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Characterizing, anticipating, and avoiding problems with crosstalk

Ransom W. Stephens, Ph.D.
Research Scientist, Teraspeed Consulting Group, LLC
ransom@teraspeed.com

Al Neves
Senior Staff Engineer, Teraspeed Consulting Group, LLC
al@teraspeed.com

Abstract

As data rates increase, interference from signals on other channels, crosstalk, increasingly affects the utility of data transmission paths. Crosstalk can cause a variety of problems for a signal that depends on both the character of the aggressors and on the geometric relationship of the signal and aggressor channels. Using examples from the Common Electrical Interface (CEI) and Fully Buffered DIMM, we demonstrate that accurate crosstalk characterization requires a combination of both jitter and amplitude noise analysis. Understanding how to characterize crosstalk provides the insight necessary to both anticipate and avoid crosstalk-related problems.

Author Biographies

Ransom Stephens is a Research Scientist with Teraspeed Consulting Group who specializes in the analysis of electrodynamics in high-rate digital systems. He has served on the electrical working groups for the PCI Express, Fully Buffered DIMM, and OIF Common Electrical Interface standards bodies. He earned the Ph.D. in Physics at the University of California, Santa Barbara.

Alfred P. Neves is a Senior Staff Signal Integrity Engineer at Teraspeed Consulting Group, LLC that focuses on measure-based model development, package characterization, high-speed board design, low jitter design, analysis, and training. He earned a B.S. in Applied Mathematics at the University of Massachusetts.

Introduction

This paper focuses on crosstalk problems at data rates greater than 1 Gb/s suffered by signals propagating on Printed Circuit Boards. At these data rates digital signals cannot be assumed to propagate independently along conducting traces etched on an insulator. Rather, the conductor behaves like a waveguide and the insulator like the dielectric medium containing the wave. In a system with several signal-carrying traces, crosstalk between different signals can degrade a system's bit error ratio (BER) beyond the critical level.

By the time the problem has been discovered through a high BER, it's too late to fix. Careful analysis signal integrity analysis at different points in the design process can illuminate problems with crosstalk and help designers solve them.

Crosstalk occurs when a signal is affected by the transmission of data on a neighboring transmission line. We refer to the affected signal as the *victim* and the affecting signal as the *aggressor*. Crosstalk is caused by the propagation of electromagnetic radiation from the conductor guiding the aggressor signal to the conductor guiding the victim.

Crosstalk can be a problem at any data rate, but especially at high data rates where the electrical signal no longer resembles a simple DC current. At high data rates electrical signals propagate as electromagnetic waves guided by conductors but radiating through the dielectric medium surrounding the conductor. In most applications the conductor is a conducting trace like a microstrip on a dielectric like flame-retardant type 4 (FR4). FR4 serves as the medium through which the signals propagate. The conducting trace behaves like a guide, but the signal is free to radiate throughout the FR4 medium. The result is that a signal on any single conducting trace includes artifacts of the signal on every other conducting trace on the board.

The use of differential signaling schemes, that is, by applying a signal and its inverse to closely neighboring traces and then subtracting one from the other at the receiver, greatly reduces crosstalk noise. At high data rates nonzero skew and separation of the differential traces allow crosstalk contamination.

The only reason we care about crosstalk is its effect on the Bit Error Ratio (BER) of a system. The nature of the crosstalk aggressor has a tremendous effect on the appearance of noise on the victim. If the crosstalk and the signal are frequency locked – that is, if they have a fixed phase relationship – then crosstalk can appear as jitter or amplitude noise on the victim. If the frequencies are not locked, then it appears as a combination of both jitter and amplitude noise. In terms of signal integrity analysis the repercussions can be profound if the relative frequency structure of the aggressor and victim are not known. For example, if a victim is analyzed under the assumption of negligible amplitude noise but appreciable jitter by measuring BER(x), the BER as a function of the sampling delay (a.k.a., bathtub plot or BERTscan), and the victim and aggressor are frequency locked and 90 degrees out of phase, then BER(x) may not be monotonic; it could look more like a double-sink than a bathtub with sloping sides but a big hump in the middle.

The characterization of a signal in the presence of crosstalk requires both amplitude and jitter analysis. Wide bandwidth equivalent-time sampling oscilloscopes and high quality Bit Error Ratio Testers (BERTs) can be used for characterization by treating crosstalk as a noise source. Vector Network Analyzers (VNAs) and Time Domain Reflectometry (TDR) equipment can be used to measure the frequency and attenuation response of the traces on a circuit board, that is, the *S*-parameters, from which crosstalk effects can be calculated.

The prudent approach to circuit design is to lay out the board in such a way that crosstalk problems are negligible. By the time a prototype is available for signal integrity analysis considerable effort and expense have already been exhausted. An effective simulation can help you design a system in a way that minimizes the effects of crosstalk before you employ a quarter million dollars worth of test and measurement equipment.

This paper begins with a review of the electromagnetic causes of crosstalk followed by a discussion of crosstalk physics, how the use of differential signaling reduces crosstalk effects and then enters a lengthy discussion of how crosstalk can be analyzed.

Electromagnetic interference

In a system with more than one signal-guiding trace, crosstalk is caused by the dielectric properties of the Printed Circuit Board (PCB), the layout of the conductor, and the voltage swing and frequency content of the signal. In other words, it’s caused by geometry, media, and electrodynamics – everything that goes into a circuit.

The fundamental frequency of a data signal is given by a clock frequency at half the data rate. Since digital data signals are based on square waves, their frequency spectrum includes harmonics well above the fundamental frequency. Data rates of 1-10 Gb/s are well into the radio, toward the microwave, realm of the electromagnetic spectrum. At these high frequencies the signal is less like a well contained current density than it is like a multi-pole oscillator – essentially an antenna. The problem is exacerbated by discontinuities and edge-effects at connectors and vias.

At high data rates the PCB behaves like a dielectric waveguide, which is why the medium plays such an important role. Digital signals are transmitted as electromagnetic waves that flow through the FR4 dielectric medium guided by the conducting trace. A PCB is a very complicated waveguide – not at all like an ideal conducting pipe of uniform geometry – and there is no closed-form analytic solution to Maxwell’s Equations [1]. But the system is cheap and, with careful design and some tricks like equalization, can be made to work.

Since electromagnetic fields are linear, crosstalk noise occurs when the field of the aggressor induces current oscillations (i.e., eddy currents) on the trace guiding the victim. The aggressor signal simply adds through superposition to the victim signal. The greatest amount of electromagnetic radiation, and therefore noise, is emitted during a logic transition; thus the rise/fall time of the signal sets the crosstalk time-scale. Here are two equivalent ways of understanding the underlying cause of crosstalk from the perspective of Classical Electrodynamics.

Maxwell’s equations show that electromagnetic radiation is caused by one and only one phenomenon: the acceleration of electric charge. In the systems of interest the acceleration is due to rapidly oscillating currents. In the context of a digital signal, the oscillating current occurs during logic transitions. The amplitude of the radiation is proportional to the acceleration of the charges or, equivalently, to the rate of change of the current, di/dt, which is proportional to the rate of change of the aggressor voltage, dV/dt.

The other argument is elegantly presented in Ref. [2] and is based on Faraday’s law of inductance. Crosstalk is caused by an electromotive force induced in the conductor that carries the victim’s current. The crosstalk electromotive force is proportional to the rate of change of the magnetic flux, $d\Phi_B/dt$, passing through the surface area enclosed by the conductor that carries the victim signal. In other words, crosstalk is caused by the mutual inductance of the conductors that carry the victim and aggressor signals. Since $d\Phi_B/dt$ is proportional to the rate of change of the aggressor current, di/dt, we again arrive at the conclusion that crosstalk is proportional to the rate of change of the aggressor voltage, dV/dt.

$$V_{Crosstalk} \propto \frac{dV_{Aggressor}}{dt}$$

Think of the trace as the source of the electromagnetic field. For a long conducting trace embedded in a large volume of dielectric, the field strength decreases linearly with the reciprocal of the distance from the conductor. The electric permittivity of the medium affects how far and with what strength an aggressor signal can reach a victim. The dielectric “constant” varies with the signal frequency, $\varepsilon(f)$, but the determining factor isn’t the variation in the response of the medium with signal frequency (i.e., dispersion), it is the value of ε near the fundamental frequency. The industry standard medium, FR-4, has a dielectric permittivity that varies weakly with the frequencies relevant in serial-data technology.

FR-4 introduces design complications through variations in its density and asymmetries in its structure. FR-4 is a laminate of primarily glass and resin. The two main constituents have different dielectric constants and, since they are not woven in a uniform fashion, exhibit specific symmetry. Signal propagation differs depending on its orientation; a signal propagating “with the grain” is a waveguide with different boundary conditions than a signal propagating “against the grain.” The variations in the dielectric properties complicate the distribution of electromagnetic fields.

More intractable complications arise from multi-path effects and reflections. In a real PCB with several traces that turn corners, have vias, connectors and components, the dielectric medium contains radiation from every signal on every trace. The farther the radiation propagates through the medium, the greater it is attenuated. Part of the signal is absorbed and part reflected at every boundary, every discontinuity and every impedance mismatch. The sum of all fields in the medium can be quite a mess.

The easiest way, in principle, to reduce crosstalk is to displace victims from aggressors by large distances or to shield them from each other with grounded conductors.

Crosstalk

Consider the simple case of two parallel traces, one a victim and the other an aggressor. If the two forward propagating signals are locked in frequency and have zero relative phase, then the bursts of noise from the aggressor that occur on each aggressor logic transition contaminate the victim signal at fixed points in the eye diagram. Figure 1 shows two examples of frequency locked crosstalk. The victim is a clock signal that corresponds to a data rate of 4.25 Gb/s, the aggressor is a clock signal at half that rate. On the left, the aggressor is in phase with the signal and, on the right, it is about sixty degrees out of phase. At each transition of the aggressor the amplitude of the victim is shifted. There are three distinct trajectories, one where the aggressor transition is from low to high, one where the aggressor has no transition and one where it is from high to low.

Two key points emerge. First, the time scale of the crosstalk pulse is given by the rise/fall time of the aggressor signal; and, second, when the victim and aggressor are in phase the crosstalk appears as jitter but when they are out of phase it appears as amplitude noise. Figure 2 shows the same system without the victim and aggressor frequency locked – crosstalk appears as a combination of both amplitude noise and jitter.

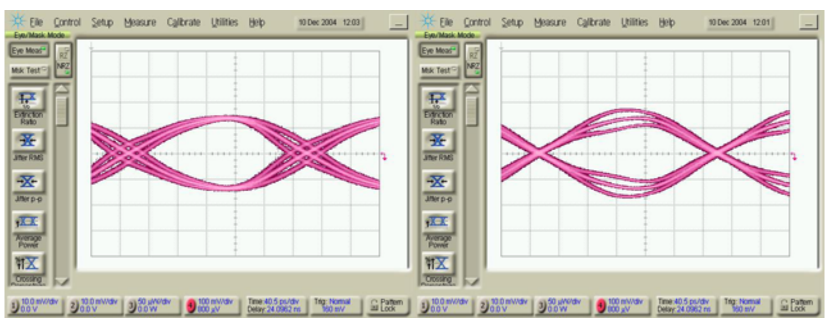

Figure 1: Frequency locked crosstalk in a nearly ideal single-ended system. The victim and aggressor are in phase on the left and about 60^o out of phase on the right.

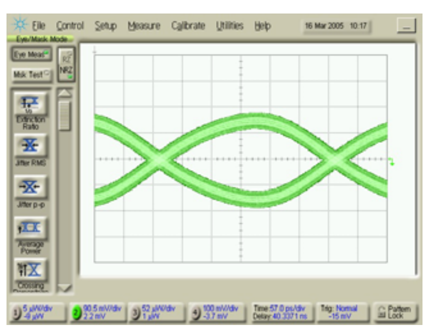

Figure 2: Crosstalk in a nearly ideal system that is not frequency locked.

In a real system traces do not run parallel; they turn corners, have different lengths, and the distance between them varies over the course of a given trace. The waveguides become more complicated with multiple paths and reflections. As the complexity grows the noise experienced in the frequency locked case becomes less distinct and resembles the unlocked case. Figure 3 shows the effect on a more realistic system with a single aggressor on a single-ended system.

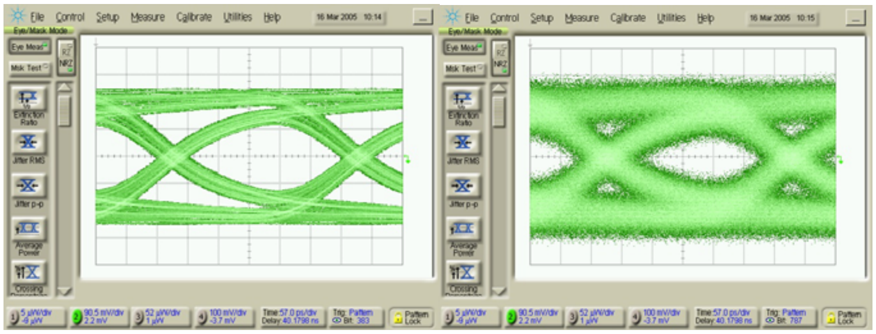

Figure 3: Comparison of a signal without crosstalk, on the left, and with crosstalk, on the right, in a realistic single-ended system.

Crosstalk caused by a victim that is propagating in the same direction is called Far End crosstalk (FEXT) and crosstalk caused by a victim propagating in the opposite direction of the victim is called Near End crosstalk (NEXT). The example above is FEXT.

Differential signals

In an ideal differential system the two differential traces have zero skew (are exactly the same length), perfectly overlap (zero separation), have ideal connectors and have identical but inverted signals on the traces. In this ideal case, the noise picked up on each trace is identical. When the two signals are subtracted at the receiver the noise is completely removed. Figure 4 shows the effect of crosstalk on a nearly ideal differential system. The eye diagram on the left has no aggressor and that on the right has an aggressor. Compare the left of Figure 4 with Figure 2 to see how using a differential-ended system reduces the effect of crosstalk.

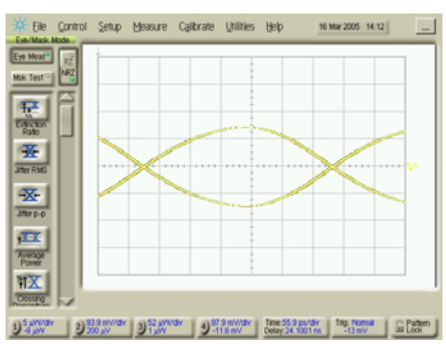

Figure 4: Comparison of a nearly ideal differential system without crosstalk, on the left, and with crosstalk, on the right.

Of course, in a real system the two traces cannot overlap without shorting the transmitter; it's impossible to manufacture a circuit with two traces of exactly the same length and the impedance of two connectors is never exactly the same. Crosstalk sneaks through differential signaling schemes for two reasons. Since the differential traces are separated by some distance, the aggressor signal is not quite the same on both traces and the difference in the aggressor at the two traces is not removed when the differential signals are subtracted at the receiver. The skew between the two differential signals is a simple imperfection in the differential scheme that causes common mode noise. In other words, rather than having the signal and its inverse subtract, over small distances – the net skew – the two signals actually add, including the aggressor.

Another advantage of using a differential signaling scheme is that the aggressor is probably also differential. An ideal differential aggressor would cancel itself through superposition over the volume of the media. The same imperfect skew and separation issues result in the aggressor signal permeating the medium.

Crosstalk analysis

Figure 5 shows the pulse shape of several different aggressors observed on a victim. Since electromagnetic fields are linear, crosstalk adds to the victim signal. Crosstalk is fundamentally amplitude noise but causes jitter. When crosstalk adds to a victim, the timing of the victim's logic transitions is changed in a way that depends on the rise/fall time of the victim signal. In the limit of zero rise/fall time, the addition of amplitude noise wouldn't affect transition timing; but adding amplitude noise to a smooth transition changes the shape of the transition which changes its timing. Consider a sine wave. The transition point – halfway between the peak and the trough – is approximately a straight line of slope *dV/dt*. Adding an aggressor pulse shape like that shown in Figure 5 pulls the whole line up or down and changes the timing of the transition point. The change in timing is what we observe as jitter in Figure 1a.

A simple rule-of-thumb for the peak to peak jitter impact of crosstalk is

$$J_{PP\ Crosstalk} \approx 2V_{PP\,Crosstalk} \times \frac{T_{rise/\ fall}}{V_{PP\,Victim}} \tag{1}$$

where $V_{PP\ Crosstalk}$ is the peak to peak voltage of the crosstalk signal like that shown in Figure 5, $T_{rise/fall}$ is the 20-80% rise or fall time of the victim, and $V_{PP\ Victim}$ is the peak to peak voltage swing of the victim signal. As the number of aggressors increases, the rule of thumb, Eq. (1), expands to include the sum of terms for each aggressor.

Here is the derivation of Eq. (1): first, approximate the waveform during a transition as a straight line with slope ½ $V_{PP\ Victim}$ / $T_{rise/fall}$; the factor of ½ comes from the 20-80% rise/fall time which only includes 60% of the transition, about ½. The waveform shape at the transition is then

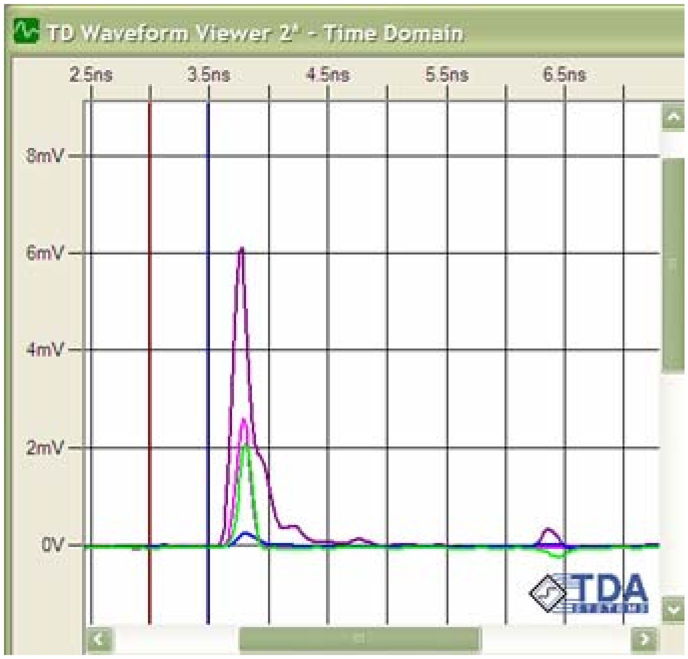

Figure 5: The pulse shapes of several different crosstalk signals.

$$V(t) \approx \frac{V_{PP\,Victim}}{2T_{rise/\,fall}}t \qquad (2)$$

where we set the nominal transition time at $t = 0$. Now add the maximum crosstalk, $V_{PP\ Crosstalk}$, to Eq.(2) and determine the new time when V = 0 to get Eq. (1).

You can improve the accuracy of Eq. (1) by substituting the reciprocal of the slope at the crossing point of the slowest rising/falling edge in a data signal, $(dV/dt|_{\min})^{-1}$, for $2(T_{rise/fall}/V_{PPVictim})$,

$$J_{PP\ Crosstalk} \approx \frac{V_{PP\,Crosstalk}}{\left.\frac{dV}{dt}\right|_{\min}} \qquad (3)$$

Never forget when analyzing crosstalk that it has the attributes of both amplitude noise and jitter:

- Random noise in the aggressor introduces both random noise and random jitter in the victim.
- Data dependent noise in the aggressor introduces deterministic noise in the victim that is correlated to the data pattern of the aggressor but not correlated to the data pattern of the victim.

- Periodic noise in the aggressor introduces periodic noise – at the same period – to the victim.

- Duty cycle distortion in the aggressor introduces noise in the victim at the aggressor's data rate and integer fractions thereof.

Let x represent the time-delay of a bit and V its voltage level. In other words, x is the horizontal axis of an eye diagram and V is the vertical axis.

Accurate characterization of the crosstalk on a signal by performing a Q analysis on the amplitude noise and jitter analysis on the timing noise is not complete. If standard serial-data jitter analysis techniques are used in the presence of crosstalk, the apparent periodic and random jitter measurements can exhibit large variations with the addition of aggressor signals and yield absurdly inaccurate results. Similarly, a measurement of the bit error ratio as a function of the sampling-point time-delay, BER(x) (a.k.a., bathtub plot or BERTscan), may not be monotonic – anomalous changes in the slope of BER(x) may occur and local maxima and minima may appear.

A time domain signal integrity analysis of a system with crosstalk should not be delegated to the realms of either jitter or amplitude noise alone, rather, the signal should be analyzed in both time-delay and voltage space.

Analysis on a BERT

The only reason we care about crosstalk is because of its impact on the bit error ratio. The direct route to measuring crosstalk's impact is to measure the BER, but a simple BER measurement doesn't reveal much information.

Instead, a BER contour measurement can tell us everything there is to know about the signal that a receiver must cope with. A BER contour measurement at the low BER levels of interest in digital systems can only be performed by a BERT but they take a very long time.

A BER contour measurement is given by the bit error ratio as a function of time-delay and voltage amplitude, BER(x, V). It is performed on a BERT by scanning the sampling point of the BERT error detector across the eye and measuring BER for each point. Figure 6 shows two BER contour measurements of data signals with moderate to large levels of jitter and very little amplitude noise measured to BER of about 10^{-6}. At 5 Gb/s a BER contour that dips down to BER ~ 10^{-6} would take a few minutes but for BER ~ 10^{-12} it would take about 24 hours. But just as a bathtub plot, $\mathrm{BER}(x)|_{\mathrm{fixed}\ V}$, can be extrapolated to predict the BER at different values of the time-delay, x [3], the two-dimensional BER contour, BER(x, V), can be extrapolated to predict the BER at different points in (x, V) down to arbitrarily low values of BER.

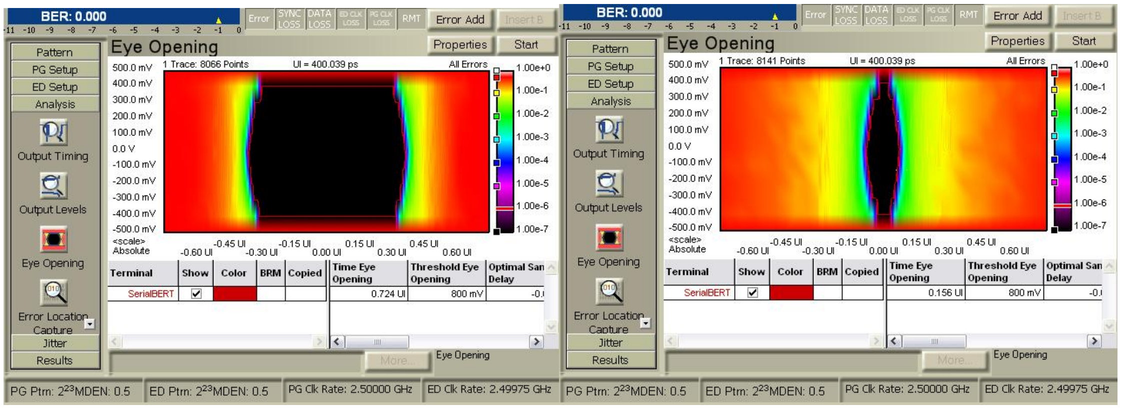

Figure 6: BER contour plots, BER(*x*, *V*), for a signal with appreciable jitter.

Analysis on a scope

The crudest form of a contour plot is a mask test. Mask tests reveal information about the eye opening in both dimensions, *x* and *V*. Just as eye-diagram analyses are most accurately performed on the low noise, high bandwidth architecture of equivalent-time sampling oscilloscopes, so are mask tests but the sampling rate of equivalent-time scopes is too low for them to accumulate sufficient statistics to convert a mask test into a useful measurement of BER(*x*, *V*). Just as developers of oscilloscopes came up with clever algorithms for jitter analysis and the prediction of total jitter defined at low bit error ratios such as JIT3 from Tektronix and DCA-J from Agilent Technologies, it is possible to extend the jitter analysis technology in the vertical direction to analyze eye closure in every direction of the (*x*, *V*) plane. Such analyses should be capable of providing fast, accurate extrapolations of BER(*x*, *V*) along with different data visualizations to separate both amplitude noise and jitter into random, periodic, sub-rate, and, perhaps, be able to extract crosstalk from other noise sources, for circuit problem diagnosis.

Extrapolation of BERT measurements should be more reliable than extrapolation of oscilloscope measurements by virtue of the fact that a BER(*x*, *V*) measurement can be performed with a thousand times more statistics than can be accumulated on an oscilloscope (either real-time or equivalent-time) in a given length of time.

Analysis with Vector Network Analyzers and Time Domain Reflectometers

Transmission system behavior can be predicted by analyzing the frequency/attenuation response of a circuit with a Vector Network Analyzer (VNA) or the impedance response of a circuit with a Time Domain Reflectometer (TDR).

Crosstalk results from coupling – whether it's capacitive or inductive, a reactance matrix can be built for its analysis. Let the matrix elements be C_m and L_m. The coupling, whether differential or single-ended, can be measured with differential TDR methods and associated software tools. When the odd and even transmission modes are equal there is no crosstalk, which provides a useful quick tool to analyze crosstalk in systems such as connectors on high data rate backplanes. Additionally, by analyzing the circuit response in both the time and frequency domains and comparing with eye diagrams it is possible to generate relationships between jitter characteristics and coupled transmission characteristics.

Consider the system shown in Figure 7 which has is significant coupling between the differential systems.

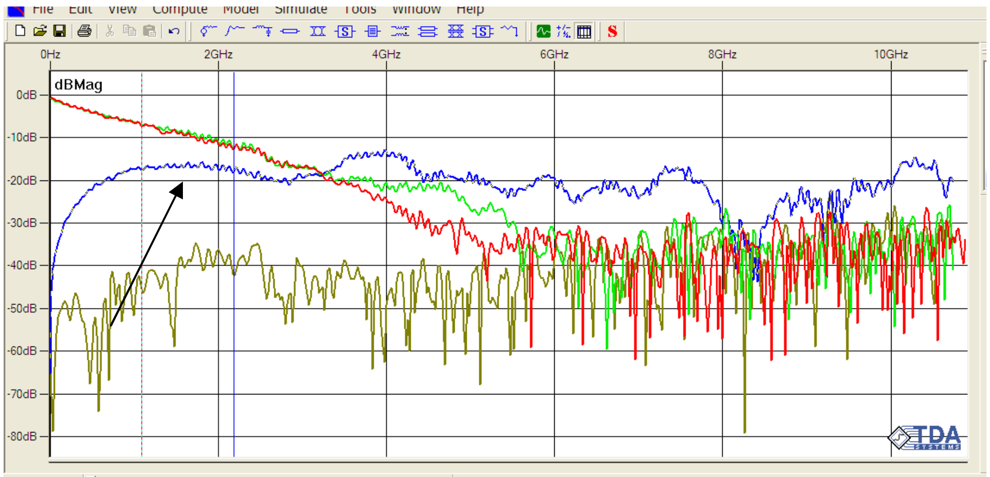

Figure 7: The arrow shows the coupling increase on adjacent signals as a function of frequency, where loss, S_{dd21}, and coupling S_{dd13} are approximately equal at 3GHz. This data was collected using time domain reflectormetry.

Figure 8 was analyzed by collecting several measurements, reference, transmission, and TDR reflection and IConnect 3.6.2 was used to analyze the data in the frequency domain.

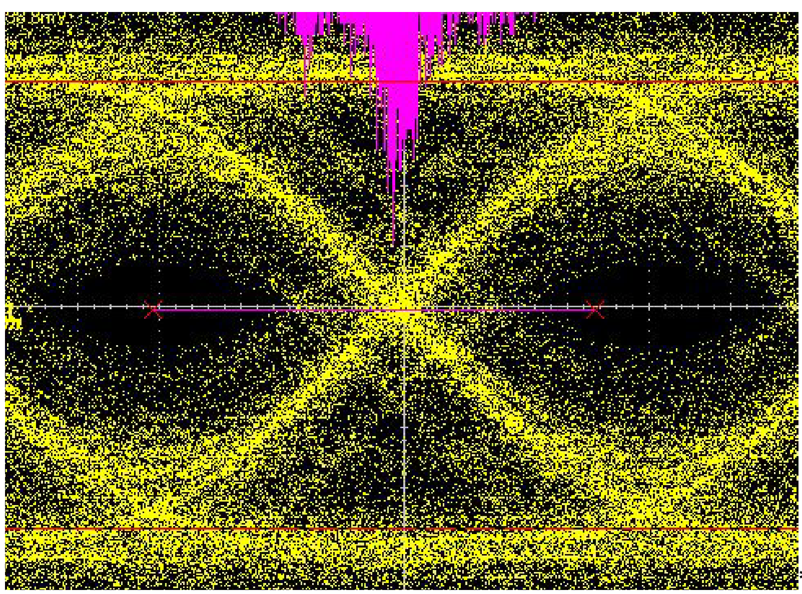

Figure 8: At 3.125 Gb/s and aggressor pair at 600 mV peak-peak with a PRBS pattern, there is considerable jitter. This figure corresponds eye degradation to crosstalk evident in the frequency domain data calculated from TDR measurements. Close examination of the jitter shows that it follows a truncated Gaussian distribution..

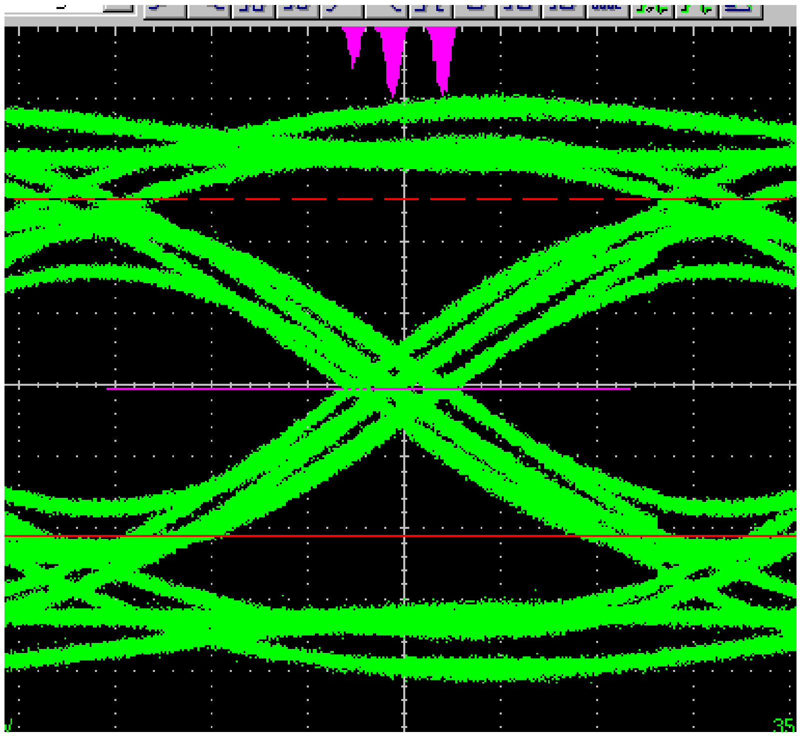

Figure 9: With aggressor output disabled, ISI is still evident and can be compared to the loss characteristics, S_{dd21}, in Figure 7.

Comparing coupling in the frequency domain and eye diagrams measured on a scope provides an easy assessment of crosstalk.

Crosstalk and device packages – A novel measurement approach

In dense circuits, such as packages, coupling between closely spaced signals causes jitter, attenuation, impedance mismatches, resonance and crosstalk which decreases signal to noise and limits interconnect performance. While this discussion is not intended to supplant 3D electromagnetic solver modeling approaches, it should provide some novel measure-based methodology that can supplement 3D analysis.

Consider the problem of coupling internal to a large BGA package and its subsequent jitter-impact on the system. The usual approach involves measuring S_{13} (Near End Xtalk, NEXT) and S_{14} (Far End Xtalk, FEXT) using a calibrated VNA but there is no forward transmission capability for measuring S_{41} making a direct measurement impossible. However, the crosstalk can be assessed by using TDR measurements to generate a model. It's important to verify that the simulation is consistent with measurements. With the model, not only can the crosstalk due to the package be determined, the sources of coupling and, therefore, crosstalk can be decomposed by de-embedding the model topology such as the ball, traces and even the launch.

The first step of this process is to acquire the data. In this case we are characterizing a package only, not a device which may have internal CML type termination, i.e., the package is open terminated. A reference waveform and both the odd and even reflected waveforms are acquired for the victim line, where the aggressor is driven both 180 degrees out of phase, odd mode, then 0 degrees, even mode. The reference waveforms are then used to calculate the impedance profiles of both modal propagations. Coupling, which generates the crosstalk, is clearly evident in the impedance profiles along with analysis of other relevant pathologies including resonance and reflections due to impedance variations. Coupling between conductors is often expressed in C_m and L_m mutual coupling.

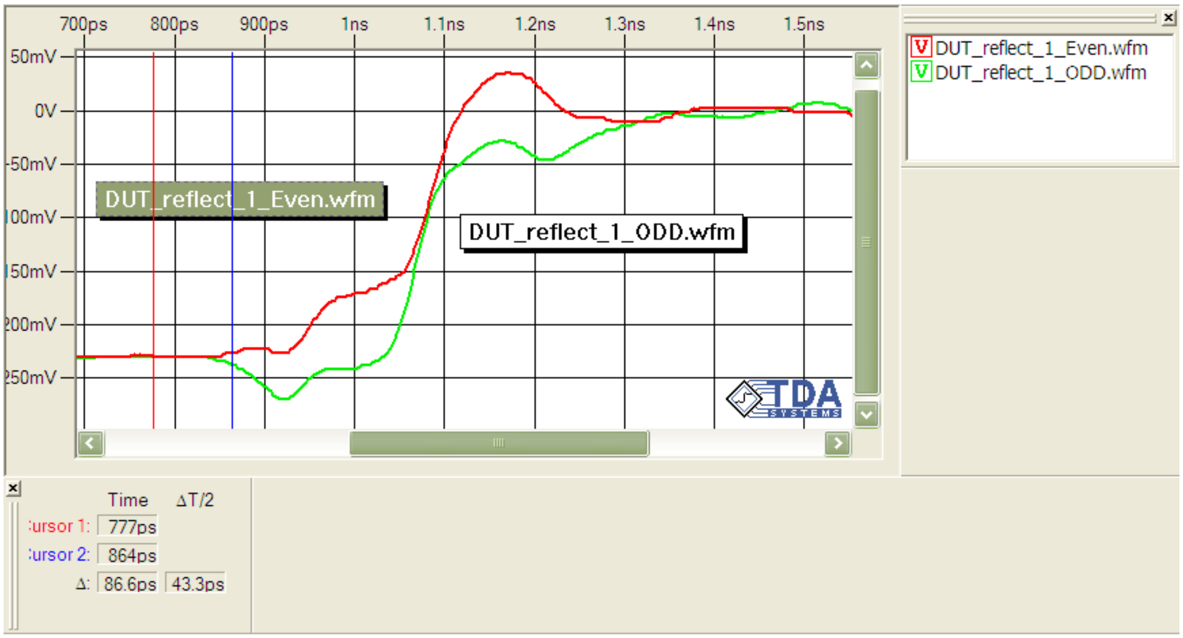

Figure 10: TDR measurements of both even and odd modes with the aggressor pair driven in phase and 180 degrees out of phase (ODD).

The impedance profile provides another perspective. The strength of the coupling is evident in the vertical separation between modes and can be separated topologically in an appropriate model.

Figure 11 shows the Impedance profile. Each part of the package structure is evident.

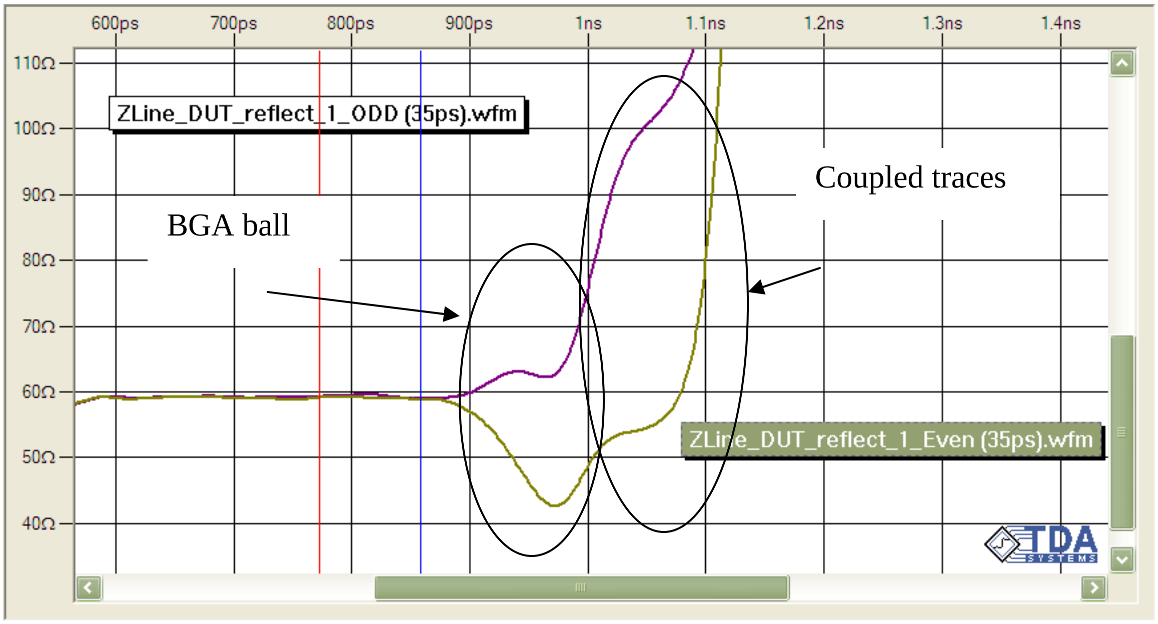

Figure 11: The impedance profile can be used to determine coupling in the system.

Using the impedance profile calculated from the voltage TDR profiles and reference calibrations, a model can be constructed. The model components are the sources, a symmetric coupled model of the ball, and one uniform symmetric coupled model of the package traces.

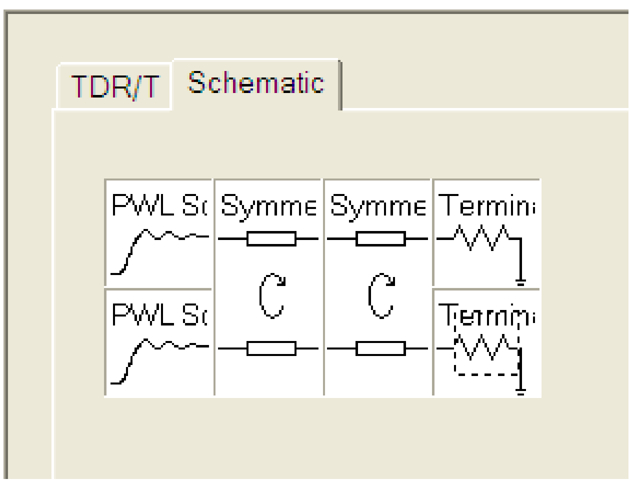

Figure 12: A simple model separating out the BGA ball and the coupled package trace as described in IConnect 3.6.2.

Figure 13: TDR of the odd reflected TDR pulse with the simulation results. The correspondence establishes the integrity of the measure-based method of determining package impact on system crosstalk.

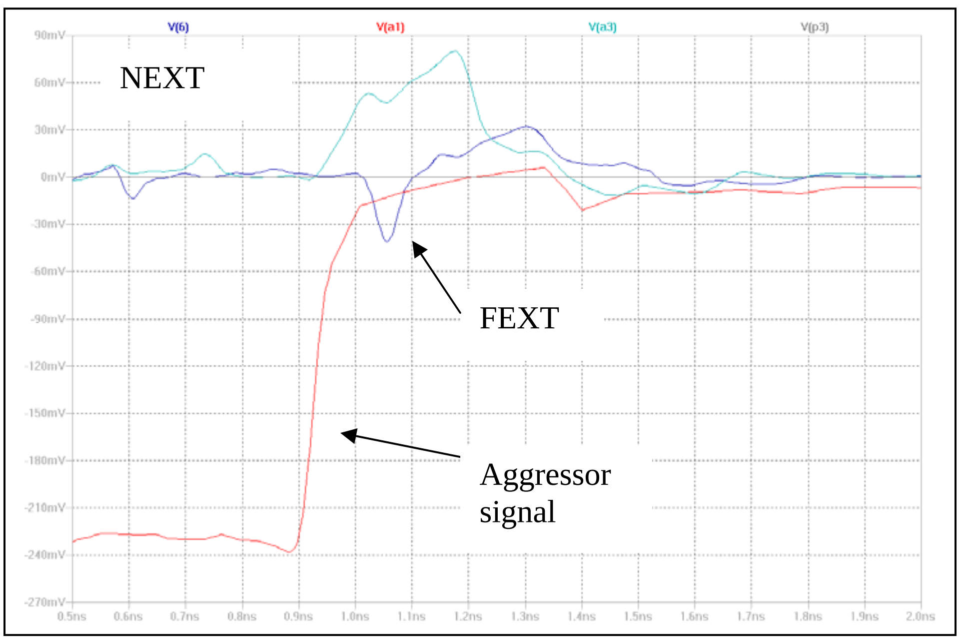

Figure 14: Using IConnect 3.6.2 a symmetric coupled model of the package was create, then directly ran the generated composite file in a freeware copy of LTSpice for a more flexible simulation platform. Crosstalk contribution is 32% for NEXT and 18% for FEXT.

Since it is a topological model, it can extend the analysis to selectively de-embed topological sections of the model. For example, it is possible to determine the worst-case crosstalk due to a 20% tolerance for a CML termination resistor. Although internal on-die terminations are more complicated than a simple real impedance, we can perform the simulation to determine approximate sensitivity to termination tolerance.

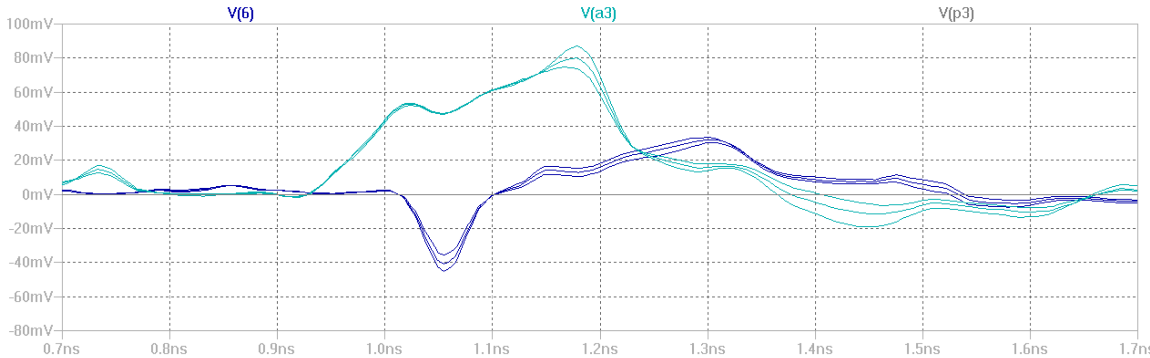

Figure 15: Family of NEXT and FEXT curves impacted by a 20% change of victim line termination impedance.

By making straightforward TDR measurements, generation of a model and simulating the results in a 50 Ohm environment, has allowed us to generate a consistent measure-based solution to assessing crosstalk.

Crosstalk compliance

High data rate electrical specifications, such as the optical internetworking forum's common electrical interface (CEI) and JEDEC's fully buffered DIMM (FBD) require that every transmitter-channel combination be tested to compliance under conditions that include active aggressor channels. The aggressor signals propagate data patterns that are designed to be aggressive and give not-too-hard to observe crosstalk on standard jitter analyzers. That is, the aggressors transmit the clock-like pattern 0xCC. The Total Jitter [3] of the transmitter-channel combination must pass the requirements of the particular standard. Similarly, receivers are tested under stress conditions that include not only active aggressor channels, but applied random jitter, periodic jitter and truncated-Gaussian jitter.

The current compliance tests are performed as if crosstalk were a jitter-phenomenon. While crosstalk is obviously amplitude noise first and jitter second, at data rates less than 6 Gb/s ignoring the amplitude problem may be adequate. More likely, though, full noise analysis like BER contour will need to be performed to assure interoperability.

Conclusion

Crosstalk is an unavoidable phenomenon in multi-signal electromagnetic systems. The magnitude of crosstalk is pre-determined by the media and geometry of the system and the voltage levels of the signaling scheme but can be reduced by use of differential signaling schemes.

There are many ways to analyze crosstalk. In this paper we have emphasized the importance of keeping in mind that crosstalk causes a combination of amplitude noise and jitter that effect the bit error ratio of a system. Since jitter analysis alone is frequently sufficient for determining acceptable levels of crosstalk, engineers may forget amplitude noise and be confused by seemingly anomalous behavior.

The crosstalk on a system can be analyzed on a BERT in a BER contour analysis, in an eye-diagram analysis on an oscilloscope and can be predicted by analyzing the frequency and attenuation response of a circuit with a VNA or the impedance response with TDR. Analysis on BERTs and oscilloscopes is from the perspective of bit-errors, i.e., effects, and analysis with VNAs and TDRs is from the perspective of coupling.

We also described a novel approach to crosstalk analysis that combines measurement and simulation with particular emphasis on coupling that can be used to de-embed different components of a system and estimate their contributions to the overall crosstalk of a system.

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