

# HIGH-SPEED SERIAL DATA LINK DESIGN AND SIMULATION

BY

EDWARD W. LEE

THESIS

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Adviser:

Professor José E. Schutt-Ainé

# ABSTRACT

This thesis describes the modeling and simulation of 10 Gb/s serial data link architectures. The first architecture uses a continuous-time receiver equalizer to equalize a channel with known frequency response. The second incorporates an adaptive decision feedback equalizer (DFE) for use in unknown or time-varying channels.

*To my Father and Mother*

# ACKNOWLEDGMENTS

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# TABLE OF CONTENTS

<b>LIST OF FIGURES</b> . . . . .	<b>vi</b>
<b>LIST OF ABBREVIATIONS</b> . . . . .	<b>vii</b>
<b>CHAPTER 1 INTRODUCTION</b> . . . . .	<b>1</b>
1.1 Motivation . . . . .	1
1.2 High Speed Data Links . . . . .	2
1.3 Challenges of Multi-Gigabit Backplane Data Transmission . . . . .	2
1.4 Pre-Emphasis and Equalization . . . . .	3
1.5 Receiver Design for AWGN ISI Channels using DFE . . . . .	6
1.6 Adaptive Equalization . . . . .	7
1.7 Thesis Organization . . . . .	8
<b>CHAPTER 2 CONTINUOUS-TIME EQUALIZER DESIGN</b> . . . . .	<b>9</b>
2.1 Methodology . . . . .	9
2.2 Channel Model . . . . .	10
2.3 Receiver Model . . . . .	10
2.4 Results . . . . .	11
<b>CHAPTER 3 DISCRETE-TIME ADAPTIVE EQUALIZER DESIGN</b> . . . . .	<b>15</b>
3.1 Methodology . . . . .	15
3.2 Receiver Design . . . . .	17
3.3 Results . . . . .	17
<b>CHAPTER 4 CONCLUSION</b> . . . . .	<b>19</b>
<b>REFERENCES</b> . . . . .	<b>20</b>

# LIST OF FIGURES

1.1	Backplane link diagram. . . . .	2
1.2	Communication link. . . . .	2
1.3	Simple pre-emphasis implementation. . . . .	4
1.4	(a) Simple DFE system. (b) DFE removing postcursor ISI. . . . .	5
1.5	Optimum receiver design with equalizer. . . . .	6
1.6	Example pulse shapes in receiver. (a) Output of sampled matched filter is always symmetric. (b) Output of noise whitening filter resulting in only postcursor ISI. . . . .	7
1.7	DFE feedback equalizer combined with a feed-forward linear transversal equalizer, which mitigates precursor ISI. . . . .	7
1.8	Adaptive receiver block diagram. . . . .	8
1.9	Alternative training system. . . . .	8
2.1	Design flow. . . . .	9
2.2	Channel model. . . . .	10
2.3	Receiver model. . . . .	11
2.4	Channel frequency response and group delay. . . . .	12
2.5	Frequency response after receiver. . . . .	13
2.6	Eye diagrams of receiver. . . . .	14
3.1	Channel model. . . . .	16
3.2	Channel frequency response. . . . .	16
3.3	System model. . . . .	17
3.4	Adaptive DFE equalization results. . . . .	18

# LIST OF ABBREVIATIONS

BER	bit error rate
EMI	electromagnetic interference
FIR	finite-impulse response
DFE	decision feedback equalizer
ISI	intersymbol interference
LE	linear transversal equalizer
LMS	least mean square
MLSE	maximum-likelihood sequence estimation
PCB	printed circuit board
VGA	variable gain amplifier

# CHAPTER 1

## INTRODUCTION

### 1.1 Motivation

The ever-increasing need for high data rate in communication systems is driving data transmission within networking equipment to multi-gigabit per second rates. Serial I/O interfaces are being widely adopted into backplane applications, short and long-haul communications, and chip-to-chip links for computing applications. A typical backplane system is shown in Fig. 1.1, consisting of chip packages soldered onto daughter cards that are then plugged into the backplane through connectors.

Looking at the many recent emerging industrial standards gives us insight into the speeds of interest. Ethernet data rates are advancing from 100 Mb/s onward to 10 Gb/s. In computing applications, serial ATA is increasing from the current 1.5 Gb/s and 3 Gb/s, and targeting 6 Gb/s. PCI Express 2 is 5 Gb/s and going on to 8 Gb/s in PCI Express 3.

Copper/FR4 backplane based serial transmission is one of the most commonly used techniques in high-speed transmission due to its lower cost, reduced complexity and high reliability. Due to these advantages, copper/FR4 PCBs are expected to continue as the material of choice for telecomm and computing applications. However, these PCBs do face some technical challenges, mainly because of distortion due to skin effect, dielectric loss and reflections. This results in reduction of transmission bandwidth, leading to closure of the transmission eye and, ultimately, high BER at the receive end.



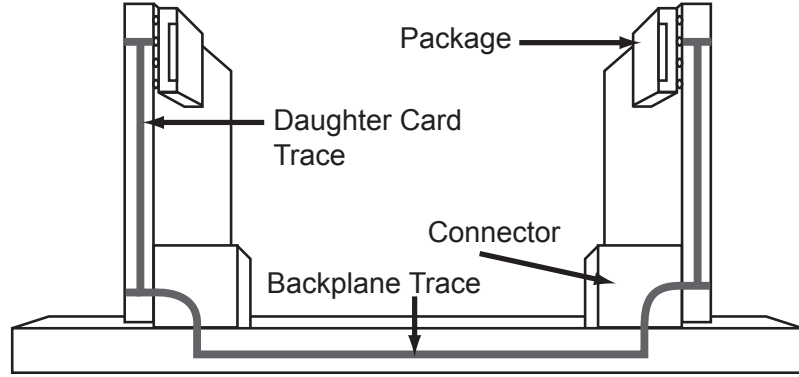


Figure 1.1: Backplane link diagram.

## 1.2 High Speed Data Links

A communication link typically consists of a transmitter, the communication channel, and a receiver (Fig. 1.2). The transmitter takes the digital data and converts it to analog waveforms on the channel. The channel is the communication medium between the transmitter and the receiver, and can have physical realizations such as free space for wireless communication, optical fibers for optical communication and PCB traces, coaxial cables or twisted pair wires for off-chip electrical communication. When a channel is implemented using electrical technologies, the implementation is often referred to as an interconnect.



Figure 1.2: Communication link.

## 1.3 Challenges of Multi-Gigabit Backplane Data Transmission

The challenges of multi-gigabit data transmission can largely be divided into two categories, the first being structural issues and the second being frequency dependent

loss due to material properties. One of the greatest structural challenges is overcoming impedance mismatches due to connector-via transitions. Improved laminates and optimized connector-via architectures can alleviate these issues, but as data rates steadily rise, the frequency dependent loss of the channel which usually manifests as low-pass nature becomes a serious threat to signal integrity.

Skin effect describes how high frequency currents tend to travel on the surface of the conductor, rather than the whole cross section of the conductor. This reduces the effective conductive area of the trace, increasing resistance which causes the signal to be attenuated. As frequencies increase into the multi-gigabit per second rates, dielectric loss emerges as the dominant factor in high frequency attenuation, since its effect is proportional to frequency, whereas skin effect is proportional to the square root of frequency. Dielectric loss can be defined as a ratio of conductivity to frequency, known as loss tangent. Materials such as FR-4 with a high loss tangent will see a large attenuation of signal at high frequency. The combined attenuation and dispersion due to these effects causes the signal to spread over to adjacent signals. This is called intersymbol interference (ISI).

From a circuit point of view, the limited bandwidth of the channel increases the rise-time/fall-time of the signal, which causes the voltage level of a single bit to depend on the bit pattern it resides in. The degradation of the signal due to ISI results in two main issues on the receiver side: reduced voltage margin and timing errors due to jitter in zero crossings. In order to reduce BER, transmitter pre-emphasis, receiver equalization, or a combination of the two is generally used to compensate the channel.

## **1.4 Pre-Emphasis and Equalization**

Pre-emphasis boosts the high frequency components of the signal at the transmitter, before the signal is sent to the channel. One simple implementation of pre-emphasis

is a 2-tap finite impulse response (FIR) filter, resulting in a boost of signal every time there is a signal transition (Fig. 1.3).

Some of the disadvantages of pre-emphasis include higher power requirements to boost the signal, aggravated crosstalk, and increased electromagnetic interference (EMI) due to the overshoot and undershoot. Also, as the channel is usually not known *a priori*, more focus is aimed at equalization design, while a simpler pre-emphasis implementation is used.

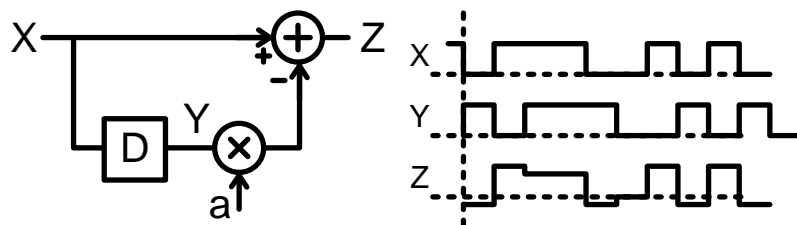


Figure 1.3: Simple pre-emphasis implementation.

Equalization is used at the receiver to flatten the channel frequency response to overcome the high frequency signal loss during transmission. Receivers can be implemented in discrete-time or continuous-time. Continuous-time receivers can be designed using an analog equalizer, while discrete-time equalizers can be designed using digital filters.

Several types of discrete-time designs can be considered. As introduced in the literature [1], using a maximum-likelihood sequence estimation (MLSE) receiver is optimum according to a probability of error criterion. However MLSE has a complexity that grows exponentially with the length of the ISI channel time dispersion, which makes implementation expensive. Other sub-optimum methods include using a linear transversal equalizer (LE) or a decision feedback equalizer (DFE).

The linear transversal equalizer (LE), which can be viewed as being the same as a FIR filter, is very versatile in that an infinite length LE can equalize any channel  $H(\omega)$  as long as  $H(\omega) \neq 0, |\omega| \leq \frac{\pi}{T}$ . Also, it can be used to implement matched filtering as well if the taps are spaced at Nyquist intervals instead of symbol intervals [2, 3].

One of the main disadvantages of using a common LE design is that the equalizer usually has a high-pass characteristic due to the low-pass nature of the channel. This can result in severe amplification of high frequency noise components, especially with channels with spectral nulls.

The decision feedback equalizer (DFE) is a non-linear filter which allows some of the noise problems associated with linear equalizers to be overcome. By using a linear combination of past decisions, the DFE compensates for the channel response while eliminating noise amplification. While the DFE is vulnerable to ‘error propagation’ where an initial decision error will feed back and cause a sequence of errors, the system will recover from the error events in relatively short order if the data is insured to be random [4].

Figure 1.4 shows the operation of a DFE canceling postcursor ISI, in which the pulse response of a channel is assumed to be a simple RC low-pass filter [5].

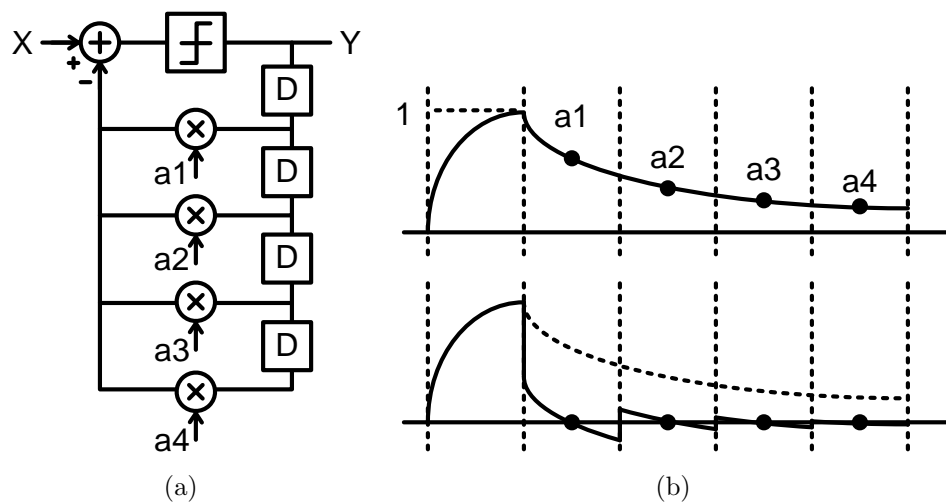


Figure 1.4: (a) Simple DFE system. (b) DFE removing postcursor ISI.

## 1.5 Receiver Design for AWGN ISI Channels using DFE

Figure 1.5 shows the optimum equalizer receiver design process for digital transmission channels corrupted by AWGN and ISI [4, 6]. A matched filter is used for maximum SNR before the sampler, and a noise whitening filter is selected so that the resulting response is minimum phase. This creates a causal and stable output  $F(z)$ , effectively eliminating precursor ISI, as shown in Fig. 1.6. The matched filter, symbol-rate sampler and noise whitening filter are collectively called the whitened matched filter (WMF). As can be seen, the DFE front-end is equivalent to the WMF [7], which ultimately operates as the precursor equalizer for a DFE (Fig. 1.7).

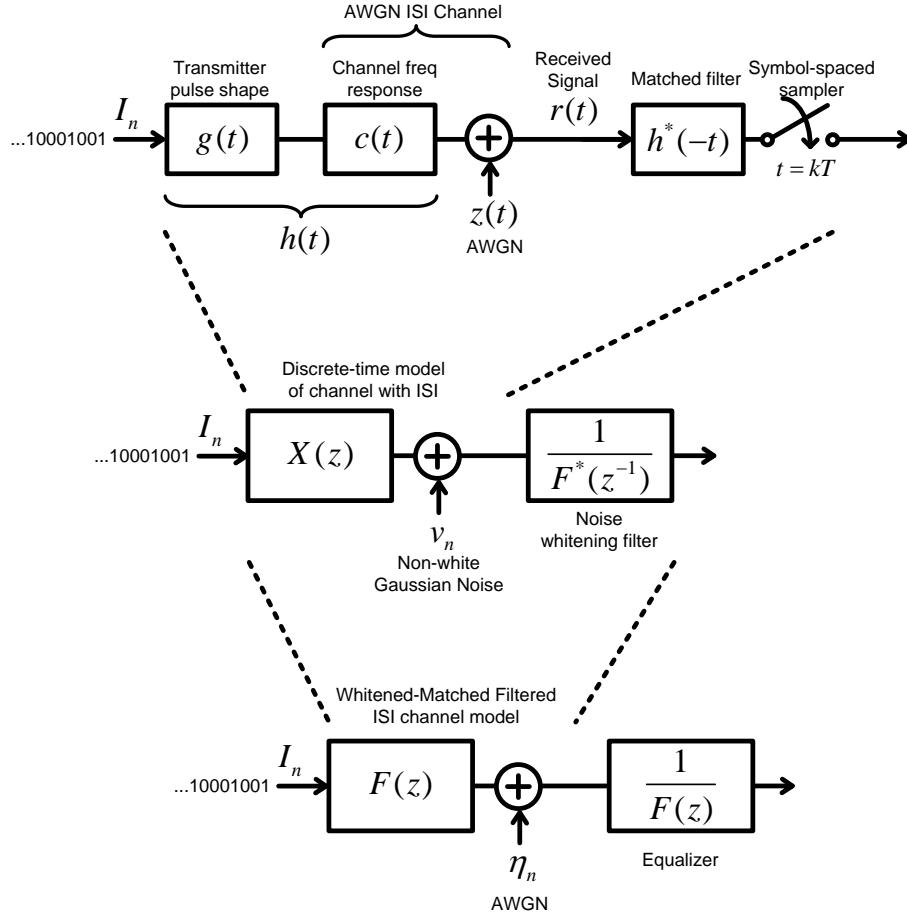


Figure 1.5: Optimum receiver design with equalizer.

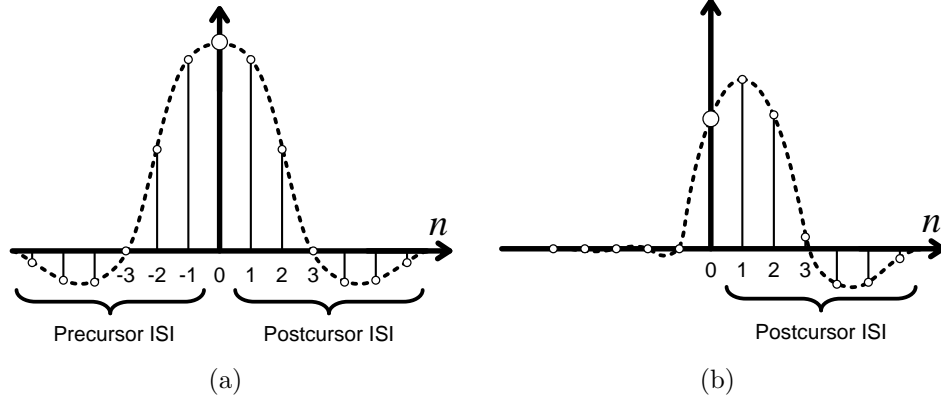


Figure 1.6: Example pulse shapes in receiver. (a) Output of sampled matched filter is always symmetric. (b) Output of noise whitening filter resulting in only postcursor ISI.

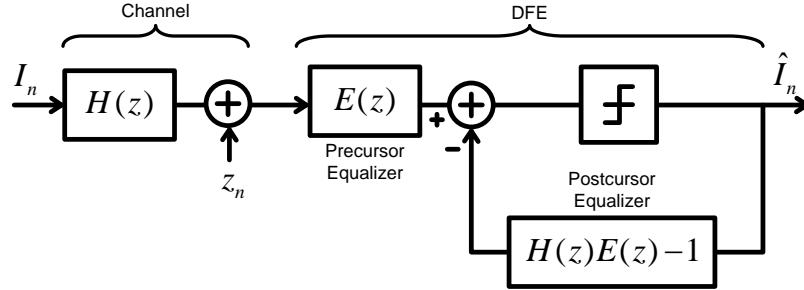


Figure 1.7: DFE feedback equalizer combined with a feed-forward linear transversal equalizer, which mitigates precursor ISI.

## 1.6 Adaptive Equalization

Adaptive equalization [2] must be considered because we cannot assume to know beforehand what the channel response is, and we are limited by implementation complexity. Coefficients can be decided using MMSE criterion [8], and the LMS algorithm is widely used due to its ease of implementation.

The simplest form of adaptive equalizer is shown in Fig. 1.8. In steady state, the decisions at the slicer output are assumed correct and are used to generate the error signal (*decision directed mode*). However, the decision directed mode is only effective in tracking slow variations that result in a low error rate at the slicer, and requires a training sequence to initially acquire the equalizer coefficients (*training mode*).

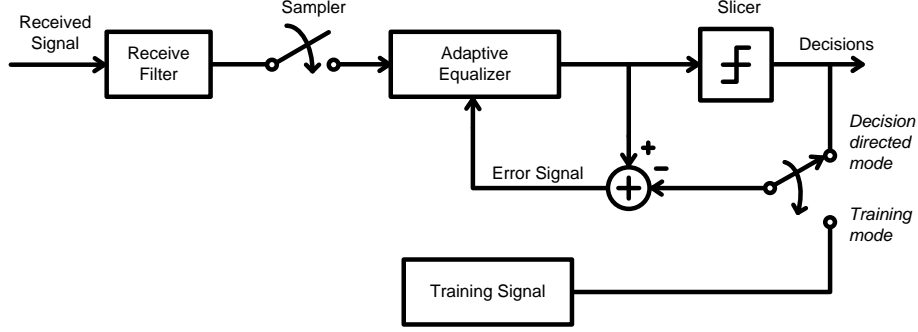


Figure 1.8: Adaptive receiver block diagram.

To avoid singularities from the zero of the channel transfer function inside the unit circle, we can select an equalizer such that  $W(z)H(z) \cong z^{-L}$ , where the result is a time-shifted version of the desired signal, as shown in Fig. 1.9. A more general  $W(z) = Y(z)/H(z)$  can also be used to reduce noise enhancement or to ease implementation, in what is known as a partial-response signaling system [9].

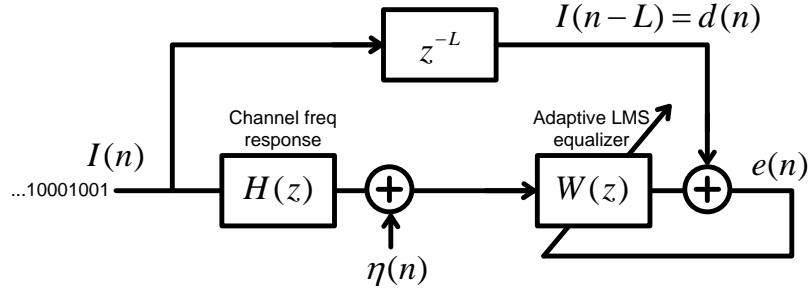


Figure 1.9: Alternative training system.

## 1.7 Thesis Organization

This thesis documents all the effort devoted to design and improve upon a high-speed serial link receiver. Chapter 2 describes the design of a continuous-time receiver equalizer for use in 10 Gb/s data links. A discrete-time adaptive DFE receiver is simulated using ADS and measured channel S-parameters in Chapter 3.

# CHAPTER 2

## CONTINUOUS-TIME EQUALIZER DESIGN

### 2.1 Methodology

Compared with discrete-time equalizers, continuous-time equalizers have some attractive properties in terms of noise, jitter and potential power dissipation. The use of passive networks is also gaining attention due to the low power and smaller size of implementation. In this chapter, an analog equalizer is designed for use in a 10 Gb/s data receiver.

The design flow is shown in Fig. 2.1. The channel is first modeled using the W-element in HSPICE where the channel AC response and transient output are simulated. The AC response is used to obtain the single zero and pole of the analog equalizer. The transient channel output is fed into MATLAB Simulink where the receiver is modeled.

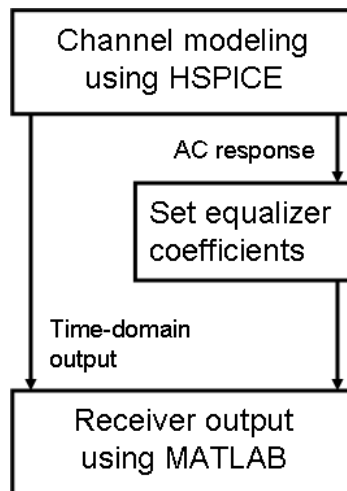


Figure 2.1: Design flow.



## 2.2 Channel Model

The channel is modeled as a differential line with a FR-4 coupled microstrip. The 20" traces on the line are copper with width 508  $\mu\text{m}$ , thickness 35  $\mu\text{m}$ , and thickness of the dielectric being 254  $\mu\text{m}$ . The width between the traces is set to be 700  $\mu\text{m}$ . The HSPICE field solver is used with the W-element to simulate the transmission line. The driver supplies 50  $\Omega$  source termination to the channel and has 10 GHz bandwidth. The receiver termination includes AC coupling capacitance, bond-wire inductance, ball-pad capacitance, and die-pad capacitance. Input amplitude of the voltage source is set to be 1 Vpp differential, with additive channel noise modeled as having infinite bandwidth and 30 dB SNR. The complete channel model is shown in Fig. 2.2.

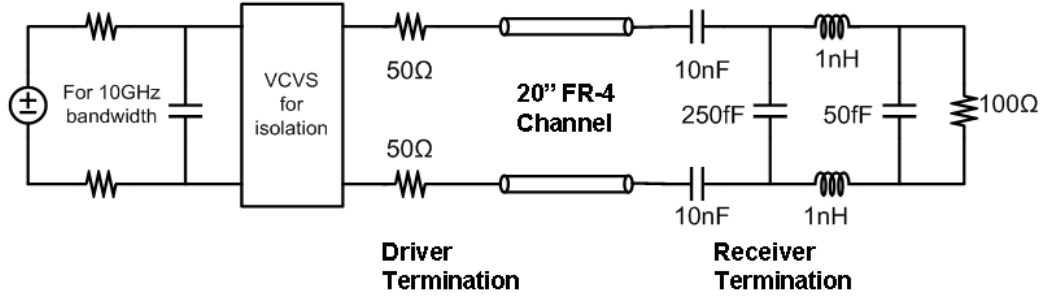


Figure 2.2: Channel model.

## 2.3 Receiver Model

The receiver model consists of a single pole variable gain amplifier (VGA) and the equalizer used to flatten the frequency response. The equalizer is modeled as a single zero, single pole amplifier. The single real zero is needed to compensate the gain loss and open the signal eye. The real pole is needed for several reasons. First, boosting the gain above a certain frequency (usually the signal bandwidth) is unnecessary because it merely amplifies more high frequency noise. Second, wide bandwidth

causes signal peaking in the middle of the eye, generating large jitter at zero crossing points of the data signal. Signal peaking in the middle of the eye also decreases the voltage margin at the middle of the eye, which increases BER. Finally, signal peaking increases crosstalk between signal lines.

Thus the real pole is needed to limit the bandwidth. The pole location determines the amount of boosting and the frequency range due to the zero. If the signal loss is low, the zero and the pole can be close together. But, if the loss is high, the zero and the pole should be separated further to get sufficient boosting. It is desirable to have the zero and pole as close as possible to avoid adding excessive non-linear phase response, for a symmetrical eye.

The receiver model which was simulated in MATLAB is shown in Fig. 2.3.

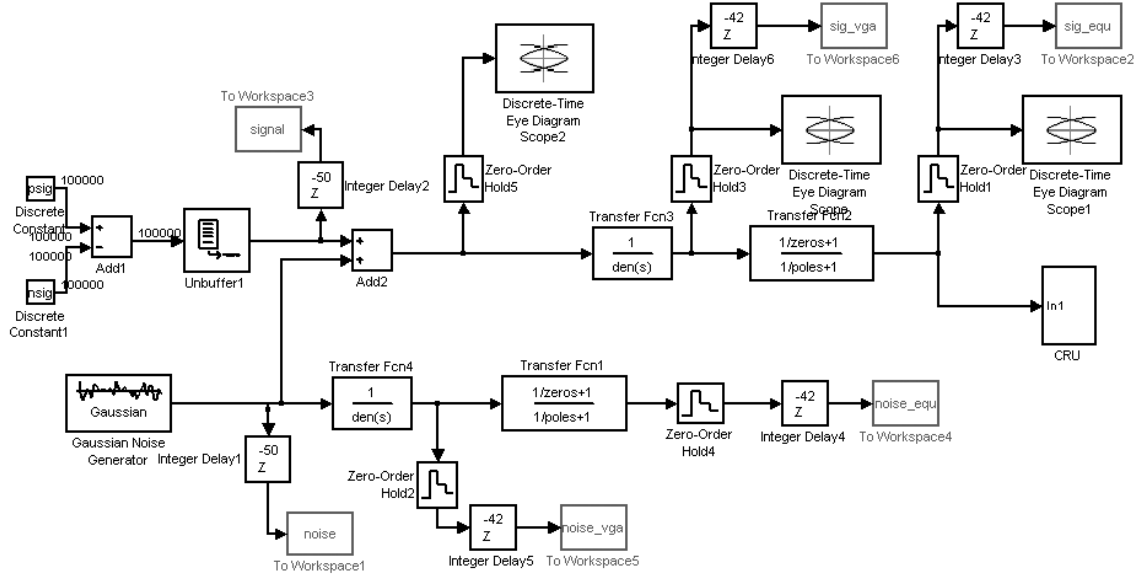


Figure 2.3: Receiver model.

## 2.4 Results

The channel model frequency response and group delay are shown in Fig. 2.4, displaying the channel, driver-channel, and driver-channel-termination, respectively. Ta-

ble 2.1 summarizes the channel characteristics. A -6 dB gain offset in the low frequency is due to the  $50\ \Omega$  output impedance of the driver, and the channel and driver losses at 10 GHz are -10 dB and -3 dB, respectively. At this frequency, the impedance mismatch between channel and termination causes additional 2 dB signal loss. The phase response is fairly linear from 100 MHz up to 10 GHz and group delay at this frequency range is about 3 ns.

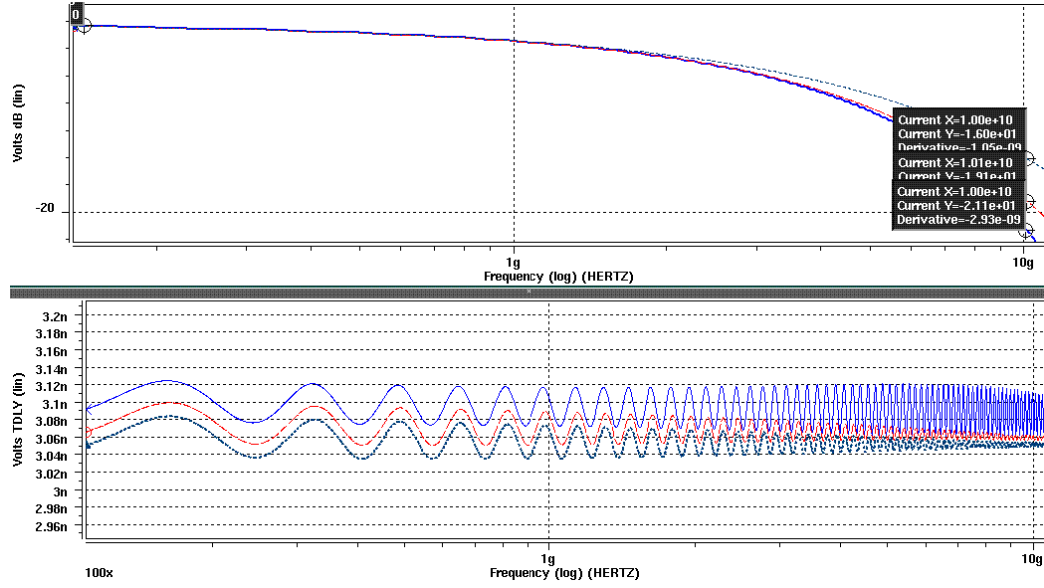


Figure 2.4: Channel frequency response and group delay.

Table 2.1: Channel characteristics.

	Channel	Driver + Channel	Drv. + Ch. + Termination
Gain @ low freq.	-6.07 dB	-6.08 dB	-6.08 dB
Gain @ 10 GHz	-16 dB	-19 dB	-21.1 dB
3 dB freq.	2.63 GHz	2.36 GHz	2.33 GHz
Group Delay	3.05 ns	3.07 ns	3.1 ns

The flattening of the frequency response after the receiver is shown in Fig. 2.5. The pole of the VGA is at 7 GHz and the equalizer has a 2.1 GHz zero and 4 GHz pole. The eye diagram of channel output has relatively good eye shape which shows that the signal loss at high frequency in channel is not severe (Fig. 2.6(a)). Slight asymmetry of the eye shows that phase response is not perfectly linear. The VGA filters high

frequency noise, increasing the SNR to 46.2 dB. However, the eye is more closed and has higher zero crossing variations due to the limited bandwidth (Fig. 2.6(b)).

The difference between the zero and the pole of the equalizer is about 2 GHz. More than 2 GHz difference causes overshoot in both eye diagram and frequency response. Less than 2 GHz difference causes a dent in the frequency response and closes the eye. The 3 dB frequency of the channel response is about 2.6 GHz, so the zero should be near that frequency. The zero is tweaked to 2.1 GHz considering the zero crossing variation, the flatness of the eye center, and the symmetry of the eye. After equalization, zero crossing deviation is reduced to 4 ps and SNR is 43.9 dB (Fig. 2.6(c)).

The SNR is degraded at the equalizer output by 2.6 dB due to the equalizer attenuating the signal as well as the noise (Table 2.2). The signal attenuation is more significant due to the signal having a low-frequency profile compared to the noise power which is distributed with a wide frequency range.

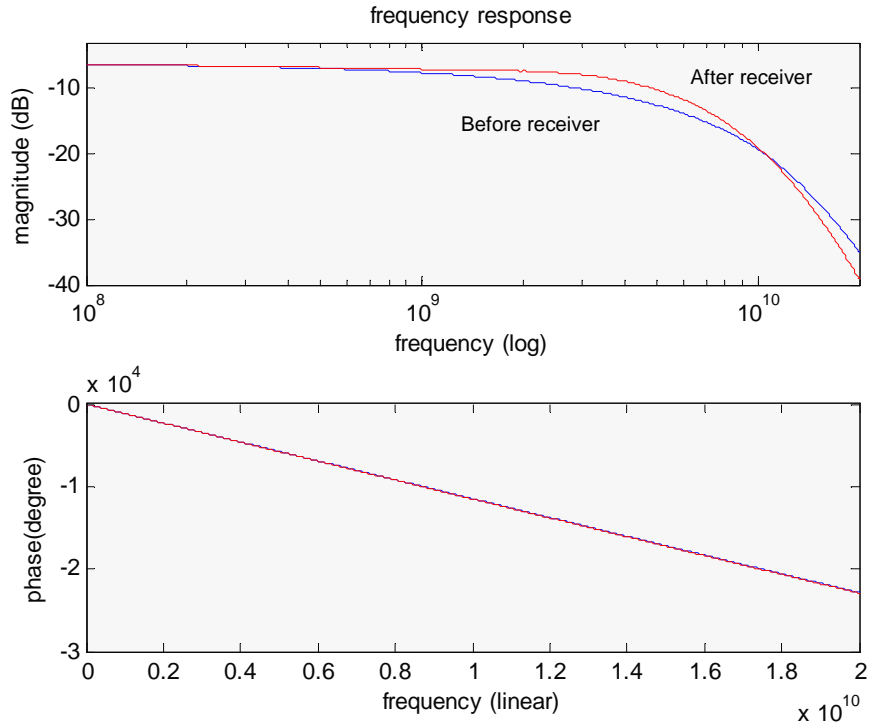
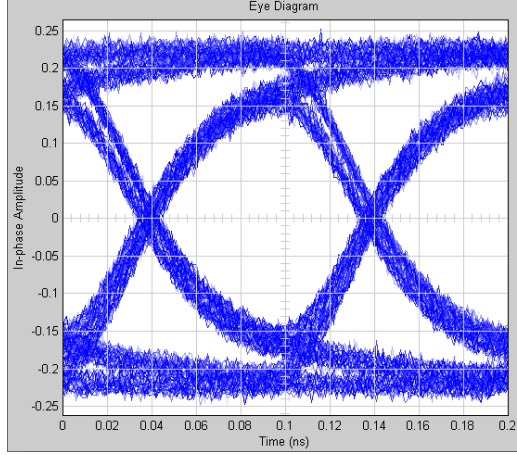
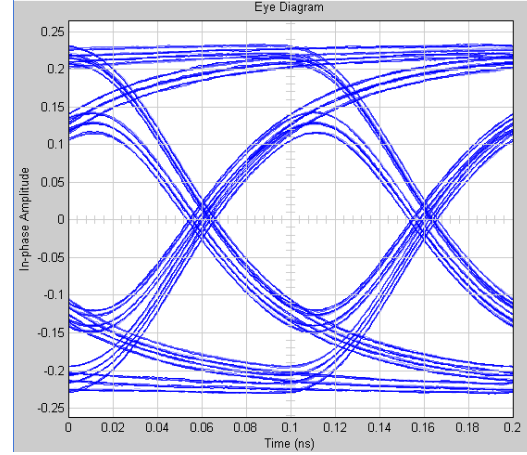


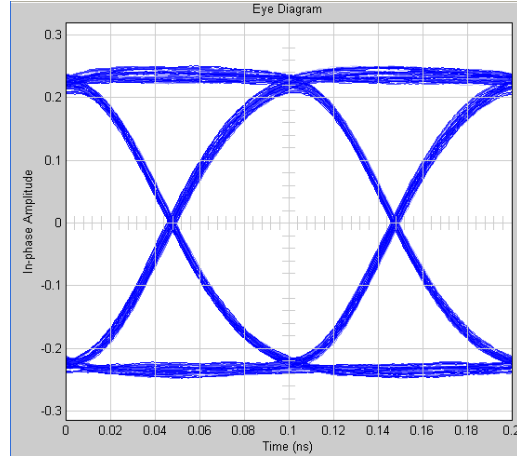
Figure 2.5: Frequency response after receiver.



(a) Channel output



(b) VGA output



(c) Equalizer output

Figure 2.6: Eye diagrams of receiver.

Table 2.2: Receiver output

	Channel Output	VGA Output	Equalizer Output
Signal Power Sum	0.0358 W·Hz	0.0312 W·Hz	0.049 W·Hz
Noise Power Sum	$3.44 \times 10^{-5}$ W·Hz	$7.46 \times 10^{-7}$ W·Hz	$2.02 \times 10^{-6}$ W·Hz
SNR	30.2 dB	46.2 dB	43.9 dB
Zero Crossing Variation	10 ps	15 ps	4 ps

# CHAPTER 3

## DISCRETE-TIME ADAPTIVE EQUALIZER DESIGN

### 3.1 Methodology

In many cases, the channel of interest is unknown or has characteristics that vary over time and temperature. This is very common in backplanes where by design, daughter cards are meant to be able to be installed or uninstalled according to the user's needs. In this case an adaptive method of updating the coefficients of the equalizers is most desirable. Given the measured S-parameters from a backplane, we design and simulate a discrete-time adaptive DFE to equalize a 10 Gb/s data link.

Using the Signal Integrity Workshop DesignGuide in Advanced Design System (ADS), an adaptive DFE is modeled and simulated. The LMS algorithm [1, 9] is used to adaptively find the equalizer coefficients for both precursor and postcursor parts of the receiver. Unlike the system described in the previous chapter, this system is simulated in discrete-time with a global clock, and so zero crossings are a non-issue in this case.

The backplane of interest is a 1 m Nelco 4000 13-SI board having 2.5" traces on each daughter card. The measured 4-port S-parameters of the backplane were obtained from the IEEE 802.3ap Backplane Ethernet Task Force [10] (Fig. 3.1). The frequency response of the channel is shown in Fig. 3.2.

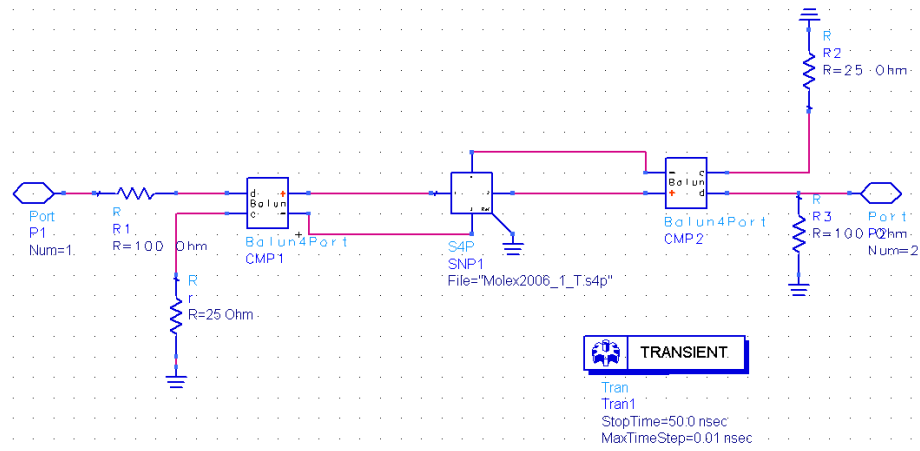


Figure 3.1: Channel model.

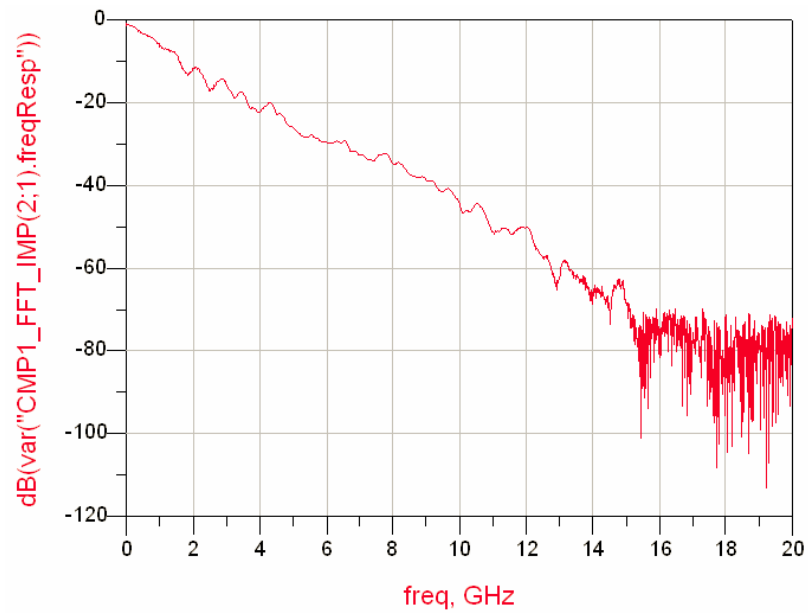


Figure 3.2: Channel frequency response.

## 3.2 Receiver Design

The receiver is modeled as a 7-tap precursor and 7-tap postcursor symbol-spaced equalizer (Fig. 3.3). Transmit bits are generated using a pseudo-random bit sequence. An initial training period using the correct transmitted data is required to converge the DFE before the receiver is able to begin coefficient tracking, where the slicer decisions are used instead.

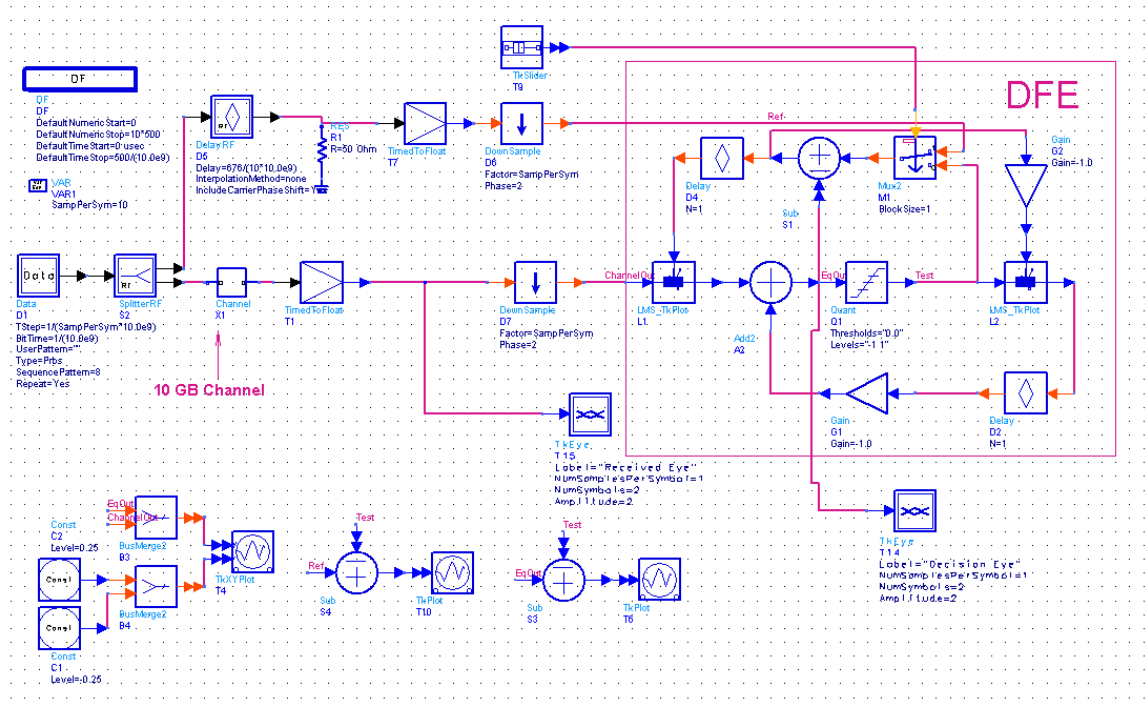
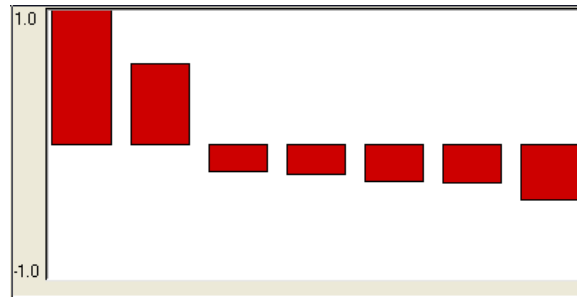


Figure 3.3: System model.

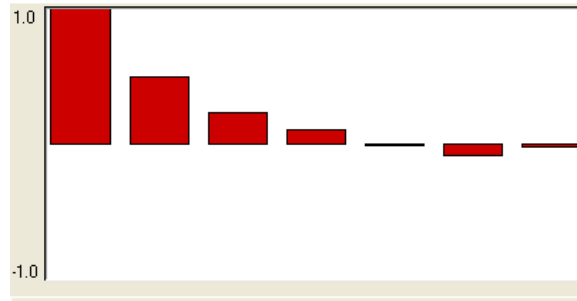
## 3.3 Results

The LMS algorithm is seen to be successfully converged and the channel equalized. As symbol-spaced sampling is used, an open discrete-time eye diagram is obtained using the data at the input of the decision slicer (Fig. 3.4).

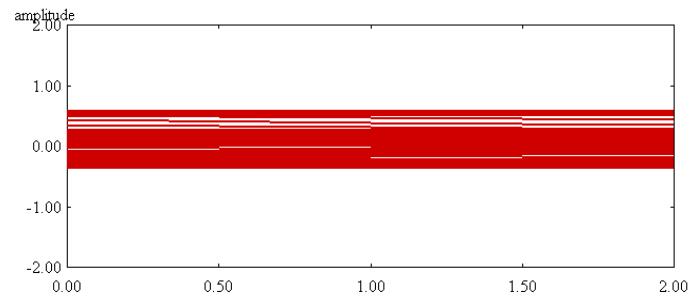




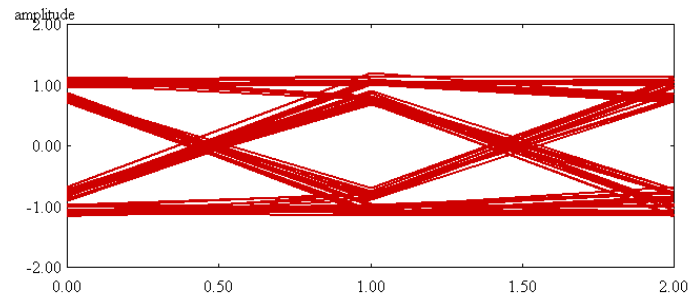
(a) Precursor equalizer weights



(b) Postcursor equalizer weights



(c) Received eye diagram



(d) Opened eye diagram

Figure 3.4: Adaptive DFE equalization results.

# CHAPTER 4

## CONCLUSION

We have presented two different equalizer architectures suitable for 10 Gb/s serial data links. Design trade-offs in both transmit-side and receive-side equalizers were described. A continuous-time equalizer was designed to equalize a channel with known characteristics. An adaptive DFE for use in unknown or time-varying channels was also presented. Future work could be extended to parallel architectures, and the use of pre-computation or pipelining architectures such as unfolding could be included for higher performance.

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