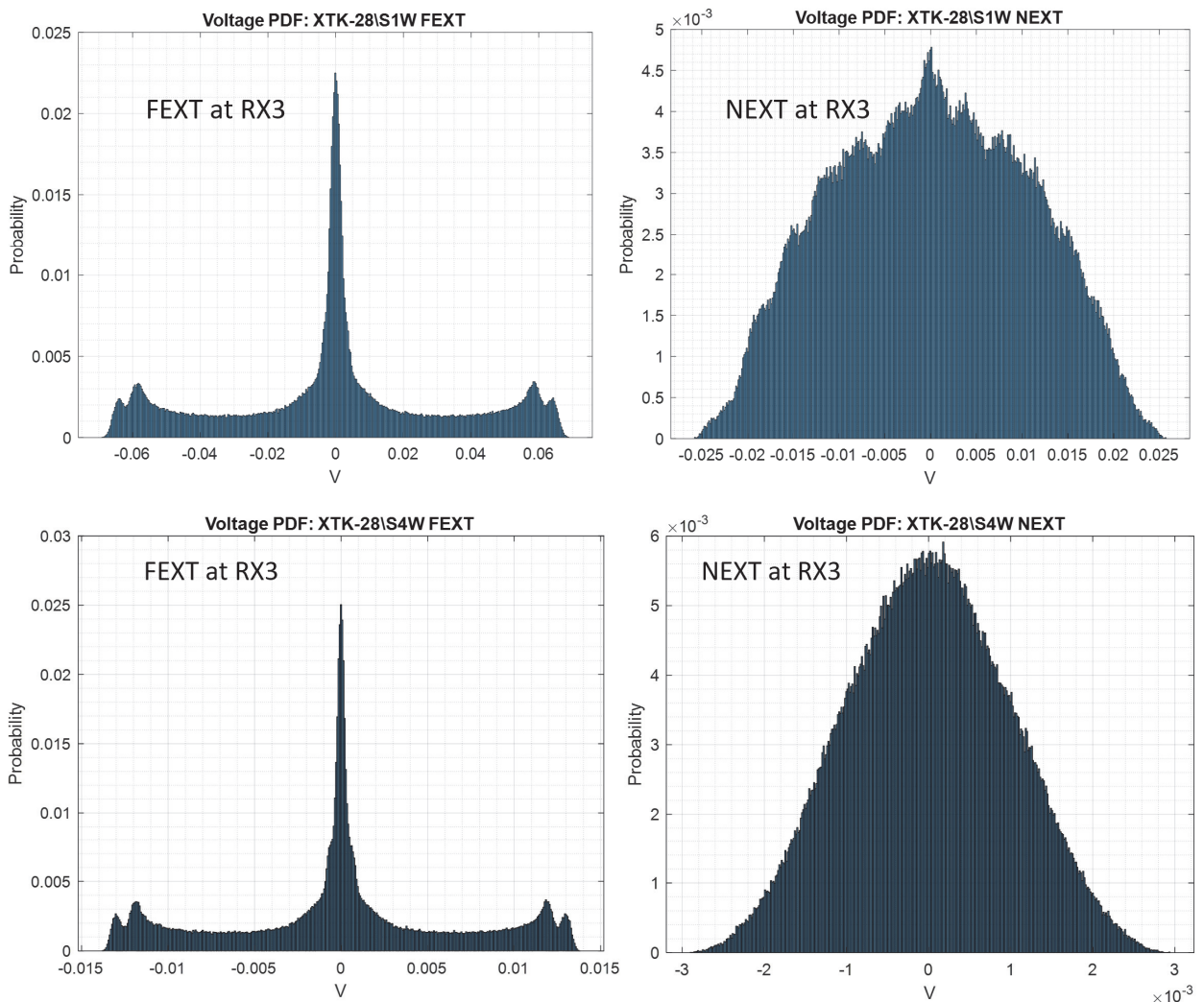


Fig. 17. Crosstalk probability density functions for 2-inch differential microstrip structures from XTALK-28/32 platform with the edge-to-edge separation 1 trace width (top plots) and 4 trace widths (bottom plots), 28 Gbps, 20 ps rise/fall time (10-90%), PRBS32, time step 2 ps.

The PDFs of the crosstalk can be used to evaluate the effect of the crosstalk on BER or on detector error rate (DER) is the channel operating margin or COM [13], [14]. This is the most “modern” way of the crosstalk quantification and may be considered as the next step in evolution of the crosstalk quantification. Channel Operating Margin [13],[14] is a signal to noise ratio defined as follows

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

Where  $A_{Signal}$  is the peak signal and  $A_{Noise}$  is the peak BER or DER noise defined through the peak signal minus the peak eye opening at a specified BER/DER level. Signal in this context includes all losses and dispersion in the link from chip to chip and the effect of equalization. It includes the reference transmitter and receiver models as well. Noise in this context includes all possible signal degradation effects with some assumptions. It includes absorption losses and dispersion, return loss, reflections and crosstalk as well as equalization by Tx and Rx. COM metric is computed in the time domain as the voltage ratio of signal available in a reference signaling architecture (Tx and Rx) to noise at the reference receiver's sampler – basically, it characterizes the complete link from chip to chip. The noise is calculated for the specified Detector Error Ratio (DER). DER is a generalization of BER for NRZ and SER for PAM4. Equalized single bit or symbol responses (SBR or SSR) of the signal link and crosstalk aggressor links are used to calculate the vertical slice of the eye diagram centered at the sampling point where DER is minimal. The crosstalk in COM is assumed to be in the middle of the eye – that is highly unlikely event.

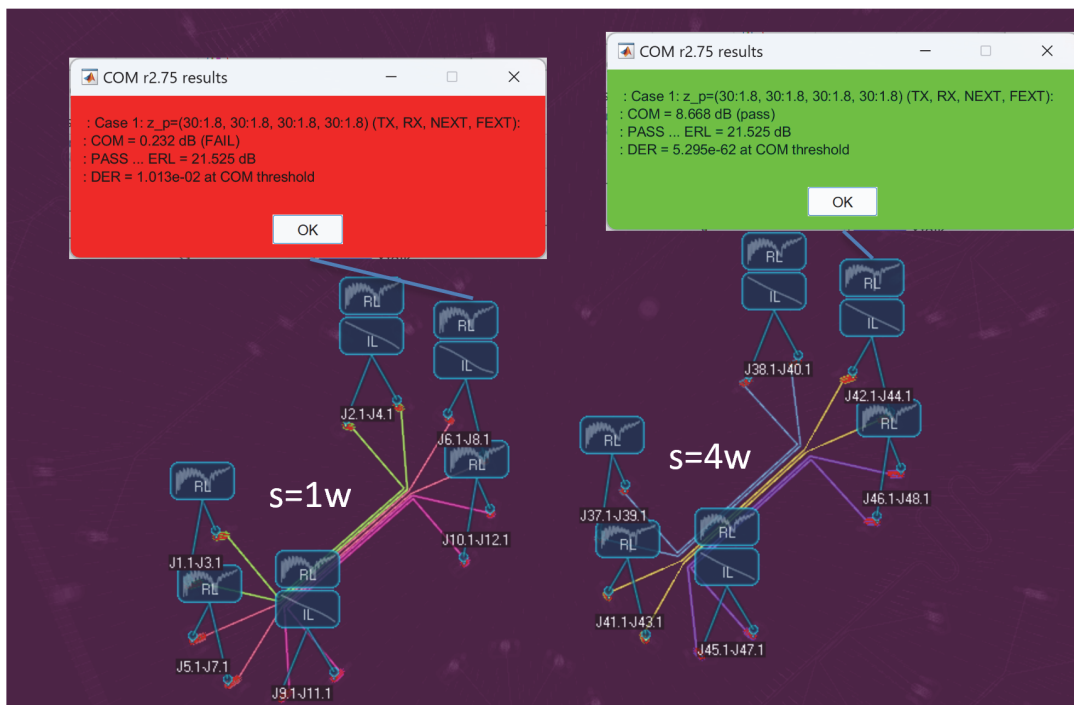


Fig. 18. Example of COM computation for 2-inch differential microstrip structures from XTALK-28/32 platform with the edge-to-edge separation 1 trace width (left) and 4 trace widths (right), 28 Gbps signal – all other parameters are from IEEE 802.3 spreadsheet.

To evaluate the crosstalk contribution, COM uses S-parameters of the crosstalk paths. The crosstalk in COM is treated as an additional bounded uncorrelated noise similar to the inter-symbol interference (ISI). For each crosstalk source COM computes probability density functions (PDFs) and convolve them with the bit PDF, to compute the overall crosstalk effect. An example of such analysis is shown in Fig. 18. The IEEE COM tool was used for this computation with default reference transmitter and receiver parameters. The limit DER is  $1e-4$  and the pass value of COM is 3 dB in this case. As we can see from Fig. 18, the middle link with 4 widths separation between differential links ( $s=4w$ ) has COM about 8.7 dB (green window in Fig. 18). Though, the eye at  $DER=1e-4$  would be almost completely closed if the links have only one width separation (red window in Fig. 17). The eye closure due to the crosstalk is highly unlikely in this case. **Thus, the COM is, probably, the most pessimistic crosstalk metric.**

## Conclusion

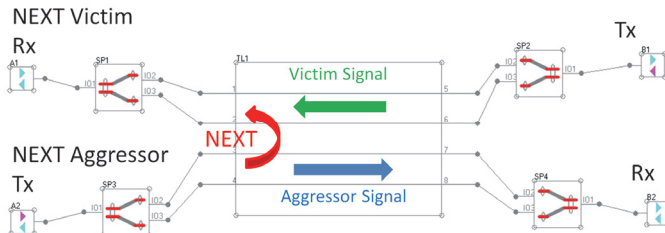
As Ransom Stephens perfectly stated in [16] - "The crosstalk problems are back." And, they are here to stay as long as we are designing interconnects as the open waveguiding systems. Thus, understanding and proper quantification of the crosstalk and mitigation of the consequences are important. This article outlines almost all possible ways to quantify the crosstalk – coupling coefficients, frequency-domain metrics, time-domain analysis of crosstalk and the modern statistical approach. The ultimate metric of the crosstalk effect is the reduction of BER due to crosstalk. The local crosstalk is deterministic but usually treated as a part of bounded uncorrelated jitter. This is because of the uncertainties in timing between the victim and aggressor signals. The most modern methods are statistical and are applicable to both local and distant crosstalk evaluation. The distant crosstalk will be covered in the third part of this paper.

## Appendix I: Terminology Relevant to Crosstalk Understanding

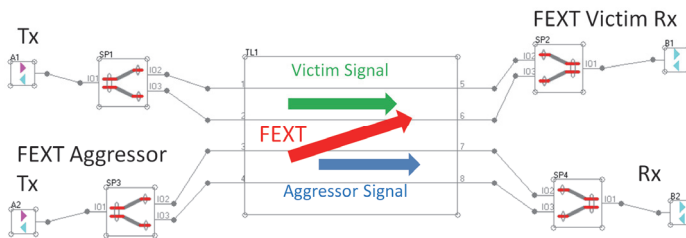
Here are some common terms used in this article:

- **Bit Rate** – number of bits transferred over one lane or link in a second, for serial interconnects usually measured in Gigabit per second or Gbps;
- **Boud Rate** – number of symbols transferred over one channel or link in second, measured in GigaBoud per second or GBps or GBoud. Equal to Bit Rate for NRZ (PAM2) and to a half of Bit Rate for PAM4 signal type;
- **Nyquist Frequency** – defines the minimal bandwidth to transmit digital signal. It is equal to half of Bit Rate for NRZ (PAM2) and to 1/4 of Bit Rate for PAM4 signal.
- **Boud Frequency (fb)** – frequency equal to the Boud Rate expressed in GHz. Used in some standards to define the compliance masks;
- **Stripline** – model for traces with two reference planes (interior layers of PCB) and mostly homogeneous dielectric;
- **Microstrip line** – model for traces routed on surface of PCB 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 the other links;
- **Victim** is a receiver (Rx) or a link that may pick up unwanted interference caused by an aggressor link transmitter;

- **NEXT** – near end crosstalk is interference observed at the link side that is closer to the aggressor link transmitter as illustrated below:



- **FEXT** – far end crosstalk is interference observed at the link side opposite to the aggressor link transmitter or driver;



- **PSXT, PSNEXT, PSFEXT** – total, near and far end power sum crosstalk for frequency-domain characterization of crosstalk;
- **MDXT, MDNEXT, MDFEXT** – multiple disturber crosstalk – total, near and far end. This is just different abbreviation for PSXTs;
- **ICR** – insertion loss to crosstalk ratio for simultaneous frequency domain characterization of crosstalk and signal losses at a receiver;
- **ICN** – integrated crosstalk noise for approximation of expected crosstalk in Volts (rms value);
- **COM** – channel operating margin;
- **BER** – Bit Error Rate;

## Appendix II: Tips for Simbeor Users

The evaluation of possible local coupling requires simulations. All cases with local coupling can be simulated with Simbeor software and the decision-making process can be further automated with Simbeor SDK and post-layout analysis automation with Lua. All examples of pre-layout crosstalk investigation are provided with description in Simbeor SDK [11]. Here are some tips for accurate analysis of crosstalk with Simbeor software.

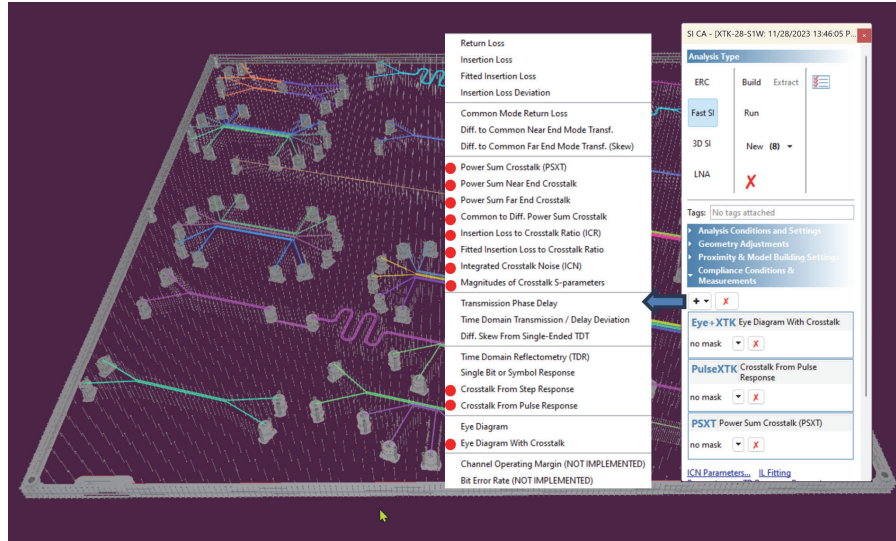
Magnitude of pulse or step signal is 0.5 V by default. It is 1 V source in series with 50 Ohm normalization impedance (or 100 differential) and about the same characteristic impedance of transmission line. Peak-to-peak voltage for Bit Stream or PRBS signal is 1 V by default (+0.5 V and -0.5 V pulses). Signal settings are very important for accurate crosstalk analysis. In particular, the rise time should be adjusted before any type of analysis. The time-domain sources can be adjusted directly in SI Compliance Analyzer in Compliance Conditions & Measurements section.

For investigation of crosstalk with step response, use “High” accuracy for rational compact models (set it as default before importing board and any type of analysis).

Coupling distance is 20 mil by default – that is typically 4-5 trace width. Increase it for analysis of traces on surface of PCB where coupling is larger at a longer distance (SI Compliance Analyzer ->

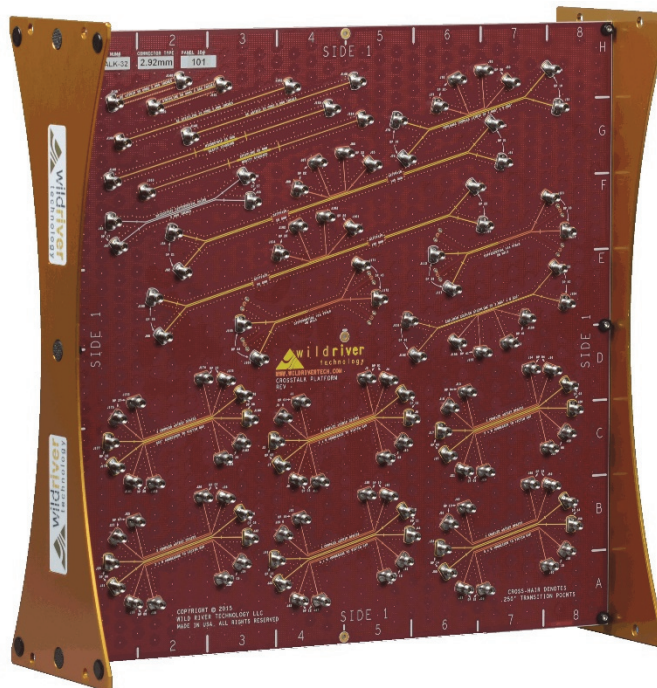
Proximity section). For XTALK-28/32 case the coupling distance should be increased to 100 mil, to simulate couplings in all cases.

To add a crosstalk metric in SI Compliance Analyzer, use “+” in the Compliance Conditions & Measurements section as illustrated below:



Next, define a mask or a simple compliance condition, select a link in Board View and click Run.

### Appendix III: XTALK-28/32 Validation Platform



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

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