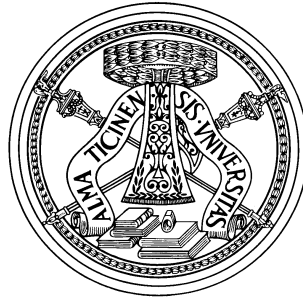


UNIVERSITY OF PAVIA

Department of Electronic Engineering



PH.D. THESIS IN MICROELECTRONICS

XXVIII CYCLE

Adaptive Analog Transversal Equalizers for High-Speed Serial Links

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October 2015

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Introduction

The growing popularity of advanced network services such as multimedia-on-demand the fast expansion of storage and computing on the cloud are powerful drivers in expanding data traffic. Every day, more users are more quickly accessing the Internet in more ways, to utilize more applications and consume more content that demands more bandwidth. Moreover, as CMOS technologies are scaled to finer dimensions and the density of digital computing cores rises, the aggregate I/O system bandwidth must be increased to harness all of the computing power available. Both technology trends and new applications have created a large demand for high-speed data communication over optical fibers and backplane channels and at all levels of the I/O hierarchy, including intra-chip, chip-to-chip, rack-to-rack and system-to-system. Consequently, fundamentals bottlenecks are appearing everywhere throughout the Ethernet networking and the future holds only more mobile, more video, more devices and more data.

In this scenario the global Ethernet system is moving now to create a plan to evolve beyond today's 100 Gigabit per second capabilities, developing four new Ethernet speeds, 2.5, 5, 25 and 400 Gigabit Ethernet (GbE), to add to the existing six speeds, Megabit Ethernet (MbE), 100MbE, GbE, 10GbE, 40 GbE and 100GbE. Over the next decade, several more speeds are being considered, including 50GbE, 200GbE and multiple speeds beyond 400GbE. Together, these speeds, define the core of the 2015 Ethernet Roadmap [1]. To address the I/O needs of future computing and network systems, single serial link data rates are now being pushed up to 25-28 Gb/s, as exemplified by standards such as OIF CEI-25G-LR, CEI-28G-SR [2] and IEEE 802.3bj (100GbE over backplane and copper cable) [3]. These standards address both short-reach (SR) and long-reach (LR) serial link channels. For short-reach links (with roughly 15 dB or less of channel loss), reliable signaling can be achieved with relatively simple and power-efficient transceivers. For long-reach

links such as backplanes, however, the channel losses are much higher, so more complex transceivers with sophisticated equalization are needed.

In the past, the interconnections were mainly parallel type but, with the growing data speed connection, clock skew and crosstalk problems in parallel transmission have shifted the attention on serial connection. In parallel transmission, multiple bits (usually 8 bits or a byte) are sent simultaneously on different wires within the same cable. As a result there is a speedup in parallel transmission bitrate over serial transmission bitrate. However, this more speed is a tradeoff versus cost since multiple wires are more expensive than a single wire and, as a parallel cable gets longer, the synchronization timing between multiple channels becomes more sensitive to distance. Today, especially for long channel, serial transmission is preferred. The bits are sent sequentially on the same wire, which reduces costs for the channel, reduces the crosstalk and, only for asynchronous transmission, no data link synchronization avoids skew problems and makes the system simpler. The main problem is that as data rates increase, the variation in channel responses becomes more severe and with the same equivalent speed, serial connection respect to parallel connection shows more insertion channel loss. Channel loss, function of frequency, results in Inter-Symbol-Interference (ISI) decreasing the Bit Error Rate (BER). The problem can be solved in two ways: the first using additional circuits equalizers in the receiver and/or in the transmitter to compensate channel loss, the second involves the use of better channels to introduce lower losses. For cost reasons, the standards and industry prefer as much as possible the first approach.

To accommodate many different interconnects, channels, backplane topologies and various configurations, adaptive equalizers are used to remove the ISI and extend the maximum I/O data rate. In general, the equalizers can be implemented at the transmitter or receiver. Adaptive receiver equalization has advantages over adaptive transmit equalization. First, transmit equalization constrains the magnitude sum of the equalizer taps which reduces the bit amplitude. Second, adaptive transmit equalization requires the receiver information be conveyed back to the transmitter.

Chapter 1 is an introduction on high-speed serial communication system. It will discuss the requirements and the reasons that led to the design of the two equalizers shown in the following chapters.

Chapter 2 is focused on description of wire-line serial link describing the typical model of a backplane communication channel, the different categories used in a

equalization system and the algorithms required to make the system adaptive.

Chapter 3 discuss about a novel design for a “A 25-Gb/s FIR Equalizer Based on Highly Linear All-Pass Delay-Line Stages”, explaining the fundamental building blocks, its behavior and finally presenting some measurements of the chip implemented in 28nm CMOS technology.

Chapter 4 shows a design of a innovative Derivative-FIR Equalizer with a remarkable improvement in the allowed input signal amplitude. An overview on the main blocks that compose the RX chain and some simulation results of the circuit are given.

Chapter 1

High-speed serial communication

1.1 Binary sequences

1.1.1 Pseudo-random binary sequence properties

High-speed communication systems usually use binary type signals to make easier the detection of the bits. The most used encoding in these systems, it means the representation of the logic levels through voltage levels, is the Non-Return-to-Zero (NRZ). However, new codes are emerging to relax the bandwidth requirements, as the PAM-4 (Pulse-Amplitude-Modulation-4) that uses four voltage levels.

A Pseudo Random Binary Sequence (PRBS) is an ordered set of numbers that has been determined by some defined arithmetic process but is effectively a random number sequence for the purpose for which it is required. A PRBS is “pseudorandom”, because, although it is in fact deterministic, it seems to be random in a sense that the probability of the one-levels is independent of the values of any of the other elements, similar to real random sequence. The sequence has a maximum length N and can be stretched to infinity by repeating it after N elements. This is in contrast to random sequence.

The knowledge of PRBS properties allows a careful evaluation of various design choices. The information inside a binary sequence is got by the alternation of two logic values that occur with equal probability. The figure 1.1 shows an example. Two different amplitude voltage levels, respectively $+V_0$ and $-V_0$, represent the two logic levels “ONE” and “ZERO”.

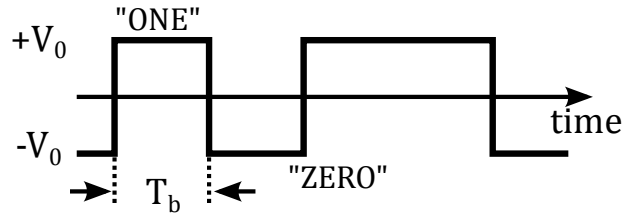


Figure 1.1: NRZ signal

If each bit lasts T_b seconds, then $BR = 1/T_b$ is the bit rate, that is the number of bits per second. The period T_b is also known as Unit Interval (UI). A binary sequence can generate several consecutive bits with the same logic value. In this case the information shows a low transition density and this may lead to some problems. Indeed, in the absence of transitions, it is difficult to maintain synchronization and for this reasons the standards typically define a maximum tolerable length of consecutive bits equal to each other. In time domain the binary sequence $x(t)$ can be expressed as:

$$x(t) = \sum_k^N b_k p(t - kT_b) \quad (1.1)$$

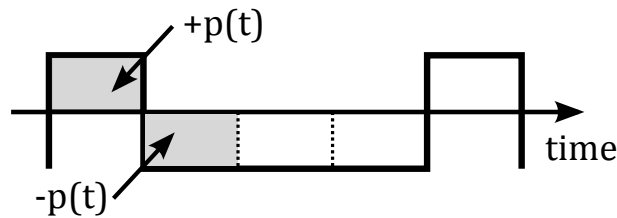


Figure 1.2: $x(t)$ signal

where $b_k = \pm V_0$ and $p(t)$ is the rectangular pulse function as shown in figure 1.2. The signal $x(t)$ is the sum of k pulses of period T_b , amplitude V_0 and with a kT_b delay. Assuming that the positive and negative pulses occur with equal probability and the amplitude $b_k = \pm 1$, you can express the Power Spectral Density (PSD) of the signal $x(t)$ as:

$$S_x(f) = \frac{1}{T_b} |P(f)|^2, \quad (1.2)$$

where $P(f)$ is the Fourier transform of the signal $p(t)$. Considering that $p(t)$ is a rectangular pulse function, the Fourier transform $P(f)$ will be a cardinal sine function:

$$P(f) = T_b \left[\frac{\sin(\pi f T_b)}{\pi f T_b} \right], \quad (1.3)$$

and finally the $x(t)$ spectrum will be given by:

$$S_x(f) = T_b \left[\frac{\sin(\pi f T_b)}{\pi f T_b} \right]^2. \quad (1.4)$$

In figure 1.3 it is shown the $x(t)$ spectrum, where it can be noted that power is zero for the frequencies $f = n/T_b$, where n is an integer number. Most of the energy is within the first lobe. However, as explained in a following section, the minimum required bandwidth in order to not impair the quality of the signal is up to the frequency $1/2T_b$ [4].

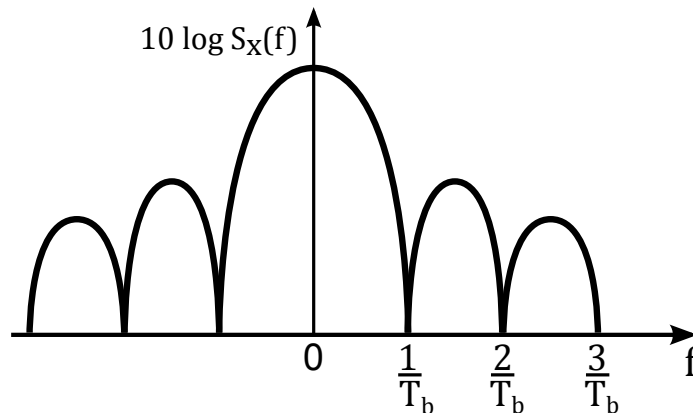


Figure 1.3: Power Spectral Density of $x(t)$

1.1.2 Inter-symbol interference (ISI)

Inter-symbol interference (ISI) is a form of distortion of a signal in which one symbol interferes with adjacent symbols. This is an unwanted phenomenon because the spreading of the pulse interferes with neighboring pulses causing errors in the decision device at the receiver. One of the principal causes of inter-symbol interference is the transmission of a signal through a band-limited channel, i.e., one where the frequency response is a low-pass transfer function. Passing a signal

through such a channel results in the attenuation of high frequency components that affects the shape of the pulse that arrives at the receiver. A NRZ signal waveform starts to spread and merge with the adjacent symbol sequence, making the data unreadable. At the receiver end, the data is wrongly decoded, because the receiver cannot predict the correct amplitude level of the square waveform, leading to the loss of information. Therefore, in the design of receiving circuits, the objective is to minimize the effect of ISI obtaining the smallest error rate possible. Error rates are minimized through the use of adaptive equalization techniques and error correcting codes. In figure 1.4a, a random binary sequence, while the figure 1.4b shown the effect when the signal crosses a generic low-pass channel, where the high frequency filtering causes slower rising and falling edges.

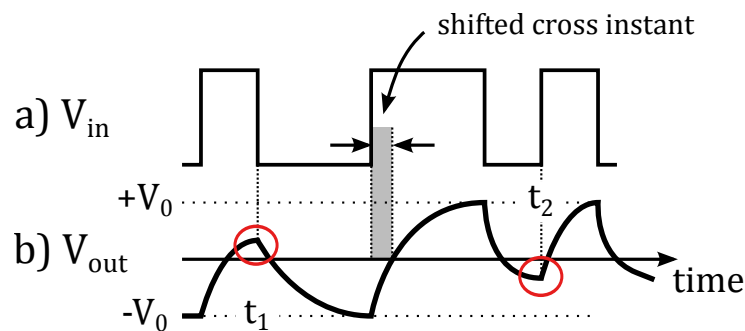


Figure 1.4: a) Ideal NRZ sequence: b) Effect of low-pass filtering on the sequence

As shown in the last figure, at time t_1 a single "ONE" between two "ZEROS" cannot reach the maximum signal level V_0 . At the other side, between t_1 and t_2 , a consecutive bits sequence allows to reach the maximum signal level. For example, with a zero voltage threshold, due to the noise, signal at time t_1 and t_2 may be misinterpreted by the receiver. Moreover, the zero crossing instants shift, introducing jitter. This phenomenon makes difficult the choice of an optimal voltage threshold for the receiver. These effects result in inter-symbol interference and since, the channel response is not known beforehand, an adaptive equalizer is used to compensate the frequency response.

1.1.3 Bandwidth requirements

To process correctly NRZ signal, the choice of an excessive bandwidth represents a penalty in terms of power consumption. It is important to identify the minimum required bandwidth for the circuits and not exceed significantly. The Nyquist theorem defines the minimum bandwidth that the communication system must possess to transmit, without ISI, a data sequence with a bit rate of BR bit/s by using a "sinc" pulse. This bandwidth is equal to $BR/2$. However, "sinc" pulses are not causal and in the practice can be only approximated. Luckily, a "square" pulse needs a bandwidth slightly higher respect to "sinc" pulse and the Nyquist frequency $BR/2$ keeps a good reference for the minimum bandwidth, which gives negligible ISI.

The optimum bandwidth is between $0.5BR$ and $0.7BR$. A lower bandwidth introduces excessive ISI, while a greater bandwidth doesn't lead to further advantages. Rather, in addition to power consumption, an oversized bandwidth introduces more noise leading to a penalty in terms of SNR [5].

1.1.4 Jitter

Jitter is the short time deviation of the edges of a signal from their ideal positions. This is one of the multiple definitions for the jitter. Jitter is a significant and undesired effect in most of communication links and can be quantified in the same terms as all time-varying signals, e.g., Root Means Square (RMS) or peak-to-peak displacement. The jitter can be organized in a diagram tree, figure 1.5, which shows the different types of jitter.

Mainly, jitter consists of Deterministic Jitter (DJ) and Random Jitter (RJ). Random jitter is caused by the combination of a huge number of sources, each of very small magnitude. Thermal processes, microscopic variations in the resistance impedance of circuit traces and so on, primarily cause RJ. Since, RJ follows an unbounded distribution, it should show a Gaussian distribution. There is a finite probability that random effects could cause a logic transition to appear anywhere and the spread is described by the standard deviation of the distribution.

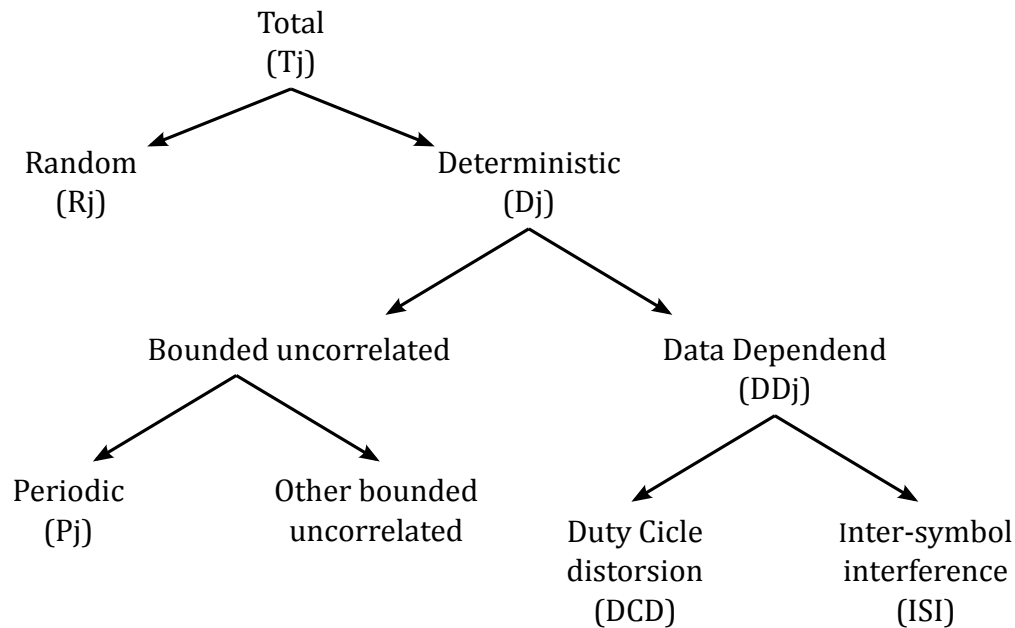


Figure 1.5: Jitter diagram tree

Deterministic Jitter is the jitter that remains after RJ has been removed. In principle, though almost never in practice, DJ can be calculated from a complete understanding of the circuit and its environment. Since DJ can be composed of all the other types of jitter, it doesn't follow a given function distribution the way that RJ follows a Gaussian. On the other hand, since DJ is composed of a finite number of deterministic processes, its distribution is bounded [6]-[7]. In subsequent rows, follow other definitions of the different types of jitter that make up the tree.

- **Data Dependent Jitter:** DDJ includes all jitter whose magnitude is affected by the transmitted data signal.
- **Duty-Cicle Distortion:** DCD is a measure of the asymmetry in the duty cycle of the TX. It is usually caused by an asymmetry in either the clock signal driving the transmitter or in a limiting amplifier within the transmitter.
- **Inter-Symbol Interference:** ISI is the primary cause of DDJ. The situation is complicated by the correlation of ISI and Duty-Cycle Distortion (DCD).
- **Periodic Jitter:** PJ includes any jitter at a fixed frequency. It's easy to measure accurately and appears in the jitter-frequency spectrum as distinct peaks.

1.1.5 Quality signal measurements

1.1.5.1 Eye diagram

The data eye diagram is a methodology to represent and analyze a high-speed signal. The signal integrity can be observed through the appearance of the eye, evaluating the amount of ISI, noise and jitter. The data eye diagram is constructed from a digital waveform by folding the parts of the waveform corresponding to each individual bit into a single graph with signal amplitude on the vertical axis and time on horizontal axis. By repeating this construction over many samples of the waveform, the resultant graph will represent the average statistics of the signal and will resemble an eye. Figure 1.6b shows an open eye diagram constructed from a received sequence sketched in figure 1.6a.

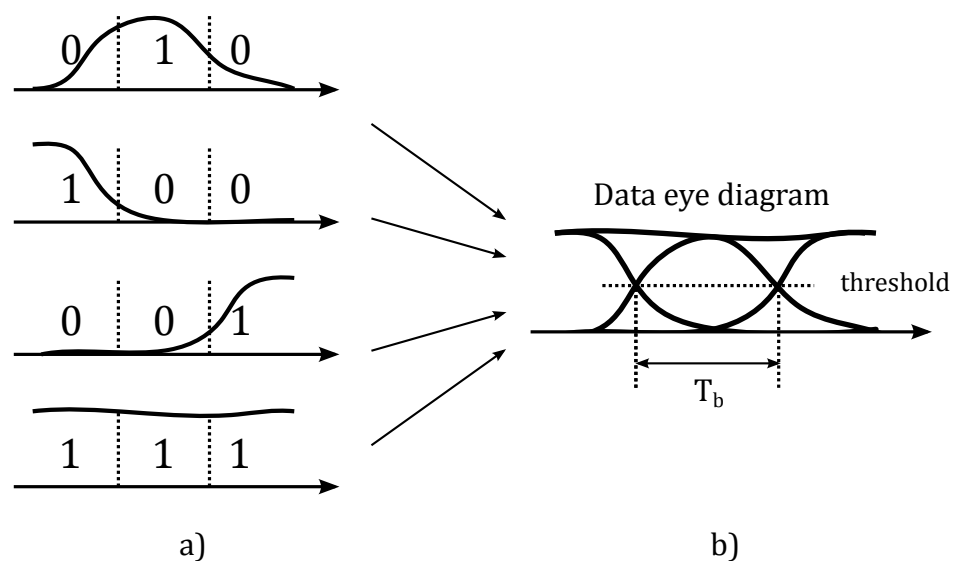


Figure 1.6: Signal representation with eye diagram

The data eye diagram can be characterized through the measurement of various parameters such as the vertical and horizontal opening, that allows quantifying the quality of the signal. The vertical eye opening is measured at the sampling instant (in the middle of the eye) and is expressed as a percentage of the full eye height (not including over or undershoots). The horizontal eye opening is measured at the slice level (threshold) and is expressed as a percentage of the bit interval. Without noise and random jitter, the opening can be determined in a simple way as shown in figure 1.7. The vertical eye closure is caused by inter-symbol interference and the horizontal eye closure is due to the deterministic jitter.

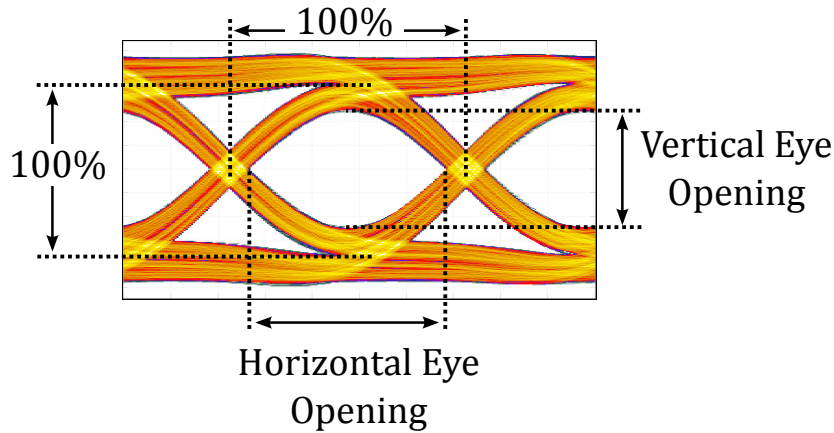


Figure 1.7: Typical measurement on an eye diagram

It is important to understand that the eye closure depends on the length of the PRBS sequence. The sequence length is typically between $2^7 - 1$ and $2^{31} - 1$ (usually named PRBS-7 and PRBS-31 respectively). If the device under test has a low frequency cutoff, the eye closure worsens with increasing the sequence length. Therefore, the sequence length must always be specified when an eye diagram is used.

Considering noise and random jitter, we have an additional complication. For a Gaussian noise distribution, we have to wait long enough to correctly measure the eye openings. Their evaluation makes use of histograms, which describe the signal distribution around a midpoint. As shown in figure 1.8 you can extract two types of histogram, horizontal and vertical, and measure the corresponding standard deviation σ . The horizontal opening is usually measured as the time interval between the 3σ points of the two horizontal distributions. The vertical opening is evaluated in an equivalent way but with the vertical histograms. We can also extract the eye amplitude as shown in the same figure [8].

1.1.5.2 Bit Error Rate (BER)

The most important way to evaluate the performance of a high-speed serial link is the Bit Error Rate (BER). The BER is defined as the ratio between the number of wrong bits received and the number of valid bits received within a certain sequence. The requirements on BER depend on the application, but generally numbers from 10^{-15} to 10^{-12} are typical values.

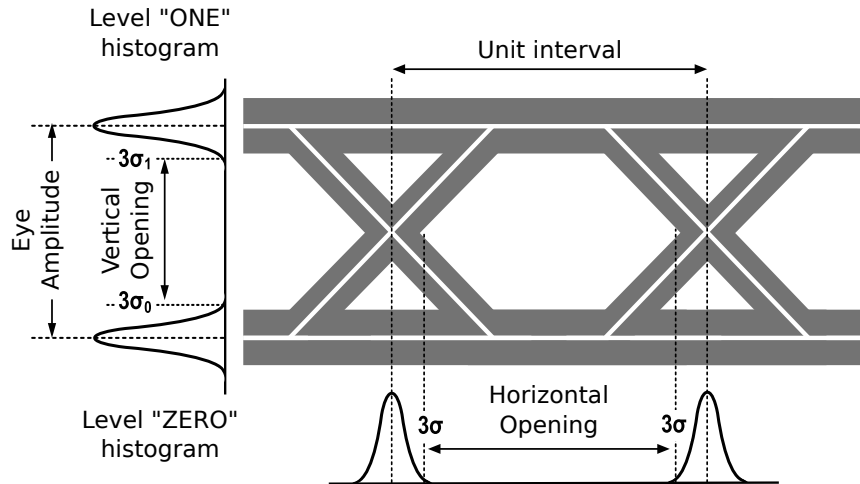


Figure 1.8: Vertical and horizontal eye diagram histograms

To derive the bit error rate expression, suppose to have a system with enough bandwidth and without distortion on the waveform signal [5]. For a generic bits sequence, as shown in figure 1.9a, the transition of the bits can be considered infinitely fast. Without noise, the signal assumes only two values: $+V_0$ for the logic level "ONE" and $-V_0$ for the logic level "ZERO". Figure 1.9b shows the corresponding Probability Density Function (PDF) of the sequence.

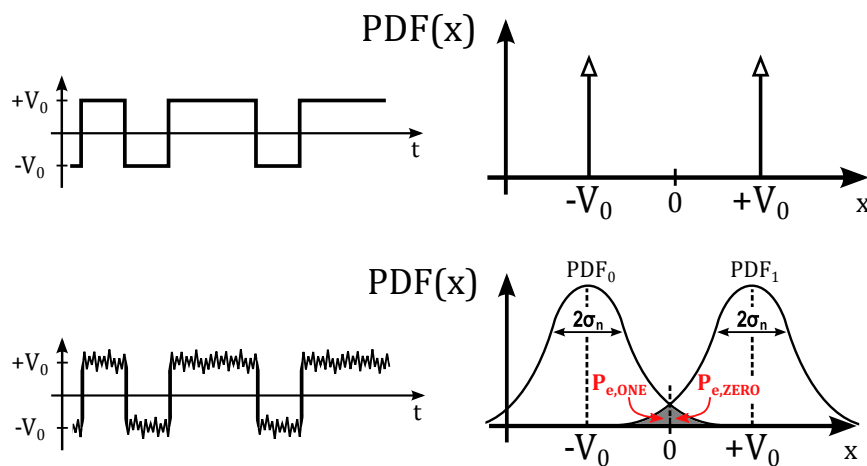


Figure 1.9: Noise effect on a generic bit sequence

Now, let's to introduce white noise on the received signal. Considering the transitions still ideal, the zero crossing position doesn't change. As shown in figure 1.9c the amplitude levels are not well defined and this leads to two different Gaussian distribution around average levels $+V_0$ and $-V_0$. These distributions can be expressed by the following probability density:

$$PDF_1(x) = \frac{1}{\sigma_n \sqrt{2\pi}} \exp \left[-\frac{(x + V_0)^2}{2\sigma_n^2} \right] \quad (1.5)$$

$$PDF_0(x) = \frac{1}{\sigma_n \sqrt{2\pi}} \exp \left[-\frac{(x - V_0)^2}{2\sigma_n^2} \right] \quad (1.6)$$

where σ_n^2 is the distribution variance. Respect to the case without noise, there is always the possibility to make an error since the PDFs extend beyond the zero threshold. Let's to define $P_{e,ZERO}$ the receiver probability to decide "ONE" when it was sent a "ZERO" and vice versa $P_{e,ONE}$ the receiver probability to decide "ZERO" when in fact it was sent a "ONE". These probabilities are the gray areas in figure 1.9d and they correspond to the integral of the PDFs:

$$P_{e,ZERO} = \int_0^{\infty} PDF_0(x) dx, \quad (1.7)$$

$$P_{e,UNO} = \int_{-\infty}^0 PDF_1(x) dx. \quad (1.8)$$

Moreover, since levels "ONE" and "ZERO" have the same probability to being transmitted, we can write:

$$P_e = \frac{1}{2} P_{e,ZERO} + \frac{1}{2} P_{e,UNO} \quad (1.9)$$

where P_e is the total error probability, i.e. the BER. Then, BER can be written as:

$$BER = Q \left(\frac{V_0}{\sigma_n} \right) \quad (1.10)$$

where $Q(x)$ is the error function and represent the above integral, while the ratio V_0/σ_n is the signal to noise ratio SNR_{noise} . The Q function defines the BER through the SNR_{noise} and a graphical representation is shown in figure 1.10. In table 1.1 instead, different values of BER as a function of SNR. Each point in the eye diagram can be interpreted as a decision point and therefore has a BER associated with it. As a result, contours of constant BERs can be plotted inside

the eye. Figure 1.11 shows only one contour with a certain target BER value. The lower the BER, the smaller becomes the area enclosed by the contour. If we make a decision inside this contour, we will find always a lower BER.

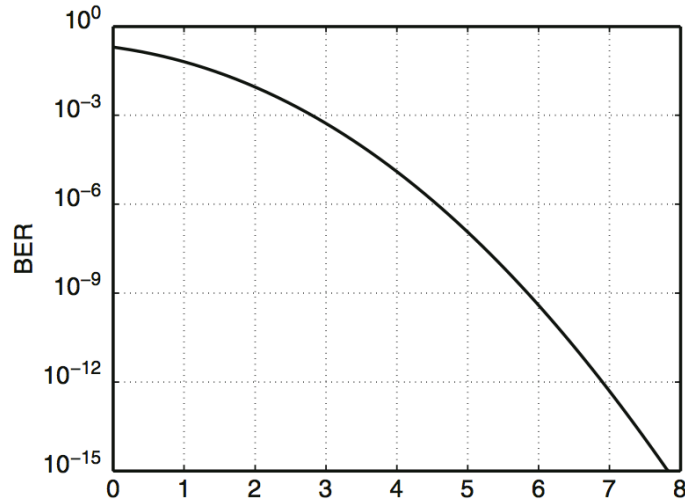


Figure 1.10: Bit Error Rate function of SNR

Table 1.1: BER as a function of SNR

SNR_{noise}	BER	SNR_{noise}	BER
0.0	0.5	5.998	10 ⁻⁹
3.090	10 ⁻³	6.361	10 ⁻¹⁰
3.719	10 ⁻⁴	6.706	10 ⁻¹¹
4.265	10 ⁻⁵	7.035	10 ⁻¹²
4.753	10 ⁻⁶	7.349	10 ⁻¹³
5.199	10 ⁻⁷	7.651	10 ⁻¹⁴
5.612	10 ⁻⁸	7.942	10 ⁻¹⁵

In figure 1.11 are also showed the vertical and horizontal eye margins. If the eye margins are larger than zero, then the decision circuit has a decision-threshold with the possibility to get the right bit and to meet the desired BER.

Eye margins are best measured with an instrument called Bit Error Rate Tester (BERT) that has a pulse pattern generator and an error detector. The instrument is connected to the Device Under Test (DUT), as shown in figure 1.12. The error detector slices the data signal at the decision threshold V_{TH} and samples it at the instant t_R . The recovered bits are compared with the transmitted bit sequence to

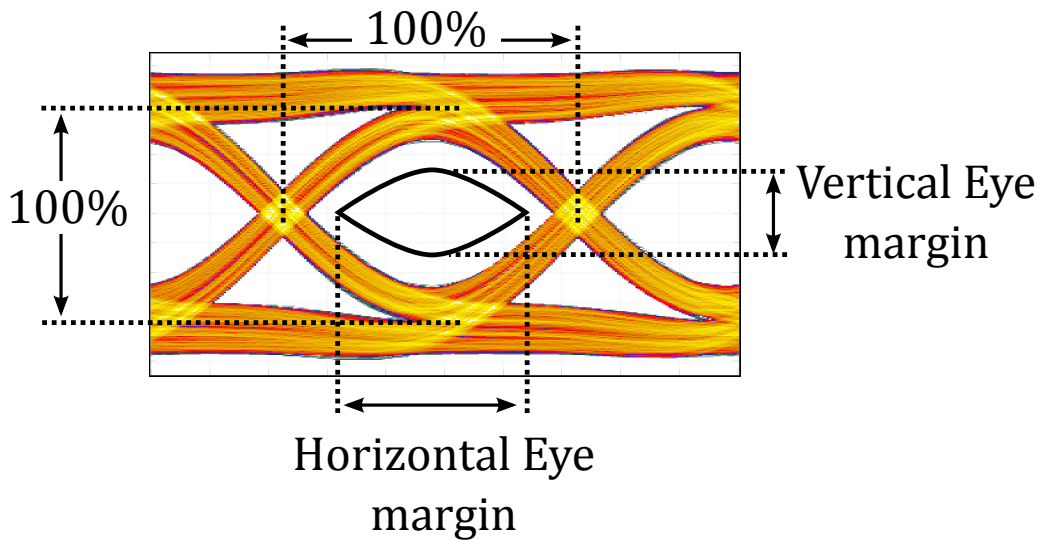


Figure 1.11: Eye margins in a noisy eye diagram

determine the BER, which is displayed on the error detector. Both the decision threshold V_{TH} and the sampling instant t_R are adjustable. A horizontal scan can be performed by setting the voltage threshold to the center of eye and scanning the instant across the eye. The resulting curve, for the particular shape, is named "bathtub" and it is shown in figure 1.13. The BER is low when the instant sampling is at the center of the eye and goes up when the sampling moves to the left or to the right. The horizontal eye margin is defined as the interval between the two points on the left and right side of the eye where the bathtub curve assumes a specified BER value. For example, in the 10GbE standard, the horizontal eye margin is specified for a BER of 10^{-12} .

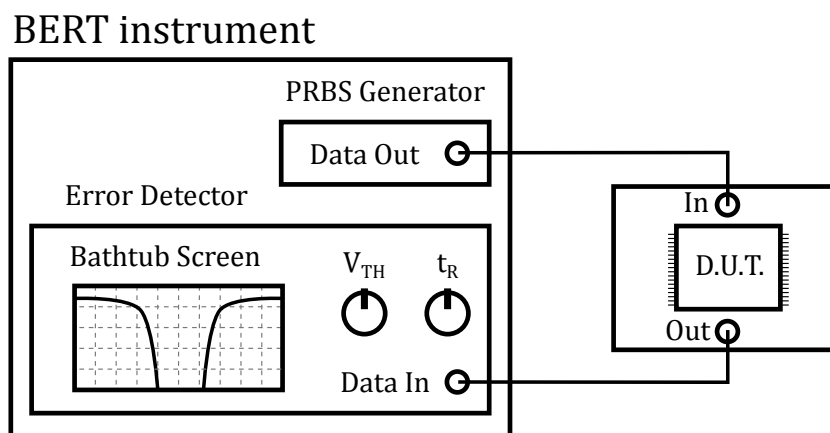


Figure 1.12: Bit Error Rate test setup

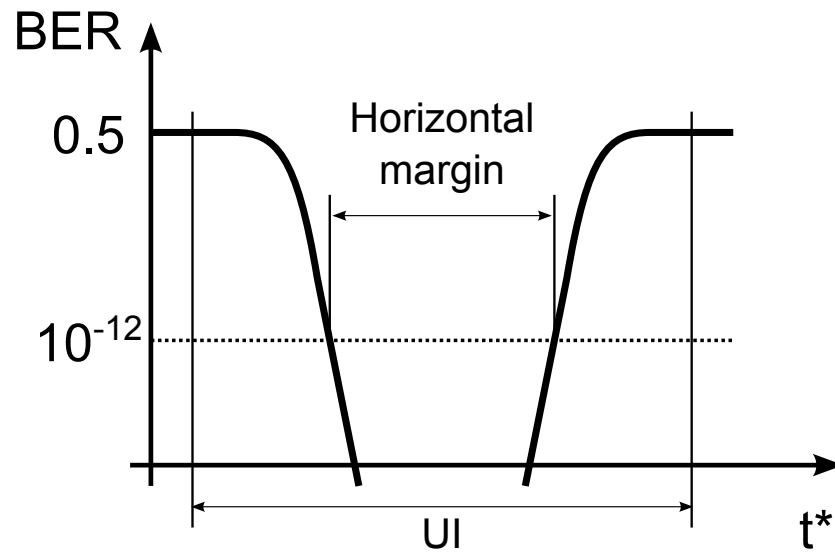


Figure 1.13: Example of bathtub

Chapter 2

High-speed serial link

Due to the density constraints on the number of wires between the chips and for the limited number of I/O pins in the packages of the chip, high-speed links usually serialize the parallel data for off-chip transmission. A simple block diagram of a high-speed serial link is shown in figure 2.1 and three fundamentals blocks compose it: the transmitter, the channel and the receiver [9].

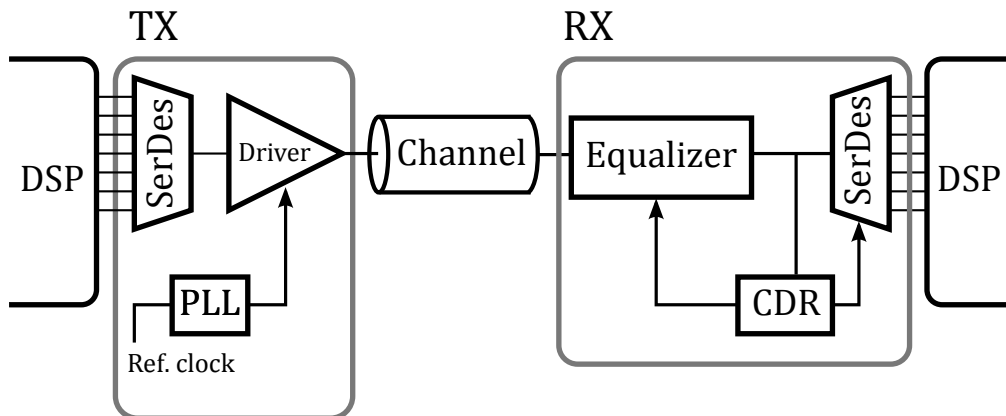


Figure 2.1: Block diagram of a high-speed serial link

The transmitter serializes, modulates and sends the data to the receiver using an internal clock generated by a PLL (Phase Locked Loop). The PLL is the timing generator in a high-speed link. It provides a high frequency clock for the system by multiplying the low frequency reference clock. The channel provides the physical connection between the transmitter and the receiver. It can be an optical fiber, a coaxial cable, a twisted pair UTP (Unshielded Twisted Pair), a PCB (Printed

Circuit Board) or a backplane. At high bit rate the channel attenuates and filters the signals by introducing noise and high inter-symbol interference. On the receiver side, in order to properly recognize the transmitted bits, a clock synchronized with the data is needed. It is convenient to generate a clock inside the receiver rather than transmit it from transmitter's PLL on a separate channel. The circuit that realizes this function is known as CDR (Clock Data Recovery). A CDR circuit incorporates a PLL and some additional circuits needed to synchronize the receiver with the incoming data stream. These timing blocks are crucial parts in a high-speed system because they provide correct spacing of transmitted data symbols and, on the receiver side, they have to sample the received signal waveforms.

2.1 Channel environment

Initially, gigabit SERDES was used in telecommunications industry and to a few niche markets such as broadcast video. Today, this kind of applications appears in every section of the electronics industry, military, medical, networking, video, communications, etc. They are also being used on printed circuit board (PCB) assemblies through backplanes and between chassis. For example there are several industry standards that use multi-gigabit transmission on different channel: Fiber Channel (FC), PCI Express, Serial-ATA, 10 GbE, etc [10].

Naturally the characteristics of the channel strongly depend on the application. However, we can divide the transmission channels into three categories. The first one includes connections for chip-to-chip accommodated on the same PCB. This type of channels is short and well controlled. The second one includes board-to-board connections, such as backplane channel, that are used, for example, to connect router on the same rack system. The last one includes channels for fast connection between computers with Ethernet and coaxial cables. In this work we use a backplane link as our design target, although the following consideration may be adapted in general to any type of wire line channel to the exclusion of the optical fibers.

The backplane shown in figure 2.2 is used to connect different line cards. Such backplanes can usually be found in large Internet routers inside data center. The high-speed serializer/deserializer chips are on the line cards and use the backplane traces as a transmission medium.

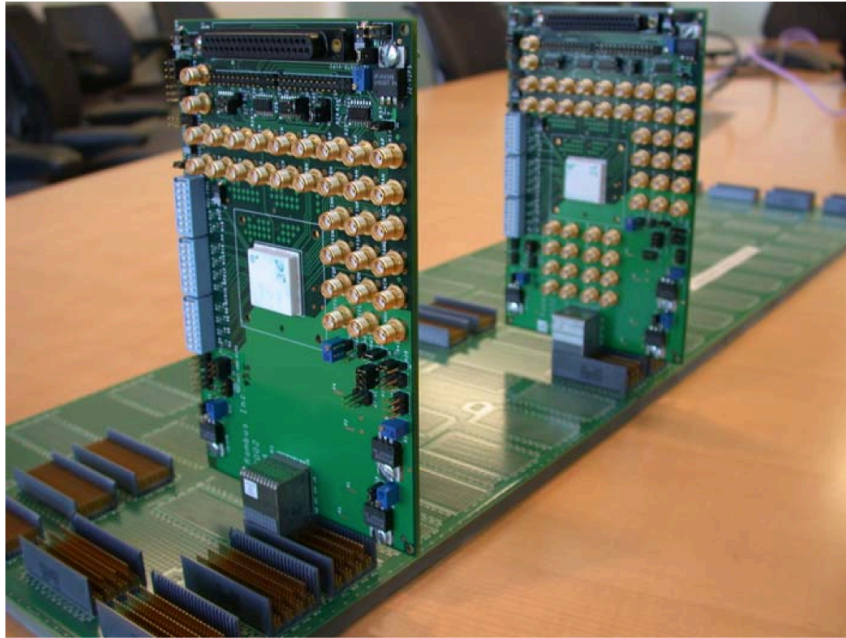


Figure 2.2: Backplane channel

Usually, the chips are mounted in packages that are soldered on the line card. The line cards communicate between them using dense through-hole connectors. The cross-section of the system shown in 2.3 is useful to see the full signaling path. The backplane is made of a dielectric material, usually flame-resistant-4 (FR-4), while the conductive traces are usually made of copper. The three primary factors that limit data transmission through backplane channels are reflections, crosstalk and loss. These three types of losses are briefly described below.

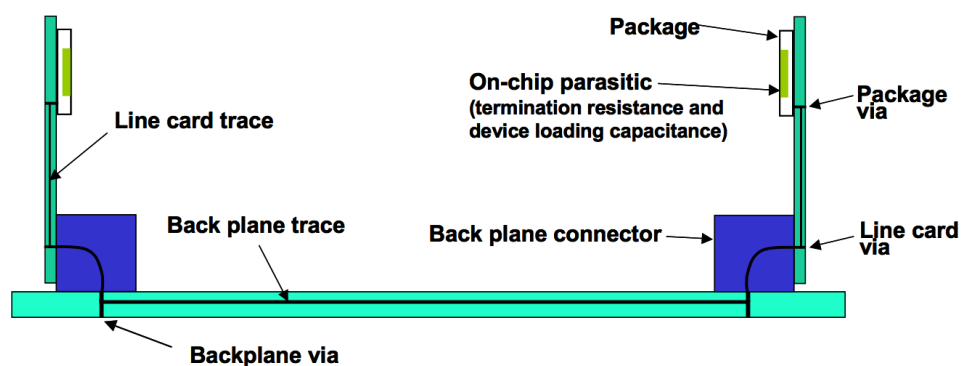


Figure 2.3: Cross-section of the system

Some attenuation effects come from the short traces (e.g. vias, or connector traces) that connect the components of the system together. For example the traces from the package into the line card and the traces to connect the line card to

the backplane. These short traces can create large impedance mismatches and cause reflections that can significantly degrade the quality of the signal. Using better impedance-controlled connectors, packages and vias minimizes loss due to reflections [11].

A second source of losses comes from undesired capacitive, inductive or conductive coupling between different signal paths; it is the crosstalk. A way to minimize the crosstalk is using better shielding for connectors, traces and vias. However, it worthy of note that, modern techniques try to compensate the crosstalk effects with active circuits but they are usually very complex.

The last source of loss is the frequency loss. The signal has to pass through a number of different traces in order to arrive from source to destination. The result is that, along the backplane, the attenuation raises with frequency due to skin-effect (conductor loss) and dielectric loss. At multi-GHz frequencies, dielectric loss dominates conductor loss.

We can now model the last source of loss to better understand the channel behavior. The distributed element model of the transmission lines is used to represent the channel through an infinite cascade of RLGC sections as shown in figure 2.4 where R, L, G and C respectively have the dimensions of Ω/m , H/m, S/m and F/m. In an accurate model of the transmission line the RLGC sections are much smaller than the wavelength of interest. The resistance R and the conductance G represent the loss, lowering the channel bandwidth. Respectively, they take into account the losses due to the conductor and dielectric that insulates and supports the connection. The inductance L and the capacitance C instead, model the inductive and capacitive behavior of the channel.

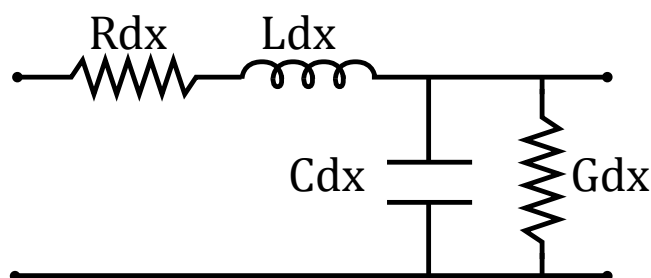


Figure 2.4: Distributed element model for the channel

The characteristic impedance of the line is a complex quantity and is expressed by the following relation [12]:

$$Z_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}}. \quad (2.1)$$

First of all, let's to consider an ideal and matched transmission line, where the characteristic impedance Z_0 is equal to the load resistance. We neglect the loss elements, $R=G=0$. The inductive and capacitive behavior introduces a phase shift in the propagation of the signal. As result an ideal channel has an infinite bandwidth and introduces only a delay τ_d :

$$\tau_d = \sqrt{LC} l \quad (2.2)$$

where l is the channel length. Now, always in matching condition, consider a lossy line where the dominant losses are due to the conductor and dielectric non-idealities. Conductor loss consists of DC loss, surface roughness loss and skin-effect loss. Since the dielectric material is not a perfect insulator, there is a loss at DC associated with current flowing through the dielectric between the signal conductor and the ground plane. Surface-roughness loss is due to the surface roughness of copper and increases with frequency. The skin effect is the physical phenomenon of the electric current to be distributed unevenly, so that the current density at the surface of the conductor is greater than at its core. Therefore, the current tends to flow at the "skin" of the conductor as shown in figure 2.5. The skin effect causes the effective resistance of the conductor to increase with the square root of signal frequency.

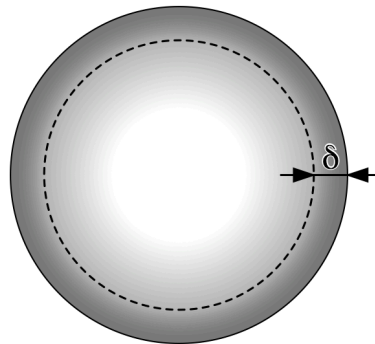


Figure 2.5: Current density in skin effect

The attenuation due to skin effect increases with increasing effective resistance. The attenuation α is given as:

$$\alpha = \frac{R}{Z_0 W}, \quad (2.3)$$

where W is the width of the conductor. The effective resistance R is given as:

$$R = \frac{\rho}{\delta}, \quad (2.4)$$

where δ is the skin depth (figure 2.5) that is given as:

$$\delta = \sqrt{\frac{\rho}{\pi \mu f}} \quad (2.5)$$

with the absolute magnetic permeability of the conductor μ and the frequency f . Therefore, at high frequencies the thickness δ is reduced, decreasing the effective cross section of the conductor. It is important to notice that the losses due to the skin effect are proportional to \sqrt{f} and typically dominate at low frequency. This is a big deal for most of the equalizer that are designed to recover the Nyquist loss (high frequency), failing to equalize at low frequency. Usually, recent works have used dedicated equalizer to recover the conductor low frequency losses [13].

Dielectric losses are due to an imperfect insulation and come at higher frequencies respect to skin effect. There are small currents into the dielectric that flow and disperse generating heat. This loss is linearly proportional to the frequency and is quantified by the loss tangent $\tan(\delta)$. The lower is tangent loss and the lower are losses. Table 2.1 shows some typical dielectric materials and the respective $\tan(\delta)$.

The expression of channel attenuation is therefore function of the frequency and length l and it is given as:

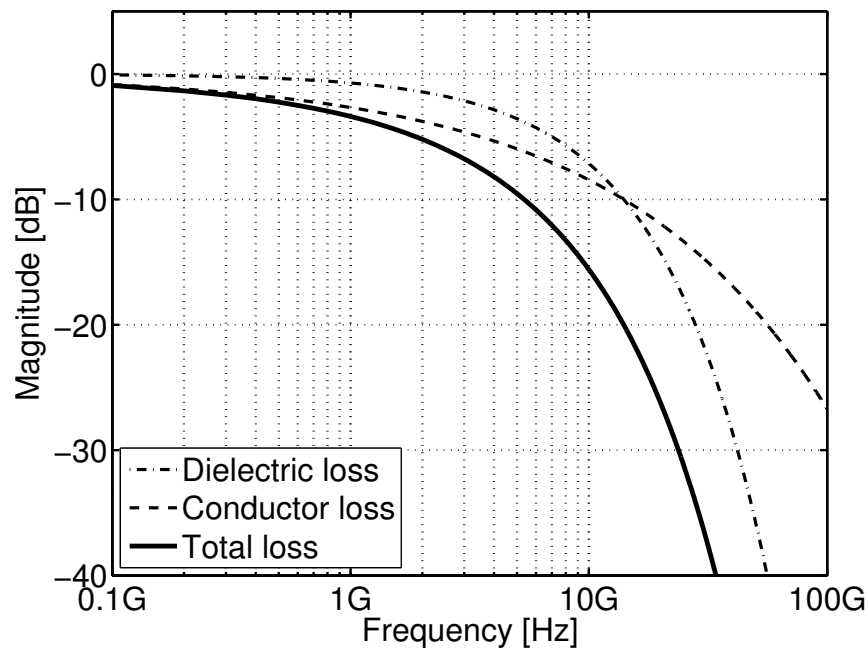
$$Loss(f) = \exp[-k_s l (1 + j) \sqrt{f} - k_d l f] \quad (2.6)$$

where k_s and k_d are coefficients related to the skin and dielectric loss respectively. Summarizing, we have seen that at low frequencies, the substrate conductance has a negligible loss compared with the skin effect. As frequency increases, the dielectric

Table 2.1: Dielectric materials

Materials	$\tan(\delta)$
FR4	0.035
Poliimide	0.025
GETEK	0.010
Teflon	0.001

loss becomes significant, leading to a more rapid drop in the magnitude. A typical channel transfer function is depicted in figure 2.6, where the loss is separated in the two contributions.

**Figure 2.6:** Crossover between skin and dielectric loss

So far we have seen the effects of skin and dielectric losses only in the frequency domain. Obviously, there will be a difference also in the time domain. The impulse responses can be obtained by taking the inverse Fourier Transform of $Loss(f)$. Skin and dielectric loss affect the overall time response, therefore is useful separate the two contributions as shown in figure 2.7. In the figure the skin impulse response on the top and dielectric impulse response in the bottom. The y-axis shows the

normalized amplitude and the x-axis shows time divided by T_b . The axes are chosen in this way to clearly show the shape of the pulses and the different time span. It can be seen that the skin impulse response is asymmetrical over time with a very long tail. Usually, for this reason, a classical skin time response is dominated by post-cursor while the pre-cursors are negligible. This long tail is directly related to the low frequency loss typical of conductor losses. In the dielectric response, instead, the pre-cursors are typically identical to the post-cursors because the time response is symmetrical. Notice that the time span, respect to previous case, is more limited.

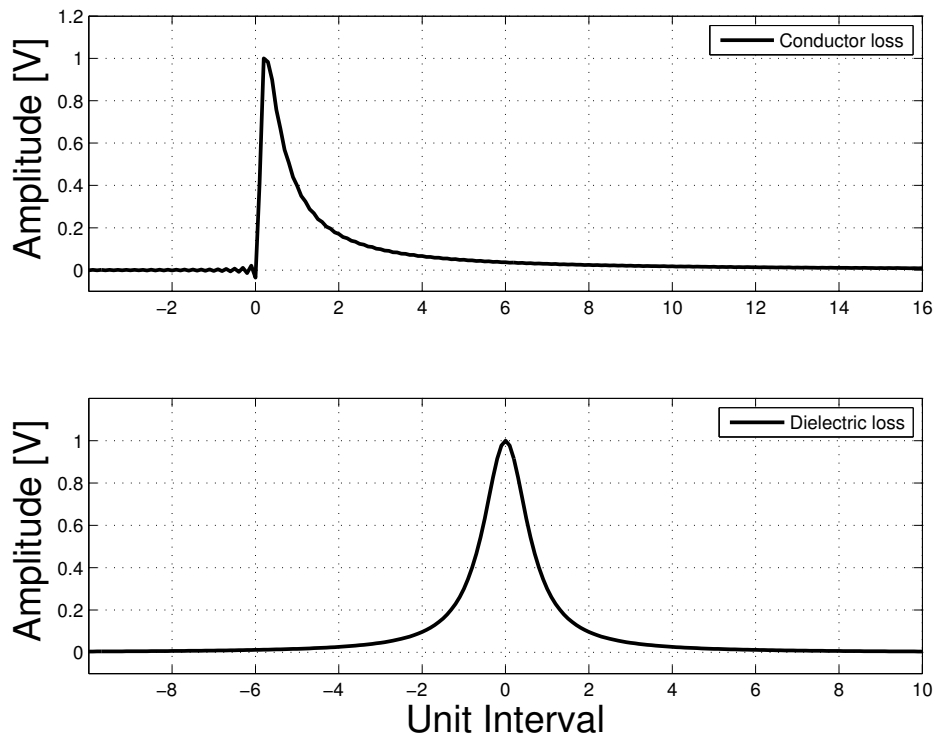


Figure 2.7: Skin and dielectric impulse responses

To conclude, by definition, a transmission line has a set and constant impedance. Actually, the impedance is not constant. The problem comes when the signals change layers, encounter pads for a component or go through a connector or cable. Every change in the channel impedance is a potential problem when operating in the multi-gigabit range. For these reasons, the accurate modeling of the complete channel (transmission line, vias, connector, etc.) is very difficult. The model of the channel is usually developed by extracting the frequency response by measures on the channel. In this way everything from transmitter and receiver can be included in the model [14].

2.2 Equalizer categories

We have seen so far that the channel losses introduce distortion on the transmitted signal. In the time domain a single pulse is spread over several unit intervals. Ideally, the purpose of the equalizer is to compensate the channel effects and it can be expressed by:

$$H_{EQ}(s) = H_C^{-1}(s) \quad (2.7)$$

where H_{EQ} is the equalizer transfer function and H_C is the channel transfer function. The figure 2.8 explains the idea.

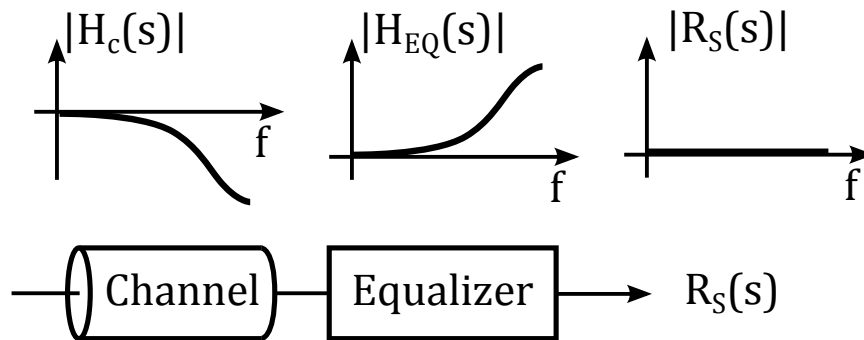


Figure 2.8: Conceptual idea for ideal equalizer

An ideal equalizer transfer function is the inverse of channel transfer function, i.e., a high pass transfer function. The H_{EQ} function is important both from the point of view of the magnitude and from the point of view of the phase because, after the equalizer, the magnitude of $R(s)$ has to be maximally flat, while the phase should introduce a constant group delay in frequency.

Depending on the application and on the maximum bit rate, there are different types of equalizer. They can be organized in several categories. The first distinction is made on the type of input signal, so we will find equalizers that work with only analog signals (Analog Signal Equalizer - ASE) and Mixed Signal Equalizers (MSE) [10]. The figure 2.9 depicts the subsequent categories in a tree diagram.

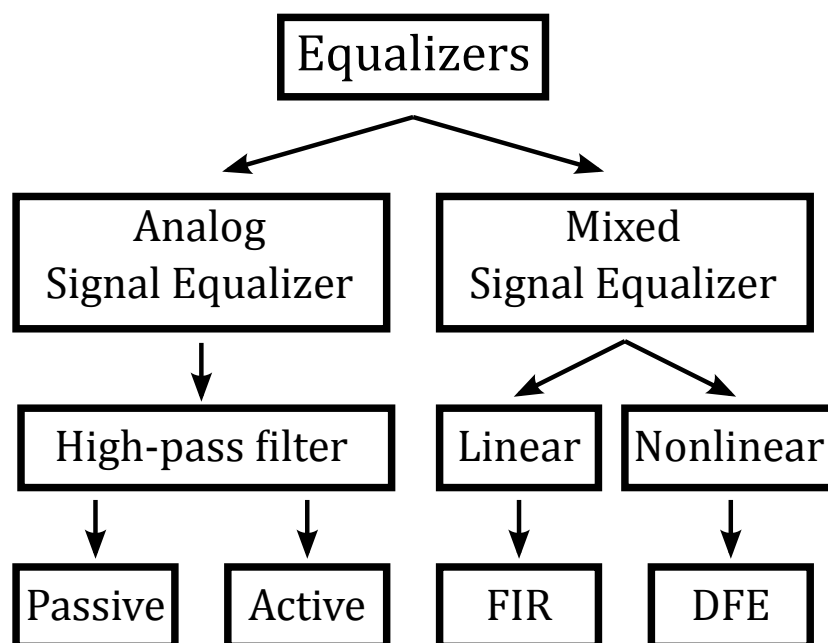


Figure 2.9: Categories of equalizers

The main advantage of the ASEs is that the clock signal is not required; therefore, the clock data recovery circuit is not necessary. Usually, they are used to equalize channels with a regular frequency transfer function, otherwise they could fail to correct any notch attenuation due for example to stub or strong reflections. In addition, this kind of equalizers have a poor adaptability, that require a custom design for each type of channel. In practice, ASEs are passive high-pass filters or active high-pass filter. The former offer excellent linearity, but the gain boost at Nyquist frequency is usually limited. The problem can be overcome through a cascade of active HP filters, but this solution requires more power consumption.

MSEs circuits are able to process both digital and analog signals. MSEs are divided into two categories: linear and nonlinear. As we shall see in the next section, their main advantage is the possibility to implement digital algorithms to adapt the equalizer to different channels. The most popular nonlinear equalizer is the Decision Feedback Equalizer (DFE). The DFE is always used after a previous equalization and it is able to remove only the post-cursors. Linear MSEs, instead, are frequently implemented through Finite Impulse Response filters (FIR).

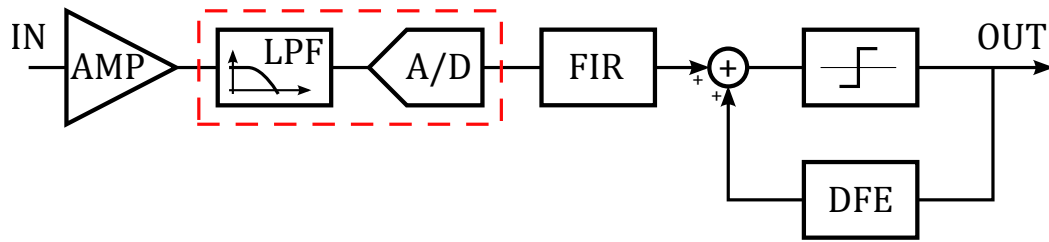


Figure 2.10: Discrete time MSE block diagram

Until about ten years ago, most of the MSEs were realized in the discrete time domain. A simple block diagram is shown in figure 2.10, where you can see an ADC after an anti-aliasing filter. However, with the increase of the bit rate transmission, an analog/digital converter in the chain has become a disadvantage. The conversion from analog to digital world requires a significant power consumption that is larger respect to equalizer power consumption.

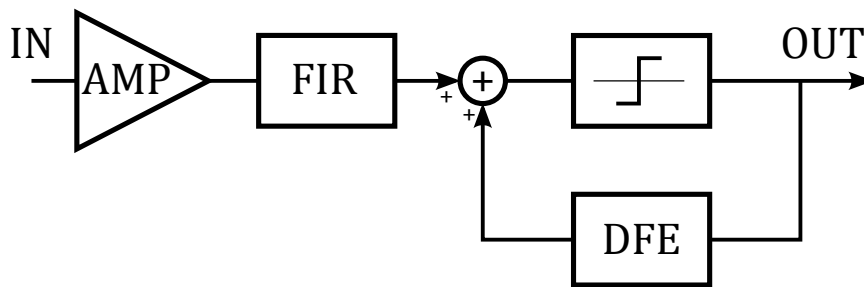


Figure 2.11: Continuous time MSE block diagram

To overcome this problem the research has pushed towards the use of hybrid techniques as shown in figure 2.11 where the A/D converter has been removed. Today, for this kind of applications, the FIR filters work in the analog domain, while the adaptation logic remains digital and works at lower speed respect to the bit rate.

Nowadays, for these reasons, FIR equalizer may be included in the ASE category and in particular in the active branch. Here, in the active class, there are many types of equalizers, but we want focus the attention on the above-mentioned FIR and on other classical solutions. Usually, they are grouped under Linear Feed Forward Equalizers (LFE, also known in the literature as FFE) category.

2.2.1 Continuous Time Linear Equalizer (CTLE)

Simple Feed Forward equalizers are implemented with a Continuous Time Linear Equalizers (CTLE) that are usually realized with a cascade of capacitive degenerated differential pair or with more complicate structures that using a split-path approach [15] - [16] - [17] . In this case the signal is fed into two parallel path: the first a unity-gain path and the second a high-frequency boost path, which are then summed to create the output. The main advantages are the simplicity, the low power consumption and less silicon area but, however, they have a limited flexibility.

The CTLE provides gain peaking in order to boost up high-frequencies to counter the channel attenuation and distortion. The peaking gain and the peaking frequency of a continuous time linear equalizer are key design parameters to improve link performance because they play a major role in shaping the CTLE transfer function.

The concept of CTLE can be explained in the frequency domain. The link channel can be viewed as a low-pass filter. To compensate for the low-pass characteristics of the channel, a high-pass filter is added at the receiver to achieve balance between the high-frequency and low-frequency components of the data stream. A typical channel transfer function, without frequency notches, has a low-pass characteristic that can be approximated by one or few poles as:

$$H_{CH}(s) = \frac{1}{s + p_{CH}} \quad (2.8)$$

where p_{CH} is the dominant pole of the channel. The CTLE has a transfer function that can be described as:

$$H_{EQ}(s) = \frac{(s + z_1)}{(s + p_1)(s + p_2)} \quad (2.9)$$

where z_i and p_i are the zero and the poles respectively. If the zero of the CTLE is at the right frequency, it cancels the dominant pole of the channel. The equalized transfer function becomes flat over a wider frequency range, and it is described by:

$$R(s) = \frac{s + p_1}{s + p_2}. \quad (2.10)$$

The zero is closely related to the dominant pole of the channel to be equalized and its location needs to be selected with care when optimizing CTLE parameters. Therefore, it is important to ensure that the peak of the CTLE is at the correct frequency and that gain boost is correct. Incorrect CTLE selection results in under-equalization or over-equalization, and thus, in suboptimal post-CTLE signal integrity.

In certain situations it is not easy to find the right zero frequency and to obtain a good equalization. To overcome the CTLE limits, recently Finite Impulse Response filters have been proposed even if they are more complex and with higher power dissipation.

2.2.2 Finite impulse response filter equalizer

FIR equalizers have different advantages respect to classical solution: they are able to remove the pre-cursor ISI and are compatible with simple adaptation algorithm. A Finite Impulse Response filter (FIR) is a causal Linear Time Invariant system with a finite impulse response. As you can see from figure 2.12, the block diagram doesn't have a feedback and therefore the system is unconditionally stable. It consists of n multipliers with a variable coefficient, $n-1$ delay cells and a summing node. A FIR can generate large types of transfer function thanks to the different values of his coefficients. Usually, since the signal input multipliers is taken along the delay line, the multipliers are called "taps".

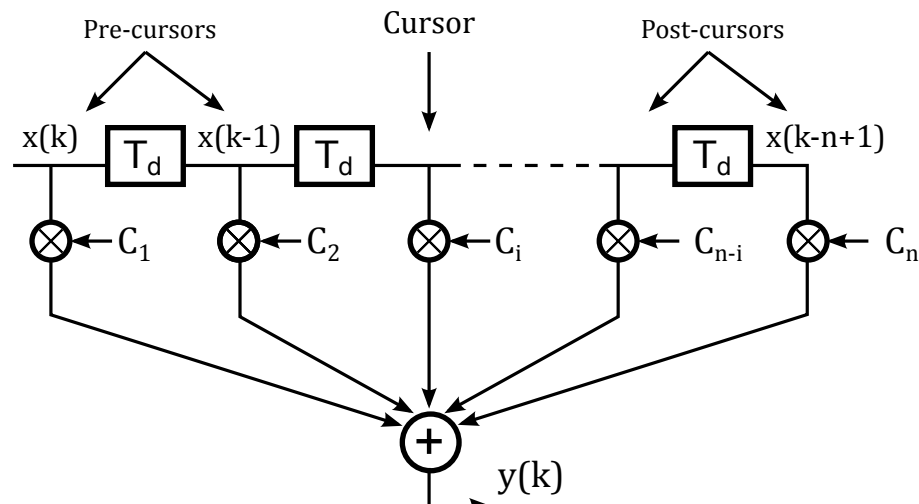


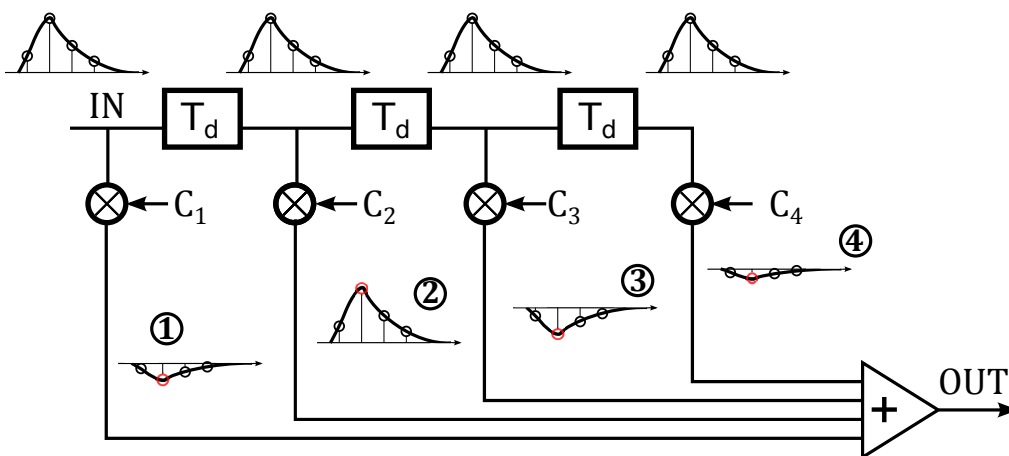
Figure 2.12: Finite Impulse Response filter block diagram

The delays are interposed between the taps and each of them provides a delay T_d . In the time domain the input-output relation of the FIR shown in figure 2.12 is given as:

$$y(kT) = \sum_{i=1}^n C_i x(k + 1 - i)T_d \quad (2.11)$$

The input signal $y(k)$ propagates along the delay chain. The delayed versions of the signal are multiplied by the tap coefficients and then summed together. The central tap is usually chosen as the main tap (cursor tap) because it has generally the highest coefficient. The input-output relation of the FIR shows that the performance, as well as from the coefficients, also depends on the delay T_d . The value of T_d can be equal to the bit period or lower. In the first case we have a Symbol Spaced Equalizer (SSE), otherwise we talking about Fractionally Spaced Equalizer (FSE).

To understand the FIR behavior is useful to take a look at the time domain operation [9]. Let's to consider a SSE structure with four taps. As shown in figure 2.13 every tap provides a delayed version of the input signal multiplied for its own coefficient. On the adder output, a destructive interference shapes the bit response removing the energy on each cursor with the exception of the main cursor. This is qualitatively shown in figure 2.13.



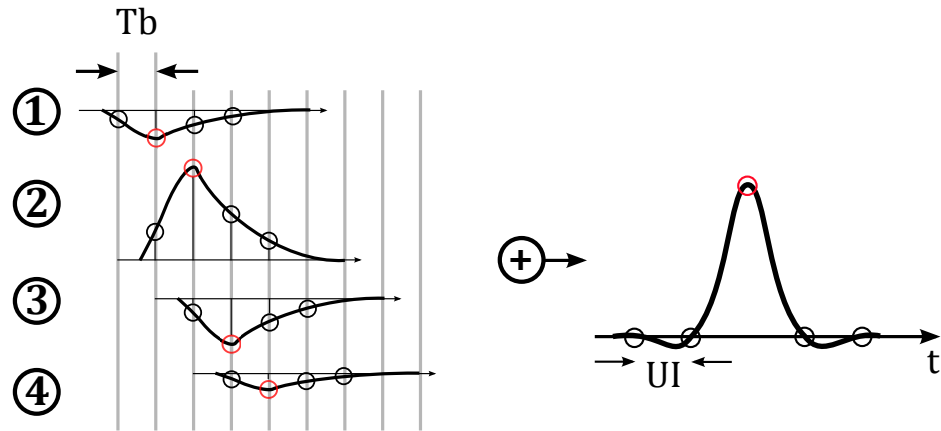


Figure 2.13: Time domain FIR block diagram with the relative time domain behavior

2.2.3 Decision Feedback Equalizer (DFE)

Feedforward Equalizers, therefore FIR and CTLE, are very simple to implement, but they generally achieve sub-optimal performance. For channels that introduce from weak to moderate ISI, their performances are often sufficient. However, they enhance the noise while suppressing ISI because they cannot distinguish between signal and noise or crosstalk. So eventually, as the channel distortion becomes severe, the performance of a linear equalizer can be limited by the noise enhancement. Figure 2.14 explains the concept.

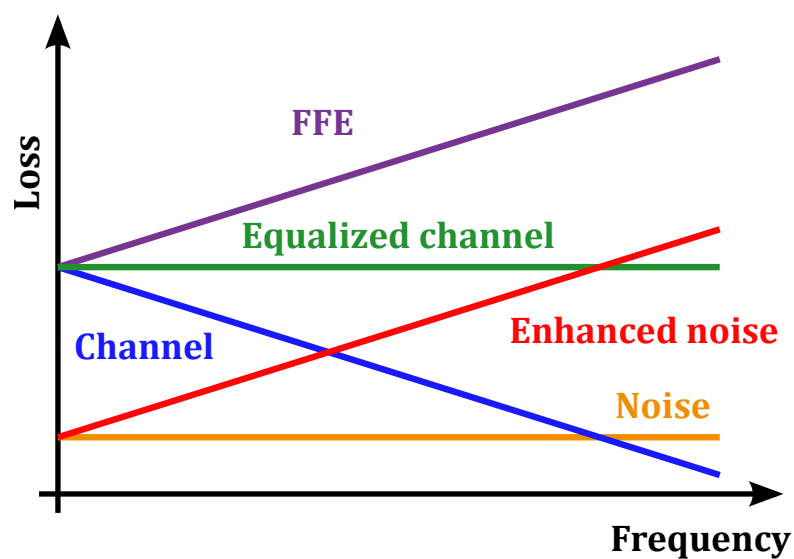


Figure 2.14: Noise limitation of Linear Feedforward Equalizer

For these reasons, another very important equalizer for today's receiver architectures is the DFE (nonlinear category). The DFE improves the performance of a linear equalizer without enhancing the noise. The DFE is a non-linear structure, where, as shown in figure 2.15, a feedback FIR filter and a decision device (slicer) are used. Assuming correct decisions, the previous ISI can then be subtracted from the current symbol by feeding back the previously decided symbols through the feedback. Since the FFE suppresses the contribution of the pre-cursor ISI, if FFE noise is not severe and there aren't decisions errors by the DFE slicer, the ISI can be eliminated without enhancing the noise.

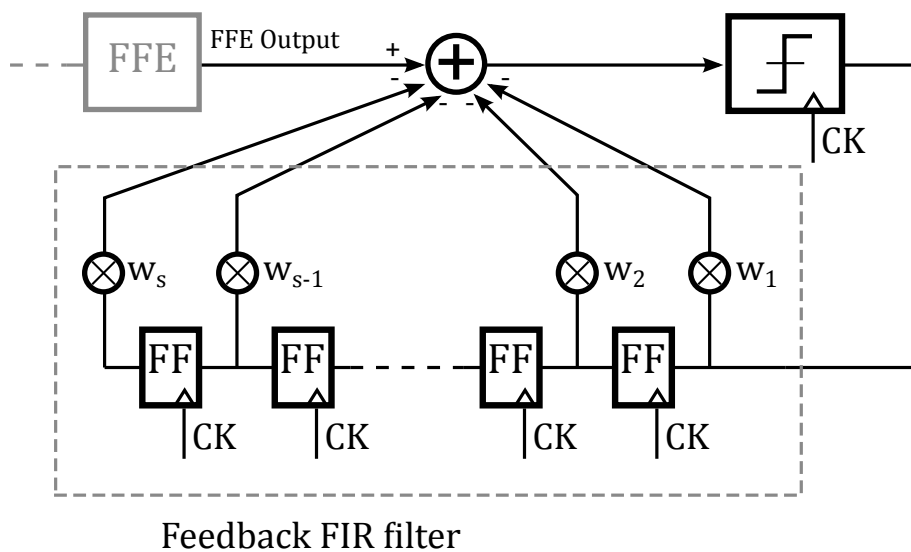


Figure 2.15: Decision Feedback Equalizer block diagram

2.3 Iterative adaptation on FIR

During the data transmission, or in more long time period, channels may exhibit wide variation in frequency-dependent loss. For that, equalizer design requires to contain a certain flexibility to set the equalizer coefficients adaptively to minimize the ISI. Such an equalizer is called an adaptive equalizer. For our purposes it is sufficient to investigate the aspects of iterative adaptation on FIR equalizers.

As you can see from figure 2.16, to recover the original signal, you have to find an equalization filter, which will minimize the error J_e between original transmitted signal and equalized signal after the FIR filter. J_e is also called “cost function” and it is a function of filter coefficients.

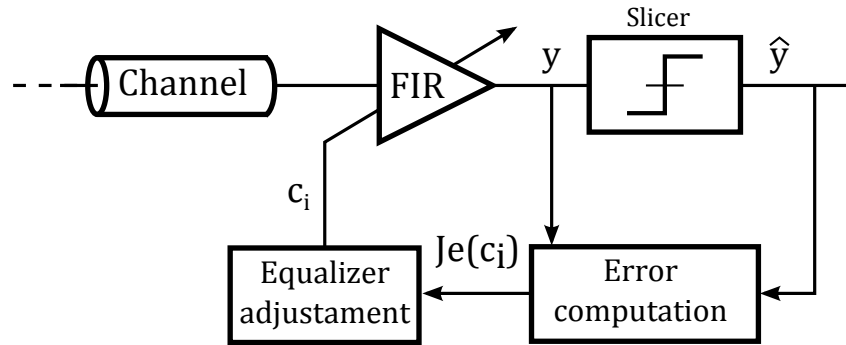


Figure 2.16: Iterative adaptation on FIR

The general approach to updating the filter tap coefficients is:

$$Coef_{new} = Coef_{old} + (stepsize)(error\ function)(input\ function) \quad (2.12)$$

where the error function J_e is typically based on the difference between the actual equalized signal y and the desired equalizer output \hat{y} . The input function is obtained from the input signal and step size is a design parameter. Designers have many options for implementing adaptive equalizers, the range of which is outside our purposes.

A widespread adaptive algorithm is the Least Mean Square (LMS) that is a particular case of a more general algorithm, the Minimum Mean Square Error (MMSE). MMSE algorithm attempts to minimize the Mean Square Error of the equalizer output at all times. However, it is not used in adaptive equalizers because it is difficult to implement. A more easy implementation is provided by LMS algorithm [10]. It operates in discrete-time domain with the sampling frequency equal to the bit time. The figure 2.17 explains the classical implementation of LMS algorithm on the equalizer. We can recognize a classical structure of a FIR filter. To describe how the algorithm works, we define the vector of filter coefficients as:

$$\bar{C}^T = [c_0, c_1, \dots, c_{n-1}]. \quad (2.13)$$

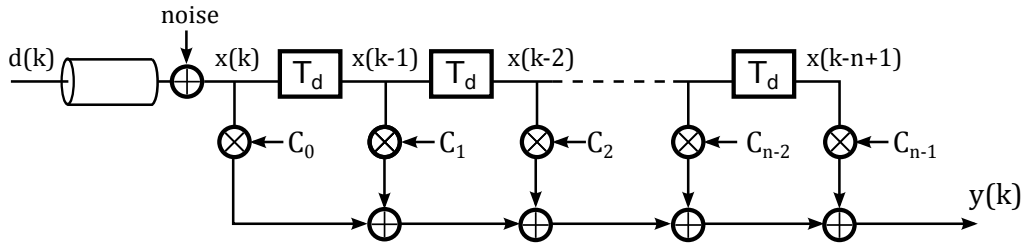


Figure 2.17: Implementation of LMS algorithm

Now, we define also the vector $x(k)$ which is the vector of input signal samples:

$$\bar{x}(k) = [x(k), x(k-1), \dots, x(k-n+1)]. \quad (2.14)$$

We have all the ingredients to write the FIR output equation $y(k)$ as the multiplication of the two vectors just described:

$$y(k) = \sum_{i=0}^{n-1} c_i x(k-i) = \bar{C}^T \cdot \bar{x}(k) \quad (2.15)$$

and finally the cost function J_e is given as:

$$J_e(k) = E[e(k)^2] = E[(d(k) - y(k))^2] \quad (2.16)$$

where the $E[\cdot]$ operator is the expected value of the argument. Minimization of the MSE minimizes the combined effect of ISI and noise. The goal of the adaptation is illustrated in figure 2.18. The figure shows the difference of the equalized output $y(k)$ from the training data $d(k)$ as a function of equalizer coefficients and it is plotted as a 3-D surface; the figure is an example with two coefficients. The surface is a convex function, so that it has a global minimum. The goal of the adaptive algorithm is to converge on a set of coefficient values that minimize the error in a small number of iterations. So, for the adaptation of the FIR filter, we have to move the FIR coefficients in opposite direction to the gradient of the cost function J_e .

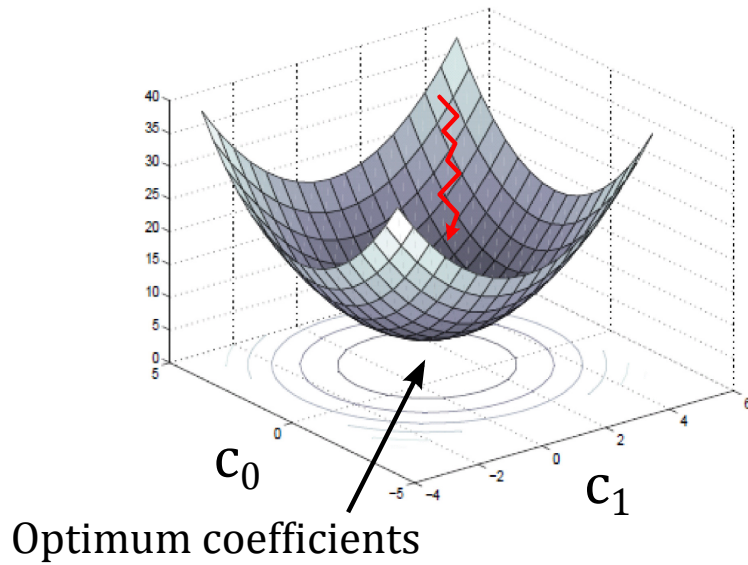


Figure 2.18: Example of error 3-D surface as function of two coefficients

Given cost function J_e with an absolute minimum, to minimize the error we have to update iteratively the coefficients in a direction opposite to the gradient of J_e :

$$\bar{\nabla}_C^T J_e(k) = \left[\left. \frac{\partial J}{\partial c_0} \right|_k, \left. \frac{\partial J}{\partial c_1} \right|_k, \dots, \left. \frac{\partial J}{\partial c_{n-1}} \right|_k \right]. \quad (2.17)$$

In this way we can write that the expression of the next coefficient, therefore, FIR coefficients are update as follow:

$$\bar{c}(k+1) = \bar{c}(k) + \mu \bar{\nabla}_C J_e \quad (2.18)$$

where μ is called step-size. It serves to act on the convergence speed, but it must be chosen carefully. If too fast, the algorithm will have too much energy and will not be accurate in finding the absolute minimum error. On the contrary, if too small, the convergence will be too slow and the algorithm will ensure the convergence for very long times.

Chapter 3

A 25-Gb/s FIR Equalizer Based on Highly Linear All-Pass Delay-Line Stages in 28-nm LP CMOS

Abstract

A continuous-time 4-tap FIR equalizer designed for loss compensation in backplane links is presented in this chapter. FIR filters are attractive to enhance the equalization performances of high-speed wire line receivers, providing high flexibility to match the channel frequency response and compatibility with simple adaptation techniques. Particular care is taken to address critical issues of continuous-time realizations, such as noise, linearity and dynamic range. To keep high SNR and not compromise equalization performances, a new all-pass stage is proposed to realize a delay line accommodating large input signal amplitude. Filter tap coefficients are realized with programmable transconductors and output currents are summed through a resistive load. Extensive experimental results, carried out on test chips realized in 28 nm LP CMOS technology, are presented. The equalizer demonstrates successful operation at 25Gbps data-rate while draws 25mA from 1V supply. Measurements with $900mV_{pk-pk}$ input signal prove equalization of a 20 – dB loss channel with 50% horizontal eye opening at $BER = 10^{-12}$. Experimental results compare favorably against state of the art [18]

3.1 Introduction

3.1.1 Motivation

To address the continuous demands for pushing speed of serial links, new standards are emerging with a data-rate of 25-28 Gb/s. Inter-symbol interference (ISI) is a severe obstacle and transceivers need to incorporate increasingly sophisticated channel equalization techniques. RX equalization combines a feed-forward linear equalizer and decision feedback equalization. Simple FFEs, realized with the cascade of peaking amplifiers, are difficult to be optimally adapted and feature limited capability to correct the pre-cursor ISI [19] (see the previous chapter). More sophisticated FFEs, implemented as transversal FIR filters, have been recently proposed in the RX [20] - [21]. They provide high flexibility in shaping the transfer function to match the channel frequency response and are compatible with simple adaptation schemes, such as the least-mean-square (LMS) algorithm. A key challenge of RX FIR equalizers is the design of a compact, wideband analog delay line withstanding a sufficiently large input signal. Gain compression of the delay line impairs the signal integrity and compromises the equalization capability, while maximizing the input signal swing is desirable to achieve high SNR. Several high-speed FIR equalizers have been proposed [21-23]. Delay lines based on lumped LC networks need high quality passive components, requiring a prohibitively large area. Active and hybrid (i.e. combining active and passive components) delay lines, occupy acceptable area but feature limited linearity, especially at low supply voltage, constraining the maximum allowed signal swing to a few hundred mV. In this work a compact active delay line featuring outstanding linear operation range is proposed to implement the fractionally spaced 4-tap FIR equalizer.

3.1.2 Proposed RX FIR equalizer

The block diagram of the proposed FIR equalizer is shown in figure 3.1. The active delay line is realized with the cascade of three stages providing a tap-to-tap delay T_d . The tap amplifiers are transconductors with a 6-bits programmable gain that can be chosen with the digital word D_i . Furthermore, a shared resistive load R_L is used to sum the tap currents, while the VGA and buffer follow the equalizer for measurement purposes. The VGA and buffer are realized with the cascade of three

degenerated differential pairs with resistive loads and shunt peaking inductors. The overall gain is controlled in the range 0-10 dB by programming degeneration and load resistors.

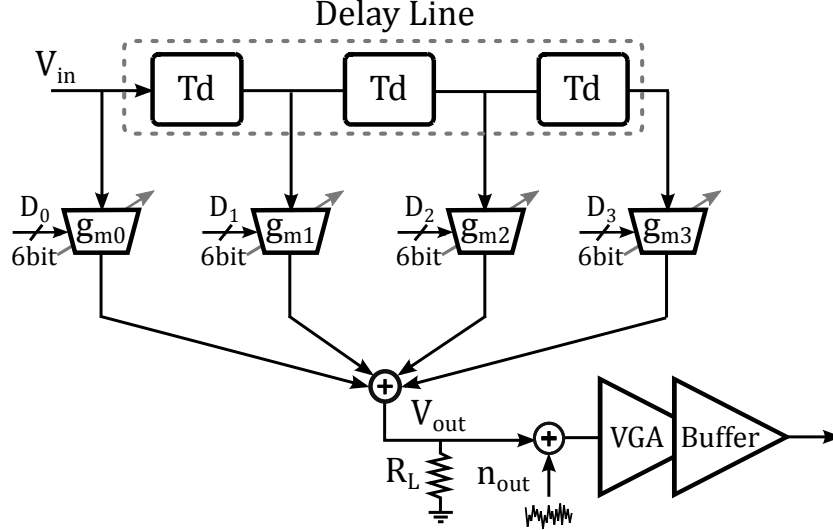


Figure 3.1: Block diagram of the proposed FIR equalizer

The transfer function $H(f) = V_{out}/V_{in}$, is given by:

$$H(f) = \sum_{i=0 \dots 3} c_i e^{-j(2\pi f \cdot i \cdot T_d)} \quad (3.1)$$

where $c_i = g_{m_i} \cdot R_L$ ($i = 0 \dots 3$) are the filter coefficients. Obviously the magnitude coefficient is limited and the trade off is between transconductance g_m and resistive load R_L because these two parameters are constrained by the current consumption and bandwidth of the output node respectively. In particular, the coefficients are bounded within ± 0.6 by the maximum transconductance of the tap amplifiers ($g_{m_i, max} = 7.5 mS$) and the value of the load resistance $R_L = 80 \Omega$. Together with the number of taps or equivalently the length of the delay line, is an important design specification and determines the equalization performances.

When considering fully digital implementations, it is well known that fractionally spaced FIR equalizers (i.e., filters with a T_d which is a fraction of the symbol time T_b) offer several advantages over symbol-spaced equalizers with $T_d = T_b$. Thanks to the higher sampling rate, they do not suffer from aliasing, may provide boost well beyond Nyquist frequency ($f_N = 1/2T_b$), are less sensitive to the clock sampling phase and allow simultaneous equalization and matched filtering [22] - [23]. On

the other hand, the shorter is the tap-to-tap delay, the larger is the number of taps to cover the same time window. When targeting an analog continuous-time realization, minimizing the number of taps is highly desirable to limit power dissipation. Moreover, differently from digital implementations, an important issue to be considered is the impact of the equalizer noise to the output signal SNR. In fact, an analog FIR equalizer impairs the SNR not only because of the enhancement of the high-frequency noise superimposed to the input signal, but also and most importantly because of the noise introduced by the equalizer building blocks, represented in figure 3.1 by the noise n_{out} added to the output signal.

To understand the FIR behavior and voltage swing requirements to keep high output SNR look the equation $H(f)$ that describes the transfer function V_{out}/V_{in} . At low frequency the gain G_{LF} is equal to the sum of filter coefficients. Substituting $f = 0$ in $H(f)$, we obtain:

$$G_{LF} = \sum_{i=0..3} c_i \quad (3.2)$$

i.e., the low frequency gain is equal to sum of the filter coefficients.

Substituting $f = f_N$ (Nyquist frequency) in $H(f)$ instead, we can find the expression that describe the high frequency gain and in particular the gain G_{HF} at Nyquist frequency:

$$G_{HF} = \sum_{i=0..3} (-1)^i c_i \quad (3.3)$$

The high frequency boost instead is the sum of filter coefficients but the sign changes with the power of i . Therefore, the low frequency gain G_{LF} and the high frequency gain G_{HF} trade each other. As results, due to the boundary of the magnitude coefficients, to provide high frequency equalization, the equalizer introduces a low frequency gain.

3.1.3 Linearity requirements

Ideally, the maximum channel loss that can be recovered by a linear equalizer is limited by the enhancement of the high frequency noise superimposed to the input

signal. However, in contrast to a digital implementation, the noise introduced by the analog equalizer itself (n_{out} in figure 3.1) may set a more stringent limitation. In this design $n_{out} \sim 1.5mV_{rms}$ is almost independent from the equalizer transfer function. To achieve $SNR \geq 30dB$, required to have good margin on the vertical and horizontal eye opening, the output voltage amplitude needs to be larger than $100mV_{pk-pk}$. Given the input swing, the amplitude of V_{out} is determined by the low frequency gain of the equalizer, i.e. by the sum of the filter coefficients. Maximum SNR is therefore achieved with all the coefficients having the same sign, but to provide high frequency peaking the filter needs positive and negative values. Because the coefficients are magnitude bounded, high frequency peaking is traded for amplitude of V_{out} and SNR. As an example the following set of coefficients $[c_0, \dots, c_3] = 0.6 \cdot [-0.35, 1, -0.16, -0.26]$ is required for equalization of a backplane channel with $\approx 20dB$ attenuation at Nyquist, determining a low frequency loss of $17dB$. As a consequence the equalizer needs $V_{in} \geq 700mV_{pk-pk}$ to have $V_{out} \geq 100mV_{pk-pk}$. This swing, easily provided by state of the art transmitters [24], mandates a wide linear operation range to the FIR filter in order to not compromise the equalization performance. Meeting this requirement at limited power consumption is a very challenging task, particularly at low supply voltage.

In conclusion, a high compression point for the equalizer is key for high SNR and for signal integrity. Moreover, the most critical building blocks in figure 3.1 are the delay line stages. For this reason a new all-pass stage amenable to high frequency operation and accommodating high input swing is introduced in the following section.

3.2 System simulation and circuit design

3.2.1 Impact of FIR filter compression

The impact of FIR filter compression on signal integrity can be estimated with computer simulations. The simulation setup is shown in the block diagram of figure 3.2, where the equalized output V_{out} is after the FIR filter. In this first case the DFE is not present. The maximum loss that can be recovered in this situation (keeping a “good” output SNR and with only a FIR equalizer) is around $20dB$. For

that, in the block diagram, you can find a weak lossy channel with an attenuation of only 14dB . A FIR filter equalizer is also present where the delay cell is modeled as a non-linear block with a certain 1dB compression point. Moreover, from the point of view of functionality and bandwidth either delay cells and tap amplifiers are ideal.

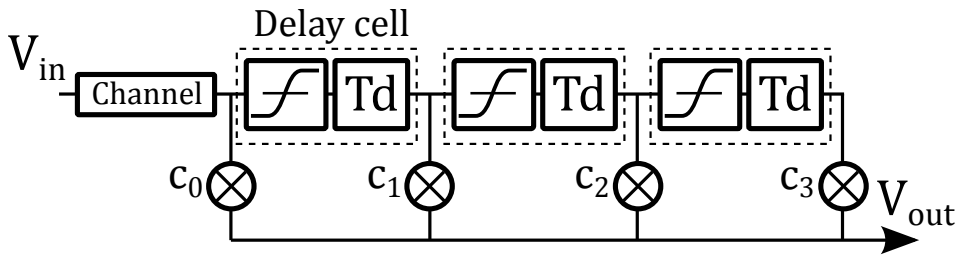


Figure 3.2: Simulation setup block diagram to evaluate the impact of FIR filter compression

The idea is to test the impact of equalizer distortion on signal integrity, when the amplitude of the input signal V_{in} grows. To compare the performances, we use the simulated eye openings. In the chart (figure 3.3), on the y-axis the horizontal and vertical eye opening are plotted while on the x-axis the input amplitude normalized to the 1dB compression point of the delay cell. When the input signal is far away from the compression point, the vertical and horizontal opening are good and eye diagram is clean as shown in figure 3.4a. However, when the input amplitude starts to reach the compression point, the eye opening worsens. The output eye diagram with the input amplitude greater than 1dB compression point is shown in figure 3.4b.

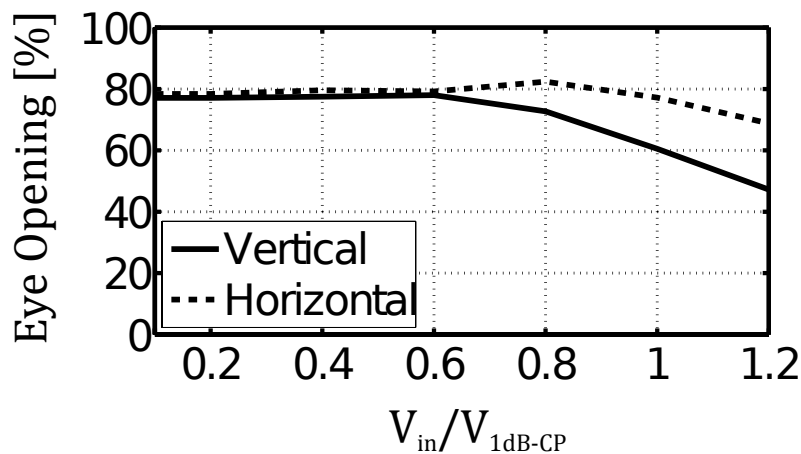


Figure 3.3: Eye opening vs input signal amplitude

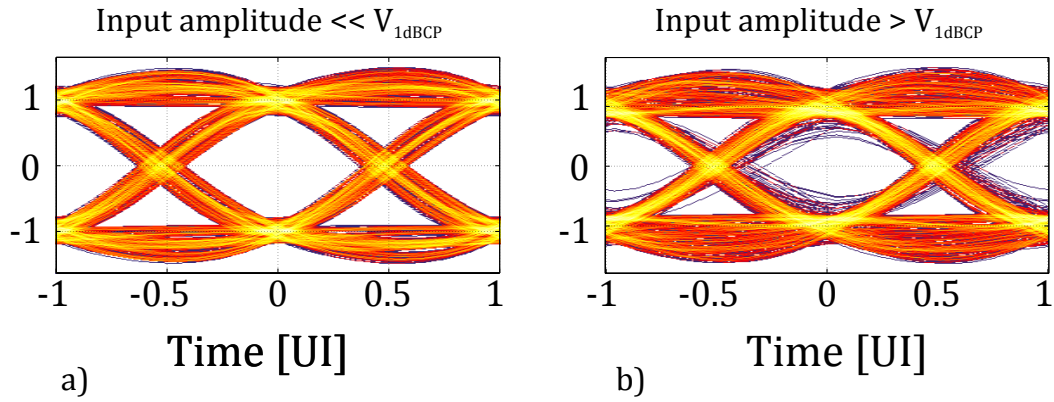


Figure 3.4: Simulated eye diagrams

The impact of FIR filter compression is even more important if we consider a DFE and a channel with higher loss. Therefore, for completeness, the case with the presence of the decision feedback equalizer is also evaluated. The simulation setup is shown in figure 3.5 and in this example the channel loss at Nyquist frequency is $34dB$.

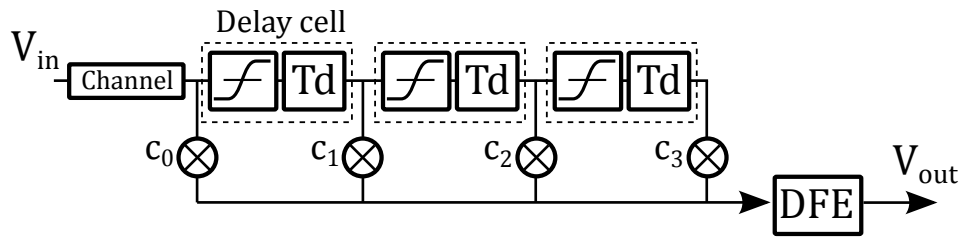


Figure 3.5: Simulation setup block diagram to evaluate the impact of FIR filter compression when a DFE is considered

As reported in figure 3.6, when the input amplitude is smaller than $1dB$ compression point, the vertical opening is very high. Unfortunately, when input amplitude grows, eye opening has a rapid roll off as you can see from the eye diagrams shown in figure 3.7.

In conclusion, these analyses prove that to keep high SNR without compromises equalization performances a highly linear FIR is required and it is even more important if a decision feedback equalizer is considered.

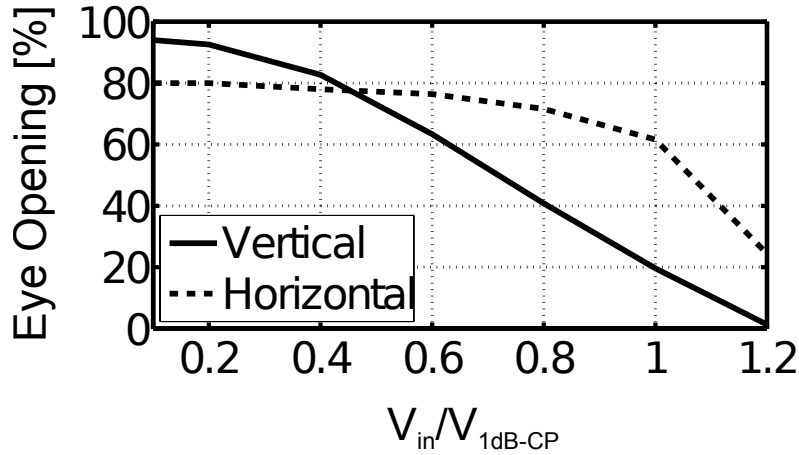


Figure 3.6: Eye opening vs input signal amplitude

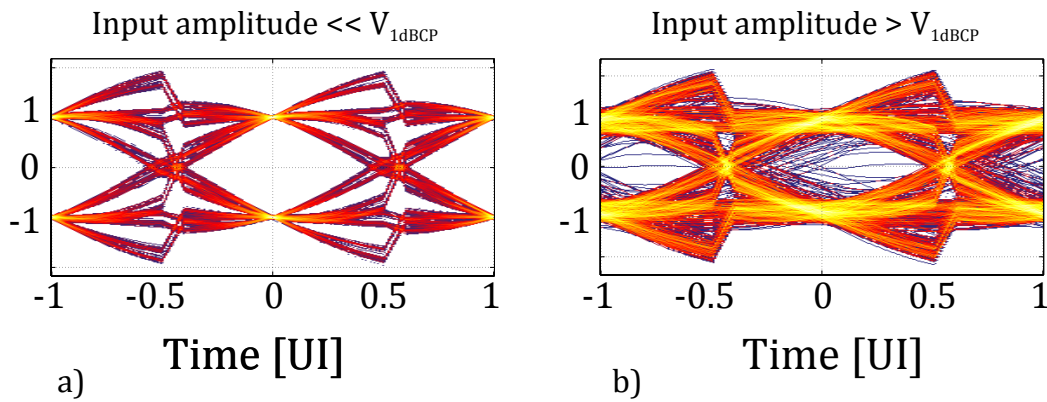


Figure 3.7: Simulated eye diagrams

3.2.2 Continuous time analog delay line

Continuous time FIR equalizers operating at more than $10Gb/s$ usually exploit lumped-elements delay lines realized with cascaded LC-ladder sections to simultaneously meet the requirements for high delay and wide bandwidth. However, as previously anticipated, LC sections suffer from narrow tuning range and require excessive silicon area due to the need for high Q inductors. To achieve a wide bandwidth delay cell and with a small size, continuous time active delay lines are potentially attractive. Several solutions have been proposed by cascading low-pass filter sections. Unfortunately, delay trades with bandwidth and to achieve a sufficiently wide bandwidth, reported FIR equalizers have a too limited number of taps and small tap-to-tap delay. Since the FIR compression is a big problem, a new delay stage is proposed to accommodating large input signal amplitude. In

particular, a CMOS all-pass stage has been investigated to implement the active delay line. Two possible solutions that provide a first-order all-pass response are shown with the block diagrams in figure 3.8.

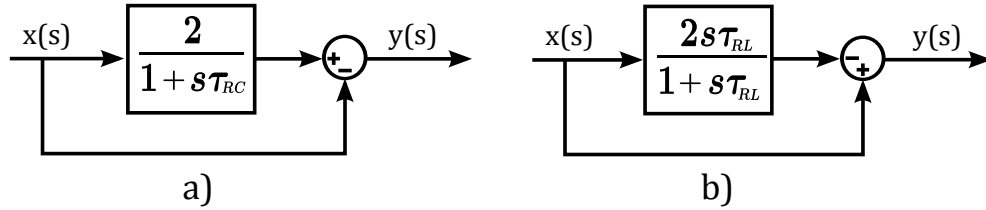


Figure 3.8: Different implementations all-pass filters: a) With a first-order low-pass filter b) With a first-order high-pass filter

The all-pass transfer function is achieved by subtracting the outputs of a first-order filter, featuring a gain of 2 and time constant τ , and a unity gain feed-forward path.

The solution in figure 3.8a has been recently exploited in a 25Gb/s FIR equalizer for multi-mode fibers. It is realized with a one-pole low-pass filter with time constant τ_{RC} , and a unity-gain path. The all-pass transfer function $H_{RC}(\omega)$ and the respective group delay $GD_{RC}(\omega)$ are:

$$H_{RC}(s) = \frac{y(s)}{x(s)} = \frac{1 - s\tau_{RC}}{1 + s\tau_{RC}} \quad (3.4)$$

$$GD_{RC}(\omega) = -\frac{\partial \angle H(\omega)}{\partial H(\omega)} = \frac{2\tau_{RC}}{1 + (\omega\tau)^2} \quad (3.5)$$

The alternative in figure 3.8, where the low pass filter is replaced by a high-pass pass-filter, is investigated in this work. For both cases, the transfer function is the same, an all-pass filter. However, when considering the transistor level realization, the last architecture provides a remarkable improvement in the maximum allowed signal swing.

3.2.3 Delay cell design

The transfer function of an all-pass filter of the first order can then be written as follows:

$$H_D(s) = \frac{1 - s\tau}{1 + s\tau} \quad (3.6)$$

where τ is the time constant relating both to the pole $p_1 = -1/\tau$ and to zero $z_1 = 1/\tau$. The magnitude and phase expressions are respectively:

$$|H_D(j\omega)| = 1 \quad (3.7)$$

$$\angle H_D(j\omega) = -2 \arctan(\omega\tau) \quad (3.8)$$

The magnitude is identical for every order of the all-pass filter, while the phase response change with the filter order. A consequence of the symmetry properties of the poles and zeros is that the phase response decreases monotonically. The group delay function of frequency can be expressed as the derivative of the phase as written below:

$$GD(\omega) = -\frac{\partial\phi(\omega)}{\partial\omega} \quad (3.9)$$

and it represents the delay between the input and the output signal. As already written above, it is given as:

$$GD(\omega) = -\frac{2\tau}{1 + (\omega\tau)^2} \quad (3.10)$$

As you can see from the equations, the ideal magnitude bandwidth is infinite, while GD_{RC} has a limited bandwidth and at low frequency is equal to $2\tau_{RC}$. As the frequency increases, the group delay starts to drop up to collapse for frequencies around the pole and the zero. This is an important aspect; ideally you would want a constant group delay at all frequencies. However, in our case, it is essential that the group delay is constant up to the Nyquist frequency.

The figure 3.9 shows a possible implementation of the all-pass function. The transconductor g_m is loaded by generic impedance. The output signal is obtained by subtracting the input signal from the filter output.

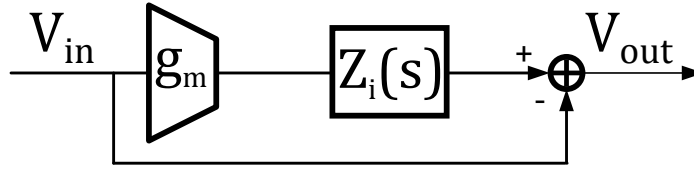


Figure 3.9: All-pass filter block diagram

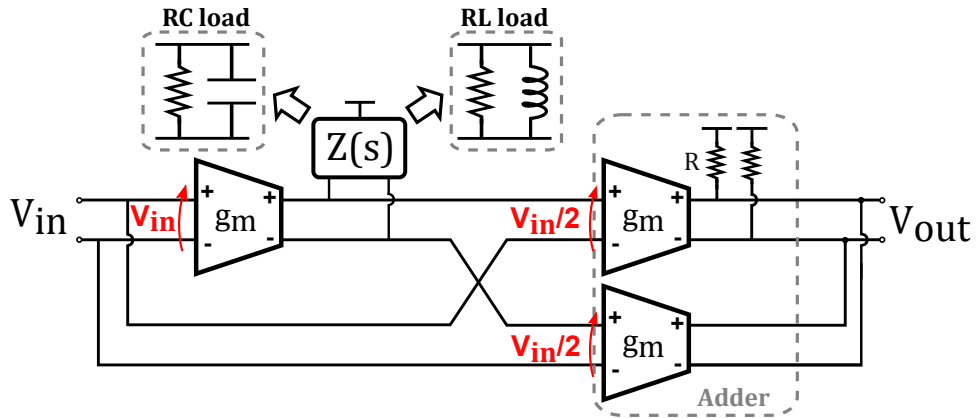


Figure 3.10: Schematic circuit of the all-pass transfer function

The circuit schematic is shown in figure 3.10. The transconductor g_{m1} and the $Z(s)$ load form the low or the high pass filter, while g_{m2} and g_{m3} are used to subtract the filter output from the input signal:

$$\frac{V_{out}}{V_{in}} = g_m Z_i(s) - 1 \quad (3.11)$$

We have seen before that the all-pass transfer function $H_D(s)$ can be obtained in two different ways based on the choice of the impedance $Z_i(s)$. If the impedance is an RC parallel, as shown in figure 3.11, we can write:

$$Z_{RC}(s) = \frac{R}{1 + sCR} \quad (3.12)$$

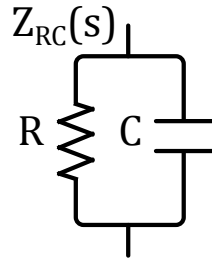


Figure 3.11: RC parallel impedance

and with $g_m = 2/R$ the all-pass transfer function can be obtained:

$$H_D(s) = \frac{2}{1 + s\tau} - 1 = \frac{1 + s\tau}{1 + s\tau} \quad (3.13)$$

where τ is the time constant RC.

In the second case instead, the impedance $Z_i(s)$ is the parallel between the resistance R and the inductance L. Therefore, as shown in figure 3.12, we write:

$$Z_{LR}(s) = \frac{sL}{1 + sL/R} \quad (3.14)$$

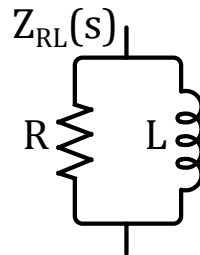


Figure 3.12: RL parallel impedance

As in the previous case, if the value of the transconductance is equal to $g_m = 2/R$, the all-pass transfer function can be obtained:

$$H_D(s) = \frac{2\tau}{1 + s\tau} - 1 = -\frac{1 + s\tau}{1 + s\tau} \quad (3.15)$$

where τ is the time constant L/R. As can be noted, in this second case, the all-pass transfer function has an additional phase shift of 180 degree.

To gain insight on the linearity performances, the voltage swings at the input of each transconductor in figure 3.10 are also reported. Let's to consider the solution with the $Z(s)$ load equal to RL parallel. The transconductor g_{m1} sustains the maximum swing, equal to V_{in} , but the low impedance of the RL load shorts the outputs at low frequency, avoiding propagation of its distortion. The compression of the stage is therefore determined by g_{m2} and g_{m3} only, driven by $V_{in/2}$. As a result, the proposed all-pass circuit implementation enjoys a very high compression at low frequency. Compared to the all-pass stage of figure 3.8a, implemented by replacing the RL load of g_{m1} with an RC load, simulations prove a 1-dB input compression point improvement of 9 dB.

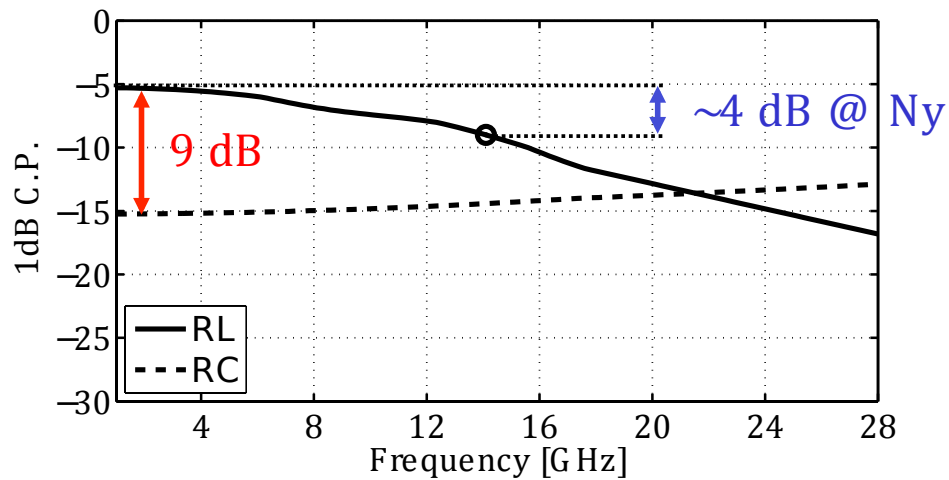


Figure 3.13: 1 dB compression point comparison versus frequency

In figure 3.13 the 1 dB compression point is reported as function of frequency and the continuous black line refers to the proposed solution. At low frequency, a 9 dB difference in the simulated 1dB C.P justifies the choice of the solution with the RL load. It is worth noticing that compression point worsens at high frequency, when the impedance of the RL load rises. However, when the equalizer is fed by a random bit stream experiencing the low-pass channel transfer function, the high frequency spectral components are significantly attenuated requiring less stringent high-frequency compression point to preserve signal integrity. Furthermore, at Nyquist frequency the 1dB compression loss is only of 4 dB, therefore very far from typical channel attenuation.

3.2.4 Group delay contributions

The circuit schematic with the parasitic capacitance is drawn in figure 3.14. The bandwidth of the delay-line stage is limited by C_{out} , determined by the input capacitance of the cascaded stage and tap amplifier. RP and the shunt-peaking inductor LP are sized to achieve $\approx 18 - GHz$ bandwidth. This network (blue color) introduces 9-ps group delay. The transconductor g_{m1} and its high-pass RL load have been sized to add 15-ps group delay.

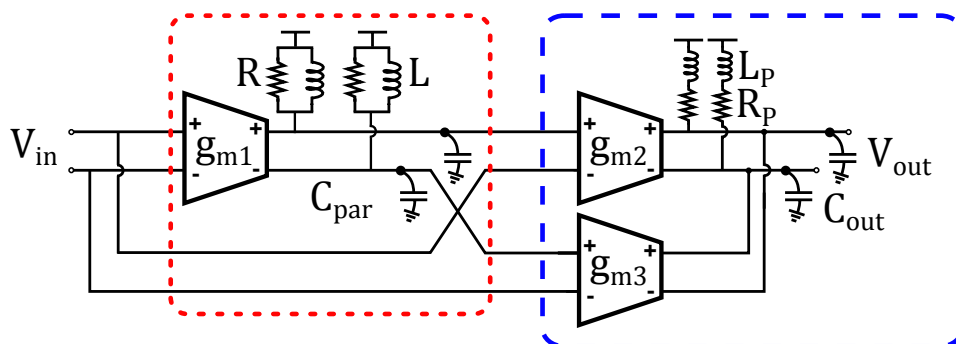


Figure 3.14: Circuit schematic with parasitic capacitance

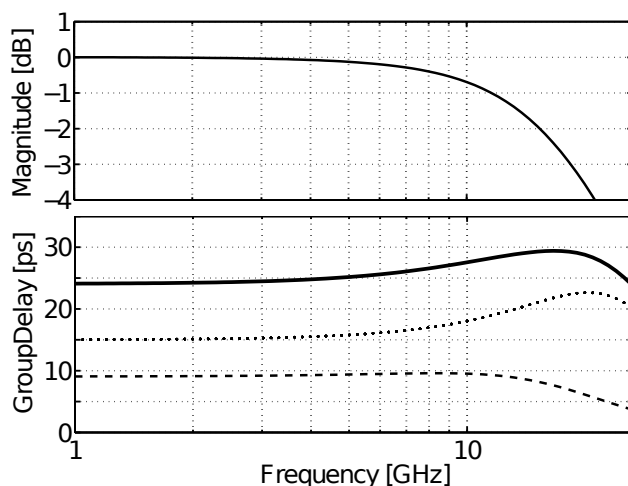


Figure 3.15: Simulated frequency response of the circuit in figure 3.14 and different group delay contribution

An optimized geometry of the inductor L has been devised to achieve high self resonance frequency while keeping compact size. The simulated AC response of a delay stage is shown in figure 3.15. The total group delay, shown in the bottom plot, peaks from 24 ps to 28 ps at 17 GHz. However, you have to pay attention to

the internal parasitic capacitance C_{par} and for this reason the two contributions to the group delay are also reported. The parasitic capacitance loading g_{m1} leads to significant peaking (dotted line) which is partially compensated by the high frequency roll-off of group delay contributed by the Cpar-RP-Lp network (dashed line).

3.2.5 TAP amplifiers design

The circuit schematic of a tap amplifier is drawn in figure 3.16. Each amplifier comprises many elements in parallel, providing a total gain programmable with 6-bits resolution. To achieve positive and negative gains, each element includes two digitally-controlled differential pairs, driven by the same input signal but delivering output currents with opposite sign.

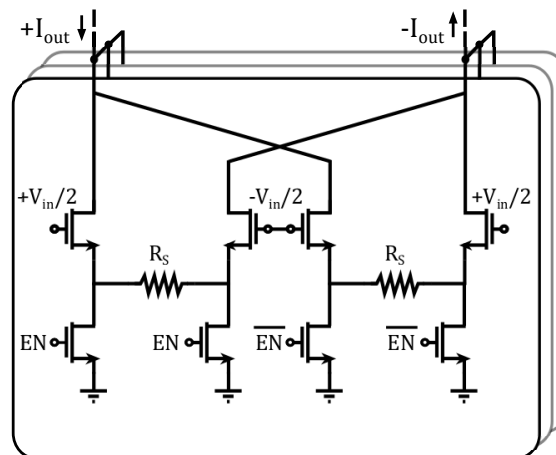


Figure 3.16: Circuit schematic of a tap transconductor amplifier

Only one pair is active at a time allowing changing the gain while keeping a fixed bias current and constant common-mode voltage at the nodes where all the tap output currents are summed. Degeneration resistors are employed in each pair to enhance the linear operation range, in order to withstand the same input signal amplitude of the delay-line stages before gain compression. In particular, the degeneration keeps the 1 dB compression point greater than -5 dBV.

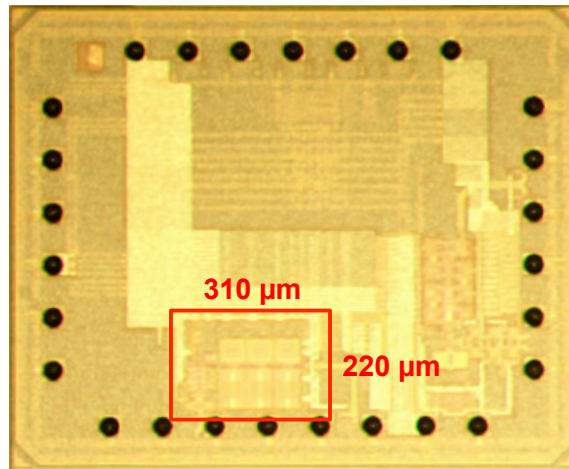


Figure 3.17: Photograph of the die

3.3 Experimental results

Prototypes of the equalizer have been fabricated in 28-nm CMOS LP process from ST Microelectronics. A photograph of the die is shown in figure 3.17. Test chips have been encapsulated in plastic flip-chip BGA packages and mounted on PCB for testing. Core power dissipation, from 1V supply, is 25 mW. First, small-signal AC measurements have been performed with a four-port S-parameter analyzer to assess the functionality of the active delay line and tap amplifiers. Then, equalization capability at 25 Gb/s has been tested: the FIR filter was fed by PRBS sequences transmitted over different backplane channels while the output was monitored on a high speed sampling scope. The scope waveforms were continuously acquired by a PC running a Minimum-Mean-Square-Error (MMSE) adaptation algorithm to control the filter coefficients. The setup used during the measure is shown in figure 3.18.

As an example, the frequency response of a typical backplane channel featuring 20-dB loss at Nyquist is shown figure 3.19. The inset in the figure reports the single-bit channel response showing a large pre-cursor and several post-cursors. The eye diagram at the output of the equalizer is reported in figure 3.20. The input amplitude was set to $900mV_{pk-pk}$. The bottom plot in figure 3.20 shows the bathtub, measured with an Anritsu 1800A BERT, demonstrating 50% eye opening at $BER = 10^{-12}$.

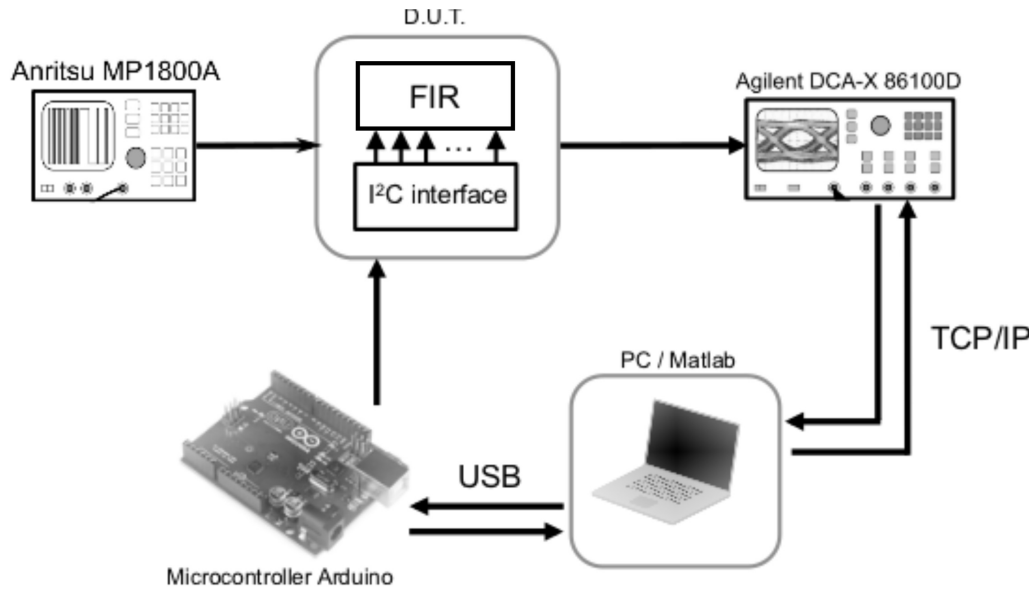


Figure 3.18: Measurement setup

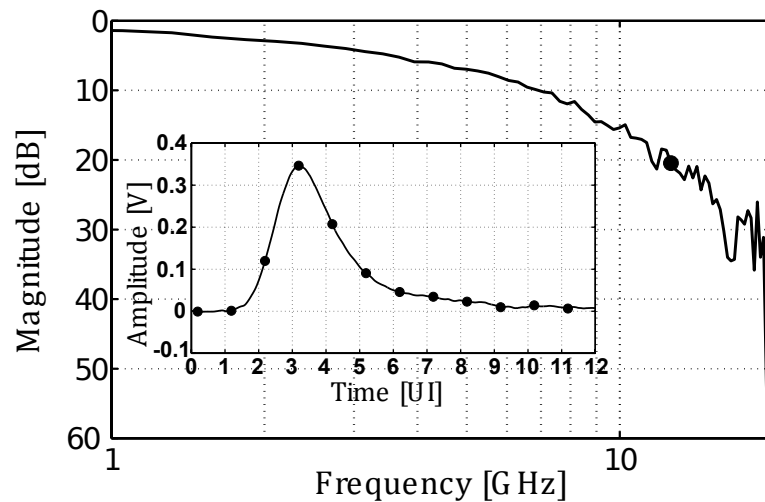


Figure 3.19: Frequency response of a typical backplane channel. In the inset the impulse response

Experimental results are summarized and compared against two recently reported FIR equalizers in Table 3.1. [20] uses a comparable supply but eye opening is worse, data rate is lower and $BER = 10^{-9}$ only. [21] supports a remarkably high speed. On the other hand the supply is much higher than what is allowed without compromising device reliability. Eye opening at $BER = 10^{-12}$ is limited to 30%. The proposed equalizer shows the widest eye opening while keeping the best power consumption normalized to data rate.

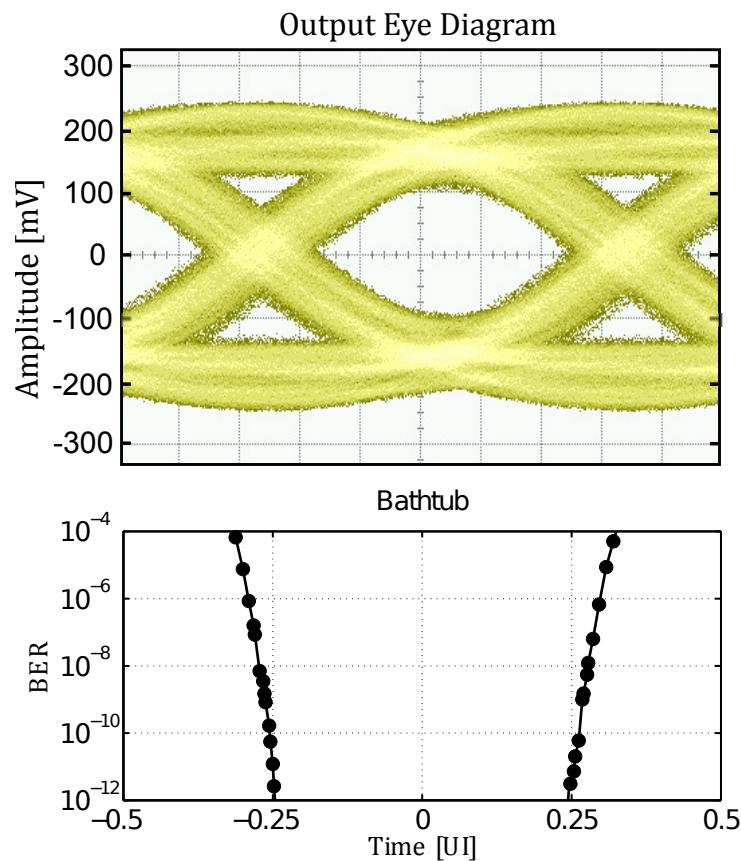


Figure 3.20: 25 Gb/s eye diagram at the output of the equalizer and measured bathtub

Table 3.1: Performance summary and comparison

	This work	[20]	[21]
CMOS node	28nm LP	45nm SOI	65nm
Supply	1V	1.1V	1.6V
# of Taps	4	4	3
Data Rate	25Gb/s	17Gb/s	40Gb/s
Attenuation @ Ny	20dB	21dB	19dB
BER	10^{-12}	10^{-9}	10^{-12}
Horizontal Opening	50%	39%	$\leq 30\%*$
Power dissipation	25mW	32mW	55.2mW
Power/DataRate	1mW/Gb/s	1.9mW/Gb/s	1.4mW/Gb/s

*estimated from plot

Chapter 4

A 28Gb/s Transversal Continuous Time Linear Equalizer in 28nm CMOS

Abstract

To approach the unstoppable demand for higher communication bandwidth, as well as new standard with a data rate of 25-28 Gb/s, multi-level signaling is under investigation to reach 56 Gb/s. The bandwidth limitation of existing backplanes makes transceiver design challenging. Channel losses at high speed are mostly determined by dielectric absorption, which, on the contrary to skin effect, enhances the pre-cursor ISI. Traditionally, RX equalization combines a continuous-time linear equalizer (CTLE), made of the cascade of peaking amplifiers, and decision feedback equalization (DFE). The CTLE needs low power consumption but is difficult to be optimally adapted and has limited capability to correct the pre-cursor ISI. Pulse shaping with a TX FIR filter solves the issues but requires a back channel for adaptation and limits signal swing due to the TX peak-power constraint, a critical issue at low supply and with multi-level signaling.

4.1 Introduction

4.1.1 Motivation

In this work a CTLE with the transversal architecture shown in figure 4.1 is investigated. The D cell refers to a derivative cell. The architecture is similar to a FIR filter, therefore to avoid confusion, we will call it D-FIR (Derivative-Finite Impulse Response) equalizer. Compared to traditional CTLEs, compared to expected simulation results, it provides superior equalization performance while keeping low power consumption. In addition, the transversal architecture makes it compatible with robust LMS adaptation algorithms.

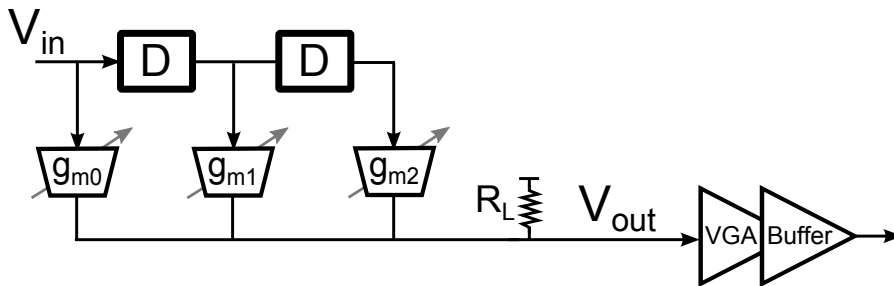


Figure 4.1: Block diagram of derivative equalizer

The new type of equalizer combines the advantages of CTLE equalizer and the advantages of FIR equalizers. As we shall see, the heart of the equalizer is the derivative cell, that is very similar to the base cell used for CTLE. The structure instead, is transversal as that used for the FIR filters implementation. The main advantage is the ability to implement LMS adaptation algorithm in a simple way.

The most critical element inside the FIR filter is the delay cell. The delay cell design is very difficult, because it is necessary to get different and stringent specifications with a reduced area and power consumption. The main limitation is the group delay-bandwidth (GDBW) product that remains constant. Therefore, it is easy to achieve a low GD with a high output bandwidth. However, a too small group delay penalizes the equalization performance because, with the same number of taps, the time span of the filter is reduced. On the contrary, the need for higher group delay, affects the bandwidth. The only way to improve the bandwidth is to use shunt peaking inductors that increase the consumed area.

Another problem is the internal high Q inductor used as load of gm1 stage in figure 2.14. A low Q inductor produces an over peaking in the group delay because it is not compensated by the GD roll-off of the external stage. This introduces a distortion on the signal decreasing the equalization performance. To prevent it, a careful design of the inductor is mandatory and usually it takes a lot of area.

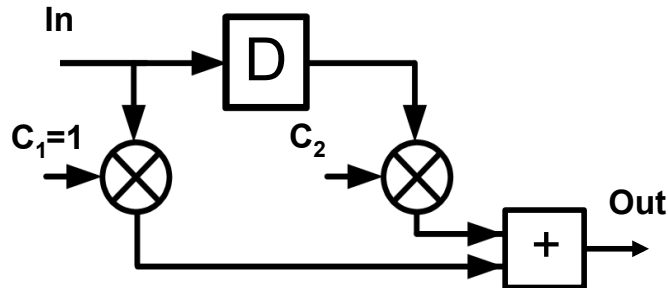


Figure 4.2: Two taps FIR equalizer

The last problem, but not least, is the typical DC loss of FIR equalizer. An example may be helpful to clarify this concept. For FIR equalizer the gain boost at Nyquist is given as:

$$G_{HF} = \frac{C_1 - C_2}{C_1 + C_2} \quad (4.1)$$

Assuming to recover 20 dB loss at Nyquist with the two taps FIR equalizer shown in figure 4.3 and with first coefficient $C_1 = 1$. As result, to recover 20 dB, the second coefficient is $C_2 = -0.82$ and DC loss is:

$$G_{LF} = C_1 + C_2 = -15dB \quad (4.2)$$

As seen before, from the point of view of the signal to noise ratio, a high input signal amplitude is required to keep high SNR on the output node. The proposed equalizer breaks the trade off between booster and loss in DC promising superior performance with a reduced area and power consumption compared to the current state of the art.

4.1.2 Proposed RX D-FIR equalizer

Equalizers must approximate the inverse of channel transfer function for good equalization. Since, a typical channel can be described by all-poles transfer function, the equalizer must be an all-zeros filter to be able to correctly invert the frequency response of the channel. The channel transfer function is therefore gives as:

$$H_{CH}(s) = \frac{1}{1 + C_0s + C_1s^2 + \dots C_{n-1}s^n} \quad (4.3)$$

and, as a consequence, following the basic idea the equalizer will be described as:

$$H_{EQ}(s) = 1 + C_0s + C_1s^2 + \dots C_{n-1}s^n. \quad (4.4)$$

Since the transfer function $H_{EQ}(s)$ is made by only zeros, a derivative cell is required to implement the equalizer.

The block diagram of the proposed equalizer is shown in figure 4.3. The structure has two derivative cells that generate the first and the second derivative of the input signal x . The transconductance of the tap amplifiers is chosen by the digital word D_i with a resolution of seven bits. The output currents of the tap amplifiers are summed together on the resistive load R_{LOAD} . A variable gain amplifier enhances the output signal amplitude and the buffer is used to drive the output pad.

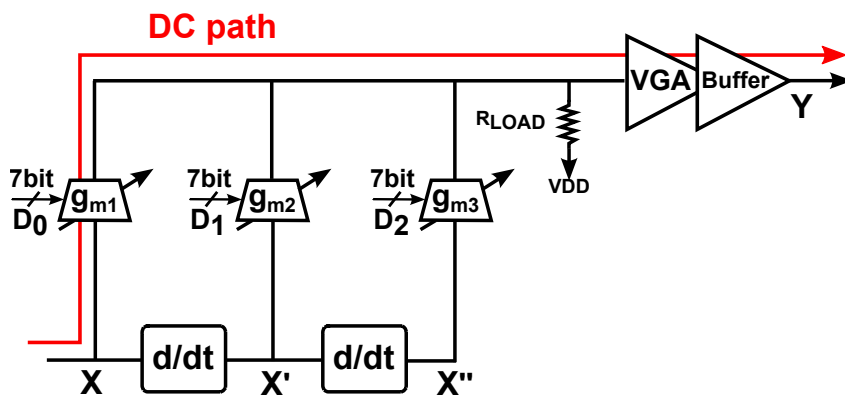


Figure 4.3: Detailed block diagram of the proposed equalizer

As shown in the figure, the red path is called “DC path”, because the DC gain is set by the first coefficient $C_1 = g_{m1}R_{LOAD}$ and also by the VGA and by the buffer. In fact, the derivative cells attenuate the low frequency components of the input

signal X and they do not contribute to the DC gain. Since, DC loss is independent from the high frequency boost, the analog equalizer boosts the high frequency components without introducing loss at DC. Actually, the DC loss is not properly zero, but it is in any case very low. In this way, a strong post-amplification is not required allowing further area and power saving.

4.1.3 D-FIR equalizer behavior

The purpose of the next part is to explain the role of the two derivatives and to justify the use of a three-tap structure. We have already described the D-FIR behavior in the frequency domain. However, for a deeper understanding, it is useful take a look to the figure 4.4. The simulated waveforms describe the behavior in the time domain.

The x signal is the input signal, while x' and x'' are respectively the first and the second derivative of the input. The figure is divided in two parts; on the left the results simulation obtained with an analytic channel modeled only with the dielectric effect, and on the right the simulation results with a an analytic channel modeled only with the skin effect. The two effects are deliberately separated to highlight the two extreme cases. Of course, real channels behavior will be in the middle.

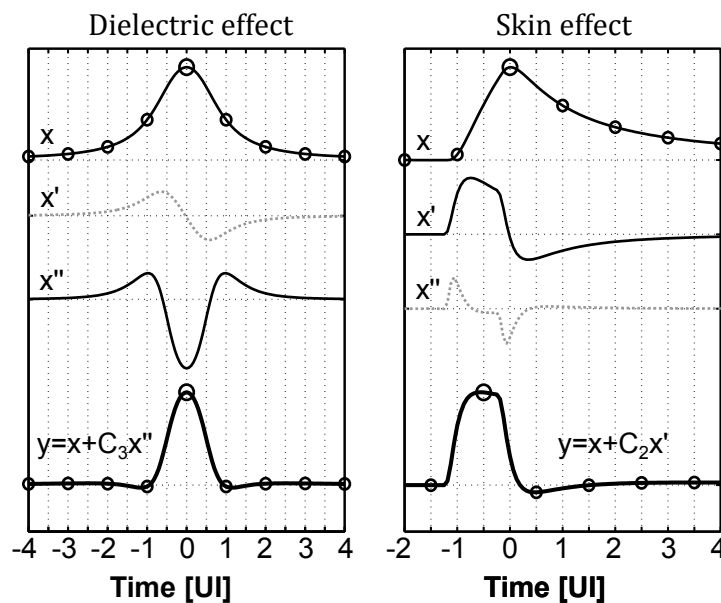


Figure 4.4: Simulated waveform to understand the D-FIR equalizer behavior

The X waveform represents the pulse response of the respective channel. In case of only skin effect the pulse can be equalized with the first derivative multiplied for the coefficient C_2 . The y signal is the equalized waveform on the output node of the equalizer. The second derivative instead is useful to equalize the dielectric effect as can be seen from the figure.

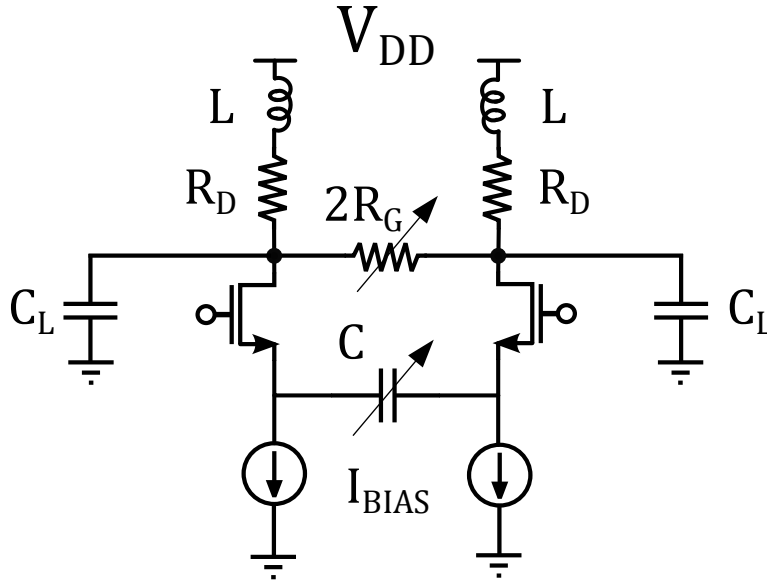


Figure 4.5: Circuit diagram of derivative cell

4.2 Simulation and system design

4.2.1 Derivative cell implementation

The simplified schematic of the derivative cell is shown in figure 4.5. The circuit is a differential pair with a capacitive degeneration, ideal current generators and a resistive load. Moreover, shunt peaking inductors are used to extend the bandwidth. The frequency transfer functions of the derivative cell is:

$$H_s = -\frac{2sCR_L}{1 + s\frac{2C}{g_m}} \frac{1}{1 + sC_LR_L} \quad (4.5)$$

where the resistance $R_L = R_D // R_G$ and, for simplicity, the inductance L is not considered. Therefore, the zero and the poles are:

$$\omega_z = \frac{1}{2CR_L} \quad \omega_{p1} = \frac{g_m}{2C} \quad \omega_{p2} = \frac{1}{C_L R_L} \quad (4.6)$$

where p_2 is the parasitic pole and C_L represent the capacitive load of the next stage. Increasing the capacitor, both the first pole and the zero moved to low frequency. The transconductance and the load R_D set the gain of the cell and the frequency separation from the pole p_1 and the zero. The zero is programmable with the capacitance C and can be moved to choose different frequency response and behavior of the derivative cell. Lastly, the resistance R_G is connected between the two nodes of the output and therefore, it must have a large value. It serves to reduce the gain of the cell to avoid signal compression of the next stage. The simulated frequency responses of derivative cell are shown in figure 4.6.

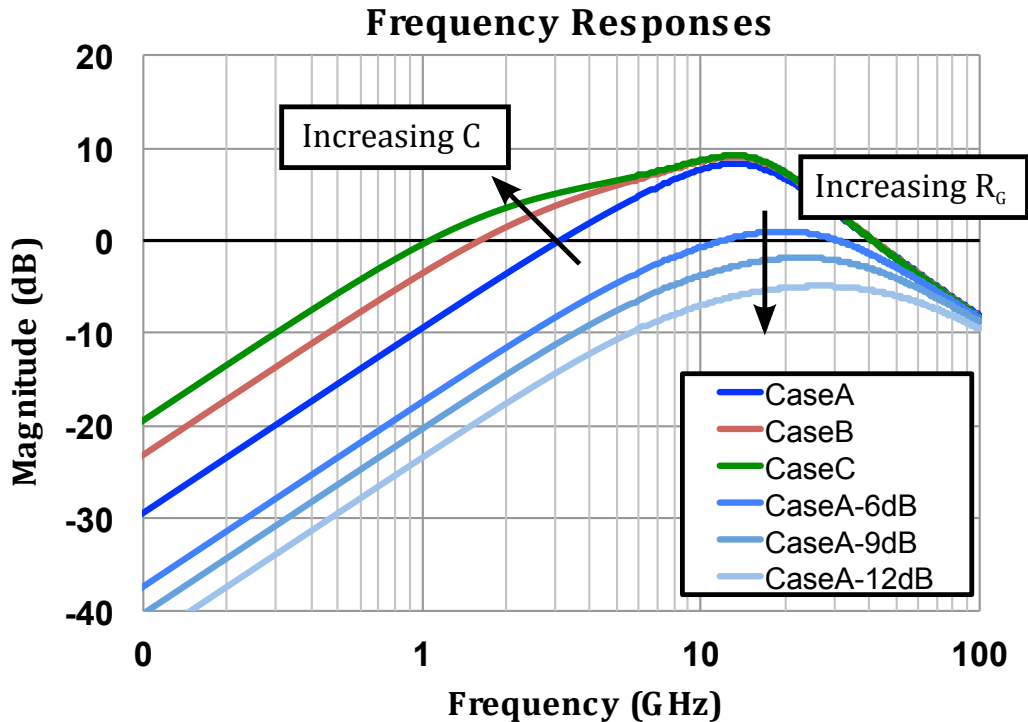


Figure 4.6: Frequency responses of derivative cell

As said before, increasing the degeneration capacitor, the zero shifts to low frequency. As shown in the figure, respect to case A, the waveforms B and C start to boost before, crossing the 0-dB axis at low frequency, while reaching the same gain. Another aspect is the shape of the waveforms. For some kind of channels, a different shape in the boost it could be useful for improving the equalization. Increasing the resistance R_G instead, the gain decreases and the net income is a

downshift of the waveform. In the figure, the cases with reduced gain are shown only for case A.

From figure 4.6, we also understand why principally the first coefficient path sets the DC gain for this equalizer. As we can see, the derivative transfer function has ideally a DC gain equal to zero. For this reason at low frequency the signal can reach the output only through the first coefficient path. As result, the analog equalizer boosts the high frequency without introducing loss at DC. Therefore, very low tap gains are required for the equalization and a very small post-amplification gain is required to preserve the output signal amplitude, saving area and power consumption.

Actually, the real case is slightly different because the ideal current source I_{BIAS} (figure 4.5) will be substitute with real current generator with a certain output resistance $R_S/2$. The schematic is shown in figure 4.7.

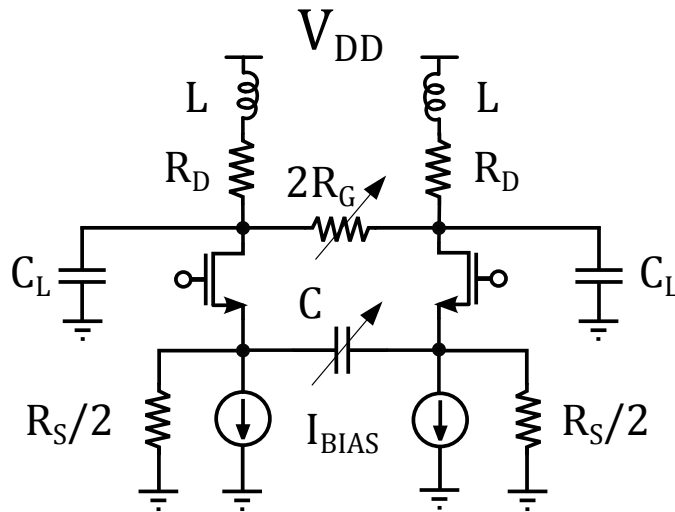


Figure 4.7: Derivative cell schematic with real current generators

To derive the frequency response of this circuit may be simpler to analyze the single ended half circuit drawn in figure 4.8. The presence of output resistance $R_s/2$ introduces a DC gain as shown in the frequency transfer function:

$$H(s) = \frac{g_m R_L}{1 + \frac{g_m R_s}{2}} \frac{1 + s/\omega_z}{1 + s/\omega_{p1}} \frac{1}{1 + s/\omega_{p2}} \quad (4.7)$$

where the zero and the poles are:

$$\omega_z = \frac{1}{R_s C_s} \quad \omega_{p1} = \frac{1 + g_m R_s / 2}{R_s C_s} \quad \omega_{p2} = \frac{1}{R_L C_L} \quad (4.8)$$

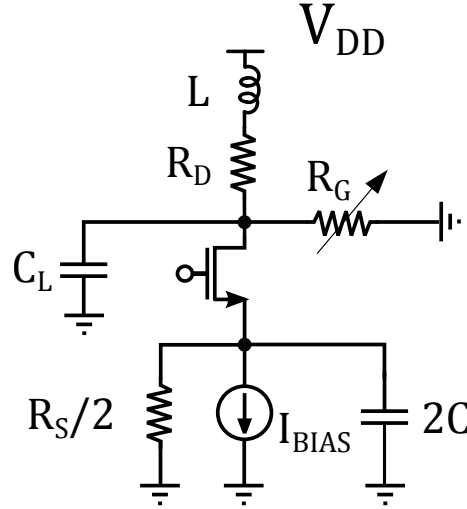


Figure 4.8: Single ended half circuit

To dimension correctly the cell, given the output capacitive load C_L , the load resistance R_L must be set for $\omega_{p2} \geq 2\pi f_N$ where f_N is the Nyquist frequency. The first pole is better to be close to $2\pi f_N$ for max gain. Now, the zero can be move with the programmable C degeneration to have the desired peaking. Moreover, it should to be noted that the DC_{gain} and the high frequency gain HF_{GM} are:

$$DC_{gain} = \frac{g_m R_L}{1 + g_m R_s / 2} \quad HF_{GM} = g_m R_L \quad (4.9)$$

The DC_{gain} value is usually low because the output resistance R_S is in the order of some unity of $k\Omega$ and it has no impact on the low frequency gain of the equalizer. To conclude in the next picture 4.9 a sketch of the magnitude and phase of $H(s)$ just described.

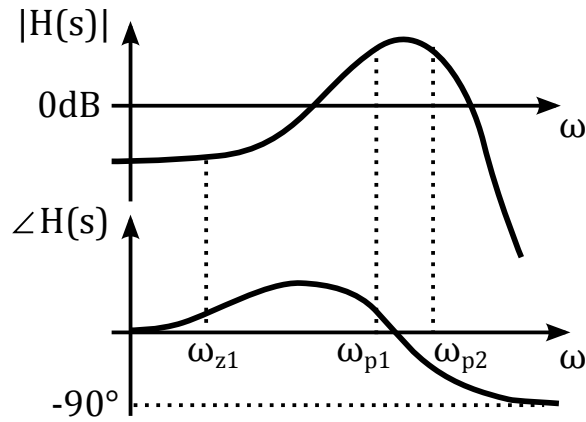


Figure 4.9: Simple sketch from derivative cell dimensioning

4.2.2 Tap transconductor implementation

Due to the natural low gain at low frequency of the derivative cell, the only way to improve the linearity of the equalizer is a careful design of the tap amplifiers. The figure 4.10 shown the parallel structure of the tap. Each amplifier comprises ten elements in parallel with the possibility to achieve positive and negative gain with 6-bits resolution. Part of the digital word is controlled in a thermometric fashion to assure a monotonic increase of the tap gain. The conversion is made by a decoder as shown in figure 4.11. Therefore, the necessary cells are ten and they are controlled by their own enable bit b_I .

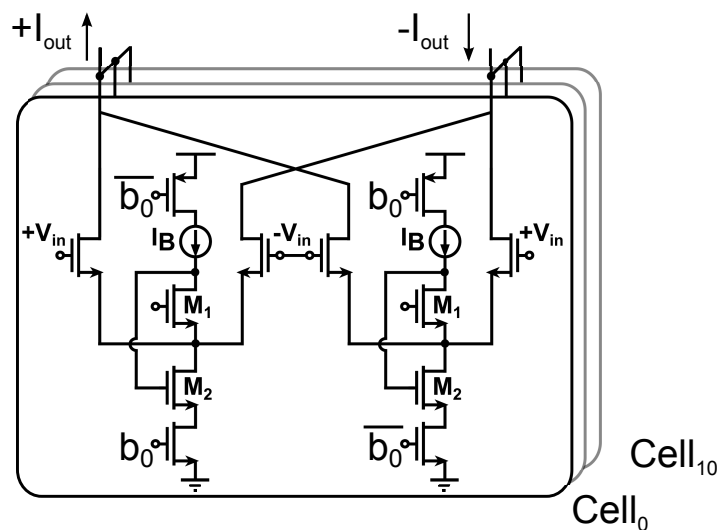


Figure 4.10: Parallel structure of tap amplifier

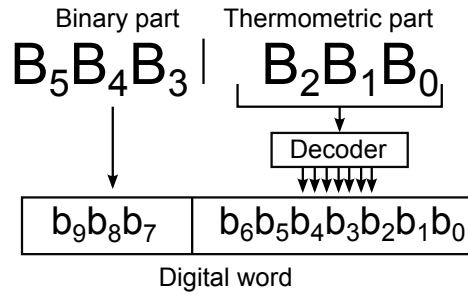


Figure 4.11: Conversion of the tap gain digital word

As in the previous chapter each element includes two digitally-controlled differential pairs, driven by the same input signal but delivering output currents with opposite sign. The main difference is the flipped voltage follower (FVF) circuit connected to the common source of each differential pair. The purpose of FVF circuit is to improve the linearity lowering the impedance seen in the source. It is essentially a cascade amplifier with negative shunt feedback where the gate terminal of M2 is used as input common mode terminal and its source as output terminal. It is characterized by very low output impedance due to shunt feedback provided by M1, high low supply requirement close to a transistor threshold voltage V_{TH} , low static power dissipation and high gain bandwidth. Output current variations are absorbed by M1, while the current in M2 remains constant. A single side of the tap amplifier circuit is proposed more clearly in figure 4.12.

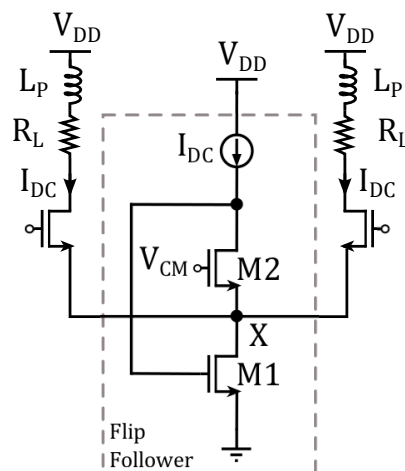


Figure 4.12: Detailed circuit of a single side tap amplifier

Unlike the conventional differential pair, the circuit in figure 4.12 is able to source a large amount of current. The large sourcing capability is due to the low impedance at the node X that is lowered by the output impedance of the FVF:

$$R_{OUT} \approx \frac{1}{g_{m1}g_{m2}r_{02}} \quad (4.10)$$

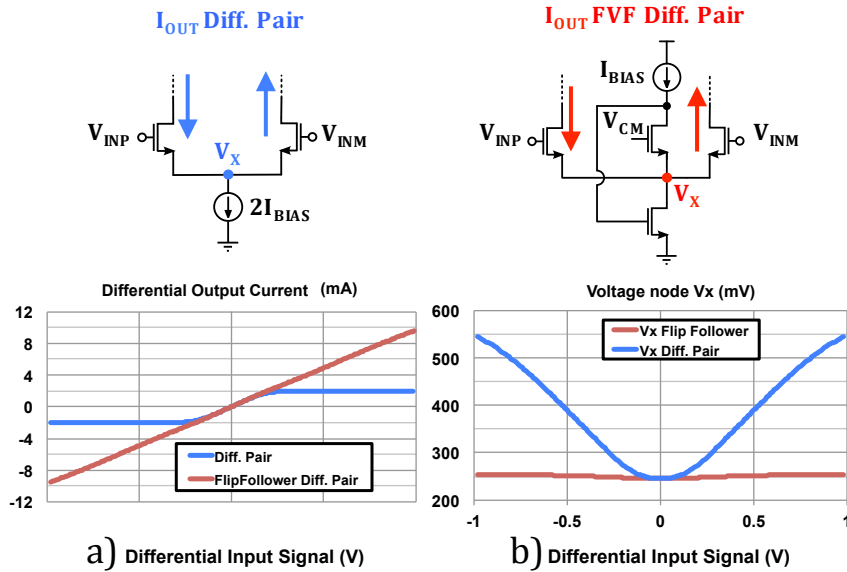


Figure 4.13: Tap DC characteristic without load

Note that M1 provides shunt feedback and, for that, detailed analyses will be necessary to ensure stability. In figure 4.13 are shown some simulation to highlight the huge linearity improvement with the use of FVF differential pair. In the graph 4.13 the output current and the voltage node V_x are plotted as function of the differential input signal. As can be see, the current of classic differential pair (blue line) is clamped to the maximum current $2I_{BIAS}$. For FVF solution instead, the rest of the current is drive by the tail transistor. This behavior is possible because, due to the very low impedance, the voltage node on x node doesn't move. Actually, the simulations have been done without resistive load on the differential pair drain.

If we consider a resistive load for the differential pair the results are slightly different and they are shown in figure 4.14.

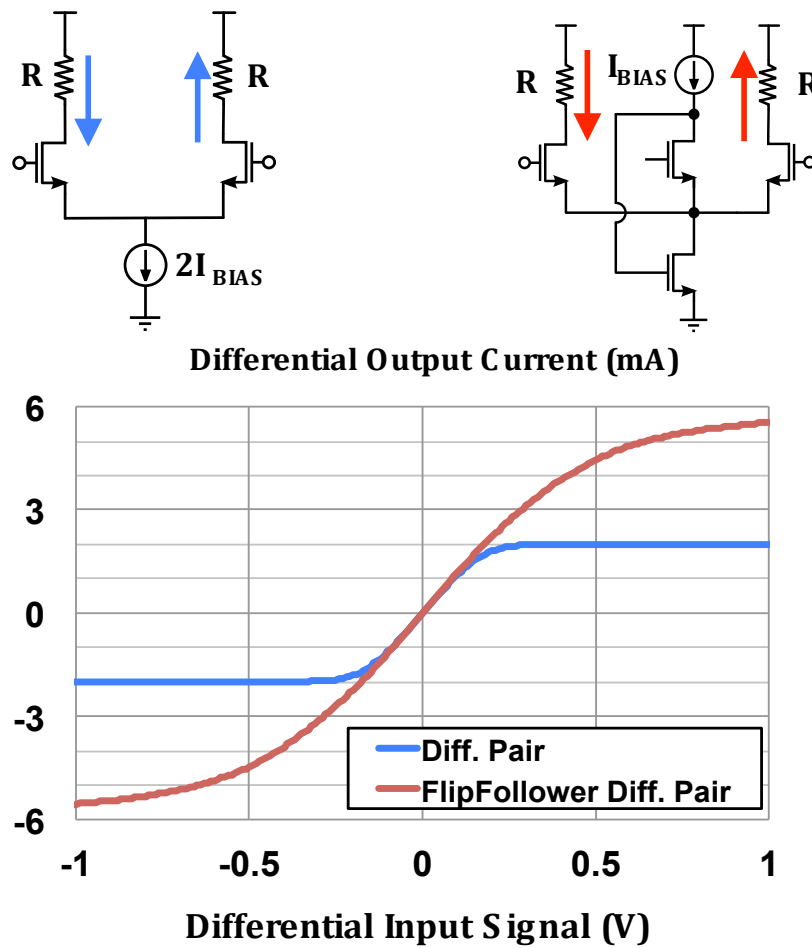


Figure 4.14: Tap DC characteristic with resistive load

In this case the FVF output current is not a straight line as in the previous figure. The little compression of the current is due to the voltage generated by the load on the drain of the differential pair. To quantify the linearity improvement is necessary a 1 dB compression point simulation. Simulating the two cases for the same size of the differential pair we obtain:

$$1dBC.P.Diff.Pair = -16dBV \quad (4.11)$$

$$1dBC.P.FVFDiff.Pair = -10dBV \quad (4.12)$$

therefore with a 6 dB improvement.

Conclusion

A 25Gb/s continuous-time linear 4-tap FIR equalizer and an 28Gb/s ultra-compact 3-tap derivative-FIR equalizer for backplane data equalization have been presented. For both equalizers, a system level analysis to derive equalizer specifications has been proposed with emphasis on the discussion of issues specifically related to continuous-time core cell implementation. The design of a test chip has been described for the first chip with a detailed analysis of the delay line and tap amplifiers. Realized in 28 nm LP CMOS technology, core silicon area is only 0.065 mm and comprehensive experimental results proved successful equalization of 25Gb/s data stream with maximum power dissipation of 25mW. Measurements are summarized and compared with published FIR equalizers in table 3.1. The presented equalizer features a state-of-the-art power dissipation normalized to data-rate. These features are particularly effective to accommodate the demand of actual transfer rates, back-compatibility and energy efficiency of emerging 100 Gb/s standards.

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