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Design Methodology of Pre-Emphasis Filtering for Digital Data Communications

by

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A thesis submitted to the Faculty of Graduate Studies and Research
in partial fulfillment of the requirements for the degree of
Master of Applied Science

Department of Electronics

Carleton University

Ottawa, Canada

April 2004

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ABSTRACT

The finite impulse response (FIR) filter as pre-emphasis technique for high-speed (Gbps) data transmission over physical layer is studied. A top-down design methodology for a high-speed backplane serializer/deserializer (SERDES) is proposed in this thesis. MATLAB scripts are programmed for optimizing tap coefficients of FIR filter. The program flow charts are given and a list of program functions are described. Simulation results of the pre-emphasis solution to a specific backplane channel are shown. Then in CMOS 90nm technology, one complete programmable multiple-tap symbol-spaced FIR filter is implemented. The tap coefficients obtained from MATLAB are realized in circuit design. Comparison is given between the MATLAB results and the HSPICE results to verify the MATLAB programming.

ACKNOWLEDGEMENTS

First and foremost, I would like to thank my thesis supervisor, Professor Tad Kwasniewski for his guidance, support and encouragement. Secondly, I wish to thank Shoujun Wang, Yuming Tao, Bill Brenza and other colleagues during I worked as intern at Ottawa IC Development, ALTERA CO. for their many great help, discussion and advices. The MATLAB program and IC implementation was first finished at the ALTERA Ottawa IC Development branch. Thirdly, I want to thank ALTERA CO. and Carleton University for the financial supports in this research. Finally, I wish to thank my parents for their love and understanding that have given me extra strength to pursue my second Master degree.

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CHAPTER 1 INTRODUCTION

1.1 The Serial Data Link

Applications such as computer-to-computer or computer-to-peripheral interconnections require gigabit-per-second (Gbps) rates over varying distances. Traditional methods using parallel buses, that require many parallel wires, are costly and power inefficient. Optical fiber is efficient for long distance communications, but not on printed circuit boards (PCB). Thus, low-cost, high-speed serial links using copper traces on PCB are an attractive solution for such applications [1, 2]. Formed in 2002, the High Speed Backplane Initiative (HSBI) launched the promise of developing serial link technology capable of sending serial data at multi-Gbps across a backplane environment up to a distance of 30 inches (including two connectors). It also announced that future work of the group would address 10 Gbps and higher data rates over the same environment [3, 4]. The need to address the improvement of the bandwidth capacity of backplane applications has also been recognized by the Optical Internetworking Forum (OIF) [5]. Within the OIF, the Physical and Link Layer (PLL) Working Group formed the Common Electrical I/O (CEI) project, whose goal is to define electrical specifications for 4.976 to 6+ Gbps and 9.95 to 11+ Gbps serial signaling for chip-to-chip and board-to-board applications. These specifications will provide the basis for the electrical interfaces that will be used to enable the higher capacities needed for the next generation of systems.

Due to the skin-effect loss and dielectric loss in conductors, copper traces show a low-pass frequency response that limits the signaling bandwidth. This will cause serious

inter-symbol interference (ISI). The backplane environment itself further complicates the ISI problem. Copper channels that are identical except for the specific layer connection will have different throughput capabilities. This is attributed to the "unpredictable" effect caused by the vias of the connector board interface [6]. ISI directly leads to affecting the bit-error rate (BER) of the system since the clock-and-data recovery (CDR) unit at the receiver samples the voltage level at what should be the center of the eye, based on the recovered clock. The key to understanding ISI is that when the pulse spreading induced by the channel spreads energy over multiple bit times, a destructive effect occurs. Since the net voltage at the point in time where the data is sampled includes the effect of energy from multiple bit times, most of the energy will be contributed from the intended bit, however contributions will occur from previous bits as well. The impact of this can be additive, with the net effect being jitter, and a closing of the eye.

1.2 Solutions to the ISI Problem

Both transmit and receive equalization have become accepted ways to deal with the loss associated with the channel. There are different opinions on how to apply equalization in order to enable higher speed serial line rates. In general, there are three possible scenarios:

- Employ only transmit pre-emphasis [7, 8, 9,19];
- Employ only receive equalization [10, 11, 12,18];
- Employ some combination of transmit pre-emphasis and receive equalization [13, 14].

When the channel characteristics can be anticipated, applying the inverse of the channel characteristics to the signal at transmitter can reduce the effects of the channel loss. This is referred to as pre-emphasis. The objective is to flatten the frequency response of the channel curve to reduce the ISI. Pre-emphasis has the effect of attenuating the low frequency components of the signal and is proven to be an acceptable mechanism for reducing ISI. Techniques also are being implemented at the receiver to help in opening the eye of a signal from a backplane connection. Boosting the high frequency components within the receiver IC will compensate for the loss induced by the channel. This is generated through analog techniques. In general, receive equalization is implemented using a high-pass filter. Digital/mixed-signal techniques can also be applied to provide receiver equalization, such as decision feedback equalization (DFE). The equalization techniques mentioned above have a net effect of flattening the channel response, which in the time domain results in the eye being widely opened. When the frequency domain effect of pre-emphasis is added with the frequency domain effect of receive equalization, the net time domain effect is that the eye is opened even further.

For multi-Gbps transmission channels, analog equalization is an attractive alternative to digital approaches. The main advantages of analog equalization are lower power, less silicon area and, most important, capability of performing equalization at high speeds. Unfortunately, analog compensation of the channel frequency response comes at the expense of reduced signal to noise ratio (SNR) [15]. For systems with analog equalization, careful system design should be performed to minimize such deterministic high-frequency noises such as crosstalk and supply induced noise.

The finite impulse response (FIR) filter is widely used for implementing transmit pre-emphasis. The inputs to this filter are the present, past and possibly future symbols. The coefficients of the filter depend on the channel characteristics. The length of the filter (i.e. number of filter taps) depends on the number of symbols that affect the present symbol. Therefore, the output of the filter is the present symbol value plus the weighted values of the N affecting transmitted symbols. Digital receiver equalizers, using FIR filters, require high-resolution sampling ADCs that run at GHz speeds. The development of GHz rated ADCs is quite a challenging task. On the other hand, the disadvantage with analog continuous time equalization is the requirement of very wide-bandwidth front-end receiver circuits that run at the same speed as the input data. This thesis will deal with pre-emphasis employing the FIR filter at the transmit side. Determining the optimal adaptive FIR filter coefficients and the optimal number of taps will be the main focus of this thesis.

1.3 Top-down Design Methodology

A BER of 10^{-17} is typically required for backplane data applications. This translates to approximately 193 days of error free transmission at a data rate of 6.25 Gbps. It is impractical to simulate such low BER at the circuit level. The estimation of BER has to rely on high-level mathematical analysis using probability density functions (PDF). However, it is not easy to mathematically comprehend the impact of circuit non-idealities on the BER. Implementation losses could come from:

- Non-ideal convergence of the equalizer taps and clock data recovery phase;
- The package and non-ideal on chip termination;

- Deterministic and random jitter;
- Common mode rejection ratio and power supply noise rejection ratio;
- On-board AC coupling capacitors and the signal via discontinuity under the package ball;
- Receiver sensitivity.

In [16], a 6.25 Gbps backplane Serializer/Deserializer (SerDes) design using a top-down design methodology is illustrated as shown in Fig. 1.1.

Fig. 1.1 Design flow of a top-to-down methodology

The design process starts with laboratory measurements of the channel response. High-level analysis using MATLAB estimates the maximum possible throughput for the targeted BER on a given channel, and identifies the best equalization and CDR architecture. Behavioral models of the analog front end (AFE) and the digital algorithmic engine translate the high level architecture into feasible CMOS circuits. The behavioral simulation, which has the same channel input as MATLAB, must reproduce the results of the MATLAB simulations. The mixed mode co-simulation step replaces the behavioral model for the AFE with a transistor netlist, and uses a SPICE like circuit simulator for the AFE and a logic simulator for the digital logic. The mixed-mode co-

simulations need to reproduce the results from the behavioral simulation. Once this has been achieved, a correlation between the MATLAB analysis and circuit performance is established. The MATLAB analysis using all the PDF based circuit information produces the final BER estimation.

Such a design flow as described in the above methodology is not openly available. The key element in the methodology is the MATLAB simulation. Another approach is the statistic eye method [17], developed by the PLL working group within the OIF. One methodology for serial data transmission over high-speed physical layer is proposed in this thesis. The MATLAB simulator for the optimization of a pre-emphasis FIR filter is designed and its flow chart is shown in Fig. 1.2. It includes two parts: FIR optimization and front-end link simulation. Each part will be discussed in detail in Chapter 4. The only way to verify the simulator is to use a silicon design and compare performance. For that purpose, a 90nm CMOS design is undertaken in this thesis.

Fig. 1.2 MATLAB simulation flow chart for pre-emphasis optimization

1.4 MATLAB Program Features and Thesis Contribution

The MATLAB program in the thesis can emulate the complete link simulation from transmit end to receive end, including frequency response, impulse response of the FIR filter, characteristics of channel, eye opening readout, and cross-talk effect. A user-friendly interface is provided. The simulation main features include:

- User selection of pre-tap and/or post-tap and number of taps;
- User selection of time and/or frequency domain presentation of the characteristics of channel and pre-emphasis;
- User adjustment of the data transmission speed;
- User provision of the specific channel of interest by selecting the measured S-parameters and measurement types;

This MATLAB program has six sub-programs according to the functions they target. These sub-programs are: FIR tap coefficients calculator, link simulator, eye opening readout, bathtub plot for bit-error rate (BER) estimation, cross-talk effect estimation, and incorporation of RC equalizer. In general with this program developed, guidance can be provided for the design of an FIR filter acting as pre-emphasis. Given a specific backplane channel, users may calculate the optimal pre-emphasis parameters that could then be used as preliminary data for their applications.

Besides solving a practical industry related problem, there are significant contributions in this thesis. There are:

- An easier approach for complete front-end backplane simulation;
- The concept of optimal number of taps for the FIR filter as pre-emphasis;
- Clarification of pre-tap effect on signal integrity;

- Effect of cross-talk on the backplane performance that can be simulated more easily and its effect can be estimated;
- Circuit design of programmable pre-emphasis to verify the MATLAB programming;
- Effect of group delay distortion on inducing ISI.

1.5 Thesis Approach

The organization of this thesis is as follows: Chapter Two introduces the backplane network of interest. Obviously, understanding its characteristics would be beneficial for pre-emphasis design. The FIR filter as a pre-emphasis technique is studied in Chapter Three. MATLAB scripts are described in Chapter Four. The program flow charts are given and a list of program functions are described in detail. Chapter Five shows the simulation results of the pre-emphasis solution to a specific backplane channel. The verification of the MATLAB simulator using HSPICE for MATLAB program is given in Chapter Six. Conclusions and future work are listed in Chapter Seven. The MATLAB source code is provided in the Appendix.

CHAPTER 2 LITERATURE REVIEW

High-speed serial communications have dealt with the interface physical layer for years. This chapter first introduces the characteristics of the high-speed interface physical layer. Then, the definition and properties of a pre-emphasis FIR filter to compensate for the channel loss is given. An adaptive algorithm, which is employed by the MATLAB program to optimize FIR tap coefficients, is also provided.

2.1 High-speed Interface Physical Layer

Fig. 2.1 shows an example of the configuration of a backplane transceiver link. It typically consists of a transmit daughter card, transmission lines, a receive daughter card and connectors. Transmission lines are routinely used for high bandwidth communication and data transfer. They come in a wide variety of geometries and sizes and operate over enormous frequency ranges. The transmission lines discussed here are typically simple PCB traces with strip or micro-strip geometries. The bandwidths of interest will be from DC to a few GHz.

Fig. 2.1 Configuration of backplane network

One key problem, which applies to all transmission mediums, but particularly related to PCB layout, is frequency dependent transmission losses, caused primarily by the skin effect of the series resistive component of copper, and the dielectric loss due to the parallel conductive component of the dielectric.

2.1.1 Skin Effect

The skin effect describes how high-frequency components tend to travel on the surface of the conductor, rather than the whole cross section of the conductor. This is caused by self-inductance of the material, which increases the inductive reactance with frequency, forcing the high frequency components to travel on the surface of the material. This reduces the effective conductive area of the trace, increasing resistance, which causes the signal to be attenuated.

Higher frequency signals propagate closer to the surface of the conductor; therefore, the resulting signal conducts within a limited depth, the skin depth, on the conductor surface, which is defined as in Eq.2.1 [20]

$$\delta = 1 / \sqrt{\frac{\pi \cdot \mu_0}{\rho} \cdot f} \quad (2.1)$$

where ρ is the resistivity of the conductor, μ_0 is the permeability of free space, and f is the frequency of the signal. Hence, the effective series resistance (R_{skin}) corresponding to this depth increases with the square root of frequency, as shown in Eq.2.2 [21]:

$$R_{skin} = R_{DC} + \frac{1}{2 \cdot r} \cdot \sqrt{\frac{\rho \cdot \mu_0}{\pi} \cdot f} \quad (2.2)$$

for a round conductor with radius r , as an example, where R_{DC} is the DC resistance at frequency equal to zero. This relation implies a frequency dependent resistive loss. For low-frequency components, the current density is nearly uniform throughout the cross section of the conductor, and the wire resistance is close to its DC resistance.

2.1.2 Dielectric Loss

All PCB materials have a specific dielectric constant, which will affect the impedance of the transmission line. The value for dielectric constant can be determined by comparing its effect on the capacitance of a conductor pair to that of a conductive pair in a vacuum. A lower dielectric constant signifies that the material will support a longer transmission line before issues occur. Some material such as FR4 will see its conductivity vary with frequency. With some insulating materials, dielectric absorption causes frequency dependent attenuation. This loss can be modeled as a conductance between the signal wire and ground.

Using a low-loss dielectric material can mitigate this effect. However, because of certain restrictions on PCB materials, the choice of dielectric is limited; thus PCB traces usually demonstrate a considerably higher dielectric loss compared to cables. Dielectric loss for each material is usually expressed in terms of a parameter, called the loss tangent defined as in Eq.2.3 [20]:

$$\tan \sigma_D = \frac{G}{\omega \cdot C} \quad (2.3)$$

where C is the capacitance per unit length. This quantity is approximately constant with frequency, therefore the conductance G between the signal wire and ground, typically increases linearly with frequency, ω .

FR4 has high loss tangent, which results in a large attenuation of the signal at high frequency. Many materials are used as PCB laminates and have both better dielectric constraints and loss tangents; however, in volume, applications become prohibitive due to cost. Table 2.1 discusses dielectric performance of various laminates and relative material costs [22].

Table 2.1 Representative comparisons of dielectric performance and cost

Material	Dielectric Constant	Loss Tangent Value	Relative Cost to FR4	Manufacturing
FR4	3.9-4.7	0.02-0.03	1X	Standard processing
GETEK	3.5-4.3	0.012	1.5X	Standard processing
Polyimide	4.0-4.5	0.01	3X	Hard processing
Speedboard N	3.0	0.02	5X	Extra Processing
Rogers (RO3003)	3.0	0.001	6.5X	Hard processing

2.1.3 Frequency Response of the Channel

Based on equations 2.1 to 2.3, the approximate frequency response of magnitude in dB accounting for both the skin effect and the dielectric absorption of the channel is given by Eq.2.4 [20]:

$$H(f, l)|_{dB} = -(h_s \cdot \sqrt{f} + h_d \cdot f) \cdot l \quad (2.4)$$

where l is the length of the channel, h_s and h_d are the skin-effect and dielectric loss coefficients respectively.

For most PCB materials h_s is marginally larger than h_d . Fig. 2.2 shows the magnitude attenuation vs. frequency of a 34" FR4 backplane. The effect is similar to a low pass filter, which decreases the amplitude of the high frequency components. The

main effects are dielectric loss and skin effect, although cross talk and stub reflections caused by poor termination can also cause issues. However, these issues are trivial and in most cases can be ignored for an FR4 PCB fabric.

Fig. 2.2 Magnitude attenuation of a 34'' FR4 backplane

2.1.4 Impulse/Pulse Response of Channel

Fig. 2.3(a) shows the impulse response of a 34'' FR4 backplane. It is the inverse Fourier transform of the frequency response of the channel. The normalized pulse responses of this channel to an ideal trapezoidal pulse with period of 320psec (3.125Gbps), 160psec (6.25Gbps), and 100psec (10Gbps) are shown in Fig. 2.3(b).

(a)

(b)

Fig. 2.3 (a) Impulse response of a 34'' FR4 backplane, and (b) pulse responses at different pulse width

The pulse suffers more attenuation of its high-frequency components (transitions) than that of its low-frequency components. The attenuation results in the long settling tail exhibited in Fig. 2.3. This long tail induces ISI that corrupts the future transmitted symbols and reduces their effective width and amplitude. This high-frequency attenuation makes it difficult, in high-speed data transmission along PCB traces, for the receiver to interpret the information. The attenuation prevents the signal from reaching its full strength within its symbol time, causing it to spread into the next signal.

2.1.5 Group Delay Effect

A further complication is caused by phase dispersion, where the losses in the medium cause the different frequency components of the signal to have different travel speed as they propagate along the transmission line. This effect further reduces the signal amplitude and adds residual error from the previous bit leading to increased ISI. Group delay is the derivative of phase to frequency (Eq.2.5).

$$GD = -\frac{d\phi}{d\omega} \quad (2.5)$$

As phase is nonlinear to frequency, group delay implies the phase dispersion.

Fig. 2.4(a) shows the group delay distortion of a 34" FR4 backplane and Fig. 2.4(b) shows its impulse response with and without group delay distortion.

(a)

(b)

Fig. 2.4 (a) Group delay distortion of a 34'' FR4 backplane and (b) its impulse response with/without group delay distortion

It can be seen that without group delay distortion, the impulse response is attenuated and symmetrically dispersed. However, the impulse response becomes asymmetrical due to group delay distortion. The resulting long tail of impulse response causes more severe post-cursor ISI, degrading the eye opening as shown in Fig. 2.5. Therefore, group delay distortion should be taken into account in channel modeling and choosing equalization techniques.

(a)

(b)

Fig. 2.5 Far-end eye diagrams (a) with and (b) without group delay distortion for a 34'' FR4 backplane at 3.125Gbps.

2.1.6 S-parameters

Backplane channel characteristics can be fully described by the measured differential S-parameters [23, 24]. Since this thesis is dealing with a backplane network, whose characteristics are based on measured S-parameters, an explanation of S-parameters should be given here.

For a network with balanced ports, each port consists of a pair of terminals. By taking a ratio of the normalized response and incident power waves, a set of mixed-mode S-parameters can be defined. For the purposes of analysis, a backplane system can be modelled as a mixed-mode two-port network. With mixed-mode S-parameters, a differential-mode stimulus is presented on the differential port terminal pair and the corresponding differential- and common-mode responses are measured on all of the device ports. Likewise a common-mode stimulus is presented to the same port and the differential- and common-mode responses again are measured. Conceptually the 4x4 matrix shown in Fig. 2.6 can be sub-divided into four quadrants that symbolize four separate modes of operation: DD, CC, CD, and DC.

Fig. 2.6 Mixed-mode S-parameters matrix

Each quadrant gives the input and output reflection characteristics and the forward and reverse transmission characteristics for that mode. For example, the

differential-mode (DD) quadrant, in the upper-left corner of the mixed-mode S-parameter matrix, describes the behavior of the circuit with a differential stimulus and differential response. In practice, there will be some amount of mode conversion. The more mode conversion, from differential-mode to common-mode or common-mode to differential-mode, that exists, the more likely there will be EMI radiation from the system or undesired ground loops.

In this thesis, the S-parameter is created in ADS by mimicking the published data of a FR4 backplane - Tyco XAUI 34" FR4 backplane [34]. This backplane is used as working subject.

2.2 The FIR Filter for Pre-emphasis

The structure of an FIR filter is shown in Fig. 2.7. For a symbol-spaced FIR (SSF) filter, the delay D is equal to one symbol period T . For a fractionally spaced FIR (FSF) filter, D is equal to a fraction of T , $D = mT/n$ (m and n are integers, and $m < n$).

Fig. 2.7 FIR filter structure

The transfer function of an FIR filter is usually written in the z -domain as in Eq.2.6 [25]

$$H(z) = \sum_{n=0}^{N-1} c_n \cdot z^{-n} \quad (2.6)$$

where c_n are the tap coefficients and

$$z = e^{j2\pi f / f_s}, \quad (2.7)$$

The sampling frequency is $f_s = \frac{1}{T}$ for SSF, and $f_s = \frac{n}{mT}$ for FSF. To illustrate the frequency response of an FIR filter, a single two-tap FIR filter is given as example. Its formula is:

$$H_1(z) = 1 - az^{-1} \quad (2.8a)$$

For 10Gbps data transmission over a backplane, the SSF has sampling frequency 10GHz. Fig. 2.8(a) shows the magnitude and group delay of Eq.2.8a, with a swept from 0.1 to 0.5.

Fig. 2.8 Frequency response of (a) $H_1(z) = 1 - az^{-1}$ and (b) $H_2(z) = 1 - az^{+1}$

It can be seen from the magnitude vs. frequency plot that the high-frequency part is boosted to compensate for channel amplitude attenuation at higher-frequency. Group delay of this FIR filter has an inverse sign to the channel group delay. Another similar SSF formula is given in Eq.2.8b

$$H_2(z) = 1 - az^{+1} \quad (2.8b)$$

Its frequency response is shown in Fig. 2.8(b). Comparing Fig. 2.8(b) with Fig. 2.8(a) shows that H_1 and H_2 have exactly the same frequency response in magnitude, and opposite frequency response in group delay. Simulation results have proven that H_2 introduces more jitter at far end of channel while H_1 can reduce the jitter. This is because, although H_2 is effective in vertical eye opening, it may deteriorate horizontal eye opening by introducing more severe group delay distortion.

The one post-tap symbol-spaced FIR filter (H_1) is commonly used as pre-emphasis in industry because of its simplicity in implementation. Writing $H_1(z) = Y(z)/X(z)$ in time domain gives

$$y(n) = x(n) - ax(n-1) \quad (2.9)$$

Fig. 2.9(a) shows the structure of the one post-tap FIR filter, and its implementation is realized through a current-mode logic (CML) circuit in Fig. 2.9(b).

Fig. 2.9 (a) Structure of one post-tap symbol-spaced FIR filter, and (b) its circuit implementation

The steering currents I_{main} and I_{post} in Fig. 2.9(b) are related to a in Fig. 2.9(a) by the formula:

$$a = \frac{I_{post}}{I_{main}} \quad (2.10)$$

The input and output signals of this FIR filter are shown in Fig. 2.10.

**Fig. 2.10 Input and output signal of one post-tap SSF
(R is the load resistance.)**

The industry pre-emphasis definition as a percentage is $a \times 100\%$. Another definition written in dB in the frequency domain is $20 \log_{10} \frac{1+a}{1-a}$. The latter one explicitly gives the amplitude boost in dB from DC to the highest amplitude, which occurs at half the sampling frequency ($f = f_s/2$).

The performance of a symbol-spaced FIR (SSF) filter is limited by aliasing, as a result of sampling at $1/T$. It cannot compensate for channel impairments beyond the sampling frequency $f_s = 1/T$. However, a fractionally spaced FIR (FSF) filter can sample the input signal at a fraction of T and thus extends the compensation beyond $1/T$ [26]. A single post-tap, one-half symbol ($T/2$) spaced FIR filter is given as an example. Its expression in the z-domain is the same as Eq.2.8a, and its expression in time domain is the same as Eq.2.9. Eq.2.8a for the SSF is periodic with a period of $1/T$,

and symmetric over $1/2T$ within one period in frequency, however, Eq.2.8a for the FSF is periodic with the period of $2/T$ and symmetric over $1/T$ within one period. This means that the FSF filter used for pre-emphasis can compensate for the channel loss to a larger extent. It is expected that the FSF pre-emphasis can perform better than SSF.

In order to optimize the tap coefficients of an FIR filter, the background of adaptive filtering and adaptive algorithms is needed.

2.3 Adaptive Filtering and Algorithm

2.3.1 The Adaptive Filter

In order to optimize the tap coefficients of the designed filter for pre-emphasis, an adaptive algorithm is required to make the filter adaptive to the channel characteristics and the specific user requirements. Fig. 2.11 shows the configuration [27, 28].

Fig. 2.11 Configuration of adaptive filter

The characteristics of the adaptive filter include:

- automatic adjustment (or adaptation) in the face of changing environments and changing system requirements;
- trainable to perform specific filtering or decision-making tasks;
- an “adaptive algorithm” is required for adjusting the system parameters.

One adaptation approach is based on the Wiener filter theory. In this theory, the optimal coefficients of a linear filter are obtained by minimizing its mean-square error (MSE). Another adaptation approach is the method of least squares. Its performance index is the sum of weighted error squares. In choosing a performance function the following points have to be considered:

- The performance function must be mathematically tractable;
- The performance function should preferably have a single minimum so that the optimum set of the filter parameters could be selected unambiguously.

The number of minima points for a performance function is closely related to the filter structure. The recursive (IIR) filters, in general, result in a performance function that may have many minima, whereas the non-recursive (FIR) filters are guaranteed to have a single global minimum point if a proper performance function is used.

With $x(n)$ as the input, the objective of the adaptive algorithm is to set the FIR filter tap coefficients in such a way that its output, $y(n)$, tries to minimize the performance function $F(e(n))$, which involves the reference signal (or desired signal), $d(n)$ by $e(n) = d(n) - y(n)$. The definitions of performance functions are listed as follows:

- Mean-square error $F(e(n)) = E[|e(n)|^2]$; (2.11)

- Least-square error $F(e(n)) = \frac{1}{n+1} \sum_{i=0}^n |e(n-i)|^2$; (2.12)

- Weighted least square $F(e(n)) = \sum_{i=0}^n \lambda^i |e(n-i)|^2$ ($\lambda < 1$); (2.13)

- Instantaneous square value $F(e(n)) = |e(n)|^2$. (2.14)

2.3.2 Wiener Filter

Wiener filters are a class of optimum, transversal (FIR) linear filters, which involve linear estimation of a desired signal sequence from another related sequence. The Wiener filter employs “mean-square error criterion” as a performance function. Fig. 2.12 shows the structure of a Wiener filter with tap coefficients $W_0, W_1, \dots, W_{N-1}$.

Fig. 2.12 Structure of Wiener filter

If the tap coefficients and input signal are written in the form of vectors,

$$\vec{W} = [W_0, W_1, \dots, W_{N-1}]^T, \quad (2.15)$$

and

$$\vec{X} = [x_n, x_{n-1}, \dots, x_{n-N+1}]^T, \quad (2.16)$$

then the current output signal is

$$y(n) = \sum_{i=0}^{N-1} w_i x(n-i) = \vec{W}^T \vec{X}(n). \quad (2.17)$$

Thus the current error signal is

$$e(n) = d(n) - y(n) = d(n) - \vec{W}^T \vec{X}(n), \quad (2.18)$$

with n being the current iteration number. Substitute Eq.2.18 into Eq.2.11 to calculate the performance function or cost function,

$$F = E[e^2(n)] = E[d^2(n)] - 2\vec{W}^T \vec{P} + \vec{W}^T R \vec{W}, \quad (2.19)$$

where $\tilde{P}$ is the defined $N \times 1$ cross-correlation vector,

$$\tilde{P} \equiv E[\tilde{X}(n)d(n)], \quad (2.20)$$

and R is the defined $N \times N$ autocorrelation matrix,

$$R \equiv E[\tilde{X}(n)\tilde{X}^T(n)] \quad (2.21)$$

To obtain the set of tap weights that minimize the performance function, we can set

$$\nabla F = \tilde{0}, \quad (2.22)$$

where ∇ is the gradient vector defined as the column vector $\nabla = [\frac{\partial}{\partial w_0}, \frac{\partial}{\partial w_1}, \dots, \frac{\partial}{\partial w_{N-1}}]^T$,

and zero vector $\tilde{0}$ is defined as N -component vector $\tilde{0} = [0, 0, \dots, 0]^T$. The solution to Eq.2.22, known as the Wiener-Hopf equation, is

$$R\tilde{W}_{op} = \tilde{P} \quad (2.23)$$

where $\tilde{W}_{op}$ is the optimum tap coefficients vector. Besides solving the Wiener-Hopf equation to get the optimum tap coefficients of a transversal (FIR) Wiener filter, an alternative way of finding the optimum tap coefficients is to use an iterative search algorithm that starts at some arbitrary initial point in the tap coefficient vector space and progressively moves, in steps, toward the optimum point in steps.

2.3.4 The Adaptive Algorithm

There are many iterative search algorithms derived for minimizing the underlying cost function with the true statistics replaced by their estimate. The method of steepest descent is one of the gradient-based iterative methods. From Eq.2.19, the gradient of F is given by

$$\nabla F = 2R\tilde{W} - 2\tilde{P} \quad (2.24)$$

With an initial guess of $\vec{W}_{op}$ at $n = 0$, the tap coefficients vector at the k -th iteration is denoted as $\vec{W}(k)$. The following recursive equation may be used to update $\vec{W}(k)$:

$$\vec{W}(k+1) = \vec{W}(k) - \mu \nabla_k F \quad (2.25)$$

where $\mu > 0$ is called the step size, $\nabla_k F$ denotes the gradient vector ∇F evaluated at the point $\vec{W} = \vec{W}(k)$. Substituting Eq.2.24 into Eq.2.25 to obtain

$$\vec{W}(k+1) = \vec{W}(k) - 2\mu(R\vec{W}(k) - \vec{P}(k)). \quad (2.26)$$

The convergence of $\vec{W}(k)$ to the optimum solution and the convergence speed are dependent on the step-size parameter μ .

The least-mean square (LMS) algorithm changes (adapts) the filter tap coefficients so that $e(n)$ is minimized in the mean-square sense. Without explicitly solving the Wiener-Hopf equation, the LMS algorithm is a practical scheme for realizing Wiener filters. The conventional LMS algorithm is a stochastic implementation of the steepest descent algorithm. It simply replaces the cost function F by its instantaneous coarse estimate

$$\hat{F} = e^2(n) \quad (2.27)$$

Substituting $\hat{F} = e^2(n)$ for F in the steepest descent recursion to get

$$\vec{W}(n+1) = \vec{W}(n) - \mu \nabla e^2(n) \quad (2.28)$$

where $\vec{W}(n) = [w_0(n), w_1(n), \dots, w_{N-1}(n)]^T$. Note that the i -th element of the gradient vector $\nabla e^2(n)$ is

$$\frac{\partial e^2(n)}{\partial w_i} = 2e(n) \frac{\partial e(n)}{\partial w_i} = -2e(n) \frac{\partial y(n)}{\partial w_i} = -2e(n)x(n-i), \quad (2.29)$$

then,

$$\nabla e^2(n) = -2e(n)\vec{X}(n) . \quad (2.30)$$

Substituting Eq.2.30 into Eq.2.28 to obtain

$$\vec{W}(n+1) = \vec{W}(n) + 2\mu e(n)\vec{X}(n) \quad (2.31)$$

Eq.2.31 is referred to as the LMS recursion.

The advantages of LMS algorithm are:

- Simplicity in implementation;
- Stable and robust performance against different signal conditions;

while the main disadvantage of the LMS scheme is:

- Slow convergence.

CHAPTER 3

A TOP-DOWN DESIGN METHODOLOGY FOR SERDES

A design methodology for a high-speed backplane serializer/deserializer (SERDES) is proposed in this thesis, as shown in Fig. 3.1(a). Its top-down design flow chart is shown in Fig. 3.1(b).

(a)

(b)

Fig. 3.1 (a) Proposed design methodology, and (b) its top-down flow chart

The characteristics of the high-speed physical layer are obtained from measured s-parameters. Optimization of an FIR filter for pre-emphasis is done through MATLAB. Guidance is provided for implementation and evaluation of the FIR filter at the transistor level. A DFE can be added at far end of channel. Its output goes to CDR for clock and data recovery. Both the DFE equalizer and the CDR can be simulated in MATLAB or through behavioral modeling. Design guidance is then provided for the implementation at transistor level. There are programs for eye opening readout and/or BER estimation for evaluation purposes: 1) at the far end of channel, 2) at the output of DFE equalizer and 3) at the output of CDR.

In this thesis, the design at the near-end of the channel, both in MATLAB and at the transistor level, is fulfilled. Due to some limitations, other blocks will hopefully be finished as a part of my future work. For the MATLAB simulator, six functions are programmed. These are: *FIR tap coefficients calculator*, *link simulator*, *amplitude and jitter reader*, *bathtub plot for bit-error rate (BER) estimation*, *cross-talk effect simulation*, and *incorporation of RC equalizer*. In the following sections, each part will be introduced in detail.

3.1 Tap Coefficients Calculator

The tap coefficients calculator is based on the Least Mean Square (LMS) algorithm to optimize the tap coefficients for the FIR filter. These coefficients are used for transmitter pre-emphasis setting. The convergence equation for the tap coefficients is:

$$\vec{W}(n+1) = \vec{W}(n) + \mu e(n) \vec{X}(n) \quad (3.1)$$

The variables will be explained later.

The programming flow chart for the tap coefficient calculator is shown in Fig. 3.2.

Fig. 3.2 Programming Flow chart of tap coefficients calculator

The channel information, including amplitude attenuation and group delay distortion, is stored in the measured S-parameters. For differential signals, differential-mode forward parameters S_{DD21} are imported. Through the inverse fast Fourier transform (IFFT), the impulse response of the channel is obtained. The impulse response is convolved with the training data to produce the output signal at far end of the channel. The output signal, which is $\vec{X}(n)$ in Eq.3.1, is used as input signals to LMS convergence block. The other input signal to LMS algorithm block is the error signal, $e(n)$ in Eq.3.1. $e(n)$ is the difference between desired signal $d(n)$ and equalized signal $y(n)$, which is obtained through the updated FIR filter inside the LMS algorithm block at each iteration. Thus, $e(n)$ can be written as

$$e(n) = d(n) - y(n), \quad (3.2)$$

where

$$y(n) = \sum_{i=0}^{N-1} w_i x(n-i) = \vec{W}^T \vec{X}(n), \quad (3.3)$$

w_i are the tap coefficients and n is the iteration step.

There is a time delay between the signal at the near end and the far end of channel. The point, at which the impulse response amplitude reaches its maximum, can determine this delay. It is approximately the average of the group delay of the channel.

The block, labeled User Inputs in Fig. 3.2, receives the specific user requirements. These requirements include data rate, channel characteristics (S-parameters), pre-tap and post-tap number and so on. Users also can decide the number of iteration. The error convergence, the impulse response, and the frequency response of the channel and the FIR filter can be chosen, by the users, for visual presentation.

Example 1: 3.125Gbps data transmission over 34" FR4 backplane. Number of post-tap = 4 and number of pre-tap = 0. A symbol-spaced FIR filter is employed for pre-emphasis. The tap coefficients are normalized with first tap set to one. Fig. 3.3 shows the optimum tap coefficients.

Fig. 3.3 Optimized tap coefficients in Example 1

3.2 Link Simulator

Once the tap coefficients are obtained, the FIR filter for pre-emphasis is also determined. Fig. 3.4 shows the programming flow chart of the link simulator, which can

be used to emulate the complete process of data transmission over the backplane with pre-emphasis included.

Fig. 3.4 Programming Flow chart of link simulator

A z-transform is applied to the tap coefficients to first obtain the frequency-domain form. The PRBS data is transformed to its frequency-domain form by the FFT. From the measured S-parameters (S_{DD21}), the channel's frequency response has been stored in the frequency domain. These three sets are then multiplied together. The result is the power spectrum of the output signal at far end of channel. After the IFFT, the output signal in the time domain is obtained and an eye diagram can be drawn.

Example 2: Fig. 3.5 shows the frequency response of channel, of the pre-emphasis (from Example 1) and their sum.

Fig. 3.5 Frequency response of channel, pre-emphasis and sum of them

Fig. 3.6(a) shows the eye diagram at far end of channel without pre-emphasis, compared with Fig. 3.6(b) after pre-emphasis is employed.

Fig. 3.6 Far-end eye diagram (a) without pre-emphasis, (b) with pre-emphasis.

3.3 Eye Opening Readout

Once eye diagrams are drawn, the problem of how to accurately read the eye opening is considered. The eye opening denotes amplitude in the vertical direction and jitter in the horizontal direction. These quantities must be read out, as shown in Fig. 3.7.

Fig. 3.7 Vertical (amplitude) and horizontal (jitter) eye opening

The programming flow chart for reading amplitude and jitter are in Fig. 3.8(a) and Fig. 3.8(b) respectively.

(a)

Fig. 3.8 Programming flow chart for readout of (a) amplitude and (b) jitter.

The parameters in Fig. 3.8 are listed and explained in the following table.

Table 3.1 Parameters in Fig. 3.8

N	Integers 1, 2, ...
Num	Sampling points within one symbol period.
SamT	Time interval between two sampling points.

The principle of amplitude readout resides in that for differential output, within one eye, minimum positive value and maximum negative value are calculated separately. The calculation is based on that for three consecutive points, the second one has the largest or smallest value among them. When iteratively drawing lines within one eye, all these points are stored. After eye diagram is drawn, the maximum and minimum points are singled out. Amplitude is the difference between these two points.

The flow chart for jitter readout is for a data pattern with one eye. A similar calculation can be obtained for the data pattern with two eyes. Generally the two eye-diagrams are not symmetrical due to duty-cycle distortion. The principle of the jitter readout is that for differential signals, the position of the jitter is labeled at the zero-crossing point. All zero-crossing points are stored and assigned for eye(s). After the eye diagram is drawn, the minimum and maximum zero-crossing points are found within one eye. The difference is the jitter.

Example 3: Continue Example 1 and 2. One-eye diagram is drawn in Fig. 3.9. The results are: vertical eye opening = 0.47 (normalized to peak-to-peak input signal amplitude), and horizontal eye opening = $1.00\text{UI} - 0.04\text{UI} = 0.96\text{UI}$.

Fig. 3.9 Eye diagram in Example 3

3.4 BER Estimation

Jitter due to inter-symbol interference (ISI) is a major part of deterministic jitter (DJ). The ISI jitter is caused by the time differences of the signal arriving at the receiver threshold when starting from different places in bit sequences (symbols). ISI is also expected whenever any bit sequence has frequency components that are propagated at different rates by the transmission media (dispersion occurs).

The following equation relates the bit-error-rate (BER) to the jitter:

$$BER(\tau, W, \sigma) = \frac{1}{4\sqrt{\pi}} \left[\operatorname{erfc}\left(\frac{\tau - \frac{W}{2}}{\sqrt{2}\sigma}\right) + \operatorname{erfc}\left(\frac{\tau + \frac{W}{2}}{\sqrt{2}\sigma}\right) + \operatorname{erfc}\left(-\frac{\tau - 1 - \frac{W}{2}}{\sqrt{2}\sigma}\right) + \operatorname{erfc}\left(-\frac{\tau - 1 + \frac{W}{2}}{\sqrt{2}\sigma}\right) \right] \quad (3.4)$$

where W is deterministic jitter, σ is random jitter, and τ is the sampling point at receive side. All variables are in unit interval (UI) unit. So far, only ISI DJ is dealt with. However, it is still a very good prediction about BER from calculated ISI.

Example 4: With Eq.3.4, bathtub plot is drawn using MATLAB for Example 3. The read ISI jitter is used as DJ, and random jitter (RJ) is swept from 0 to 0.04UI with 0.01UI as step. Fig. 3.10 shows this bathtub plot.

Fig. 3.10 Bathtub plot for Example 3

3.5 Cross-talk Effect

Cross talk occurs when a victim line (a trace on a circuit board or two adjacent wires in a cable) is affected by the magnetic field from a driver (aggressor) line, as shown in Fig. 3.11. The incremental inductance of a victim conductor converts an induced magnetic field into an induced current. The induced current adds (positively or negatively) to the current in a victim line, thereby increasing or decreasing voltage. This causes jitter.

Fig. 3.11 Cross talk example

The programming flow chart for the cross talk effect is shown in Fig. 3.12. With this programming, the cross talk effect on the LMS outputs (tap coefficients) are calculated in quantity. In Fig. 3.12, there are three switches 1, 2, and 3. Their values “Y” or “N” are explained in Table 3.2.

Table 3.2 Switches explanation in programming flow chart for cross talk

Switch 1	Switch 2	Switch 3	Function
N	-	-	There is no cross talk from neighboring channels.
Y	N	Y	The Data Source X in neighboring channel is delayed. The delayed time in UI is user input.
Y	Y	N	The Data Source X in neighboring channel has a different data rate.
Y	Y	Y	If delayed time is 0, and data rate is unchanged, Data Source X is in phase with Data.

The data transmitted in the neighboring channel can be delayed or be at a in different rate. Both can be implemented by manipulating the sampling points within one symbol period. Due to FR4 fabrication, the cross-talk effects are minimized. However, the cross talk mainly occurs within the backplane connector because of pin density and routing complexity. Only the case of delaying neighboring data within one UI is studied. The results of cross-talk effect on LMS outputs are shown in next chapter.

Fig. 3.12 Programming flow chart for cross-talk effect.

3.6 Modeling of RC Equalizer

The traditional transistor source degenerated resistor/capacitor (RC) equalizer can be added at the receive end of channel for higher gain boost requirement. The values of R and C are programmable. At a set of fixed R and C values, the frequency response of the equalizer can be obtained by running a HSPICE simulation. The frequency responses in the forms of magnitude/phase or real/imaginary are then imported into MATLAB simulator. Thus, one block, which has the same characteristics of RC equalizer, is set up in MATLAB program. Since the link simulator as explained in 3.2 is also processed in frequency domain, in this way, the equalizer is combined with the pre-emphasis FIR filter in the same domain.

The implementation of a traditional RC equalizer is shown in Fig. 3.13(a). Since this kind of equalizer obtains a gain boost from DC to the high frequency of interest by decreasing the DC level, one limiter is needed for the equalizer output signal. A CML structured limiter is proven to satisfy a high gain bandwidth product for bandwidth of interest (around 7.5GHz). Its implementation is shown in Fig. 3.13(b). Simulation results will be given in section 4.4.

Fig. 3.13 (a) Implementation of RC equalizer; (b) implementation of CML structured limiter

3.7 User Interface

There is an user interface in the MATLAB program. Users are prompted to provide information to the program. Understanding these prompt questions and choices is the convenient way for users to utilize the MATLAB program to the best. In this section, these prompt questions and choices are explained in detail. An example is given for clarity.

The MATLAB user interface is:

- *Measurement network format: 1 – four-port; 0 – two-port*
- *Port number arrangement: 1 – input port # 1, 2 and output port # 3, 4; 0 – input port # 1, 3 and output port # 2, 4*
- *S-parameters full file name:*
- *S-parameters frequency unit: 1 – MHz; 0 – Hz*
- *S-parameters data format: 2 – complex; 1 – linpolar; 0 – dBpolar*

- *Data baud rate of interest (unit Gbps):*
- *Number of iterations (default 3000):*
- *Number of pre-tap of the pre-emphasis FIR filter:*
- *Number of post-tap of the pre-emphasis FIR filter:*
- *Are you interested in watching the LMS error convergence? 1-YES, 0-NO*
- *Are you interested in watching the frequency response and impulse response? 1-YES, 0-NO*

Next, each prompt question and choice are explained:

- *Measurement network format: 1 – four-port; 0 – two-port*

For unbalanced backplane networks, there are two common measuring single-ended formats: four-port network and two-port network. In Fig. 3.14, these two formats are depicted.

Fig. 3.14 Single-ended network formats: (a) two-port; (b) four-port.

- *Port number arrangement: 1 – input port # 1, 2 and output port # 3, 4; 0 – input port # 1, 3 and output port # 2, 4*

If the user chooses a four-port network, this choice question is prompted to ask for choice. According to port number arrangement when measuring the s-parameters, there are generally two types as depicted in Fig. 3.15.

Fig. 3.15 Four-port networks: (a) input port number 1&2, output port number 3&4; (b) input port number 1&3, output port number 2&4.

- *S-parameters full file name:*

If program and file are in the same directory, user only inputs the full file name; otherwise, user needs to input the path name and file name.

- *S-parameters frequency unit: 1 – MHz; 0 – Hz*

When s-parameter data are measured, the frequency unit is often in MHz or Hz. Users should check the choice of frequency unit for their measured s-parameters.

- *S-parameters data format: 2 – complex; 1 – linpolar; 0 – dBpolar*

The S-parameter data are often measured in three formats. The complex format includes real-part and imaginary part; the linpolar format means the magnitude and phase in angle; and the dBpolar format is magnitude in dB and phase in rad.

- *Data baud rate of interest (unit Gbps):*

User inputs the data baud rate of interest. Unit has been set as Gbps.

- *Number of iterations (default 3000):*

User should give the number of iterations number for the purpose of LMS convergence. The user can input this default number and choose to watch the LMS error convergence in a later step. If the default number is not enough for the convergence process, the user can increase the number.

- Number of pre-tap for pre-emphasis FIR filter:

User should input the pre-tap number of FIR filter as pre-emphasis. This number can vary from 0 to tens. Its range is large enough to cover the pre-cursor of the impulse response of a channel.

- Number of post-tap for pre-emphasis FIR filter:

User should input the number of post-tap for pre-emphasis FIR filter. This number can vary from 0 to tens. Its range is large enough to cover the post-cursor of the impulse response of a channel.

- Are you interested in watching the LMS error convergence? 1-YES, 0-NO

This question allows the user to observe the status of the error convergence.

- Are you interested in watching the frequency response and impulse response? 1-YES, 0-NO

If choice 1 is checked, the magnitude frequency response in dB of channel, pre-emphasis FIR filter, and sum of both are depicted in one figure. In another figure, the impulse response of the channel and impulse response with pre-emphasis employed are depicted. All are shown when the program is finished running.

An example is the best way for explanation. For each step above, the following answer or choice is made:

- 1; (four-port backplane network)
- 0; (input port # 1, 3 and output port # 2, 4)
- thru_bpa.s4p; (S-parameter file name, in current directory)
- 1; (frequency unit in MHz)
- 2; (S-parameter data format is in complex)
- 6.25; (data baud rate is 6.25Gbps)
- 3000; (LMS iteration number is default number)
- 0; (FIR filter as pre-emphasis has zero pre-tap)
- 4; (FIR filter as pre-emphasis has four post-taps)
- 1; (Interested in watching the LMS error convergence)
- 1; (Interested in watching the frequency response and impulse response)

In a UNIX system, when running program, the following results are shown at prompt sign.

- $\mu_{\text{uon}} = 0.0012$
- $\sigma = 0.0358$
- $\text{coeff.} = 1.0000 \quad -0.4943 \quad 0.0637 \quad -0.0600 \quad -0.0062$
- Drawing eye diagram with obtained tap coefficients ...
- Horizontal = 0.9123
- Vertical = 0.2215

Muon and sigma are mean and standard deviation of error signal, which is the difference between the desired signal and the output of the LMS convergence engine. Tap coefficients are shown and the eye openings in horizontal (UI) and vertical (normalized) directions are read out.

Also, there are four plots are depicted. They are in Fig. 3.16.

Fig. 3.16 (a) Eye diagram at far-end of channel; (b) error convergence; (c) impulse response; (d) frequency response.

CHAPTER 4 SIMULATION RESULTS AND ANALYSIS

Based on MATLAB programming, this part will give simulation results and analyses, including the optimal post-tap number, optimal tap coefficients, pre-tap effect, as well as cross talk effect on tap coefficients. Horizontal eye opening and normalized vertical eye opening are read out. A 34" (30" trace and 2*2" connector) FR4 backplane channel is used as working subject. From the measured S-parameters (both channel and cross talk), its characteristics are shown in Fig. 4.1.

Fig. 4.1 (a) Frequency response and (b) impulse response of channel and cross talk

4.1 Optimal Tap Coefficients and Post-tap Number

PRBS data is transmitted at rates of 3.125Gbps, 6.25Gbps, and 8Gbps respectively. Symbol-spaced FIR (SSF) filter is employed as pre-emphasis technique in order to counteract inter-symbol interference (ISI) due to channel loss (no cross talk). Pre-tap number is set as zero. The optimal tap coefficients of SSF are obtained at above data rates with the post-tap number varied from 1 to 9 (total tap number is from 2

to 10). The far-end eye diagrams are drawn after applying the obtained SSF pre-emphasis to the link. Horizontal and vertical eye opening are read out and the data is noted down in Table 4.1.

Table 4.1 Eye opening at far end of channel with post-tap number varying

Horizontal (UI) Vertical	1-tap	2-tap	3-tap	4-tap	5-tap	6-tap	7-tap	8-tap	9-tap
3.125Gbps	0.92 0.48	0.93 0.48	0.94 0.48	0.95 0.48	0.95 0.48	0.95 0.49	0.95 0.49	0.95 0.49	0.95 0.49
6.25Gbps	0.85 0.18	0.85 0.18	0.91 0.22	0.92 0.21	0.91 0.22	0.91 0.22	0.92 0.22	0.92 0.22	0.91 0.21
8Gbps	0.81 0.09	0.81 0.09	0.82 0.09	0.87 0.12	0.88 0.12	0.89 0.13	0.89 0.13	0.89 0.13	0.89 0.13

The eye opening versus tap number in Table 4.1 can be depicted in Fig. 4.2 (a) and (b).

(a)

(b)

Fig. 4.2 Eye opening vs. tap number: (a) horizontal; (b) vertical.

For both horizontal and vertical eye opening, there exists a tap number beyond which eye opening cannot be notably improved by simply increasing the number of taps. This tap number is optimal in terms of performance and complexity trade-off for FIR implementation. Fig. 4.3 shows the pulse response of channel at different data rates.

Fig. 4.3 Channel pulse responses at different data rates

It is not surprising to find that the optimal tap number obtained through LMS optimization is close to the number of symbol-spaced points (circle for 3.125Gbps, square for 6.25Gbps and triangle for 8Gbps in Fig. 4.3.) that cover the most part of the tail of the pulse response. For the 34" FR4 backplane used here, 3 post-taps is optimal for 3.125Gbps and 6.25Gbps, and 4 post-taps is optimal for 8Gbps. SSF pre-emphasis with the obtained optimal tap number is applied to the link at different data rates. Far-end eye diagrams with and without SSF pre-emphasis are compared in Fig. 4.4. The effectiveness of SSF pre-emphasis at optimal tap number can be clearly seen.

Fig. 4.4 Comparison of eye diagrams without and with pre-emphasis:
(a) 3.125Gbps, 3 post-taps; (b) 6.25Gbps, 3 post-taps; (c) 8Gbps, 4 post-taps.

4.2 Pre-tap Effect on Eye Opening

Pre-tap effect is worthwhile for feasibility of using pre-tap in pre-emphasis design. In the above section, pre-emphasis only employs multiple post-taps to counteract inter-symbol interference due to post-cursor effect of impulse response. At a rate of 10Gbps, this experiment is repeated and it is found that the application of only the post-tap is not enough to achieve a vertical eye opening. ISI caused by the pre-cursor effect of impulse response also needs to be counteracted.

Without pre-emphasis, at 10Gbps, the eye diagram at the far end of the channel is totally closed, as shown in Fig. 4.5 (a). Then set number of post-tap to 10, which is large enough to counteract most ISI due to the post-cursor effect. The eye diagram at the far end of the channel is drawn in Fig. 4.5 (b). It can be seen that the eye opening is very small. In comparison, set number of post-tap to 10, but number of pre-tap is set as 4 other than 0. The eye diagram is drawn in Fig. 4.5 (c) and it can be seen that the eye is opened vertically, even though the jitter distortion is not improved largely.

Fig. 4.5 Eye diagrams at rate of 10Gbps (a) without pre-emphasis; (b) 10 post-tap FIR pre-emphasis; (c) 10 post-tap plus 4 pre-tap FIR pre-emphasis.

The eye opening values in Fig. 4.5 are read out and listed in the following table.

Table 4.2 Eye opening in Fig. 4.5

	No Pre-Emphasis	10 Post-tap	4 Pre-tap+10 Post-tap
Horizontal (UI)	Closed	0.55	0.63
Vertical (normalized)	Closed	0.03	0.08

Table 4.2 shows that the eye is opened both horizontally and vertically. In most cases, group delay due to pre-tap may have the same sign as the group delay due to the channel. Thus it would introduce more jitter, other than help cancel the group delay, even though it is helpful in opening eye vertically. However in the case of 10Gbps, it can be explained that the ISI due to the pre-cursor cancelled out by pre-tap is more than ISI due to group delay caused by pre-tap. Therefore the final eye diagram displays better performance after neutralization.

This pre-tap experiment is repeated for data rate less than 10Gbps. 6.25Gbps is selected as another example. Fig. 4.6 (a) shows the eye diagram at far end of the channel without any pre-emphasis. Fig. 4.6 (b) shows the eye diagram with 9 post-taps pre-emphasis. Then set pre-tap number equal to 3, and post-tap number still equal to 9, eye diagram is drawn and is shown in Fig. 4.6 (c). Table 4.3 shows the eye opening values in Fig. 4.6.

Fig. 4.6 Eye diagrams at rate of 6.25Gbps (a) without pre-emphasis; (b) 9 post-tap FIR filter pre-emphasis; (c) 9 post-tap plus 3 pre-tap FIR filter pre-emphasis.

Table 4.3 Eye opening in Fig. 4.6

	No Pre-Emphasis	9 Post-tap	3 Pre-tap+9 Post-tap
Horizontal (UI)	Closed	0.91	0.87
Vertical (normalized)	Closed	0.21	0.24

From Table 4.3, it can be seen that for a lower data rate such as 6.25Gbps, pre-tap shows itself in degrading the horizontal eye opening. Therefore, pre-tap is a choice, based on how strong the pre-cursor detrimental effect is.

4.3 Cross-talk Effect on LMS Outputs

All above results and analysis is based on the LMS outputs, which are the optimized tap coefficients of FIR filter, without considering cross talk. The characteristics of cross talk can also be described by measured s-parameters. They can be depicted in frequency domain (frequency response) or in time domain (impulse response). Fig. 4.1 at the beginning of this chapter has included the cross-talk characteristics of the 34" FR4 backplane system under study. In the frequency domain, the cross talk is around –

30dB and in the time domain impulse response due to cross talk is about 1/10 of that of channel. Therefore, cross-talk effect can be neglected. The following case in quantity calculation confirms this point.

Data is transmitted at a rate of 6.25Gbps over 34" FR4 backplane which acts as the affected line. Inside the neighboring channel (aggressor/driver line), data is also transmitted at the same data rate, but in cases of synchronous transmission (delay 0 UI) with data inside the affected line, and delayed transmission (delay from 0.2UI to 1UI with 0.2UI as step) to the affected line. In these cases as well as the case of no cross talk, tap coefficients (LMS outputs) are calculated for a 3 post-tap symbol-spaced FIR filter used for pre-emphasis. Table 4.4 gives the values of tap coefficients and the eye opening at the far end of the channel in each case.

Table 4.4 Tap coefficients/eye opening without/with Xtalk effect

	<i>Tap 1</i>	<i>Tap 2</i>	<i>Tap 3</i>	<i>Tap 4</i>	<i>Horizontal (UI) /Vertical</i>
No Xtalk	1	-0.47	0.06	-0.06	0.91/0.22
0 UI	1	-0.47	0.06	-0.07	0.90/0.21
0.2UI	1	-0.47	0.06	-0.06	0.90/0.21
0.4UI	1	-0.47	0.06	-0.07	0.89/0.21
0.6UI	1	-0.47	0.06	-0.07	0.92/0.22
0.8UI	1	-0.47	0.06	-0.06	0.90/0.21
1.0UI	1	-0.47	0.06	-0.06	0.91/0.22

Fig. 4.7 shows the values in Table 4.4 graphically. It can be concluded that cross talk has almost no effect on LMS outputs and eye opening at the level shown in Fig. 4.1.

(a)

(b)

(c)

Fig. 4.7 Eye opening against synchronous and delayed cross-talk, (a) tap coefficients, (b) horizontal eye opening, and (c) normalized vertical eye opening.

4.4 Incorporation of RC Equalizer

Pre-emphasis at the transmit end and equalizer at the receive end can be used in combination as shown in Fig. 4.8.

Fig. 4.8 Pre-emphasis and equalizer are in combination

RC equalizer is popularly used. The source followed resistance R and capacitance C can vary in order to compensate for specific channel loss at specific data speed. Fig. 4.9(a) shows that when R is fixed, equalizer gain reaches its maximum point at frequencies from lower to higher, with C varying from larger value to smaller value. This characteristic can be utilized for RC equalizer adaptive to data speed while channel length is fixed. For a PRBS data pattern, most power is concentrated within the bandwidth of half data speed. Letting the maximum gain point target to the half data speed or a bit larger, can compensate for most channel loss.

Fig. 4.9 RC equalizer frequency response with (a) R fixed, C varying and (b) C fixed, and R varying

Another RC equalizer characteristic can be seen in Fig. 4.9(b). In this figure, C is fixed while R varies. The frequency at which the maximum gain is reached remains almost the same while the DC gain level varies from low to high and R changes from a larger value to a smaller value. The fixed maximum gain point means that data speed

can be fixed. Thus, this characteristic can be utilized for an RC equalizer adaptive to channel length, which represents channel loss.

In the data path, after the RC equalizer is a limiter, which is needed to increase the DC gain level. Otherwise the signal swing is too small to interpret. A high gain bandwidth product is required for the limiter in high-speed data transmission. Most data power should be covered within the bandwidth. CML structured limiter can satisfy the gain bandwidth tradeoff. A three-stage CML structured limiter is designed. These three stages are arranged in inverse scaling. Its frequency response is shown in Fig. 4.10. The DC gain level can reach 20dB while -3dB bandwidth is about 7.5GHz.

Fig. 4.10 Frequency response of 3-stage CML structured limiter

As an example of the incorporation of an RC equalizer with pre-emphasis, one RC fixed equalizer is designed in HSPICE. Its frequency response is shown in Fig. 4.11(a) and Fig. 4.11(b) for magnitude and group delay respectively.

Fig. 4.11 Frequency response of RC equalizer: (a) magnitude; (b) group delay

The data shown in Fig. 4.11 is then imported from HSPICE into MATLAB. Thus, one RC equalizer is set up in MATLAB. For data transmitted at 8Gbps through 34" FR4 backplane, a one post-tap FIR filter is employed for pre-emphasis at the transmit end. The eye diagram at the output of the equalizer is shown in Fig. 4.12. It is an integrated result of FIR pre-emphasis and RC equalizer in combination.

Fig. 4.12 Eye diagram at the output of equalizer

CHAPTER 5 CMOS PRE-EMPHASIS IMPLEMENTATION AND EVALUATION

A channel model has been set up in HSPICE. Its frequency response and impulse response can emulate the characteristics of an FR4 backplane trace. The tap coefficients of an FIR filter for pre-emphasis are calculated in the MATLAB environment first. Optimal tap coefficients and number of tap for symbol-spaced FIR (SSF) filter are obtained. Eye diagrams at receive end are drawn and eye openings, horizontal and vertical, are read out. Then in HSPICE one complete programmable multiple-tap SSF filter is implemented. The tap coefficients obtained from MATLAB are realized in circuit design. Comparison is given between the MATLAB results and the HSPICE results to verify the MATLAB programming.

5.1 Channel Model in HSPICE

Fig. 5.1 (a) shows the channel attenuation of one FR4 channel model in HSPICE. Its impulse response in time domain is shown in Fig. 5.1 (b).

Fig. 5.1 Channel model in HSPICE (a) attenuation; (b) impulse response.

The characteristic data of this channel model are imported into MATLAB from HSPICE. Thus same channel characteristics can be seen in MATLAB. Programming can proceed to calculate tap coefficients for FIR filter as pre-emphasis.

5.2 Results of MATLAB Modeling

The tap coefficients calculator and eye opening reader discussed in Chapter 3 are employed to deal with the channel model obtained from HSPICE. Data rate is set at 8Gbps. A symbol-spaced FIR filter is chosen for pre-emphasis. ISI due to the pre-cursor can be neglected. This would be seen from the impulse response and MATLAB simulation. Thus only post-taps are used to counteract ISI due to post-cursor. With number of post-tap varying, optimal tap coefficients are calculated and the eye openings at the receive end are read, as listed in Table 5.1.

Table 5.1 Optimal Tap Values and Eye Opening at Different Post-tap Number

Tap #	Horizontal (UI)/Vertical	Tap Values
1	0.88 / 0.11	1.00 -0.41
2	0.91 / 0.11	1.00 -0.39 -0.04
3	0.95 / 0.11	1.00 -0.39 -0.03 -0.03
4	0.96 / 0.11	1.00 -0.39 -0.03 -0.03 -0.02
5	0.96 / 0.11	1.00 -0.39 -0.03 -0.03 -0.01 -0.01
10	0.96 / 0.11	1.00 -0.39 -0.03 -0.03 -0.01 -0.01 0 0 0 0 0

From the eye opening values, it can be seen that at a data rate of 8Gbps, four post-taps are enough for pre-emphasis. The eye diagrams at receive end are shown in Fig. 5.2.

(a)

(b)

(c)

(d)

(e)

(f)

Fig. 5.2 Eye diagrams at far end of channel with FIR post-tap number of (a) 1, (b) 2, (c) 3, (d) 4, (e) 5, and (f) 10.

5.3 Implementation of Pre-emphasis in CDE

5.3.1 Test Bench Setup

In the Cadence Design Environment (CDE), CMOS 0.09um technology is employed to implement the FIR filter pre-emphasis. All HSPICE simulation results are obtained from Slow-Slow (SS) process corner. The total power is about 20mW. A test bench as shown in Fig. 5.3 is set up. The major blocks in Fig. 5.3 are listed in Table 5.2.

Table 5.2 Blocks in Fig. 5.3

A	B	C	D	E	F	G	H
Retiming Data Block	Tap Sign Block	Pre- Emphasis	Tap Values Decoder	Sign Control Signals	Value Control Signals	Channel Model	Receive Limiter

If an equalizer is used in combination with FIR filter pre-emphasis, it can be added in front of block H, the limiter.

Fig. 5.3 Test bench setup in CDE - see Table 5.2

5.3.2 Blocks

The limiter is made of a three-stage current-mode logic (CML) circuit as shown in Fig. 3.13 (b). At SS corner, its DC gain can reach 20dB at a bandwidth of 7.5GHz. Fig. 5.4 (a) shows the inside of Block A, the retiming data block. DFF is employed to retime data and get the non-delayed and symbol-delayed signals (up to 4 symbols delay). One DFF is made of two D latches as shown in Fig. 5.4 (b). Fig. 5.4 (c) shows the structure of D latch. The retimed outputs of DFF are buffered first before they go into the next block. The outputs of the buffers are shown in Fig. 5.4 (d).

(a)

(b)

(c)

Fig. 5.4 (a) Block A; (b) DFF; (c) D latch; (d) from bottom to top: non-delayed signal, 1-bit delayed, 2-bit delayed, 3-bit delayed and 4-bit delayed signals

Fig. 5.5 (a) shows the inside of Block B, the tap sign block. Buffer chain as shown in Fig. 5.5 (b) is used to reach high driving capability for the following pre-emphasis block. The +/- sign of the tap values can be realized by exchanging the differential input to a MOS gate. This is controlled by an XOR logic circuit as shown in Fig. 5.5 (c). Four inputs are differential input signals and differential control signals.

Fig. 5.5 (a) Block B; (b) buffer chain; and (c) XOR logic circuit.

Block C, the pre-emphasis block, is shown in Fig. 5.6 (a). Four post-taps are designed. As example, tap 2 is shown in Fig. 5.6 (b), and other taps have the similar structure. Tap value can be varied by changing the number of parallel transistors, inside of black box in Fig. 5.6 (b). In Fig. 5.6 (b), decoupling capacitor is used to reduce the transient switching noise and reduce the variation of gate voltage. Taps 1 and 2 have wider adjusting ranges, from 0.2 to 0.48 and 0.02 to 0.14 respectively. Taps 3 and 4 are made the same, with values from 0.02 to 0.06. Logic gates are used to activate or inactivate the taps.

(a)

(b)

Fig. 5.6 (a) Block C; (b) tap 2.

Block D, as shown in Fig. 5.7 (a), is the decoder block designed for decoding tap value control signals to the corresponding tap values. As example, one 2 to 4 decoder is shown in Fig. 5.7 (b).

Fig. 5.7 (a) Block D; (b) decoder 2 to 4.

The logical calculation can be explained in the following equation (decoder 4 to 16 here):

$$Q_0 = \overline{\overline{ABCD}}, \quad Q_{i+1} = Q_i \cdot X \quad (5.1)$$

and

$$X = \overline{\overline{ABCD}} \quad \overline{\overline{ABCD}} \quad \overline{\overline{ABCD}} \quad \dots \quad \overline{\overline{ABCD}}, \quad (5.2)$$

with A, B, C, D as inputs, and Q_n ($n = 0, 1, 2, \dots, 15$) as outputs. The truth table for Q is shown in Table 5.3.

Table 5.3 Truth table for Eq.5.1

$ABCD$	$Q_0Q_1Q_2 \cdots Q_{15}$
0000	0000000000000000
0001	1000000000000000
0010	1100000000000000
...	...
1111	1111111111111110

5.4 Results of HSPICE Simulation

If there is no pre-emphasis technique employed, the eye diagrams at near end and far end of channel model are shown in Fig. 5.8 (a) and (b) respectively.

Fig. 5.8 No pre-emphasis, eye diagrams: (a) near end; (b) far end of channel.

If one four post-taps FIR filter is used for pre-emphasis, its expression, written here, assuming a, b, c, d are positive, is:

$$y(n) = x(n) - ax(n-1) - bx(n-2) - cx(n-3) - dx(n-4) \quad (5.3)$$

For four post-tap FIR filter circuit, tap values listed in Table 5.1 are depicted in Fig. 5.9.

Fig. 5.9 Tap coefficients of four post-tap FIR filter, with $a=0.39, b=0.03, c=0.03$ and $d=0.02$ corresponding to plots (a), (b), (c) and (d).

The eye diagrams are drawn in Fig. 5.10. The eye opening at far end of channel is listed in Table 5.4, compared with the eye opening read in MATLAB as shown in Table 5.1.

Table 5.4 Eye opening at far end of channel

	Horizontal (UI)	Vertical
HSPICE	0.94	0.09
MATLAB	0.96	0.11

(a)

(b)

(c)

Fig. 5.10 Eye diagrams at: (a) near end; (b) far end of channel model; and (c) output of limiter with four post-tap FIR filter employed

5.5 Variation in Tap Values

In the implementation of an FIR filter for pre-emphasis, there are all kinds of factors that could affect the accuracy of the tap values. In the case that there is a $\pm 2\%$ variation in the tap values, it is meaningful to redraw the eye diagrams and watch the variation in eye openings, especially the horizontal eye opening at the receive end.

In order to explain the variation in tap values in quantities during its implementation, one example is given here. For a two post-tap FIR filter, its expression is:

$$y(n) = x(n) - ax(n-1) - bx(n-2). \quad (5.4)$$

In Table 5.1, for the same channel used before, $a = 0.39, b = 0.04$. After implementation and simulation of this two post-tap FIR, Fig. 5.11 shows the eye diagrams at near end, far end of the channel and the output of the limiter. The horizontal eye opening at the far end of the channel is $0.90UI$.

Fig. 5.11 Eye diagrams at: (a) near end; (b) far end of channel; and (c) output of limiter with two post-tap FIR filter employed

In the case of $\pm 2\%$ variation in tap values, the new values of a and b are listed in

Table 5.5:

Table 5.5 Variation in values of tap a and b

a =	0.37	0.37	0.41	0.41
b =	0.02	0.06	0.02	0.06

With new values of tap a and b, simulations are run and the eye diagrams for each case from Table 5.5 are drawn in Fig. 5.12.

Fig. 5.12 Eye diagrams at near end, far end of channel and limiter output in cases of (a) $a=0.37$, $b=0.02$; (b) $a=0.37$, $b=0.06$; (c) $a=0.41$, $b=0.02$; and (d) $a=0.41$, $b=0.06$.

The horizontal eye opening in each case are also read out and listed in Table 5.6.

Table 5.6 Horizontal Eye Opening in Variation of Tap Values

	$a = 0.37, b = 0.02$	$a = 0.37, b = 0.06$	$a = 0.41, b = 0.02$	$a = 0.41, b = 0.06$
Horizontal Eye Opening (UI)	0.85	0.87	0.86	0.88

From above results, it can be seen that there is about 0.05UI variation in the horizontal eye opening in the case of a $\pm 2\%$ variation in tap values of the two post-tap FIR filter. Therefore, in the implementation of an FIR filter, although the variation in tap values cannot be avoided, it should be suppressed to its minimum, catering to the clock data recovery circuit design and bit-error-rate requirement.

CHAPTER 6 CONCLUSIONS AND FUTURE WORK

6.1 Conclusions

In industry, pre-emphasis at the transmit end and equalizer at the receive end are two widely used techniques to counteract inter-symbol interference caused by channel amplitude attenuation and dispersion of phase. These two techniques can be employed independently or used in combination. In the implementation of programmable pre-emphasis and equalizer at the circuit level, the realistic question of how much pre-emphasis or equalization should be provided in order for perfect link performance is well answered in this thesis. Backplane system designers can find the guidelines in this thesis for designing FIR pre-emphasis and RC equalizer.

A top-down design methodology for a high-speed SERDES is proposed. As one part of the design methodology, one MATLAB simulator has been developed to emulate the complete link simulation from transmit end, including an FIR filter for pre-emphasis, to the receive end (may include RC equalizer) through a backplane network. The functions it can realize include almost all topics of interest, like frequency response and impulse response of the FIR filter and channel, characteristics of the channel, eye opening readout, cross-talk effect and so on. Given a specific backplane channel, this program allows users to calculate the optimal FIR pre-emphasis coefficients, and set equalization parameters that could then be used as preliminary data for their applications. With enhanced features of jitter read-out and BER estimation, this program will help the user to predict BER in real application.

Besides solving a practical problem faced by industry, contributions from a theoretical point of view in this thesis include that: ISI due to group delay distortion is treated seriously; the role that pre-tap can play is studied, and its contribution for ISI deduction is calculated in quantity; and in which way the optimal post-tap number can be determined for the best pre-emphasis performance. Circuits have been designed in Cadence design environment. HSPICE simulation has verified the results from MATLAB simulation.

6.2 Future work

As described in the proposed methodology flow chart, future work can include the design of equalizer (DFE preferred [29,30]) and clock-data recovery (CDR) [31,32,33]]. The MATLAB simulation and behavioral modeling are used to optimize their performance, and provide guidance for transistor-level design. Currently one CDR circuit, including half-rate phase-frequency detector (PFD), and ring voltage-controlled oscillator (VCO), has been designed in CMOS 0.18um technology. Next work is its behavioral modeling in MATLAB or HSPICE method.

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ACRONYM AND ABBREVIATION

ADC	Analog digital converter	ISI	Inter symbol interference
AFE	Analog front end	LMS	Least mean square
BER	Bit error rate	OIF	Optical internetworking forum
CDE	Cadence design environment	PCB	Printed circuit board
CDR	Clock data recovery	PDF	Probability density function
CEI	Common electrical interface	PFD	Phase frequency detector
CML	Current mode logic	PLL	Physical link and layer
DFE	Decision feedback equalizer	RC	Resistance/capacitance
DFF	D flip flop	RJ	Random jitter
DJ	Deterministic jitter	SERDES	Serializer/deserializer
FSF	Fractionally spaced FIR	SI	Signal integrity
Gbps	Giga-bit per second	SNR	Signal noise ratio
HSBI	High speed backplane initiative	SS	Slow-slow
IFFT	Inverse fast Fourier transformation	SSF	Symbol spaced FIR
IIR	Infinite impulse response	UI	Unit interval