Application Note

THE IMPORTANCE OF ACCURATE S-PARAMETERS FOR PAM-4 APPLICATIONS

Products:

- R&S®ZNA
- R&S®ZNB

Ransom Stephens Ph.D. | GFM355 | Version 1e | 01.2021

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Contents

1 Overview

PAM-4 (4-level pulse amplitude modulation) has been introduced in high-speed serial data technology to reduce the bandwidth demands of ultra-high data rates. It uses only half the bandwidth per bit that would be required of conventional NRZ (non-return to zero) modulation. But at the same time, PAM-4 signaling complicates design and test at every turn. Evaluation of channels is now just as important to system development as serializer / deserializer (SerDes) testing and the challenges presented require a higher level of test and measurement performance than ever before. This paper investigates the evaluation complexities of PAM-4 interconnects at high data rates.

2 Introduction

PAM-4 (4-level pulse amplitude modulation) has been introduced in high speed serial data technology to reduce the bandwidth demands of ultra-high data rates. By encoding two bits per UI (unit interval), Figure 1, PAM-4 uses half the bandwidth per bit that would be required of conventional NRZ (non-return to zero) modulation.

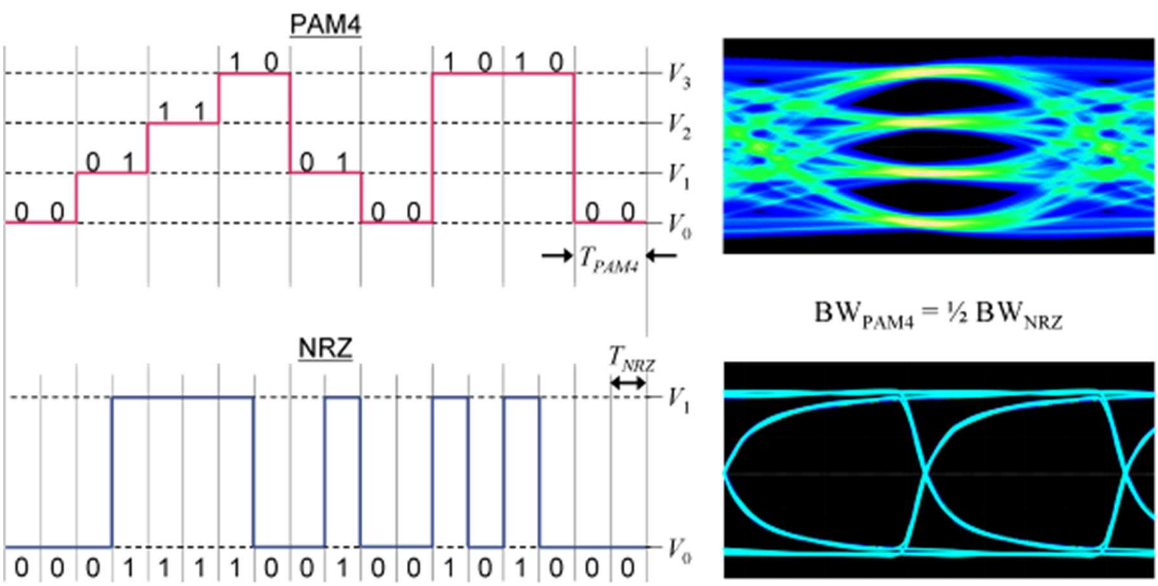

Figure 1: A PAM-4 signal and (b) the equivalent NRZ signal.

PAM-4 has so far been adopted in the 100 and 400 GbE (Gigabit Ethernet) standards 802.3bj, 802.3bs, and 802.3cd, OIF-CEI (Optical Internetworking Forum-Common Electrical Interface) 4.0, and is expected to be adopted by 64 and 256 GFC (Gigabit FibreChannel) and InfiniBand HDR (high data rate) at 50, 200, and 600 Gb/s. Extreme data rates like 400 Gb/s are achieved by combining multiple lanes at lower rates, like the 8×53 Gb/s example shown in Figure 2.

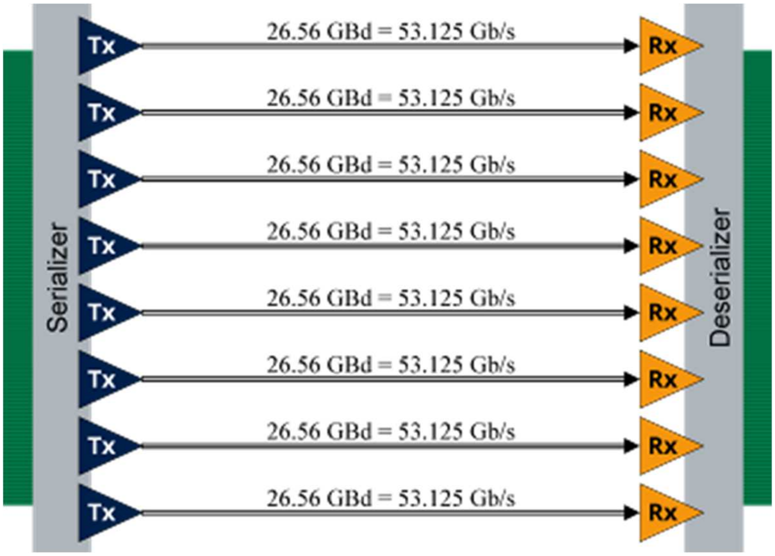

Figure 2: Example of a 400 Gb/s system, 400GAUI-8.

PAM-4 signaling complicates design and test at every turn: PAM-4 signals have 12 distinct symbol transitions, each with its own rise and fall times; the average transition density of PAM-4 signals is 75% compared to NRZ's 50%; a PAM-4 symbol error can result in either one or two bit errors; a system's BER (bit error ratio) must be distinguished from its SER (symbol error ratio); and a bit rate of 112 Gb/s corresponds to a 56 GBd PAM-4 symbol rate.

The key problems solved by PAM-4 are illustrated in Figure 3, the frequency response of a typical interconnect as illustrated by the absolute value of the differential S-parameter, Sdd21. First, Figure 3 shows that loss increases exponentially with frequency on conventional PCB (printed circuit board). Switching from NRZ to PAM-4 reduces the frequency of the signal's fundamental harmonic by a factor of two. For the interconnect measured in Figure 3, insertion loss is reduced from more -60 dB to -27 dB.

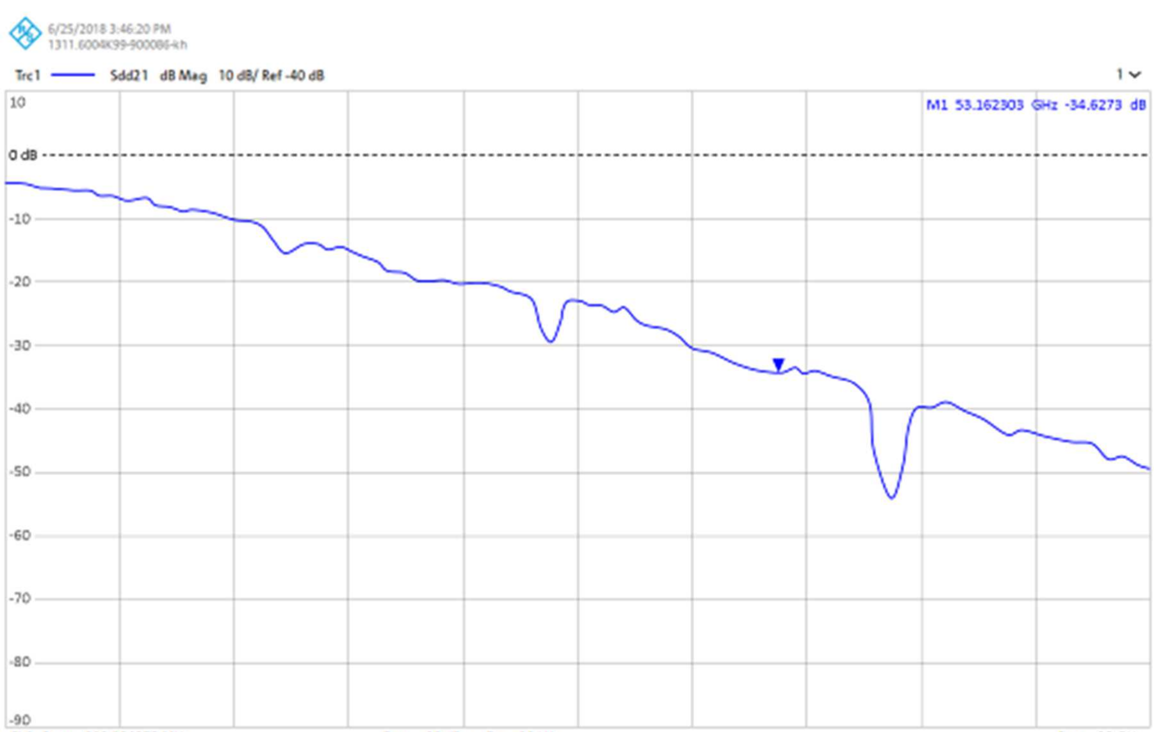

Figure 3: |Sdd21| vs frequency, the insertion loss response of a typical differential high-speed serial circuit.

Second, the nonuniform frequency response in Figure 3 causes ISI (inter-symbol interference). The waveforms of digital signals have rich frequency spectra, Figure 4. The frequency components experience different levels of attenuation: Low frequencies are barely attenuated and high frequencies are attenuated a great deal. Since the signal is composed of a sum of Fourier components whose amplitudes and frequencies give the digital waveform its structure, nonuniform loss degrades the relationships between frequency components and causes symbols to interfere with their neighbors, hence the term "inter-symbol interference." The peaks and valleys in the frequency response, which indicate impedance mismatches in the signaling path at connections between circuit boards, chips and traces, vias, etc., cause reflections that exacerbate ISI.

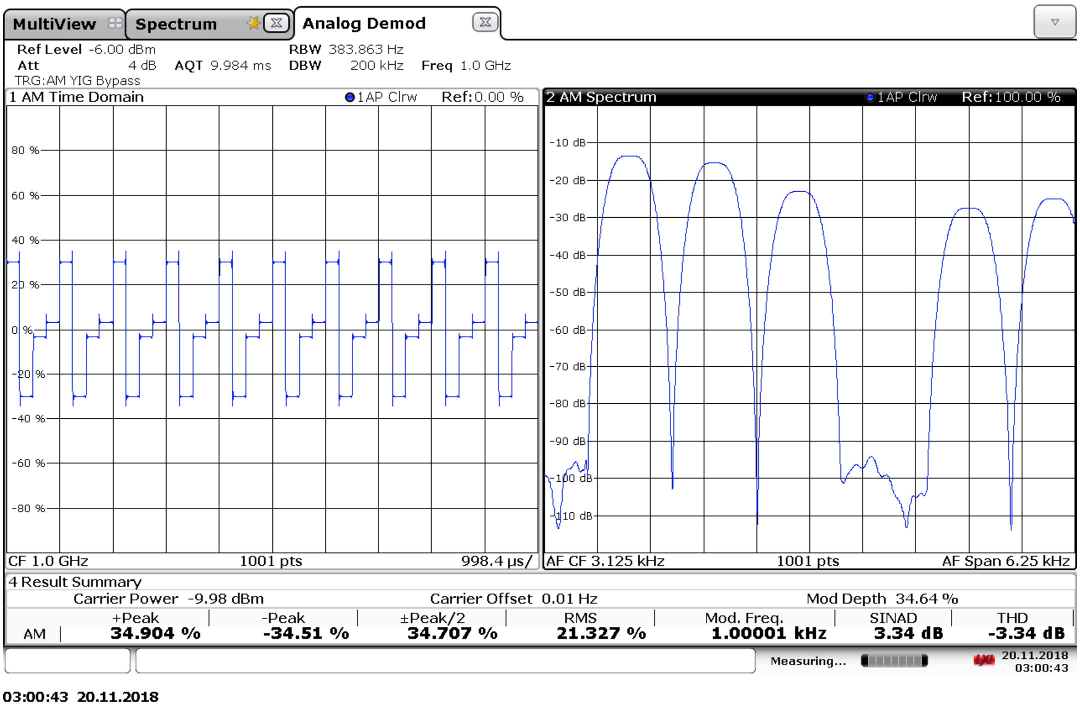

Figure 4: Frequency spectrum of a PAM-4 waveform.

That an essential cause for the introduction of PAM-4 signaling is illustrated by a measurement of S-parameters, demonstrates the need for accurate S-parameter measurements in the development of PAM-4 SerDes, interconnects, and systems.

Equalization at both the transmitter and receiver have been effective in correcting ISI sufficiently for NRZ systems to operate at BERs (bit error ratios) of less than 1E-12. The need for equalization persists with PAM-4 signaling and, like everything else, gets more complicated. Three types of equalization are common for NRZ signaling: transmitter FFE (feed-forward equalization), which includes de-emphasis, receiver CTLE (continuous time linear equalization), and DFE (decision feedback equalization), also at the receiver. The efficacy of DFE relies on the better than one in a trillion accuracy of NRZ logic decoders, but when the decoder makes a mistake, DFE feeds back an error. An errored feedback term deteriorates the quality of subsequent bits at the decoder and can cause burst errors.

The most obvious complication in going from NRZ to PAM-4 is that the SNR (signal to noise) drops by at least 9.5 dB. The standards address the SNR drop by introducing FEC (forward error correction). The maximum allowed pre-FEC BER (bit error ratio) is increased from typically 1E-12 or 1E-15 to as high as 1E-4; FEC then corrects the operational BER back to the standard low levels. At such high pre-FEC BERs, the propensity for DFE to cause burst errors can be a liability in PAM-4 systems. As a result, instead of pairing CTLE with DFE at the receiver, many PAM-4 receivers follow CTLE with DSP (digital signal processing) that incorporates FFE at the transmitter.

The SNR drop impacts every measurement. Test instrument noise floors must be kept lower than ever for signal analysis, for determining SerDes and transceiver performance, and, to address the primary topic of this paper, for evaluating interconnects.

At data rates below 25 Gb/s, components could be designed to specification without careful consideration of their impact on the system as a whole. Interconnects could be specified by their insertion loss, return loss, insertion loss deviation, and insertion loss to crosstalk ratio. The performance of interconnects at data rates of 25 Gb/s and higher, both NRZ and PAM-4, now must be evaluated according to their SER and BER performance in the presence of noise, jitter, crosstalk, and both transmitter and receiver equalization.

This paper investigates the evaluation complexities of PAM-4 interconnects at high data rates. The next section is a synopsis of S-parameters that includes conceptual tools for understanding the nuances of differential signaling and crosstalk. Sections 4 and 5 describe interconnect evaluation in the context of the latest PAM-4 standards and then through modeling and simulation. Section 6 concludes with discussion of the critical PAM-4 test equipment requirements.

3 S-parameters

S-parameters describe the scattering properties of network elements. They provide the necessary information to model and calculate the inter-dependent effects of ISI, equalization, and crosstalk. S-parameters are a measure of frequency-dependent impedance across the bandwidth over which they are measured; they capture a network element's amplitude and phase response, loss and reflectivity, and the coupling between separate network elements.

The S-parameters of a network element include the frequency domain version of its impulse response. For any given signal, the S-parameters of an interconnect provide sufficient information to calculate the resulting waveforms both transmitted through and reflected by that interconnect.

For a one-dimensional device like a transmission line, there are four S-parameters, Figure 5a: one characterizes transmission properties from left to right, S21, another from right to left, S12, and the other two, S11 and S22, characterize reflection properties at each end. For a LTI (linear time invariant) device, the S-parameter matrix should be symmetric, S21 = S12. S-parameter notation follows a simple convention: Snm characterizes the response at port n from a stimulus at port m.

The S-parameter matrix for a pair of single-ended transmission lines, Figure 5b, yields a 4×4 S-parameter matrix that includes coupling between the two lines that can result in crosstalk. For example, S41 can be used to calculate the crosstalk that would emerge from port 4, the right end of the lower channel, due to an aggressor signal transmitted into port 1, the left end of the upper channel.

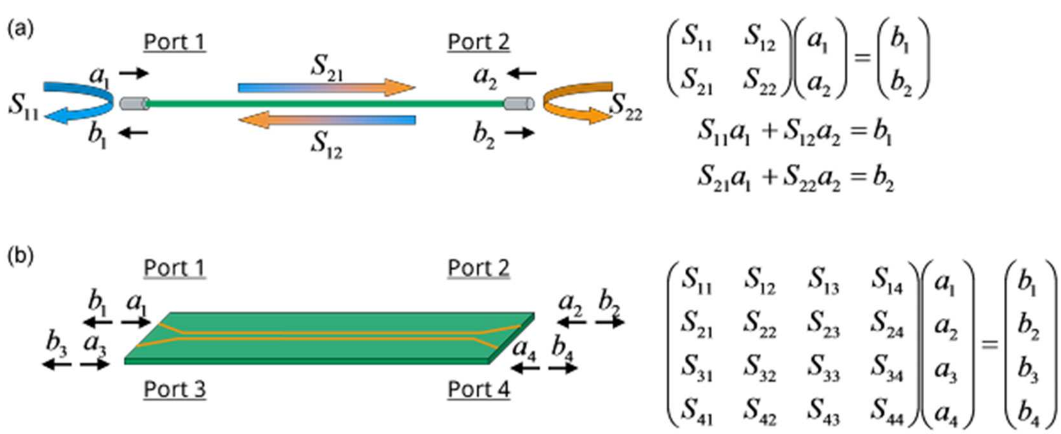

Figure 5: S-parameters: (a) 2×2 S-parameter matrix for a single trace or cable and (b) 4×4 S-parameter matrix for two channels.

3.1 Differential S-parameters

Of more importance in the context of high speed serial data technology, the 4×4 S-parameter matrix also describes a differential pair. A unitary transformation rotates the 4×4 matrix into what are called differential S-parameters, Figure 6. The transformation preserves all information about the pair but presents it in a more useful form. The upper left quadrant, Sdd, describes the transmission and reflectivity of the differential pair as a unit; Sdd21 characterizes the transmissivity of a differential signal propagating into the left differential port and emerging from the right.

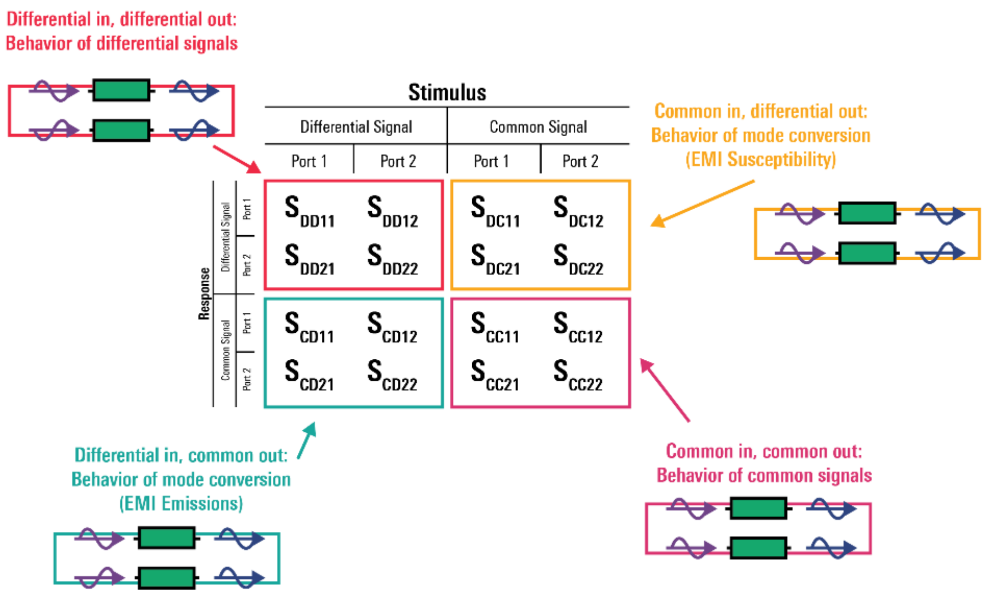

Figure 6: Differential S-parameters.

The subscript c in the differential S-parameter matrix indicates the response of the pair to common signals. Since differential receivers use a comparator, signals that are common to both conductors cancel each other. The CMRR (common mode rejection ratio is given by:

$$\mathrm{CMRR} = 20\log\frac{|Sdd21|}{|Scc21|}.$$

The off-diagonal, mixed-mode matrix elements describe how the channel converts differential signals to common signals and vice versa. In the limit of an ideal, perfectly symmetric differential pair, two identical channels with identical impedance properties and zero differential skew, the mixed-mode elements are all zero. Generally, the mixed-mode elements like Scd21 can be used to calculate the extent to which a differential signal applied to port 1 results in a common signal at the output of port 2. Conversion of differential signals to common signals is called mode conversion. It causes emission of EMI (electromagnetic interference). Common to differential conversion results from the channel’s EMS (electromagnetic susceptibility).

3.2 Crosstalk S-parameters: NEXT and FEXT

S-parameter matrices can be measured for any number and combination of LTI network elements. The number of matrix elements increases exponentially: 2N where N is the number of ports. Multiport VNAs or switch matrixes can facilitate measurements of multi-lane systems.

Figure 7 shows the S-parameters for a multi-lane system with the transmission, reflection, NEXT (near end crosstalk), and FEXT (far end crosstalk) terms indicated. NEXT comes from the sum of all aggressor signals that couple to the victim and emerge at the "near end" closest to the victim transmitter. FEXT is the sum of aggressors coupled to the victim that emerge at the end far from the victim transmitter.

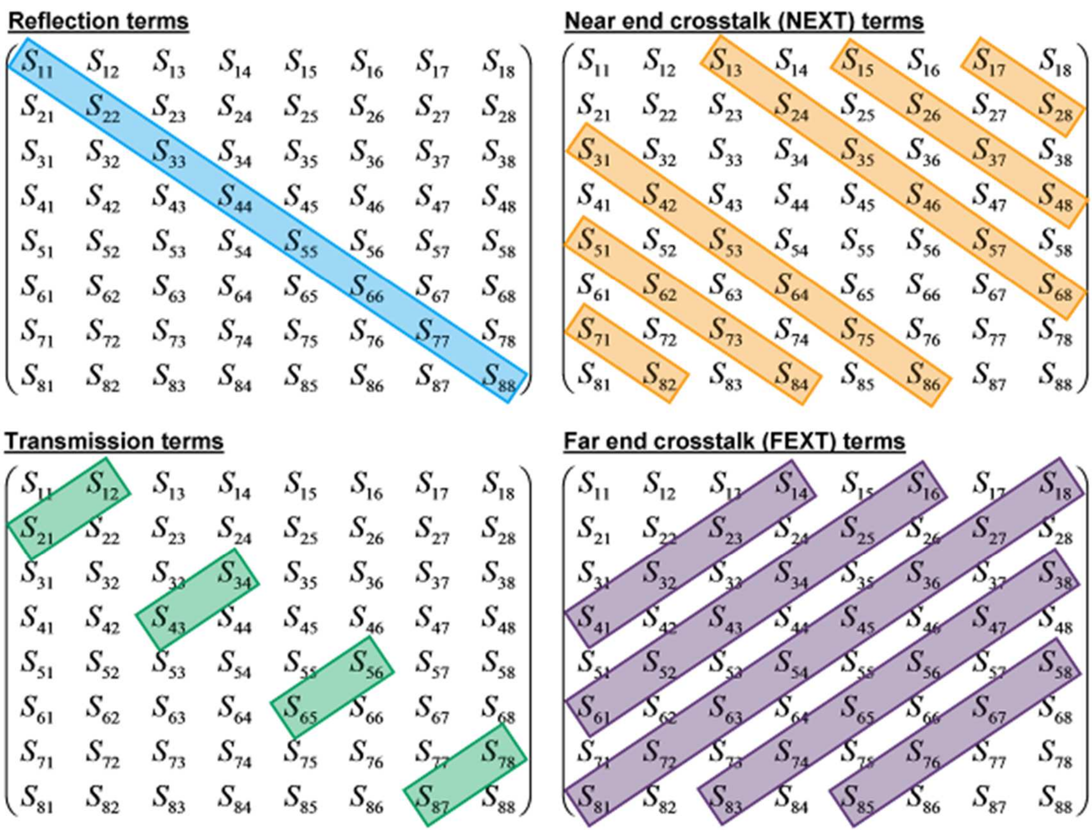

Figure 7: The matrix of S-parameters for a 4 lane system with transmission, reflection, NEXT and FEXT terms indicated.

S-parameters of separate components—BGAs (ball grid arrays), traces, cables, connectors, vias, etc—can be combined into S-parameters for the whole channel by "cascading." Cascading is a linear algebra transformations of S-parameter matrices into so-called T-matrices. The product of the T matrices for separate network elements yields the T matrix for the whole. The T matrix of the whole is then transformed into the S-matrix of the whole.

3.3 S-parameters measured by Vector Network Analyzers

Since S-parameters are determined solely by the properties and geometry of the conducting and dielectric media that compose the device, they are independent of modulation scheme; they're the same for PAM-4 and NRZ. However, where S-parameters derived from TDR/TDT (time domain reflectometry/time domain transmissivity) oscilloscope measurements are sometimes acceptable for NRZ signaling, the SNR drop experienced in PAM-4 signaling demands S-parameters with the low noise floor and accuracy that can only be obtained by measuring them with a VNA (vector network analyzer).

VNAs measure S-parameters frequency by frequency. They applying sinusoidal waves with precise amplitudes, frequencies, and phases to each port of a network element one-by-one and then record the amplitude and phase of the transmitted and reflected waves. Instruments like the R&S®ZNA or the R&S®ZNB VNA feature phase coherent generators and can also provide a true-differential stimulus to the device. Since sine waves are the easiest signal to generate with precision, VNAs can ascend to extremely high bandwidths with both large dynamic range and extraordinarily low noise floors—typically for the Rohde and Schwarz vector network analyzers R&S®ZNA series or the R&S®ZNB —provided that the VNA is carefully calibrated prior to measurement.

4 Interconnect Evaluation for PAM-4

Current PAM-4 technology standards specify interconnect requirements through a combination of S-parameter masks of insertion loss as a function of frequency, |Sdd21|, return loss, |Sdd11|, and by requiring a minimum value of COM (channel operating margin) and ERL (effective return loss). Figure 8 and Figure 9 show medium and long reach configurations that illustrate the entirety of the channel.

The channel extends from the BGA (ball grid array) of a transmitting chip, across circuit and backplane traces, through connectors and vias, to the receiver. The S-parameters of the package are often provided by the chip vendor. In many cases the less accessible sections of the path, like between the BGA and interconnect, must be calculated from the S-parameters of a PCB coupon. The S-parameters for each component of the channel can then be cascaded into the whole from TP0 to TP5, Figure 9.

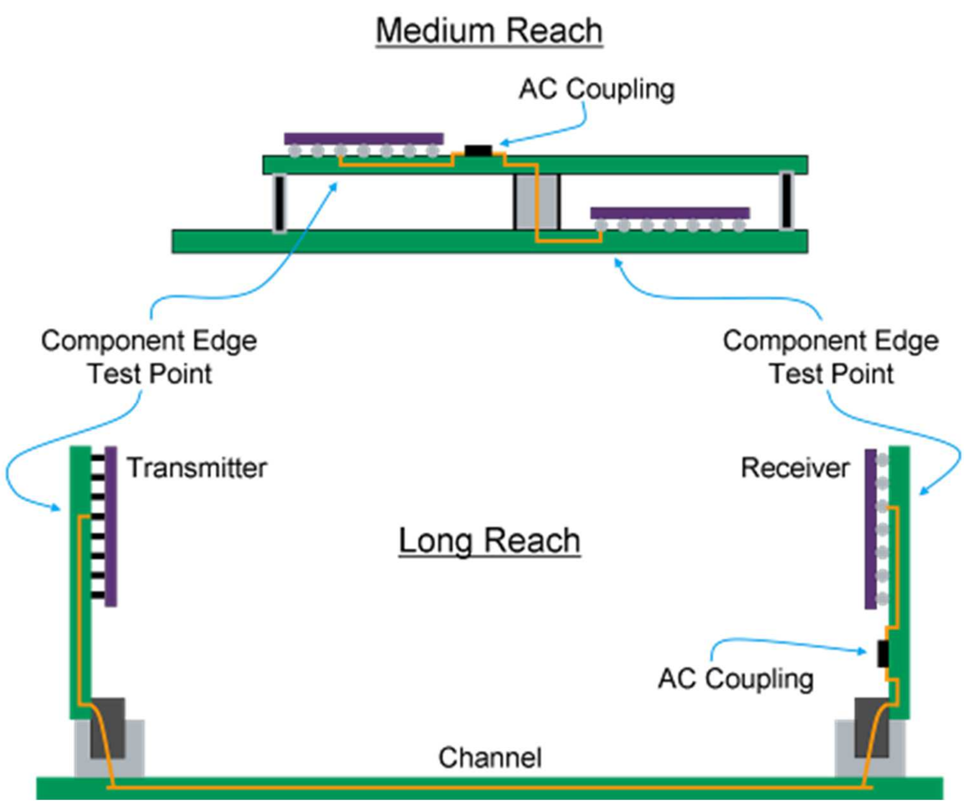

Figure 8: Medium and long reach configurations.

Figure 9: The channel model and test points.

4.1 De-embedding

The reference plane is that point where a signal is analyzed. The most desirable reference planes are at the i/o points of the DUT, not at the input to the test instrument. Test fixtures cables, connectors, traces and/or probes are often necessary to connect a DUT (device under test) to a test instrument. The effects of the test fixtures should be removed from the measurement by de-embedding them.

De-embedding begins with the measurement of the test fixture S-parameters. The R&S vector analyzer software records S-parameter matrices in “Touchstone” files that are stored with the extension sNp, where N indicates the number of ports. With their sNp files, test fixtures can be de-embedded by using the Offset De-embed function through keypad or software commands.

To analyze the effects of different circuit elements, the reference plane can be moved through the circuit by de-embedding segments of the channel. The ability to examine signals at inaccessible locations within a circuit can be powerful, but if taken too far, it can also be misleading. Since de-embedding corrects for loss, de-embedding is equivalent to raising the instrument’s noise floor.

Accurate de-embedding requires instrumentation with extremely low noise, especially in the PAM-4 environment due to its degraded SNR.

4.2 IL(f) and RL(f) requirements

IL (insertion loss) obviously must be limited to prevent the signal from being attenuated to the point where no reasonable voltage slicer could resolve the symbols. It also has to be limited to prevent extensive ISI. However when compliant signals are transmitted with FFE the IL must meet predefined lower limits.

Typical IL(f) requirements for medium and long reach configurations, like those depicted in Figure 8, are shown in Figure 10. Of course, longer reach application must be permitted greater insertion loss, though at the expense of more extensive equalization at both the transmitter and receiver.

IL(f) is given by |Sdd21| or equivalently, for an LTI device, |Sdd12|.

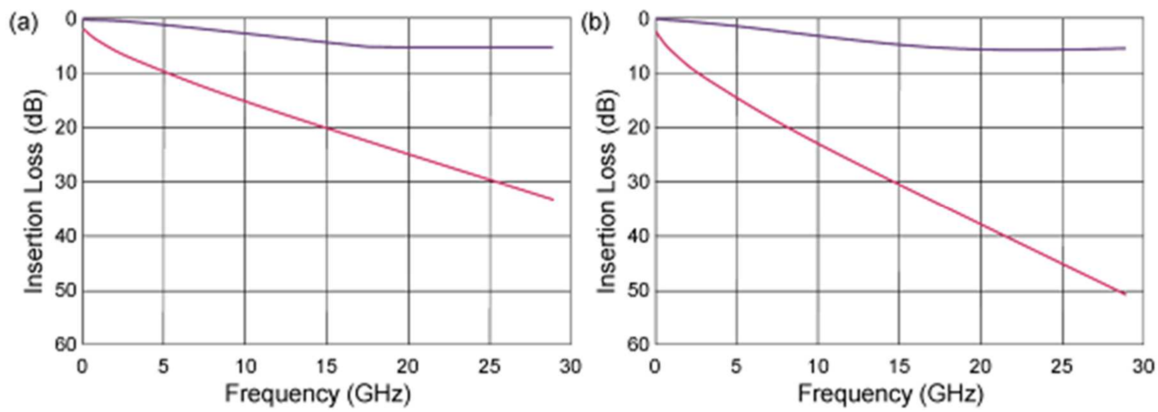

Figure 10: PAM-4 (a) medium and (b) long reach IL(f) masks for 28 GBd.

RL (return loss) must be limited to prevent reflections from causing ISI that extends over so many symbols that no equalization scheme can correct it. RL(f) (return loss) is equivalent to |Sdd11| or |Sdd22|.

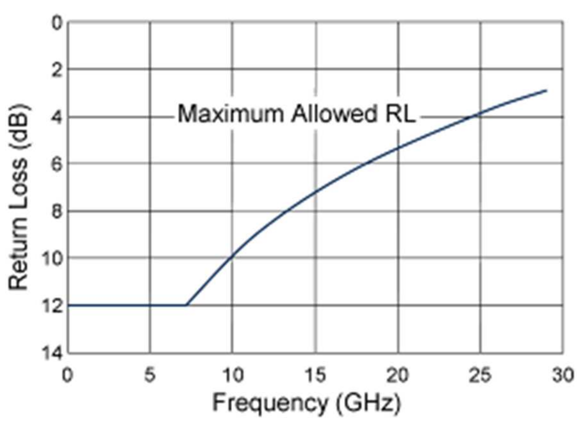

Figure 11: Typical PAM-4 RL mask for 28 GBd.

4.3 Calculating eye diagrams from S-parameters

Eye diagrams for a given signal on a given interconnect can be calculated directly from the interconnect's S-parameters. The process is essentially a transformation from the frequency domain to the time domain: The signal is convolved with the interconnect response to get the waveform at the output of the channel. Each UI (unit interval) of the waveform is overlaid to form the eye diagram.

R&S®ZNA VNAs plus installed option ZNA-K20 or R&S®ZNB VNAs plus installed option ZNB-K20 provide software that automates eye diagram calculations including the ability to add jitter and noise. The user sets the signal pattern, baud rate, symbol voltage levels, and rise/fall times of a prospective signal. The resulting eye diagram is calculated and displayed, as shown in Figure 12. Equalization schemes can be implemented with a few clicks of the user interface and their effects observed. Segments of the channel can also be de-embedded or modified. The resulting eye diagrams and waveforms can be downloaded in CSV (comma separated variable) format for analysis on a PC.

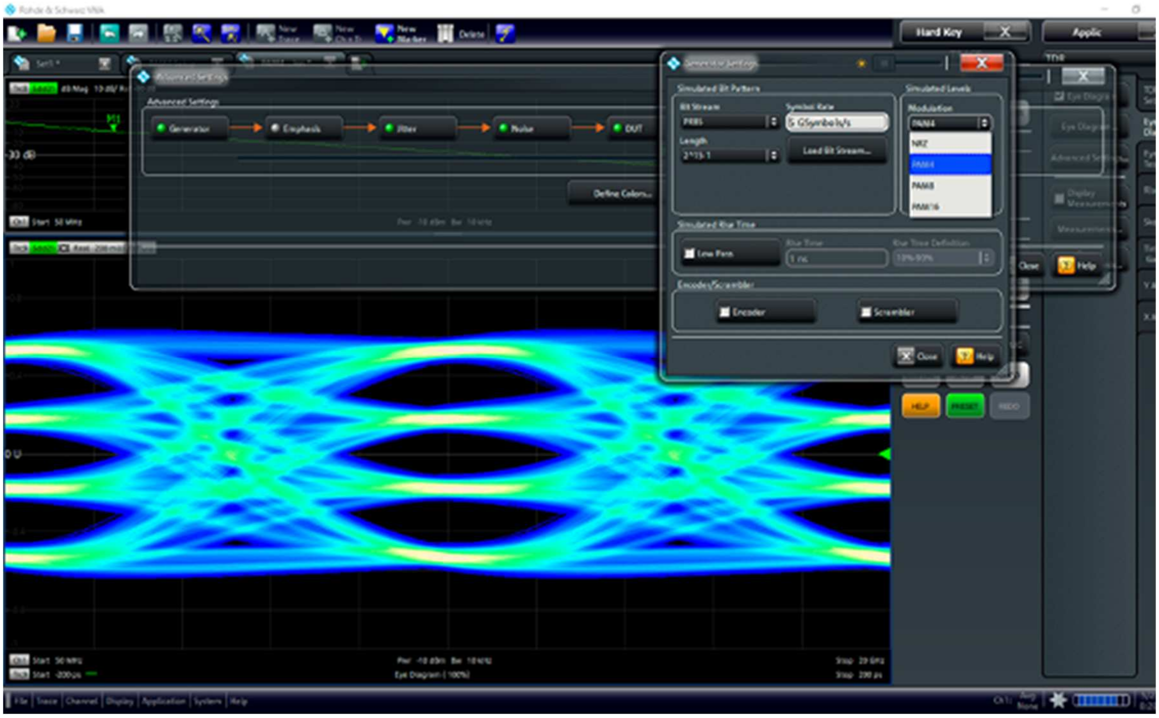

Figure 12: PAM-4 eye diagram calculation.

4.4 COM—Channel Operating Margin

As its name implies, COM is a measure of an interconnect's performance margin. COM is calculated from a model like that illustrated in Figure 13. The only measurements that go into the calculation are S-parameters. The rest of the model is assembled through assumptions about the signal and receiver that are specified by the technology standard.

The transmitter model includes:

- ► The signal's differential peak-to-peak voltage, SNR, and one-sided spectral noise density.
- ► Maximum allowed level-separation mismatch ratio, RLM. The level-separation mismatch ratio characterizes the asymmetry of the three PAM-4 eye diagrams; RLM = 1 indicates perfect symmetry, RLM = 0 indicates collapse of at least one of the eyes. Typically, the COM calculation requires RLM ~ 0.92.
- ► Transmitter FFE including maximum and minimum bounds on FFE parameters.
- ► RJ (random jitter) and uncorrelated DJ (deterministic jitter), that is, DJ independent of ISI.
- ► A transmitter device package model including capacitance, reference and termination resistances, and characteristic impedance.
- ► Peak-to-peak voltages for both near and far end crosstalk aggressors.
- ► The channel model is given by the full system S-parameter sNp file, including the coupling of the DUT and all crosstalk channels.
- ► The reference receiver model includes:
- ► The receiver device package, including capacitance, reference and termination resistances, input impedance, and 3 dB bandwidth.
- ► Two pole CTLE filter with gain step size and maximum allowed gain.

► Definition of a DFE: number of taps and limits on the tap magnitudes.

Figure 13: Elements of the COM calculation.

The transfer function is assembled from the model and the pulse (or single-bit) response is calculated. A vertical slice of the eye diagram is calculated from the pulse response. The parameters of the equalizers—FFE at the transmitter, receiver CTLE gain and DFE taps—are adjusted to minimize the SER in the presence of crosstalk. The peak signal amplitude, A_{Signal}, is calculated from the modeled eye diagram. The combined amplitude of noise and crosstalk, $A_{NoiseXtalk}$, is given by the vertical eye closure at the specified target DER (detector symbol error ratio) as illustrated in Figure 14. COM is given by

$$\mathrm{COM} = 20\log\frac{A_{Signal}}{A_{NoiseXtalk}}.$$

Typically, standards require COM ≥ 3 dB.

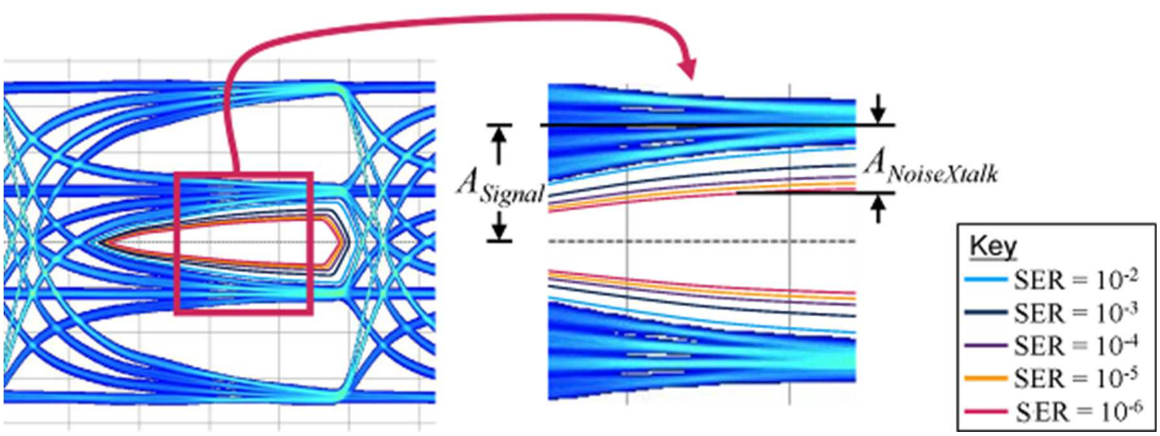

Figure 14: COM is an SNR-like parameter calculated from an eye diagram derived from the pulse response of a system model that is defined by the channel S-parameters.

While it is calculated from the convolution of many different quantities, the only measurements involved in the calculation are the multi-port S-parameters of the configuration, e.g., Figure 2. Since it is the key interconnect specification, small uncertainties in the calculation of COM can have large consequences in channel evaluation; another example of the importance of accurate S-parameter measurements for development and evaluation of PAM-4 components and systems.

4.5 ERL—Effective Return Loss

An example for a long reach configuration with typical discontinuities is shown in Figure 15. At these discontinuities, a part of the signal travelling to the receiver is reflected back to the transmitter before it gets reflected again at another discontinuity back to the receiver. The reflected signal finally reaches the receiver at a later time and with lower amplitude and is causing inter-symbol interferences (ISI) with the wanted signal. While COM describes the impairment of noise and xtalk on the modelled eye diagram, the effective return loss ERL models the impact of these reflections, based on the measured S-parameters.

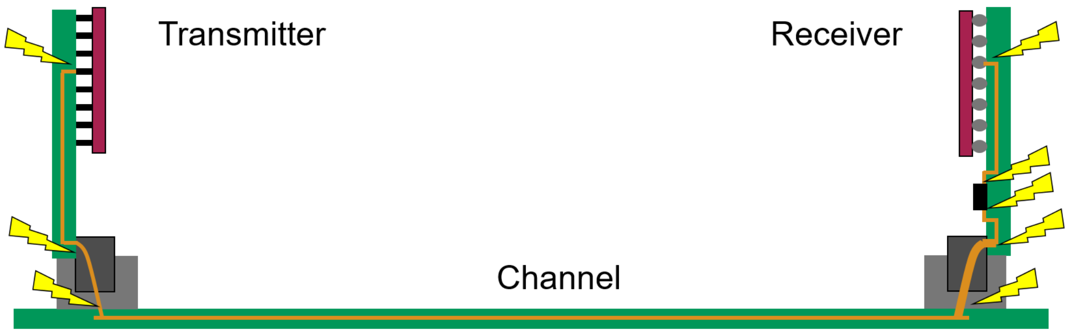

Figure 15: Long reach configuration with multiple discontinuities and corresponding reflections.

As DFE equalizers efficiently compensate inter-symbol interferences within their number of taps, the ERL calculation only considers reflections with a delay longer than the length of the DFE. ERL is given by

$$\mathrm{ERL} = 20 \log \frac{\mathrm{ASignal}}{\mathrm{AReflections}}$$

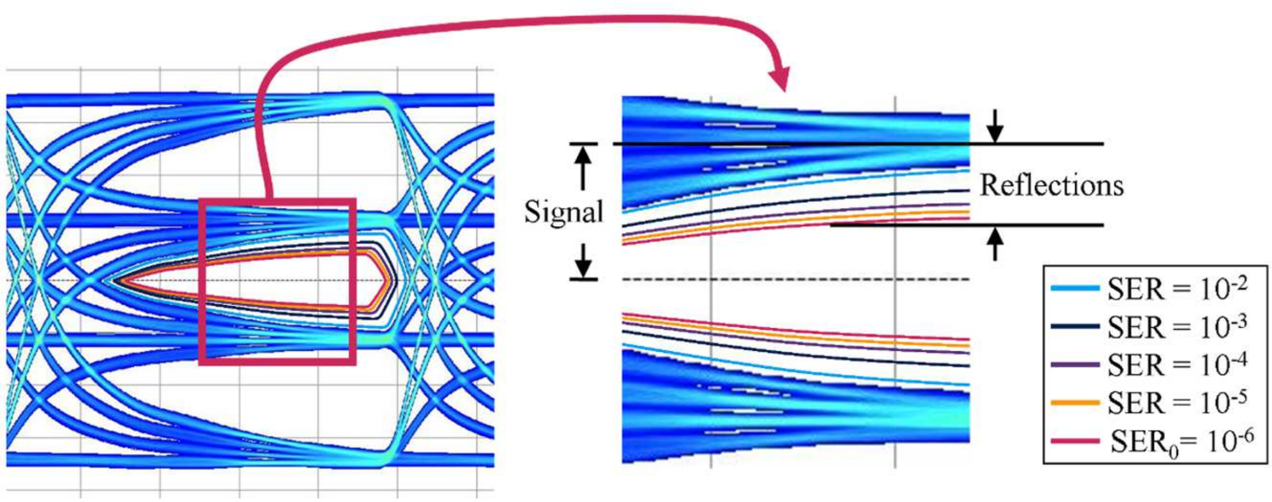

Figure 16: Like COM, also ERL is an SNR-like figure of merit, calculated from the measured S-parameters based on a defined model of the overall transmission system.

5 Simulation and modeling of SerDes, interconnects, and circuits

The complexity of PAM-4 signaling requires in-depth evaluation of every high speed serial data component prior to spinning PCBs or fabricating wafers. Simulation and modeling are now standard tools for evaluating SerDes, interconnects, and complete circuits before they are developed. Simulation accuracy relies on accurate S-parameter measurements.

There are three levels of modeling complexity: Electromagnetic field solvers calculate Maxwell's equations at the relevant points in conductors and media, in sequential time increments. They are the slowest but most accurate simulations and can calculate S-parameters from descriptions of the geometry, conductivity, and dielectric properties of a circuit. SPICE (Simulation Program with Integrated Circuit Elements) circuit simulators calculate the voltage, current, impedance, etc from equivalent R, L, C, G circuits. SPICE simulations are faster than field solvers and quite accurate. However, SPICE models require information from component vendors that is usually considered proprietary and is therefore difficult to obtain.

IBIS (Input/output Buffer Information Specification) models are fast, easy to modify, ubiquitous, and often free. IBIS-AMI models use ASCII input files that are supported by chip manufacturers. Inputs include system parameters like symbol voltage levels, rise/fall times, amplitude noise, jitter, transmitter and receiver equalization schemes, clock recovery, voltage slicer sensitivity, and, most importantly, the channel S-parameters.

S-parameters are provided to the IBIS model through sNp Touchstone files. The Touchstone files for any configuration are automatically formatted by Rohde and Schwarz VNAs and can be used in any EDA (electronic design automation) tool.

The AMI (Algorithmic Modeling Interface) extension of IBIS simulations, as the name implies, allows for evaluation of algorithmic signal processing.

IBIS-AMI simulators assume that the analog and digital aspects of both transmitters and receivers can be analyzed independently. That is, the analog nature of a transmitter's output buffer and termination is separated from digital signal conditioning properties like equalization; similarly, the analog nature of the receiver's input termination and filtering is separated from clock recovery and receiver equalization, as indicated by Figure 17. Under these assumptions, the waveform can be calculated at any point from the transmitter to the output of the receiver.

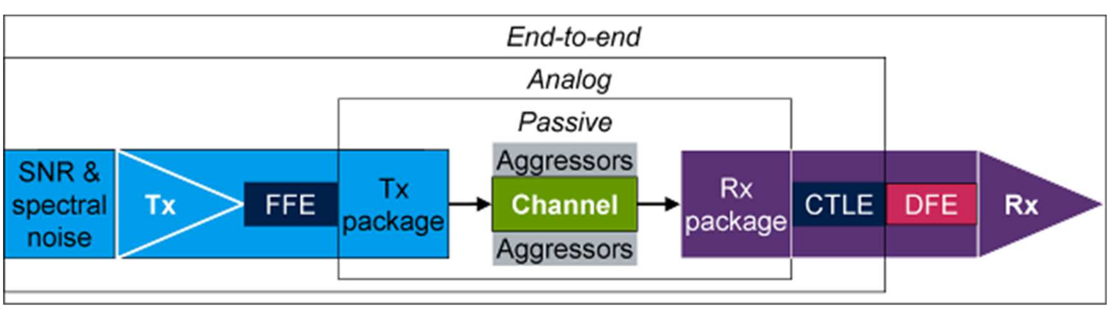

Figure 17: Separating the elements of an IBIS-AMI simulation.

5.1 Evaluation of transmitter and receiver equalization schemes

There are two types of IBIS-AMI simulators: statistical and bit-by-bit.

Statistical models calculate eye diagrams by convolving the transmitted signal with the impulse response and then imposing jitter and noise, as the name indicates, with statistical methods. Since they don't calculate the waveform that results from a transmitted signal, statistical models require fixed equalization parameters and do not permit adaptive equalization or link training.

Bit-by-bit modeling is slower than statistical. It is a time-domain simulation that calculates the waveform bit-by-bit, developing eye diagrams by overlaying each unit interval. The waveform can be processed by clock recovery and adaptive equalization algorithms. Since they provide waveforms for analysis, bit-by-bit simulations allow evaluation of link training schemes.

In both cases eye diagrams can be calculated at any point in the signal path, prior to or following each equalization algorithm. When the entire multi-channel configuration is simulated, the effects of crosstalk and how it's affected by different equalization schemes can be evaluated. The resulting eye diagrams can also be translated into corresponding SER-contours.

Statistical simulation is fast enough to calculate BERs and SERs (symbol error ratios) for different equalization schemes down to 1E-15 in seconds. On the other hand, bit-by-bit simulations take minutes to simulate millions of bits but allow evaluation of non-LTI features.

6 PAM-4 demands high fidelity VNA measurements

6.1 S-parameter and frequency limits

S-parameters are frequency domain measurements. In principle, the frequency and time domains are equivalent. In practice, the accuracy of a transformation from time to frequency is limited by the duration of the time domain measurement; similarly, the accuracy of the transformation from frequency to time is limited by the bandwidth of the frequency measurement. While VNAs have the widest bandwidths of any equipment used in development and manufacture of high speed serial technology, it's important to keep in mind that the low end of the bandwidth is also important. Obviously one cannot generate a DC sine wave.

The accuracy of VNA measurements is limited by two factors: the high frequency limit and the low frequency limit. As long as the high frequency limit is above the highest signal harmonic, the high frequency limit shouldn't pose problems. On the other hand, inaccurate low frequency S-parameters can create the illusion of effects preceding their causes in conversion to the time domain. The DC limit has to be obtained by extrapolated from the lowest measured frequency.

The high bandwidth of the R&S®ZNA VNAs, Figure 18, matches well with the requirements from the previous paragraphs. It measures S-parameters from 10 MHz to as high as 67 GHz. Regarding the low frequency limit an interesting instrument feature helps further. If the DC value of the channel is known, a user can manually input a DC. This can be an open, a short, a match or anything in between. If the DC value is not known, the VNA can automatically extrapolate to the best matching DC value, anyhow the user can override the calculated DC value at any time.

Figure 18: R&S®ZNA VNA.

6.2 Key features and specifications of Rohde & Schwarz VNAs

The complexity and low SNR of PAM-4 signals makes accurate, low noise, large dynamic range S-parameter measurements critical for channel evaluation, simulation accuracy, and the calculation of COM and ERL. The typical dynamic range of R&S®ZNA VNAs exceeds 140 dB and the typical RMS noise floor is -132 dBm/Hz.

To evaluate channel performance, the effects of the test fixture must be de-embedded. By measuring the test fixture S-parameters prior to evaluating the channel, the R&S®ZNA VNAs will automatically de-embed it and report the actual channel performance in quasi-realtime.

Eye diagrams serve as intuitive measurements of performance. The R&S®ZNA VNAs can calculate eye diagrams for many modulation schemes including NRZ, PAM-4, PAM-8, up to PAM-16. The eye diagrams can be downloaded, point-by-point in CSV files for further analysis.

Since high speed serial technology achieves data rates over 400 Gb/s by combining parallel differential lanes, crosstalk is added to the already hostile PAM-4 environment. Modeling the system requires measurements of every aggressor-victim pair. Sixteen separate measurements are required on a 4-port VNA to measure every pair-wise combination of a four lane differential system. In a practical sense, measurement speed is nearly as important as accuracy.

R&S®ZNA VNAs are uniquely configured to make fast measurements. The notoriously cumbersome calibration process has been automated through use of an optional Calibration Unit, Figure 19.With a wide range of IF-bandwidth, selectable filter types and an extremely high sensitivity, R&S®ZNA VNAs provide a high degree of freedom to find the golden medium between sensitivity, accuracy, trace noise and short test time. Since they are designed with multiple sine wave sources, pairs of two port channels can be measured simultaneously, corresponding to a 4x speed increase. Switch matrices can also be used to automate the process and reduce the effort required to connect and disconnect every channel, though at the expense of raising the noise floor,

Figure 19: R&S Automatic Calibration Unit.

6.3 Conclusion

The challenges presented by PAM-4 high speed serial technology require a higher level of test and measurement performance than ever before. Since the core problem that caused the switch from NRZ modulation is caused by the channel, evaluation of channels is now as important to system development as SerDes testing. R&S®ZNA VNAs have all the qualities, benefits, features, and performance specifications, necessary to meet the PAM-4 challenge.

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