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Effect of the Maximum Frequency and Frequency Resolution of S Parameters on Channel Simulation

Sameh Y. Elnaggar, Semtech Inc.
selnaggar@semtech.com

Bill Kirkland, Semtech Inc.
wkirkland@semtech.com

Eduard Frlan, Semtech Inc.
efrlan@semtech.com

Abstract

For reliable Extreme Short Range (XSR) and Very Short Range (VSR) system modelling and equalization, accurate channel representation is of paramount. Channels are usually measured or simulated in the frequency domain with certain resolution frequency f_{res} and up to a maximum frequency f_{max}. As Baud rates keep increasing, f_{max} and f_{res} are generally limited by the measuring instrumentation and/or computational resources. From a practical point of view, a channel is assembled via the cascade of different components (traces, packages, connectors, etc) that may be characterized using different f_{res} and f_{max} values. In this contribution, we investigate the effect of such limitations on the overall system characteristics. It is demonstrated via theory and numerical analysis of typical channels that f_{max} and f_{res} can severely result in unreliable, sometimes unobvious, channel simulation results. Additionally, the paper discusses the effect of the DC point on the time response. We provide rules of thumb and possible remedies. An important aim of the contribution is to highlight that the impact of the limitations is of a fundamental nature and may add significant artifacts that influence the decisions of the signal integrity and/or the system engineer.

Author(s) Biography

Sameh Elnaggar has been a Signal Integrity Engineer with Semtech since 2017. Prior to that, he was a postdoc at the School of Engineering and Information Technology, University of New South Wales, researching the properties of nonlinear metamaterials. Throughout his career, he held several engineering jobs that span automation, optics, power systems, RF and signal integrity. He holds a Ph.D degree in Electrical and Computer Engineering from the University of New Brunswick. He is a registered Professional Engineer at the Association of Professional Engineers and Geoscientists of New Brunswick.

Ed Frlan is a Senior System Architect within the Signal Integrity Products Group of Semtech responsible for the definition of next generation 100G, 200G and 400G physical layer products. He has co-authored several of the Open Eye MSA PAM-4 optical standards including the 200G-FR4 and 50G per lane MMF physical interfaces and is presently involved in the specification of several electrical interfaces including the OIF's CEI 112G VSR and XSR clauses. Ed holds a Ph.D. degree in Electrical Engineering from Carleton University.

Bill Kirkland has been a Signal Integrity Engineer with Semtech for the last four years. Prior to that, seven years in RF engineering with BlackBerry/RIM and 15 years with Nortel/BNR. He spent 10 years being educated in Electrical and Communication Engineering at McMaster University, PhD, M. Eng, B. Eng.

Introduction

Having channel models that accurately describe real channels became essential in today's Extreme Short Range (XSR) and Very Short Range (VSR) communication systems. Indeed, the demand for reliable online services keeps increasing [1]. In particular, due to the COVID-19 pandemic many services are partially or fully moving online with the expectation that many of the services will be part of our lives post COVID-19. With the soaring use of social networks, virtual meetings, online learning, cloud computing and e-commerce, datacenters are becoming central to modern life. Hence, there is an escalated demand on higher baud rates and better quality of services that current and future systems must handle.

In today's technology, baseband data is transmitted in a PAM-4 format with Baud rates that can go beyond 50 GBd (with a period or unit interval (UI) of less than 20 ps) per lane. This implies that today's challenges are mainly limited by the available signal to noise ratio and the short UI (UIs are equivalent to a length of 4 mm, at 50 GBd assuming a microstrip transmission line with an effective dielectric constant of approx. 2.25, or less trace length).

From a system level perspective, performance quality translates to lower symbol error rates (SER), higher signal to noise ratio and better repeatability. Such system parameters are ultimately functions of impairments such as inter-symbol interference (ISI), noise and jitter. The effects of ISI and jitter strongly depend on the channel; in particular the degradation in channel transfer function with frequency. The inevitable limitation of channel bandwidth extends the width of the transmitted symbols beyond one UI and hence increases ISI. On the other hand, the rate of change of the channel pulse response amplifies the effect of jitter; a larger rate of change results in a wider jitter probability distribution function [2].

Different channel metrics such as Channel Operating Margin (COM), Transmitter Dispersion Eye Closure, Quaternary (TDECQ), Symbol Error Rate (SER), Vertical Eye Closure (VEC) and Eye Height (EH) are calculated from data as *seen* in the time (natural) domain. The time domain provides a straightforward perspective of the system evolution as we, humans, can easily interpret. On the other hand, the frequency domain gives an alternative perspective that can augment our perceptions and allows complex systems to be simplified. This is made possible because the frequency domain view represents the response of linear time invariant systems in the system basic characteristic (eigen) functions: the sinusoids. Therefore, the frequency domain representation reduces the response to a direct scalar multiplication at each sinusoid, rather than the "convoluted" convolution operation in the time domain. In another words, the response to an input sinusoid is another sinusoidal with possibly a different magnitude and phase, where the multiplication factor is nothing but the transfer function. Moreover, a complex system can be assembled from its cascaded subsystems via the multiplication of the individual transfer functions. Going between the time (human natural) and the frequency (system natural) domains is a skill that electrical engineers develop throughout their career.

Fourier (Inverse Fourier) transform is used to go from the time (frequency) domain to the frequency (time) domain. Although the different channel metrics are calculated from the time domain data, most often, the channel behavior is determined in the frequency domain; this is true whether the channel is measured via the use of a vector network analyzer (VNA) or computed by an electromagnetic full wave solver. Practical considerations may limit the engineer's ability to collect enough data points. For instance, the maximum and resolution frequencies f_{max} and f_{res} may be the limited by the VNA or computational resources.

In this paper, we discuss the impact of both f_{max} and f_{res} on the accuracy of channel models; possible remedies and rules of thumb are also proposed. We anticipate that the discussion will help engineers to better understand (and hopefully avoid) the subtle pitfalls that may arise when these frequency parameters are not wisely selected.

Channel Model Extraction and DC point

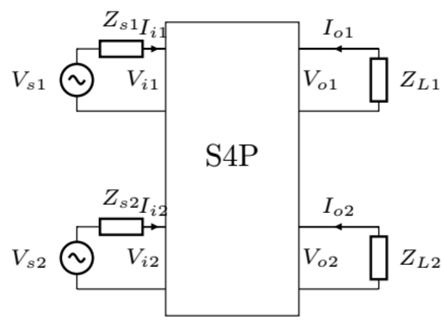

Figure 1. A 4-Port network. One is usually interested in $H_{CH}(f) = (V_{o1} - V_{o2})/(V_{s1} - V_{s2})$.

An electric channel is usually a four port (differential) network as shown in Figure 1. The channel behavior can be determined via measurements or electromagnetic computation. The differential voltage transfer function $H_{CH}(f)$ is calculated from the 4-port network parameters. A convenient way is by converting the S parameters to the ABCD parameters and hence representing the voltage and current at one side of the network in terms of the corresponding parameters at the other side. The accuracy of the voltage transfer function at each frequency is limited by measurement and numerical errors. However in the time domain, the inevitable upper limit on f_{max} and the lower bound on f_{res} can significantly affect the reconstructed time domain impulse response $h(t)$, regardless of measurement and numerical errors. The two most important factors that affect the ability of the model to reflect the channel response are the maximum and resolution frequencies, f_{max} and f_{res} respectively.

In most cases, the network S parameters are measured using a Vector Network Analyzer (VNA). The VNA RF specifications impose an upper limit on f_{max}. Additionally, the resolution bandwidth fixes f_{res}. The smaller the resolution bandwidth, the more the time needed to collect the frequency domain data during measurement.

Additionally, electromagnetic simulation can be used to predict the channel terminal parameters. The S parameters are computed given the channel input and output ports. Hardware, computational resources and delivery time limit f_{max} and f_{res}. Whether the S parameters are collected via measurements or simulation, they are converted to a voltage transfer function that connects the output to input.

The VNA does not measure the network at DC, however it is crucial to include the behavior at DC. Although DC is just one data point in the frequency domain, its effect is distributed over the whole time domain response, which can be seen from the relation below

$$H(0) = \int_{-\infty}^{\infty} h(t)\,dt.$$

The above relation is nothing but the Fourier transform of $h(t)$ at $f = 0$. If for instance a DC blocking capacitor is present, the channel transfer function $H(0)$ is zero and hence $h(t)$ will have a zero average value. In this case if $H(0)$ was not set correctly, a sustained steady state error in $h(t)$ may be present that will artificially alter the ISI interaction and consequently change the eye height and width. To demonstrate the effect of DC, consider the channel shown in Figure 2, where a channel has been extrapolated to two different values at DC. In Figure 2(a) (Figure 2(b)) the channel is extrapolated to one, hereafter Transparent (zero, here after DC blocked).

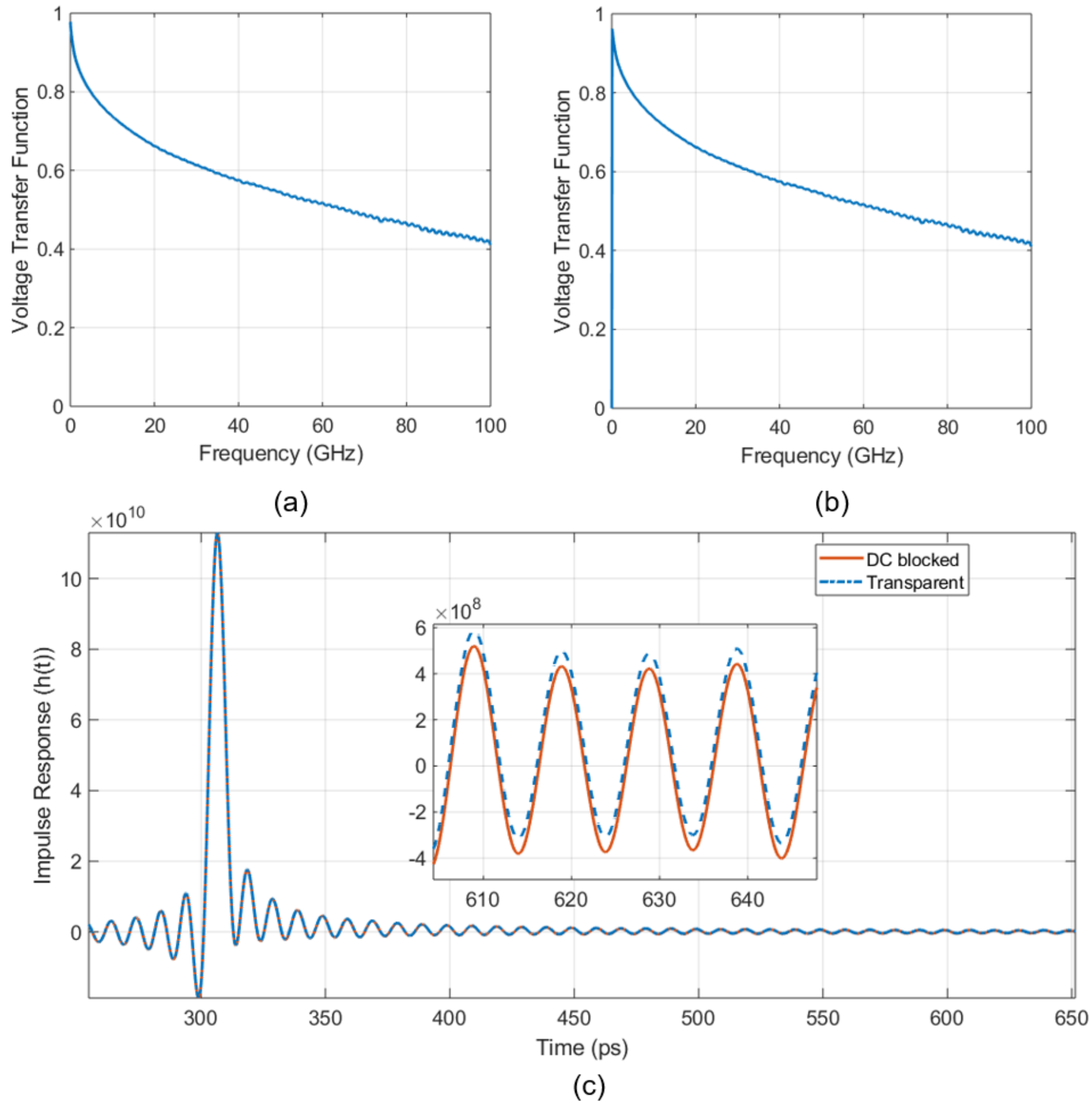

Figure 2. A channel that is extrapolated to (a) unity gain (transparent) and (b) zero (DC blocked) value at DC. (c) The impulse response.

The impulse response of both transfer functions are very close as witnessed by the plot in Figure 2(c). However, there is a small difference between the responses, which sustains over an extended period as the inset illustrates. As the above equation dictates, the total area under the transparent and DC blocked scenarios case must be one and zero, respectively. Forcing the response of the DC blocked scenario to be zero implies that over an extended period the signal height must be of a lower magnitude as the inset shows. Therefore, although the effect is minute, it is spread over the whole response and potentially will accumulate in statistical data as will be shown later.

Frequency Limitations on $h(t)$

As we have mentioned in the previous section, f_{max} and f_{res} limit the accuracy of $h(t)$. The purpose of this section is to elaborate on how both values affect the accuracy and what the possible remedies are.

Limitation due to max frequency

The impulse response $h(t)$ of a real passive lossy channel is practically time limited. Hence its spectrum extends from $-\infty$ to $+\infty$. As has been shown earlier, the response is determined up to f_{max}. Truncating the spectrum is equivalent to passing the channel response through an ideal low pass filter (LPF) with a cut-off frequency f_{max}. Although, within measurement errors, the determined transfer function $H(f)$ is accurate for each frequency over the measured or computed range, the impulse response $\hat{h}(t)$ can be quite different as Figure 3(b) and (d) demonstrate. To explain why $\hat{h}(t) \neq h(t)$, one can refer to Figure 4. Truncating the frequency range is equivalent to a convolution with a sinc function $s(t) \sim sinc(f_{max}t)$ [3]. The left panel in Figure 4 shows that at initial times (close to $t = 0$) if the tails of the sinc function extend where the $h(t)$ is non-zero, non-physical pre-cursors may appear in the reconstructed response $\hat{h}(t)$. The larger the value of f_{max}, the narrower the sinc function becomes and hence the less its tail interacts with $h(t)$. Additionally for $t \geq 12$ (i.e, after $h(t)$ becomes negligibly small), the sinc function can still interact with $h(t)$, leading to high order post cursors (resulting in an ISI with symbols sent in the future several UI away).

From a signal integrity point of view, these artificial cursors will deteriorate the eye metrics and may eventually lead to incorrect implications. For instance, the presence of an artificial precursor may naturally suggest the use of a feedforward equalizer (FFE) that has larger than necessary precursors' weights. Additionally, the presence of post cursors at the pulse tail may suggest that it is necessary to use a FFE with longer taps. Clearly if f_{max} is sufficiently large, the amplitude of the artificial cursors will be small. Later, we will precisely define limits that the system and testing engineer must consider to avoid such nuances.

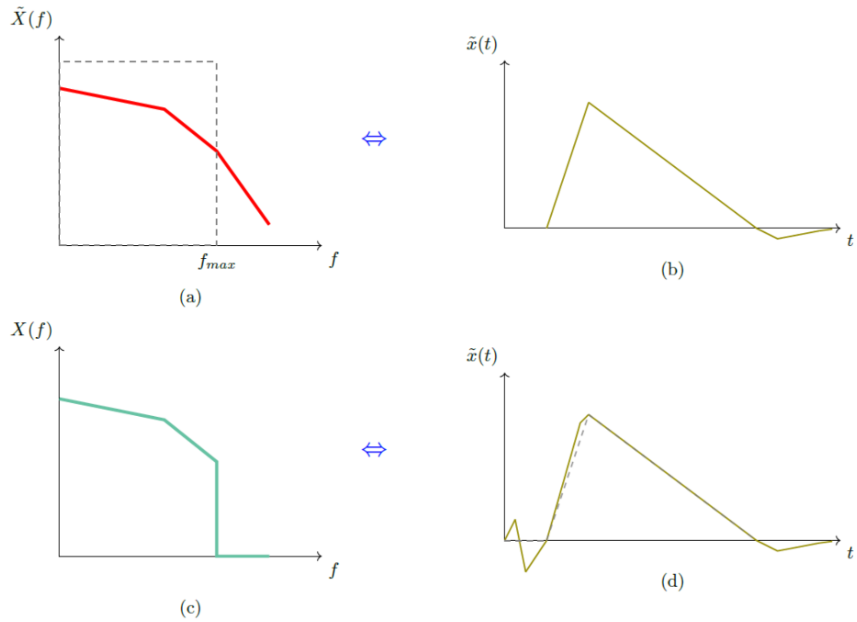

Figure 3. Effect of truncating the spectrum. (a) and (b) Original Channel response in frequency and time domain, respectively. (c) Truncated spectrum. (d) Effect of truncation on Channel impulse response.

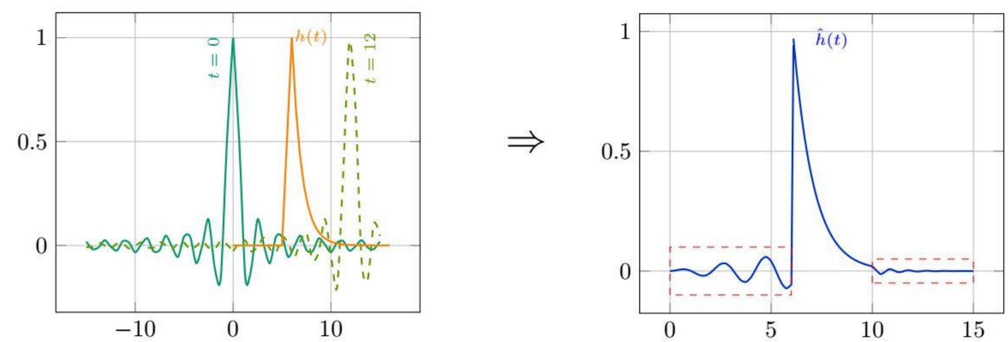

Figure 4. (Left) Orange: Channel Impulse Response. Green: sinc function at two different times. The response is the integral of the impulse response with the sinc. (Right) impulse response of the truncated system. Note the highlighted pre and post cursors

Limitation due to resolution frequency

Figure 5, Top panel shows a hypothetical band limited channel transfer function. The green curve represents the continuous response. Note that for a real channel $H^*(f) = H(-f)$. The dashed vertical lines are the sampled channel frequency response. Invoking the Fourier and inverse transforms duality, and noting that the Fourier transform of a *time* sampled signal is nothing but a repeated copy of the original continuous signal one can infer that the time domain representation of the *frequency* sampled data is a repeated copy of the original signal that repeats every $1/f_{res}$. For causal systems ($h(t) = 0, t < 0$), the Nyquist condition guarantees that time aliasing will not occur [4] [3] whenever,

$$f_{res} < \frac{1}{T_{span}}$$

where T_{span} is the impulse response width.

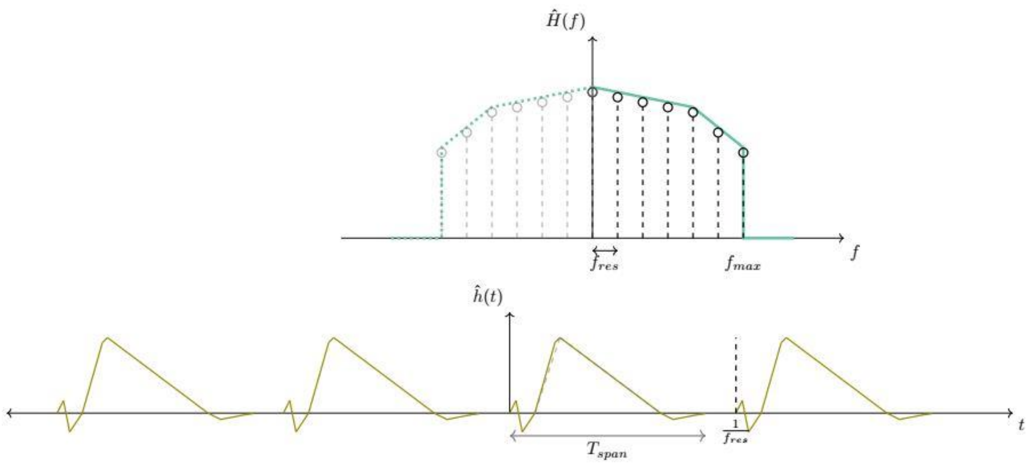

Figure 5. (Top) A hypothetical channel transfer function. The time domain of the frequency sampled data.

Figure 6 shows a signal with a 4 arbitrary unit (a.u) span and a variable delay τ that assumes the values 1, 5, 9 and 25 a.u. The signal is frequency sampled with $f_{res} = 10\ a.u$. From the Fig. $T_{span} \approx \tau + 4\ a.u$. Therefore for $\tau = 1, 5$, $f_{res} < 1/T_{span}$ and, as the Figure shows, the frequency sampled signal can be unambiguously restored. For the other two cases, aliasing occurs. The response when $\tau = 25\ a.u$ is particularly interesting. In this case, the signal shape is correctly restored albeit at a different delay, The inverse Fourier Transform algorithm restores the signal copy from 0 to approx. $1/f_{res}$. In the case shown in the last panel of Figure 6, this is equivalent to a delay of 5 a.u; or equivalently a considerably shorter channel.

As will be shown in the next section without suitable measures, f_{res} limits the number of symbols that can be simulated using a symbol-by-symbol approach.

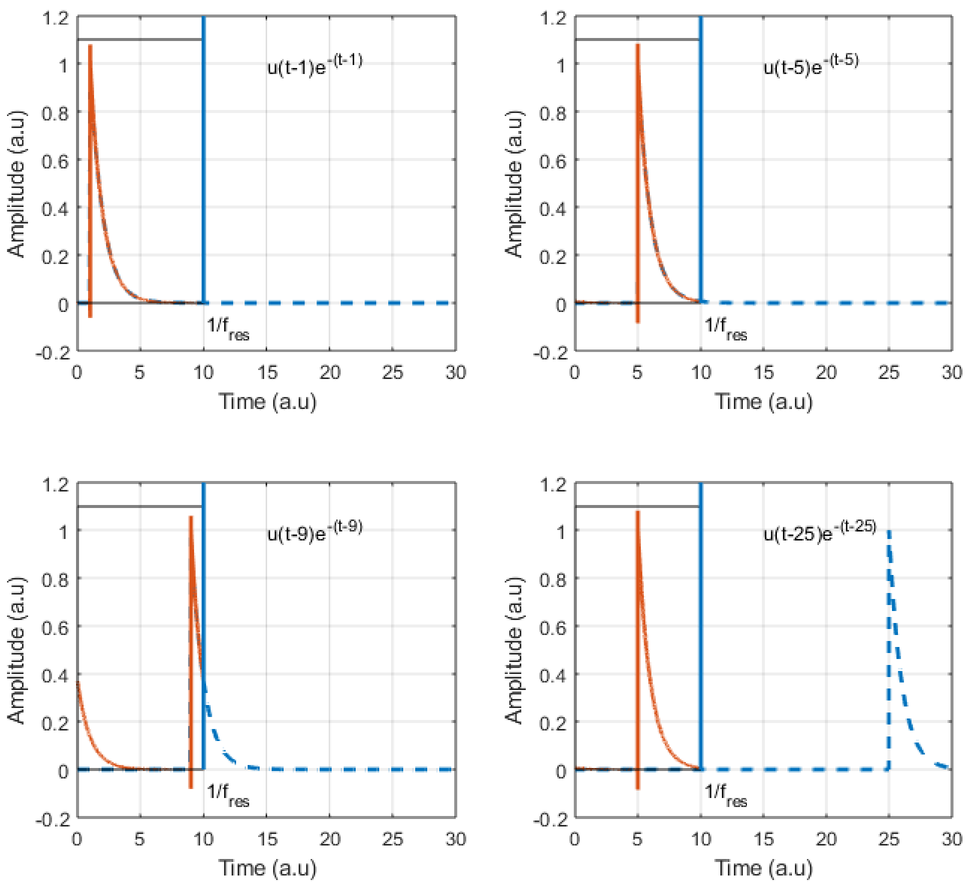

Figure 6. Resolution frequency is fixed, while response delay is changed. Dashed curve represents the original signal. Solid line represents the signal recovered from the inverse transform of the frequency sampled data.

Application to PAM-n Communication Links

The previous section discussed two important factors that, if overlooked, may lead to erroneous calculation of the time domain response of a frequency-sampled channel. In the next two subsections, we will determine the minimum f_{max} and maximum f_{res} that can safely be used to reconstruct the channel impulse response.

Communication System and DC offset

Figure 7 shows a typical VSR or XSR communication system. In general, the input is the source of symbols that pass through a wave shaper (Gaussian, Rectangular or Raised Cosine filters) and possibly a Tx feed forward equalizer. The Rx front end may include a LPF to limit noise and possibly a Continuous Time Linear, Feed-forward and/or Decision Feedback equalizers (CTLE, FFE and/or DFE). As the input symbols pass through the channel they get distorted. The different equalizers strive to restore the signal such that at the sampling point (OUT), the symbols can be determined with an error rate that is below a certain symbol error rate (SER).

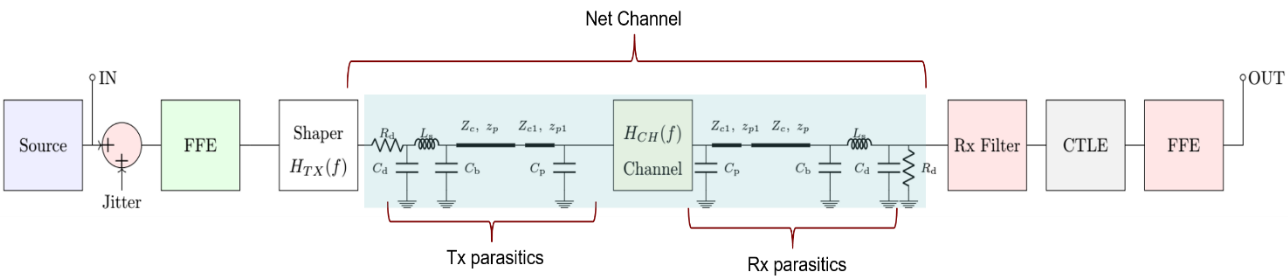

Figure 7. A typical Very Short Range Communication System, including Transceivers parasitics and possible equalizers.

Previously, we have shown that incorrectly setting the DC value results in a small discrepancy that sustains over the whole impulse response to guarantee that the total area matches the newly added DC value. This minute change will accumulate and may affect the symbols probability distribution. Quantitatively, this can be described by noting that at a given sample time $t = 0$, ISI results from the stochastic accumulation of different cursors

$$s(t) = a_0 c(0) + \sum a_k c(kT),$$

where a_k is the k^{th} symbol amplitude uniformly drawn from a set of symbols, T is the UI duration, $c(t)$ is the pulse response, and is the amplitude of the a_0 symbol resulting from ISI.

The ISI term presented in the above equation can be considered as a distortion added to the main signal part $a_0 c_0$. Hence, we can assess the system quality using a signal to distortion ratio (SDR) that is defined as follows

$$SDR = \frac{c^2(0)}{{\sigma_X}^2 \sum c^2(kT)},$$

where

$$\sigma_X^2 = \frac{L^2 - 1}{3(L-1)^2}$$

is the variance of the symbols distribution function and L is the type of modulation ($L = 2$ for NRZ, $L = 4$ for PAM-4). Since SDR lumps the effects of different cursors, it can be used as a metric to quantify the effect of different DC extrapolations.

From a communication system perspective as the one shown in Figure 7, the channel transfer function $H_{CH}(f)$ combines with the other subsystems $H_{sub}(f)$. This means that

$$H_{CH}(0)H_{sub}(0) = \int_{-\infty}^{\infty} h_{CH}(t) * h_{sub}(t)dt,$$

where '*' stands for the convolution operation. This means that the effect of DC still propagates to the communication system output.

To explore the impact of DC extrapolation, we consider the channel whose impulse response is presented in Figure 2. The unequalized and equalized pulse responses are computed, and the SDR is calculated in each case over one UI. Figure 8 shows the pulse response for the two studied cases: Transparent and DC blocked.

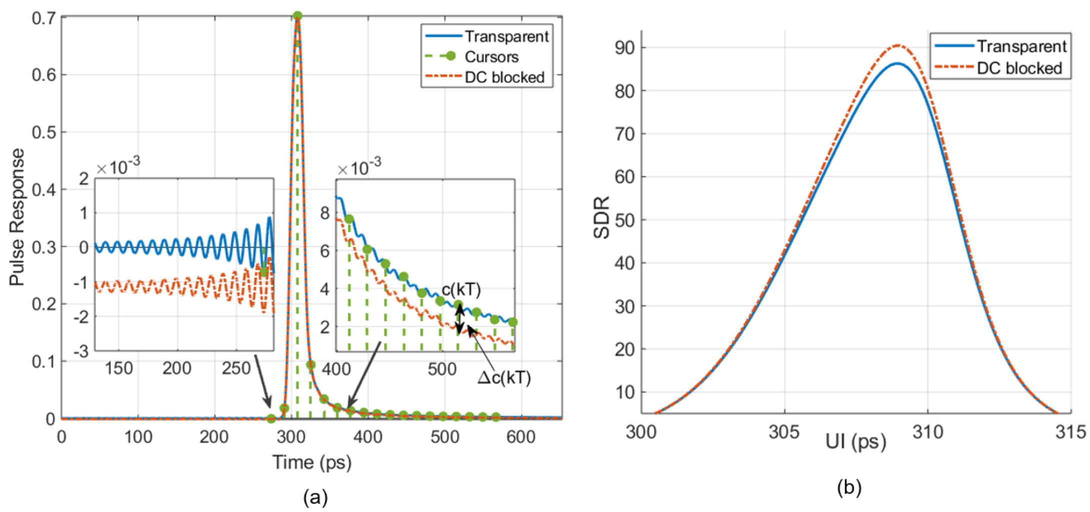

Figure 8. Two possible DC extralpolation. (a) Unequalized Pulse Response. (b) Unequalized SNR.

Recall that the area under the pulse response must match $H_{CH}(0)H_{sub}(0)$, which is a finite positive number for the transparent case and a zero for the DC blocked one. Hence, the response for the DC blocked case is below the other one and can be negative when the response of the transparent case decays to zero (Figure 8(a), left inset). This means that for the DC blocked scenario the SDR can be slightly smaller (the summation in the SDR expression was taken over 80 UIs). Note that, as the left inset in Figure 9(a) shows, the response of the DC blocked is negative. If more precursors are considered, SDR of both scenarios will become closer. Note that due to the fact that $c_k \gg \Delta c_k$, the effect of the DC extrapolation on SDR is not significant.

On the other hand, when the pulse is equalized using an Rx CTLE and 3-taps Tx-FFE c_k becomes smaller and it may be comparable to Δc_k, resulting in a significant change in SDR. Figure 9 presents the results for an equalized pulse, where it is clear that for the DC blocked situation the SDR degrades by almost 50%. SDR degradation leads to a reduction of the eye height and an increase of VEC as Figure 10 shows. To accurately consider the effect of far away cursors, the simulation was done using 40 K randomly transmitted symbols and a relatively high SER of 1E-2.

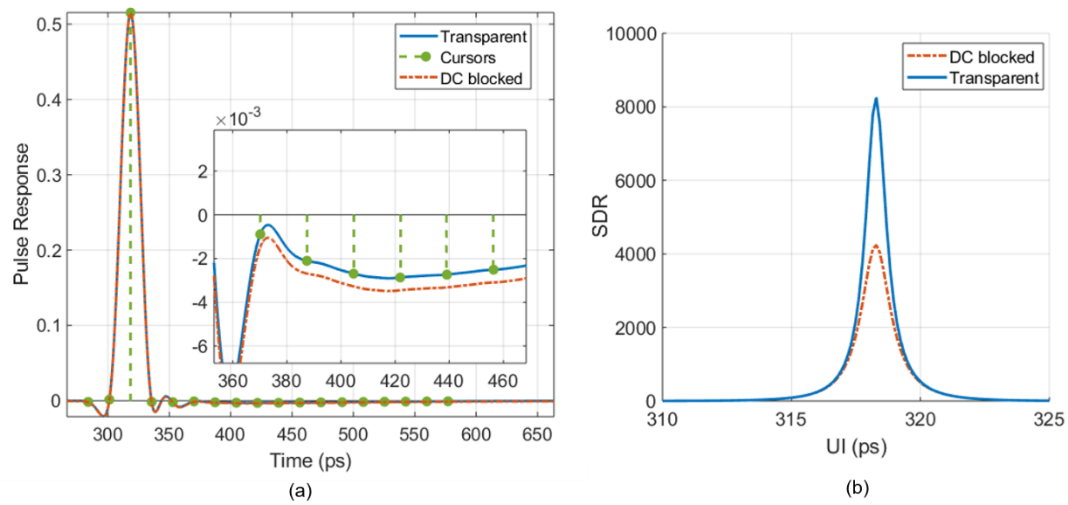

Figure 9. Two possible DC extralpolation. (a) Equalized Pulse Response. (b)Equalized SNR.

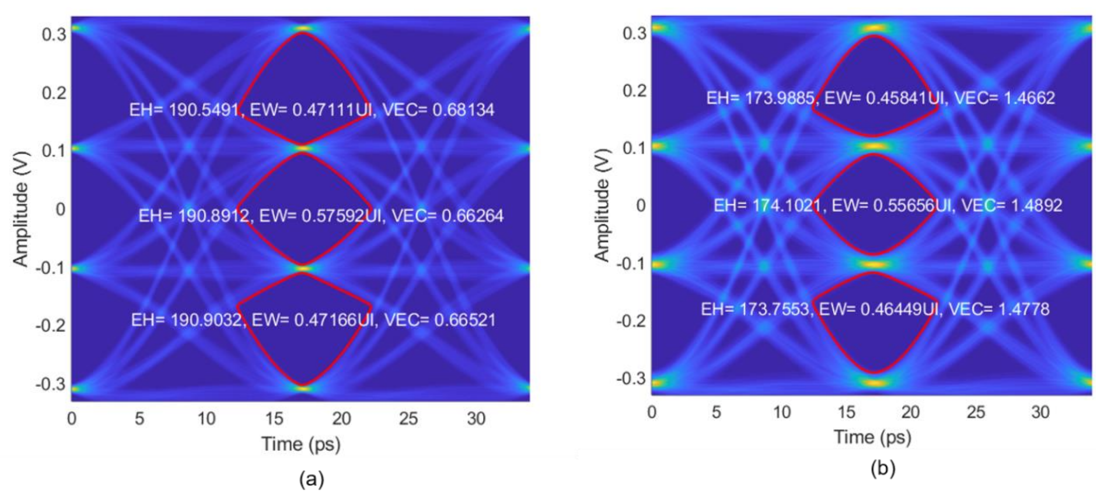

Figure 10. Eye plot and SER=1E-2 Contour: (a) Transparent, (b) DC blocked.

Fortunately, at DC the S parameters or the voltage transfer function can be determined by either direct multi-meter measurement or inspection. In the transparent case, the DC resistance R_{DC} of the channel can be measured, where the DC voltage transfer function can be calculated from

$$H_{DC}(0) = \frac{R_L}{R_{DC} + R_L} \approx 1,$$

where R_L is the DC load resistance, usually 100 Ohm and $R_{DC} \ll R_L$. Using a DC blocking capacitor,

$$H_{DC}(0) = 0.$$

Effect of f_{max}

From the aforementioned discussion it is clear that the better the channel (including parasitics) is, the less the distortion and consequently the better the eye become. From a characterization point of view, however, the more transparent the channel is the higher the value f_{max} must be; which may require a sophisticated and expensive experimental setup. Fortunately for the typical system shown in Figure 7, f_{max} of the total transfer function is approximately equal to the baud rate R_b. This is mainly due to the bandwidth limitation imposed by the pulse shaper and other filters in the system. To see how this is the case, consider the situation, where parasitics are ignored (or lumped inside $H_{CH}(f)$) and all filters except $H_{TX}(f)$ are bypassed. Therefore, the system transfer function $H(f)$ becomes

$$H(f) = H_{TX}(f)H_{CH}(f).$$

Hence, the bandwidth of $H(f)$ is determined by (at most) the minimum bandwidth of $H_{TX}(f)$ and $H_{CH}(f)$. Figure 11 depicts a typical VSR system, where $R_b = 58\ GBps$. A rectangular pulse shaper is used such that $H_{TX}(f) = sinc(f/R_b)$. Note that the channel transfer function $H_{CH}(f)$ has non-negligible components above R_b. However the limited bandwidth of $H_{TX}(f)$ limits the bandwidth of $H(f)$ to essentially R_b.

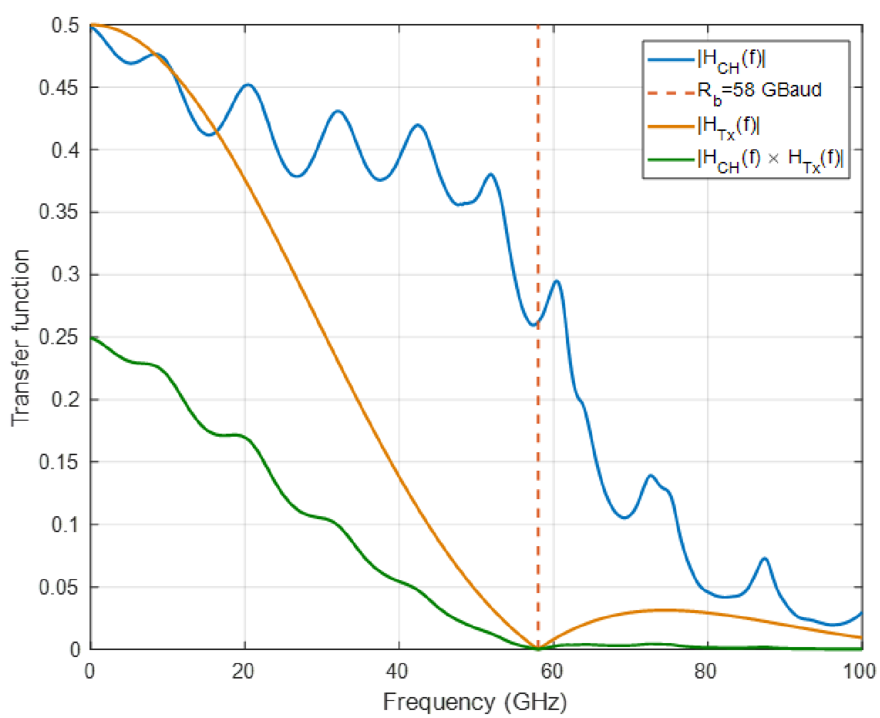

Figure 11. Bandwidth of a typical VSR channel and of the shaping filter.

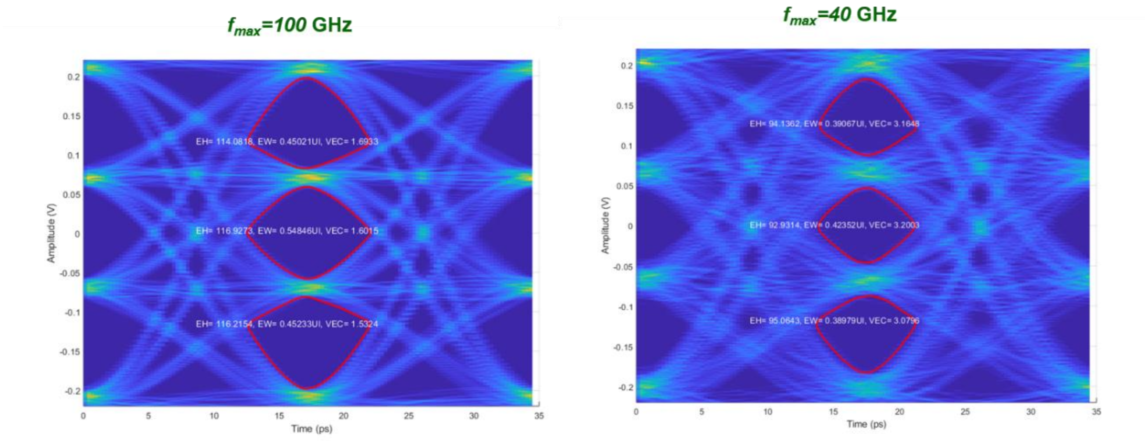

Figure 12. Eye diagram with SER=1E-2, no noise, no jitter for two different maximum frequency values. The channel is equalized using a Tx 3-taps FFE and an Rx CTLE.

Figure 12 presents the eye diagram of a typical equalized XSR channel, running at 58 GBd. Equalization was achieved using the full bandwidth (100 GHz) via the application of a zero forcing algorithm. A relatively high SER of 1E-2 was used to reduce the simulation time. The frequency f_{max} is changed from 100 GHz down to 40GHz, where all other system parameters are kept fixed. Note how the truncation of the bandwidth deteriorates the eye. For instance, the eye height is reduced by approx. 20%. To correlate this with the emergence of artificial cursors similar to the ones shown in Figure 4, please refer to Figure 13 that shows the equalized pulse response for three different cases $f_{max} =$ 40, 50 and 100 GHz. Note that the pulse response does not significantly change when f_{max} changes from 100 GHz to 50 GHz ($\approx R_b$). However as f_{max} drops to 40 GHz, we observe stronger ripples due to the interaction with the sinc function representing the bandwidth truncation.

Furthermore, Figure 14 shows the optimal equalizers' parameters when one considers the frequency domain data up to 40 GHz and 100 GHz. It is clear that using a severely truncated bandwidth may result in different equalization, where it may even sway the engineer's decision in the wrong direction. As a rule of thumb, it is recommended to set $f_{max} \approx R_b$ or larger.

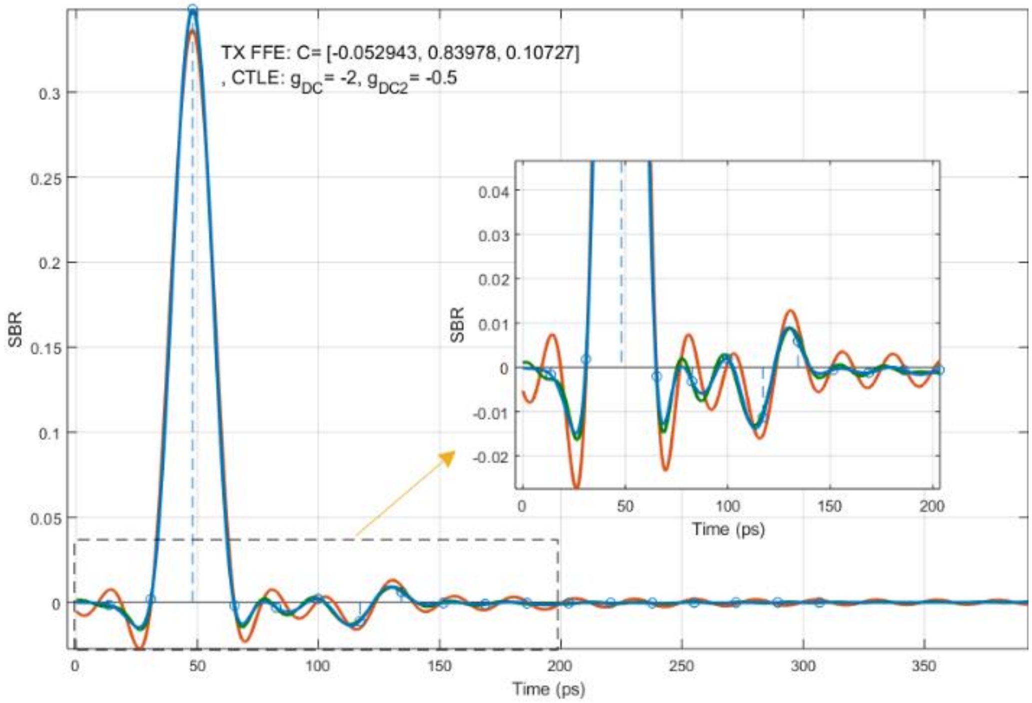

Figure 13. Equalized Pulse response for three different maximum frequencies (Blue 100 GHz, Green 50 GHz, Red 40 GHz).

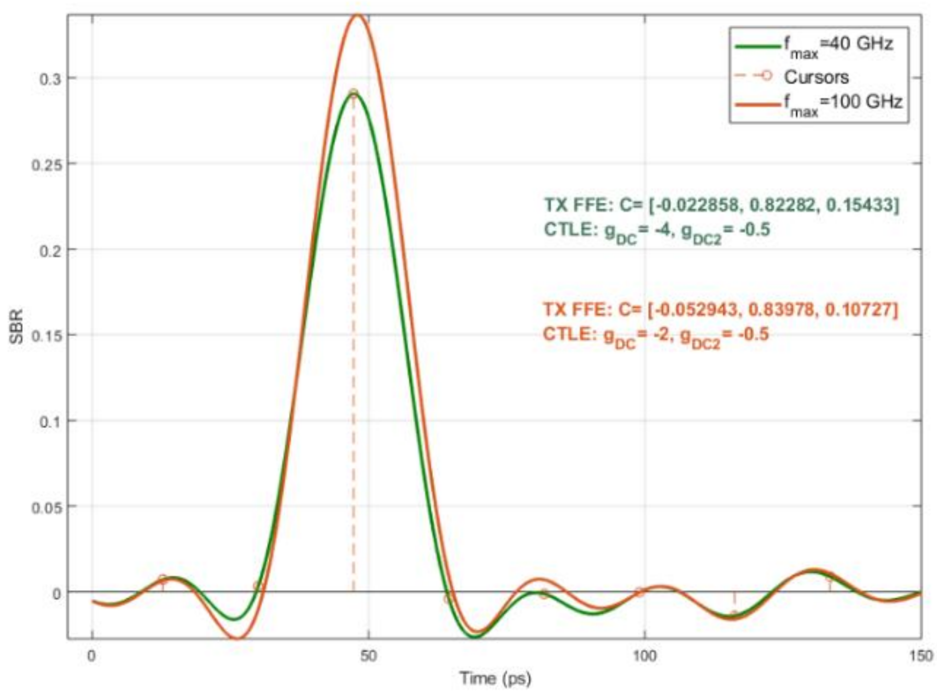

Figure 14. Equalized Pulse response at two different max. frequencies.

Effect of f_{res}

As will be seen, a wrong selection of f_{res} can lead to a dramatic change in the eye plots. Fortunately, this is a sign that can be detected and rectified. The time span of the symbol streams at the output consists of three main components

$$T_{span} = T_D + NT_b + T_{relax},$$

where T_D is the delay due to channel length. N is the total number of symbols used in the simulation, NT_b is the total length of the pulse stream, and T_{relax} is the relaxation time (i.e, the duration at which the last pulse decays). Quite often, N is large enough to warrant

$$T_{span} \approx NT_b.$$

Therefore, the resolution frequency must be bounded by R_b/N, i.e.,

$$f_{res} < \frac{R_b}{N}$$

Typically for 56-58G-VSR channels f_{res} is set to be 10-20 MHz. Therefore, the maximum number of symbols that can be simulated is between 2900 and 5800; not enough for estimating SER contours less than 1E-3.

Figure 15 shows the eye diagram of a typical channel, where $f_{res} = 20\ MHz$. At 58 GBd, beyond $N = 2900$ aliasing will occur (compare Figure 15(a) with Figure 15(b)). Fortunately, the correct response can be restored by interpolating over a frequency grid, as shown in Figure 15 (c), such that $f_{res} < R_b/N$. In our case this means that the f_{res} must be less than approximately 14 MHz. For SER less than 1E-6, f_{res} will be less than 6 KHz. Therefore as a rule of thumb interpolation should be done over a grid with a frequency spacing (i.e, resolution frequency) such that

$$f_{res} < \frac{R_b}{N}$$

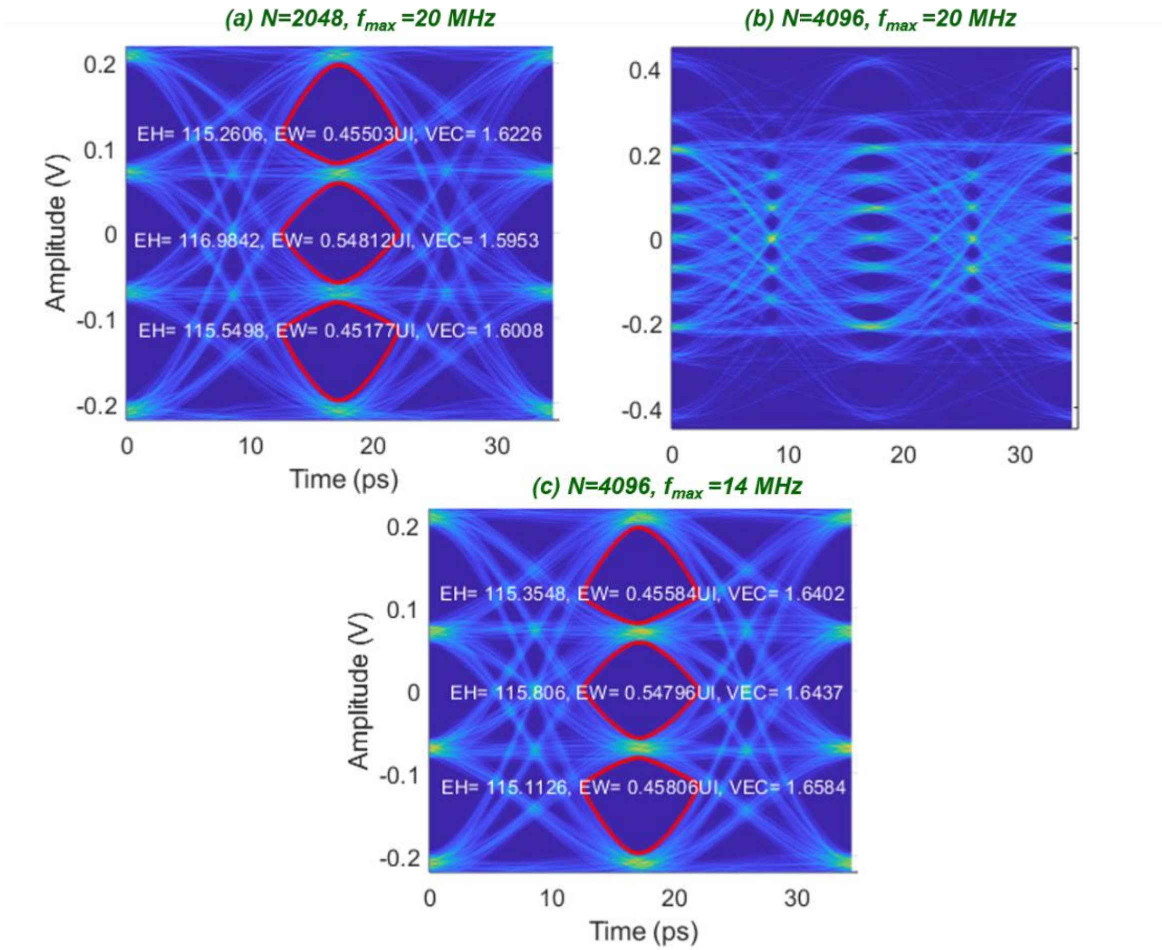

Figure 15. Eye diagram for different number of symbols and resolution frequency.

Conclusion

The impact of a channel measured or computed maximum and resolution frequencies is evaluated, where it is shown how it can affect the system simulation and hence signal integrity decisions. Rules of thumb are recommended. The channel maximum frequency should be at least equal to the intended baud rate. For meaningful channel simulation results the resolution frequency must be less than R_b/N to avoid aliasing effects. Additionally, DC extrapolation must be carried out such that it accurately describes the system at DC.

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