Digital Equalization of Fiber-Optic Transmission System Impairments

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BY
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Abstract

In the past half century, numerous improvements have been achieved to make fiber-optic communication systems overweigh other traditional transmission systems such as electrical coaxial systems in many applications. However, the physical features including fiber losses, chromatic dispersion, polarization mode dispersion, laser phase noise, and nonlinear effect still post a huge obstruction in fiber-optic communication system. In the past two decades, along with the evolution of digital signal processing system, digital approach to compensate these effects become a more simple and inexpensive solution.

In this thesis, we discuss digital equalization techniques to mitigate the fiber-optic transmission impairments. We explain the methodology in our implementation of this simulation tool. Several major parts of such digital compensation scheme, such as laser phase noise estimator, fixed chromatic dispersion compensator, and adaptive equalizer, are discussed. Two different types of adaptive equalizer algorithm are also compared and discussed. Our results show that the digital compensation scheme using least mean square (LMS) algorithm can perfectly compensate all linear distortion effects, and laser phase noise compensator is optional in this scheme. Our result also shows that the digital compensation scheme using constant modulus algorithm (CMA) has about 3~4db power penalty compare to LMS algorithm. CMA algorithm has its advantage that it is capable of

blind detection and self-recovery, but the laser phase noise compensator is not optional in this scheme. A digital compensation scheme which combines CMA and LMS algorithm would be a perfect receiver scheme for future work.

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Last, but not least, I would like to dedicate this work to my dear parents and beloved for their continued and unconditional love and support.

Notation and abbreviations

3D-IC	Three-Dimensional Integrated Circuit
ADC	Analog-Digital Converter
ASE	Amplified Spontaneous Emission
AWGN	Additive White Gaussian Noise
BER	Bit Error Rate
CD	Chromatic Dispersion
CMA	Constant Modulus Algorithm
DCF	Dispersion Compensating Fiber
DSF	Dispersion-Shifted Fiber
DSP	Digital Signal Processing
EDFA	Erbium Doped Fiber Amplifier
GVD	Group Velocity Dispersion
IC	Integrated Circuit
ISI	Inter-Symbol Interference
LMS	Least Mean Square Algorithm
PDL	Polarization Dependent Lose
PLL	Phase Lock Loop
PM-QPSK	Polarization-Multiplexing QPSK

PMD	Polarization Mode Dispersion
QPSK	Quadrature Phase-Shift Keying
SOC	System-On-A-Chip
SMF	Single-Mode Fiber
WDM	Wavelength-Division Multiplexing

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Chapter 1

Introduction

1.1 Introduction to Fiber-optic Communication

Optical fibers have been employed in telecommunication system for decades and the market is still growing rapidly nowadays. It becomes more and more popular at both long-haul and local wired transmission area owning to its many advantages such as high capacity, low cost, no electromagnetic interferences, and huge bandwidth. While the demand of fiber-optic system increases dramatically, the relative technology advances significantly.

By the invention of laser in 1960s, fiber-optic communication system first found a suitable optical source. Later in 1970, Corning Incorporated demonstrated titanium doped silica glass with 17dB/Km attenuation and soon the attenuation is reduced to 4dB/Km. Now there are both suitable optical source and transmission media for the first generation of fiber-optic communication system. The first generation system operated around 0.8 μm and 45Mbit/s with repeater spacing (repeater is used to regenerate an optical signal by converting it to an electrical signal, processing that electrical signal and then retransmitting an optical signal based on the processed electrical signal pattern) in excess of 10Km. The major limitation of this generation system comes from chromatic dispersion (CD) which causes significant pulse broadening and thus inter-symbol interference (ISI).

In order to reduce fiber dispersion effect, single-mode fiber (SMF) was introduced.

Comparing to multi-mode fiber, SMF is designed to support only one mode of propagation light and thus has lower dispersion dramatically. This allows a higher system symbol rate and longer repeater distance. In 1981, single-mode fiber-optic system achieved 2Gbit/s at 1.3 μm with 44km spacing [1].

Later in 1990, along with the development of silica fibers (low loss) and dispersion-shifted fiber (DSF) (low dispersion while operating at low loss wavelength 1.5 μm) the third-generation fiber-optic communication system emerged. This generation system was characterized by operating at 1.5 μm wavelength region (for latest 0.2dB/Km loss) and 10Gbit/s with repeater spacing of up to 100Km [2].

The fourth-generation fiber-optic communication system evolution started from 1992 till 2001, owning to the invention of practical optical amplifier (erbium-doped fiber amplifier (EDFA) more specifically) to replace repeaters and the wavelength-division multiplexing (WDM) method to increase data capacity. Since 1992, fourth-generation fiber-optic communication system bit rate start doubling every six months till a bit rate of 10TB/s was achieved by 2001.

Recently, owning to the revolution of integrated circuit (IC) design and the emerge of system-on-a-chip (SOC) and three-dimensional integrated circuit (3D-IC) level IC, the cost of high speed DSP system dropped tremendously while the available maximum processing speed skyrocketed. Researchers started trying to use DSP device to replace the complex and expensive optical device for channel compensation. This approach was first proved to be feasible in 2003[3], and then various DSP algorithms have been proposed and tested. Empowered by these technology advancements, fiber optic communication system has become a versatile, reliable, and cost-efficient wired communication system solution.

1.2 Distortion effects in Fiber-optic Communication System

As the fiber manufacturing technology advances, many distortion and impairment effect in fiber-optic communication system decreased significantly. These effects include but not limit to amplified spontaneous emission (ASE), laser phase noise, chromatic dispersion (CD), polarization mode dispersion (PMD), and fiber nonlinearity. The first effect, ASE, sets the hard constraints of any practical fiber-optic communication system. It imposes the upper limit on system performance. The following three effects, laser phase noise, CD, and PMD, are the distortion effects that can be compensated. They are the soft constraints of fiber-optic communication systems. Depending on different compensator performance, the overall system performance might vary a lot. The last distortion effect, fiber nonlinearity which is increased as signal power increased, imposes an upper power constraint to system.

Similar to all other transmission media, optical signal that transmitted in optical fiber also suffers fiber loss and decays as it is transmitting through the fiber. Although the loss of a fiber is extremely low compared to other transmission media (usually at 0.2dB/km for 1.55 μm wavelength optical signal nowadays), optical signal decays along fiber link and will become indistinguishable from noise. Generally, all fiber-optic communication systems, especially long-haul systems, require optical amplifiers or optical repeaters to amplify or reconstruct optical signal to ensure its signal-to-noise ratio stays above a certain recoverable limit. Amplified spontaneous emission noise is the side effect of employing such optical amplifier in system, and it can be modeled as additive white Gaussian noise (AWGN).

Laser phase noise comes from the transmitter laser and local oscillator laser (used in coherent receiver). It can be modeled as a Wiener process, a random walk like white noise. This effect is hard to compensate by optical device. However, a DSP compensation device can easily handle this effect. This is first of several prime motivations of using DSP compensation system in fiber-optic communication system.

Chromatic dispersion (often referred as group velocity dispersion (GVD)) is a frequency dependent dispersion which causes the optical signal that traveled through fiber arrived at receiver at different time. This will cause the pulse broadening and inter-symbol interference which was the major limitation of early fiber-optic communication system. Although SMF, DSF, and dispersion compensating fiber (DCF) have been invented to reduce and/or counteract this effect, these compensation approaches are still quite expensive overall. This is another prime motivation of using DSP compensation system in fiber-optic communication system, to reduce the system cost.

Polarization mode dispersion (PMD) effect is a fiber only impairment that comes from the random imperfections and asymmetries of the fiber cross section shape. As the imperfections exist throughout the whole length of fiber which cause the optical signal in two polarization travel at different speed and thus cause a random distortion of optical signal. Because the imperfections are random and slowly time varying (due to thermal stresses, mechanical stresses, and etc), the PMD effects are random and time-dependent. Thus, in order to compensate this effect, a compensation device with a feedback mechanism is required. This is also a prime motivation of using DSP compensation system in fiber-optic communication system, since an optical device that provides such function is very expensive and complex.

Fiber nonlinearity describes the optical nonlinearity effect which exhibit at high optic launch power. When the electric field intensity of electromagnetic wave (light) is comparable to the inter-atomic electric field of fiber, optical signal will interact with atoms of fiber and generate new frequency components. Fiber nonlinearity increases as optical signal power increases. Thus, it enforces an upper limit to the launch power of fiber-optic communication system.

Our major focus in this thesis is about the electronic compensation method for laser phase noise, chromatic dispersion, and polarization mode dispersion of fiber-optic communication system. We will explore and provide insight for different DPS compensation approaches that replace all traditional optical compensation fiber, time lens, phase lock device and/or optical repeater in once.

1.3 Thesis Layout

The thesis is organized as follows:

In chapter 1, a brief introduction to our topic system is given with the content of each chapter. A general explanation of system distortion effects is given in section 1.2.

In chapter 2, some background of digital coherent receiver and digital compensation techniques are discussed.

In chapter 3, a comparison of different laser phase noise compensation scheme will be shown. The basic concept of laser phase noise model is reviewed. Various compensation techniques is introduced and compared.

In chapter 4, chromatic dispersion will be introduced to system along with its compensation approaches. Polarization mode dispersion is also discussed. Both LMS and CMA adaptive equalizer is reviewed, and their performance is compared.

In chapter 5, the work in this thesis is concluded.

Chapter 2

Background

In this chapter, we will review the background of the digital signal processing algorithm for coherent receivers, especially focusing on the coherent QPSK phase and polarization diversity receiver. We will discuss the several laser frequency drift (laser phase noise) compensation techniques, LMS and CMA adaptive equalization technique, and digital coherent receiver arrangement.

2.1 Laser Frequency Drift Estimation

Estimate and compensate the laser frequency drift is an important step in digital coherent receiver. Different algorithms were proposed to remove signal phase and thus estimate laser phase noise [4], [5]. Afterwards, many digital frequency estimation schemes were proposed such as wiener filtering phase noise estimation[6], symbol-by-symbol phase noise estimation[7], and block phase noise estimation[8][9]. Accompanied the prevailing M-th power laser phase estimation algorithm, many different phase unwarping techniques were also proposed to handle phase wrap effect in M-th power method [6], [8], [10]. The laser frequency drift estimator is discussed in chapter 3 in detail.

2.2 Chromatic Dispersion

The basic physics of chromatic dispersion is the frequency dependence of the refractive index $n(\omega)$. Since $v_g(\omega) = c/n(\omega)$, the velocity of light at frequency ω is dependent on its refractive index in the media. The group velocity delay (GVD) of different spectral components of emitted signal light is determined by the difference of their refractive index. This frequency dependent dispersion causes the different spectral components that sent at the same time to arrive at receiver at different time. In a common communication system where signal pulse is sent consecutively, spectral components of different pulses overlap due to such spectral broadening effect, and thus causes inter-symbol interference effect (ISI).

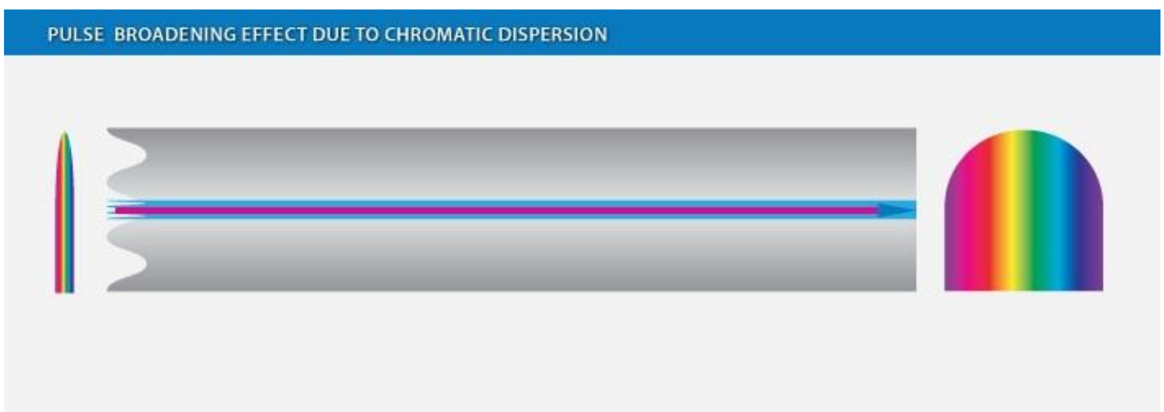


Figure 2.1 Illustration of pulse broadening effect due to chromatic dispersion

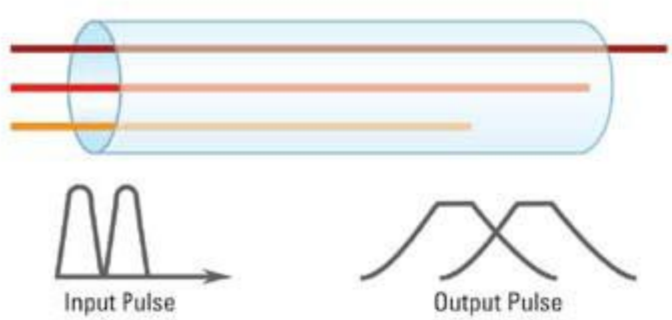


Figure 2.2 Illustration of GVD and pulse broadening effect

The effects of chromatic dispersion can be modeled by expanding the mode-propagation constant β in a Tayler series about the center frequency ω_0 as following

$$\beta(\omega) = n(\omega)\frac{\omega}{c} = \beta_0 + \beta_1(\omega - \omega_0) + \beta_2(\omega - \omega_0)^2 + \cdots,$$

where

$$\beta_i = \left.\frac{d^i\beta}{d\omega^i}\right|_{\omega=\omega_0}$$

thus

$$\beta_1 = \frac{1}{c}\left[n + \omega\frac{dn}{d\omega}\right]$$

$$\beta_2 = \frac{1}{c}\left[2\frac{dn}{d\omega} + \omega\frac{d^2n}{d\omega^2}\right]$$

where c is the speed of light in vacuum, β_1 and β_2 is the first and second order mode-propagation constant.

Chromatic dispersion is an accumulating effect as signal propagates along fiber. Usually the propagation distance of the communication channel is known, and thus the total chromatic dispersion effect can be modeled as following:

$$\Delta T = \frac{dT}{d\omega}\Delta\omega = \frac{d}{d\omega}\left(\frac{L}{v_g}\right)\Delta\omega = L\frac{d^2\beta}{d\omega^2}\Delta\omega = L\beta_2\Delta\omega \tag{2.1}$$

Equation 2.1 can also be expressed in terms of wavelength as following:

$$\Delta T = \frac{d}{d\lambda}\left(\frac{L}{v_g}\right)\Delta\lambda = DL\Delta\lambda$$

where

$$D = \frac{d}{d\lambda}\left(\frac{1}{v_g}\right) = -\frac{2\pi c}{\lambda^2}\beta_2$$

D is called the dispersion parameter and is expressed in units of $ps/(km - nm)$ [27].

2.3 Polarization Mode Dispersion

Another source of pulse broadening is related to fiber birefringence. Any real fibers have considerable variation in the shape of their core cross section along the fiber length. They may also sustain non-uniform stress that break the cylindrical symmetry of the fiber. These random imperfection and asymmetry cause fiber to lose degeneracy between the orthogonally polarized fiber modes and gain birefringence. The degree of modal birefringence is defined by

$$B_m = |\bar{n}_x - \bar{n}_y|$$

where $\bar{n}_x$ and $\bar{n}_y$ are the mode indices for the orthogonally polarized fiber modes. [27] Birefringence leads to a periodic power leakage between the two polarizations. The period, referred to as the beat length, is given by

$$L_B = \lambda / B_m$$

where λ is the wavelength of the optical signal. In most single mode fibers, birefringence changes randomly along the fiber in both magnitude and direction. Thus, optical signal travelled through fiber with linear polarization reaches a state of random polarization. Furthermore, if both polarization components are excited by input signal pulse different frequency spectral component of the signal pulse acquire different polarization states, and thus lead to pulse broadening. [27]

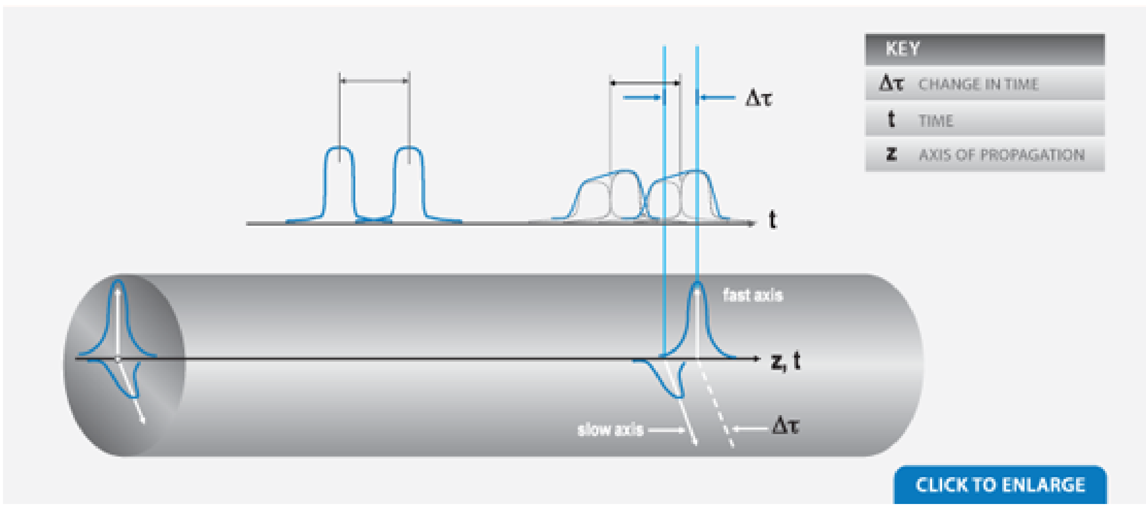


Figure 2.3 Illustration of pulse broadening effect due to PMD

In a fiber with constant birefringence, pulse broadening effect can be modeled as following

$$\Delta\tau = \left|\frac{L}{v_{gx}} - \frac{L}{v_{gy}}\right| = L\left|\beta_{1x} - \beta_{1y}\right| = L\Delta\beta_1 \, .$$

It is a bit different for conventional fiber in which birefringence varies randomly along the fiber. In this case, the pulse broadening effect cannot be estimated by previous formula, and only statistical mean dispersion can be calculated as follow

$$\langle\Delta\tau\rangle = D_p\sqrt{L}$$

where D_p is the PMD parameter. Measured values of D_p vary in the range between $0.01 ps/\sqrt{km}$ to $10 ps/\sqrt{km}$.

These distortion effects are called the polarization mode dispersion and have been studied extensively because it is a major performance degradation factor. In our simulation, we have taken both polarization power leakage and differential group delay variations into consideration.

2.4 Adaptive Equalization

In order to use digital coherent receiver to replace the former optical compensation device and receiver, a suitable equalization scheme is required to compensate other linear impairment effects such as PMD/PDL, CD. Several single channel (PMD/PDL ignored, no polarization multiplexing is used) compensation schemes using adaptive equalizer were introduced [6] [11]. For polarization multiplexed system, a modified scheme was proposed theoretically in [12] with fixed tap weight. Later, adaptive adjusted tap weight has been used in [7] (theoretical analysis) and [13]-[15] (experimental work). LMS algorithm was used to adaptive update tap weight in order to compensate time-varying impairment effects in [7], [13] and [15]. In [14] constant modulus algorithm (CMA [16] [17]) is employed in steady of LMS. Afterwards, several slightly modified schemes were described in [18]-[20]. These schemes add a complex FIR filters to compensate bulk CD before adaptive equalizer. In [19] and [20], CMA algorithm is used in training stage for initial tap weight conversion, and then LMS algorithm is used in receiving stage.

Chapter 3

Phase Noise Compensation

3.1 Phase Noise Model

Laser phase noise is come from the non-ideal oscillators used at both transmitter and receiver side. Laser phase noise is a continuous-time stochastic Wiener process, which is like a random walk in phase domain. The instantaneous frequency of a laser is modeled as a white Gaussian noise. Therefore, the phase noise becomes a Wiener process.

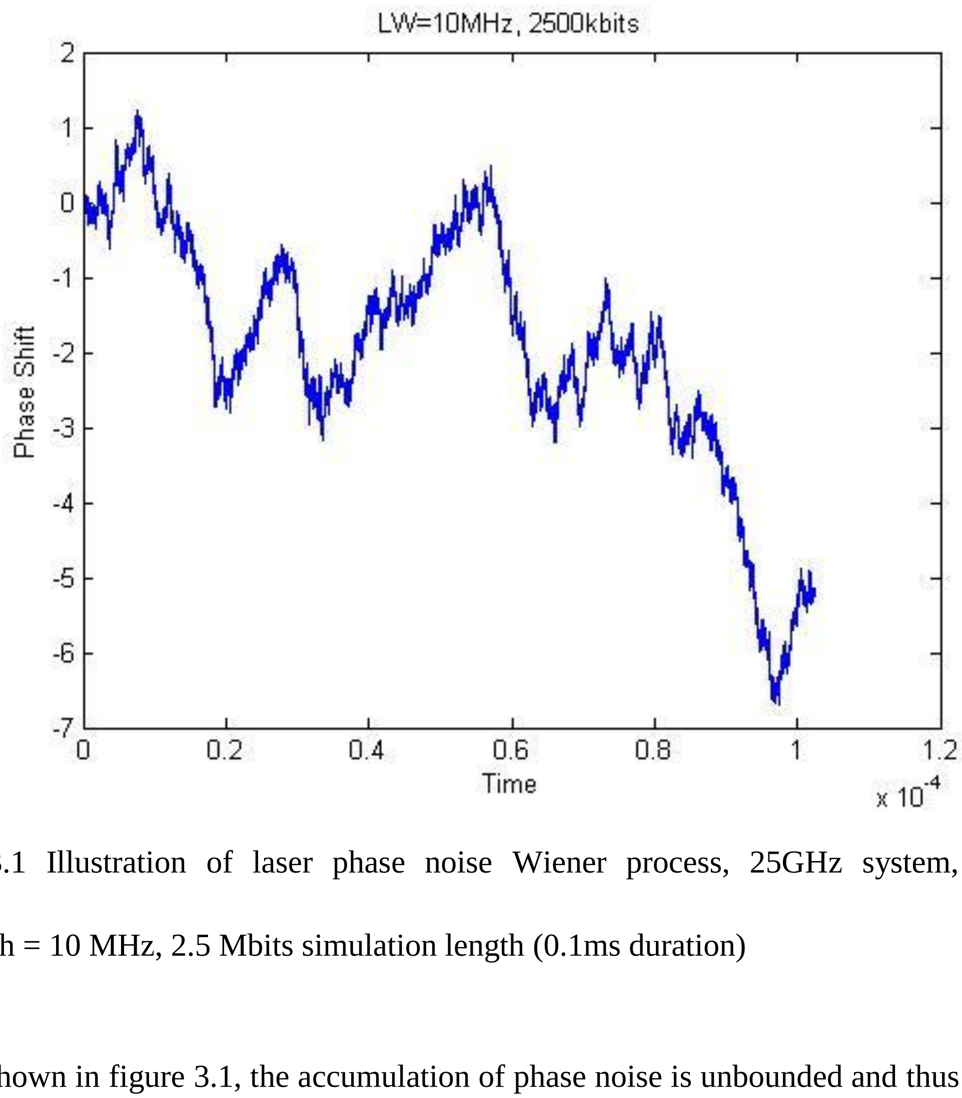


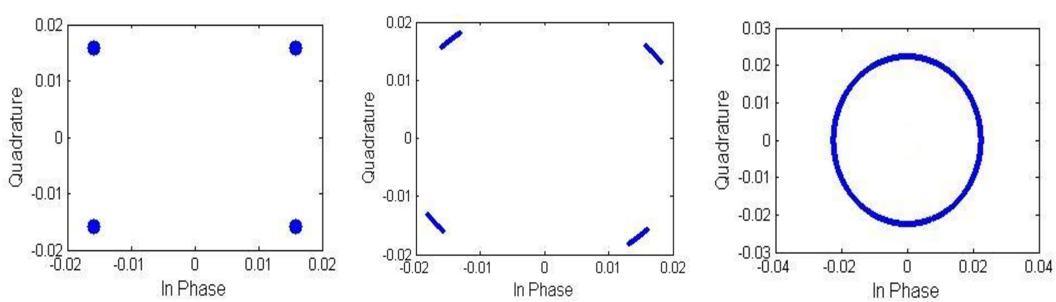
Figure 3.1 Illustration of laser phase noise Wiener process, 25GHz system, laser line-width = 10 MHz, 2.5 Mbits simulation length (0.1ms duration)

As shown in figure 3.1, the accumulation of phase noise is unbounded and thus could cause a complete loss of signal phase information. In most fiber-optic communication schemes, this effect is vital and must be compensated.

In this chapter, we assume all other impairments other than laser phase noise and optical amplifier noise have been perfectly compensated. In this case, we will have following formula:

$$R_k = S_k * exp\big(j * (\theta_{k-1} + \Delta\theta_k)\big) + N_k \qquad (3.1)$$

Where R_k is the received signal symbol at $k\Delta T$, ΔT is the sampling period, and S_k is the modulated signal. Total phase noise off-set of previous symbol is θ_{k-1}, and additive phase noise in k-th time interval is $\Delta\theta_k$. The additive channel noise(ASE) is represented as N_k.



(a) (b) (c)

Figure 3.2 Illustration of laser phase noise effect, 25GHz system, (a) laser line-width = 0 kHz, 300kbits length (1 ms) (b) laser line-width = 10 MHz, 25kbits length (1 μs) (c) laser line-width = 10 MHz, 2.5 Mbits length (0.1 ms)

Comparing figure 3.2 (b) to (a), we find out that laser phase noise will cause MPSK signal phase shift and thus degrade the system performance. From the difference of figure 3.2 (b) and (c), we observed the impact of the phase shift accumulation effect over time. It is obvious that laser phase noise has to be compensated or eventually the receiver will lose track of correct signal-symbol mapping.

3.2 Phase Noise Compensation Algorithms

There are two different approaches to compensate laser phase noise, by using phase lock loop (PLL) device to track signal phase in optical domain or using coherent detection with DSP device to compensate it in electrical domain. In this chapter, we will focus on DSP approach in electrical domain and discuss several different possible algorithms.

3.2.1 M-th Power Method

The first step of any DSP compensation algorithm is to isolate phase noise from signal message phase. Assuming that we are using M-PSK modulation scheme, the signal message phase is $S_k = A * \exp\left[l\frac{2\pi}{M}\right]$, where S_k is modulated signal, l=0,1,2, … ,M-1, M is the number of phase levels. Since $\exp(2\pi * m * i) = 1$, where m=0,1,2,$\cdots$, and $i = \sqrt{-1}$, we can only remove the signal message phase and thus isolate phase noise. This is achieved by multiplying the received signal phase by M and then take exponential function of the phase,

$$\frac{\angle \exp\{[M * \angle(R_k)] * i\}}{M} = \frac{\angle\exp\{M * \angle\{S_k * \exp(j * \theta_k) + N_k\} * i\}}{\mathrm{M}} \qquad (3.2)$$

$$= \frac{\angle\exp\left\{M * \left[l * \frac{2\pi}{M} + \tilde{\theta}_k\right] * i\right\}}{\mathrm{M}} = \frac{\angle\exp\{[l * 2\pi + M * \tilde{\theta}_k] * i\}}{\mathrm{M}} = \frac{\angle\exp(M * \tilde{\theta}_k * i)}{M}$$

$$= \tilde{\theta}_k ,$$

where θ_k is total laser phase noise, $\tilde{\theta}_k$is total effective phase noise. When N_k=0, $\tilde{\theta}_k = \theta_k$.

However, the computation cost of this method is relatively high (70% higher in computation time than equation 3.3) which lead us to introduce a small modification to this method (equation 3.3) [21].

$$\tilde{\theta}_k = \frac{\angle R_k^M}{M} \qquad (3.3)$$

Thus, from equation 3.3, we can estimate current phase noise from lasers and channel. Assuming that

$$(\tilde{\theta}_k - \tilde{\theta}_{k-1}) \ll \frac{2\pi}{M} \qquad (3.4)$$

to ensure that we can track signal phase noise increment sample by sample, this method can perfectly remove signal message phase from received signal and track signal phase noise increment at each sample. It enables the DSP device to reconstruct the phase noise model and thus compensate it.

Although M-th power method can perfectly remove signal message phase, it will also remove any $\frac{2\pi}{M}$ component in phase noise and cause a phase wrapping effect to wrap phase noise into $(-\frac{\pi}{M}, \frac{\pi}{M}]$ range. We need a phase unwrapping routine to track this effect and reverse it [22], [23]. Let current total phase noise to be

$$\tilde{\theta}_{original} = p * \frac{2\pi}{M} \pm \tilde{\theta}_{wrapped} \qquad (3.5)$$

, phase wrapping effect will crop the $p * \frac{2\pi}{M}$ part, which is similar to signal message, and output $\theta_{wrapped}$. p is a integer that will minimize $abs(\theta_{wrapped})$. Obviously, this effect needs to be undone to prevent undesirable phase jump when p changes.

3.2.2 Phase Unwrapping

As we discussed in section 3.1, laser phase noise is an accumulating effect and its absolute value might and will eventually become larger than $\frac{2\pi}{M}$. In order to still keep track of phase noise after using M-th power method, we also need to monitor any change of value p in equation 3.5. In order to achieve this objective, we employed the maximum likelihood principle. First we assume that $\Delta\tilde{\theta}$, which equals to $\tilde{\theta}_k - \tilde{\theta}_{k-1}$, will always lesser than $\frac{2\pi}{M}$ before taking M-power and being wrapped. Based on this assumption, with following formula

$$p = floor\left[\frac{1}{2} + \frac{2\pi}{M}\left(\tilde{\theta}_{k-1\ wrapped} - \tilde{\theta}_{k\ wrapped}\right)\right] \qquad (3.6)$$

$$\tilde{\theta}_{k\ unwrapped} = \tilde{\theta}_{k\ wrapped} + \frac{2\pi}{M} * p \qquad (3.7)$$

We are able to unwrap phase noise to its original phase index and avoid phase jump error.

However, the first assumption that we made is not necessary to be true in practical system, especially looking it over large time span or in an extremely noisy system. As stated in section 3.2.1, total phase noise in received signal comes from not only laser phase noise, but also additive channel white noise. In fact, in a noisy system, total phase noise in received signal mainly comes from channel noise. The combined effect of laser phase noise and channel noise will almost for sure result in a violation of our assumption within every millisecond or even lesser. We want to ensure that our unwrapping routine will not introduce extra error when our first assumption is violated. This goal is achieved by intentionally introducing a maximum compensation limit within a certain time interval. This prevents the fast-varying zero-mean component of phase noise, which comes from channel noise, from destroying our signal reception. In our software and simulations, this limit is set to two times of the average of total laser phase noise that will be additively introduced at current laser line-width configuration.

3.2.3 Phase Noise Compensation

Since we can track the phase noise change sample-by-sample, it is very easy to counter rotate this effect and remove it from the received signal. We first implemented this sample-by-sample compensation scheme on our system that has only laser phase noise, and obtained a perfect compensation result. The compensated received signal perfectly matched the transmitted signal, but when we start adding other impairment effects into our simulation software the disadvantage of this scheme emerged. Owing to its rapid update and precise tracking of any phase noise, this scheme was found out to be relatively unstable when comparing to our next candidate scheme, block phase noise compensation scheme.

Block phase noise compensation scheme was introduced to increase the stability of phase noise compensation device and improve its performance [24], [25]. The major difference between block phase noise compensation scheme and sample-by-sample compensation scheme is how we evaluate $\tilde{\theta}_{k-1\,wrapped}$ or $\tilde{\theta}_{k-1\,unwrapped}$. In sample-by-sample compensation scheme, these two values are directly taken from the measured result from last sample and thus will fluctuate promptly with additive Gaussian white noise. Since we only intend to compensate laser phase noise and in fact only laser phase noise can be effectively compensated, block compensation is employed to average out additive white noise over certain number of previous samples. Block compensation scheme works quite well in terms of removing additive channel white noise, but it also introduce an extra parameter, block size, to optimize. We will discuss the optimization of this parameter in section 3.3. By compensating phase noise block-wise and canceling a large proportion of phase noise that come from channel white noise, we can force our phase noise compensator to focus on laser phase noise tracking. The following figure shows how it works.

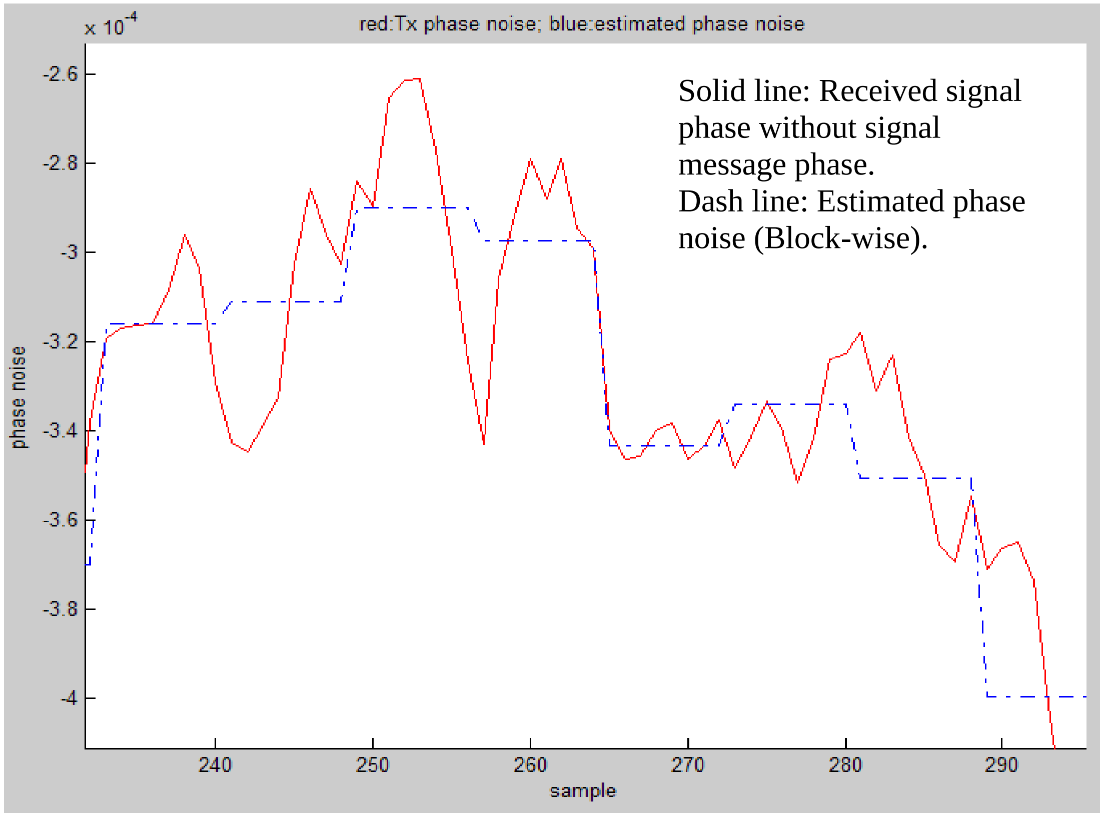


Figure 3.3 Illustration of block phase noise compensation scheme, 25GHz system, laser line-width = 1 Mhz, phase noise compensator block length = 8 samples, M-th Power method.

Block phase noise estimator will try to “ignore” the rapid varying part of total phase noise and only track the overall trend of laser phase noise. After estimation we will counter rotate this effect and compensate the received signal.

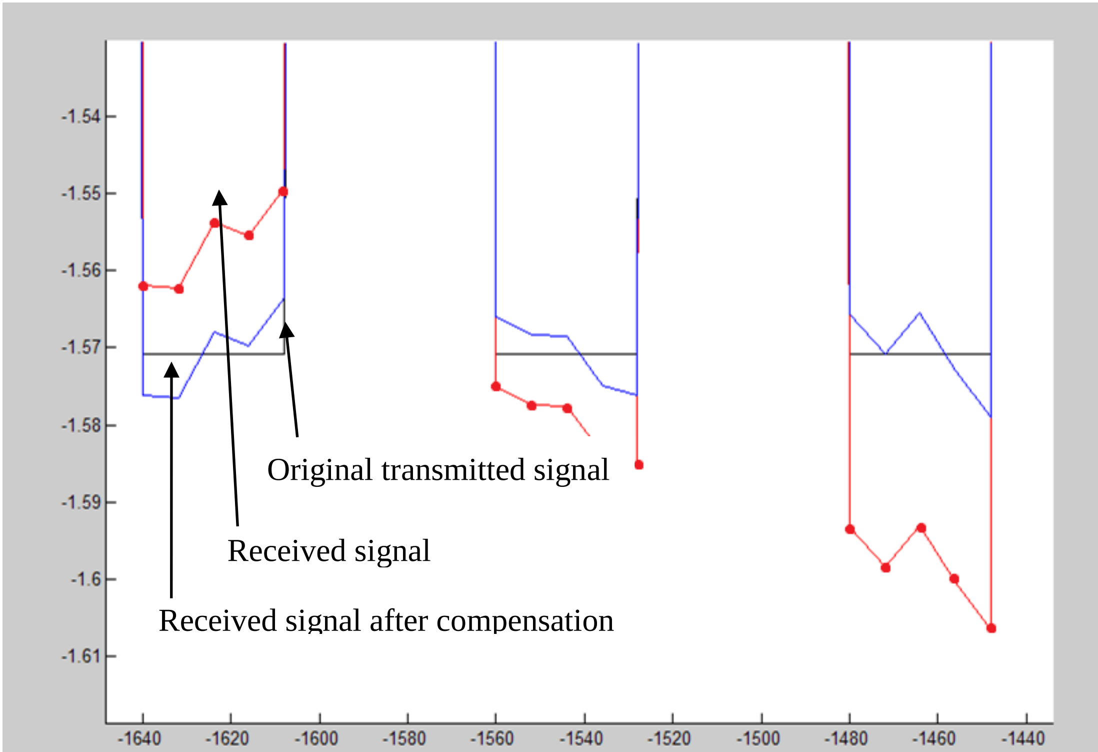


Figure 3.4 Illustration of compensated received signal, 25GHz system, laser line-width = 1 Mhz, phase noise compensator block length = 10 samples, M-th Power method.

3.3 Performance

Before testing the performance of our phase noise compensator, we need to optimize of its block size. A suitable block size should neither be too large which will degrade its ability to closely track slow varying laser phase noise, nor too small which will reduce the averaging effect on white noise. At our testing laser line-width, 1 MHz, we found out that a block size of 5~10 would be a suitable.

Testing and comparing the performance of laser phase noise compensator is quite difficult. Before we introduce the phase noise update limit, the simulation result is either perfect or totally ruined. Due to the nature of M-th power method, phase noise compensator without update limit has almost zero error tolerance. One error will ruin the signal tracking and result in wrong detection of all following symbols. Thus we introduce update limit to our compensator. After this, we have many performance tests and can generally summarize our conclusion in table 3.1.

Have Phase Update Limit				Don't Have Update Limit	
Block size=30	Block size=10	Block size=2	Sample by Sample	Block size=10	Sample by Sample
Perfect compensation. But isn't able to closely track laser phase noise drift.	Perfect compensation. Never yield an extra error in all simulation. Never lose track of signal phase.	Good. Slightly better than sample by sample scheme.	Normal. Performance drop rapidly when system noise increased.	Poor. Usually don't yield any error but not always the case, especially in noisy system.	Worst. Almost never able to stay in track in a 100kbit simulation in noisy system.

Table 3.1 Phase noise compensation scheme conclusion

Chapter 4

Dispersion Compensation

4.1 Chromatic Dispersion Model

Chromatic Dispersion or group-velocity dispersion (GVD) exhibit in any dielectric medium due to the fact that $v = \frac{c}{n}$, and n is the frequency dependent refractive index. Optical signal with different frequency will travel at different speed in media and arrive at receiver at different time. The time difference ΔT can be calculated by

$$\Delta T = L\beta_2\Delta\omega = LD\Delta\lambda \qquad (4.1)$$

where L is the propagation length (distance travelled), $\beta_2 = d^2\beta/d\omega^2$ is group-velocity dispersion parameter, $\Delta\omega$ is the frequency spread between two reference optical signals, D is the dispersion parameter and is usually expressed in units of ps/(km*nm), and $\Delta\lambda$ is wavelength difference between two reference optical signals.

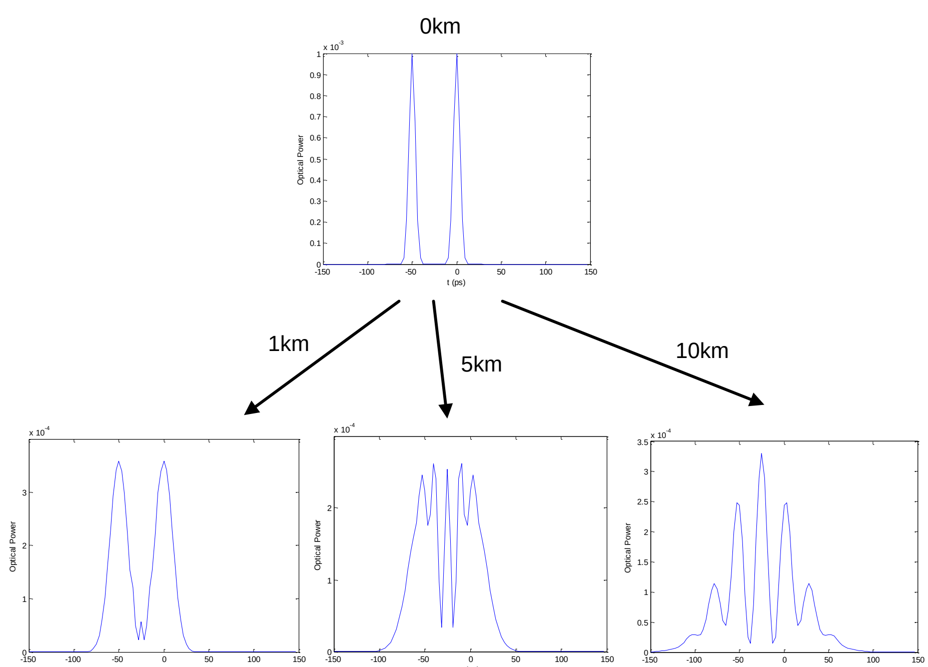


Figure 4.1: Illustration of a Gaussian pulses pass through a dispersive fiber, dispersion

length is 5 km, T_0 = 10 ps (20 Gbits/s 20% Gaussian RZ-OOK) [26]

Chromatic dispersion leads to pulse broadening effect as the different frequency components of the pulse arrive at receiver at different times, and thus causes inter-symbol interference (ISI). In our optic-fiber communication scheme, chromatic dispersion is one of the most significant distortion effects on optic signal, since we compensate all impairment effects at receiver side and thus allow the broadening effect to accumulate. Although chromatic dispersion is a major distortion effect, it is not a major performance limitation of our optic-fiber communication system. Due to the fact that chromatic dispersion is almost a fixed and predictable distortion effect, it can be compensated effectively via many methods. Initially, we do not compensate this effect individually and leave it to adaptive equalizer which will be covered in section 4.4 to 4.7. However, when we are simulation long-haul and ultra-long-haul fiber transmission system, we found out that the tap length of our adaptive equalizer has to be increased linearly to match up the chromatic dispersion. We know that keep updating an excessive long tap weight for adaptive equalizer is a big computational waste if it is avoidable. So we introduce an individual fixed chromatic dispersion compensator to compensate majority of chromatic dispersion (fixed part) and let the adaptive equalizer to handle the small fluctuation. In this scene, we are able to keep a small adaptive equalizer tap length.

4.2 Fixed Chromatic Dispersion Compensator

A general long-haul fiber system will have a severe chromatic dispersion, resulting in intensive inter-symbol interference (ISI). For example, a Gaussian pulse with T_0= 10ps, where T_0 is the half-width at 1/e intensity point, travels through a 1000 km long fiber with $\beta_2 = -21\, ps^2/km$. At the end of fiber, this Gaussian pulse will be broadened 200 times of the initial pules width. In order to compensate it, we need an equalizer, adaptive or fixed, with at least 200 taps. Thus, with a fixed CD compensation, we do not have to make our adaptive equalizer over 200 taps long. It will reduce the overall computation cost dramatically due to the fact that a fixed tap weight do not need to be adjusted nor updated at each iteration. By estimating the effective $\beta_2 \times L$, we can introduce an inverse dispersion to compensate the original dispersion just like dispersion compensating fiber (DCF).

4.3 Polarization-Mode Dispersion Model

In our communication scheme, polarization-multiplexing QPSK (PM-QPSK), polarization mode dispersion (PMD) is also an important distortion source that has to be compensated properly. As we stated in section 1.2, compensating PMD requires an active device that is able to self-adjust based on real time feedback, and the implementation of such device in optical domain is very expansive and complex. However, the implementation of such device in electrical domain is relatively much simpler and easier. A tap and delay equalizer will able to handle it alone with many other distortion effects.

4.4 Adaptive Equalizer

Generally speaking, inter-symbol interference exists in all optical communication system. A general tap and delay equalizer is the perfect device to compensate it and in fact all linear distortion can be compensated. In most cases, we will operate optic-fiber communication system in linear region where nonlinearity can be ignored. This is due to the fact that the nonlinearity is hard to compensate while it comes and interacts with other distortion effects and noises.

In a simple case where polarization multiplexing is not employed, PMD/PDL can be neglected, and polarization alignment is done, we can illustrate the adaptive tap and delay equalizer for this case in figure 4.2.

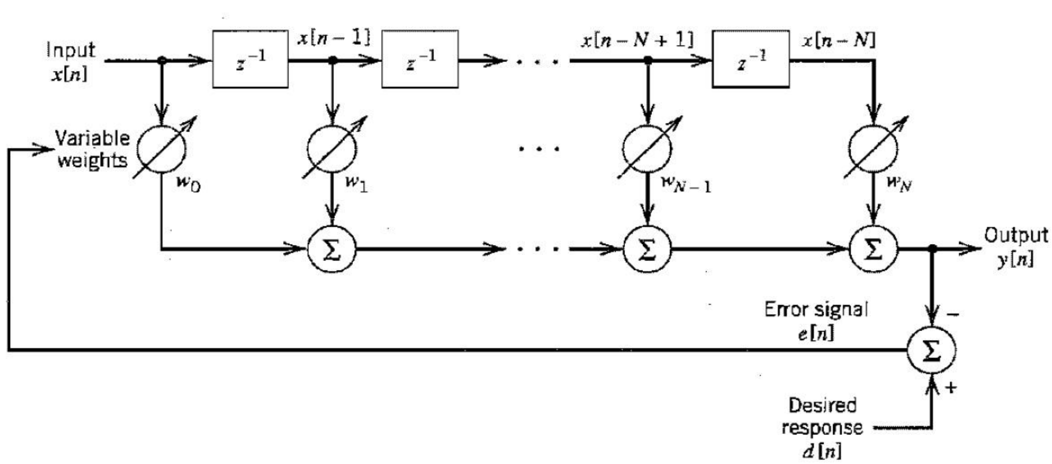


Figure 4.2 Structure map of a tap and delay equalizer with feedback.

$$y[n] = X^T[n]W[n] \tag{4.1}$$

$$e[n] = d[n] - y[n] \tag{4.2}$$

$$W[n+1] = F\{W[n], \mu, e[n], X[n]\} \tag{4.3}$$

where X[n]=[x[n-1], x[n-2], …, x[n-N]] is the received signal, N is tap length of the equalizer, y[n] is the equalized output signal, e[n] is signal error vector for tap weight update, d[n] is the expected signal waveform, W[n] is the tap weight at n-th iteration, the function F is different based on different update algorithms, μ is the step size of the equalizer.

In our case, PM-QPSK system, an adaptive tap and delay equalizer is implemented in such way as in figure 4.3.

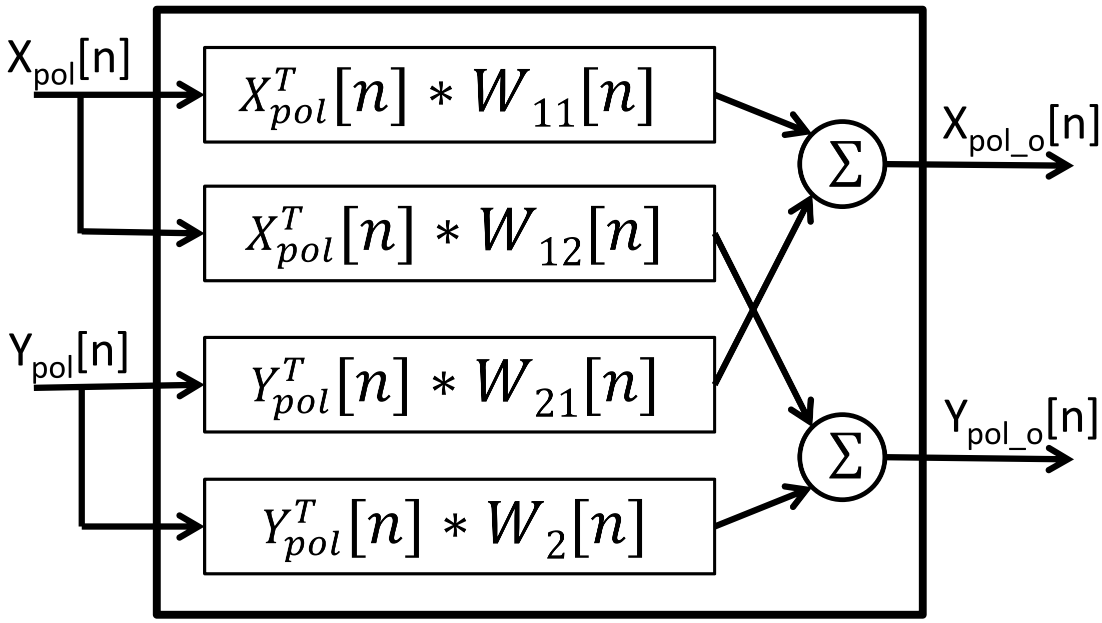


Figure 4.3 Block diagram of adaptive equalizer for coherent system.

Such scheme will also able to create an inverse jones matrix of fiber and thus able to compensate cross-polarization interference. It will also able to handle any polarization related distortion effect such as random polarization rotations.

4.5 System Overview

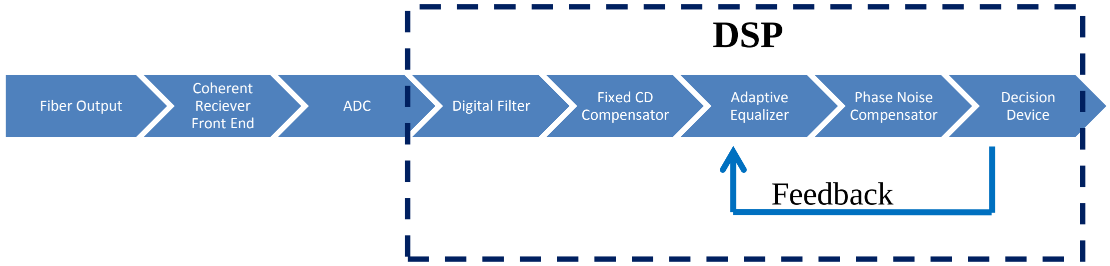


Figure 4.4 Block diagram of receiver layout using DSP compensation scheme

4.5.1 System with LMS adaptive equalizer

In our program, we implemented our DSP receiver according to figure 4.4. We employ an ideal electrical filter right after the analog-digital converter (ADC) to reduce total noise power. A fixed chromatic dispersion (CD) compensator is used after filter to compensate bulk CD in long-haul communication system. It will remove majority of the chromatic dispersion effect from signal and pass a relatively low distortion signal to adaptive equalizer. The adaptive equalizer will then handle everything including time-varying chromatic dispersion, laser phase noise, polarization- mode dispersion (PMD), and other linear slow varying distortion effects. After equalizer, a phase noise compensator is usually used to further tighten the signal from nearby constellation point. However, phase noise compensator is not a must-have compensation device in the LMS scheme since LMS adaptive equalizer also has the ability to handle phase noise and phase rotation. In fact, via many different system simulations, we can conclude that if the LMS adaptive equalizer is optimized, a phase noise compensator is not necessary. In CMA equalizer system, this statement is not valid, and we will cover this in more detail in section 4.5.2. In optimal state, phase noise compensator may or may not improve system bit error rate (BER) performance. Thus, phase noise compensator should be considered as an optional device that ensures system performance when LMS adaptive equalizer optimization is not guaranteed.

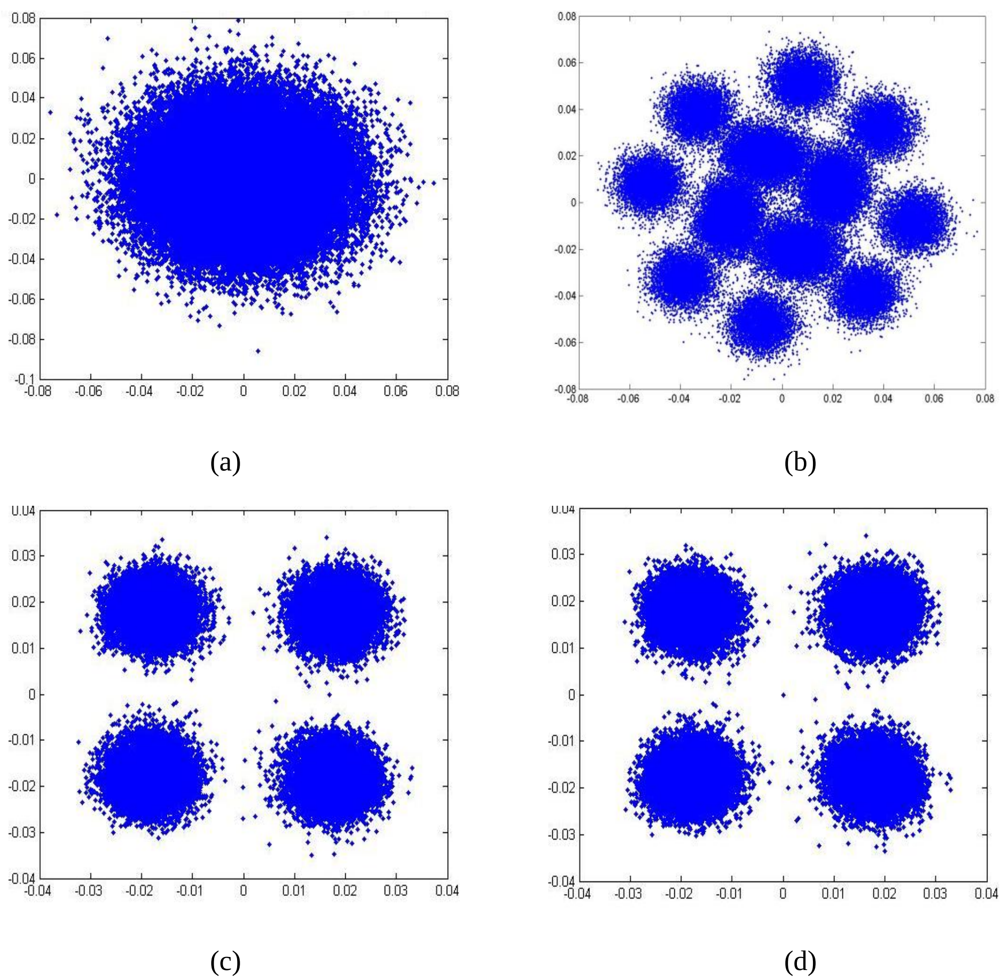


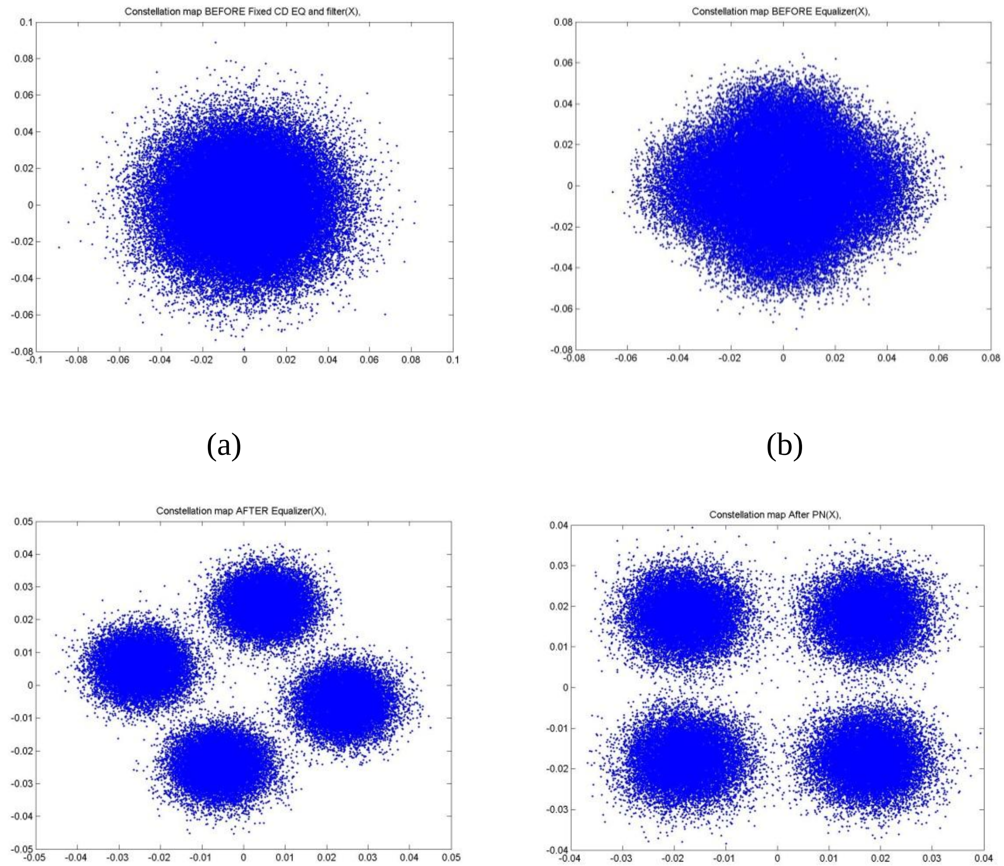
Figure 4.5 Constellation map of received signal at different LMS receiver stage.

Figure 4.5 (a) shows the signal constellation map after digital filter (before any compensation). At this stage, the received signal is totally unreadable. Figure 4.5 (b) shows the signal constellation map after fixed CD compensation. At this stage, major chromatic dispersion effect has been removed and the signal becomes grouped and somewhat distinguishable. Figure 4.5 (c) shows the signal constellation map after adaptive Least Mean Square (LMS) equalizer. Most linear impairments have been compensated now, and the signal is almost ready for detection. Figure 4.5 (d) shows the signal constellation map after phase noise compensator. The signal constellation map is more tightened and further away from nearby constellation point now. This will ensure overall better system performance. Simulation parameters for figure 4.5: 30Gbit/s each channel, PM-QPSK (4 channels), LMS adaptive equalizer used, launch power 0dbm, transmitter and receiver laser line-width 1Mhz, ASE nsp=1.5, CD $\beta_2 = 21e - 27 ps^2/km$, nonlinearity $\gamma = 1.1\,W^{-1}/km$, PMD $D_p = 1\ ps/\sqrt{km}$, lose $\alpha = 0.2 dB/km$, 96k bits, $\mu = 0.2$, adaptive equalizer tap length =10, fiber length L = 10*115km. Figure 4.5 only shows the constellation maps for X polarization. Launch power is the total power of two polarization signal unless otherwise stated.

After phase noise compensator (or adaptive equalizer if PN compensator is not employed), signal is ready to pass through decision module. In decision module, signal will be read based on maximum likelihood principle and the result will send back to adaptive equalizer to generate reference signal d[n]. However, in our LMS algorithm, we use soft decision result to generate reference signal d[n] instead of hard decision result from output. Since the LMS algorithm is not capable of blind-detection, a training sequence is required to train adaptive equalizer till the initial equalizer tap weight converged. This training sequence is usually pre-defined and known. Receiver will use this training sequence as desired signal d[n] in training stage. After a certain length of training sequence have been applied (15k bits length in our program), the receiver change to receiving stage and start using soft decision result to generate desired signal d[n].

4.5.2 System with CMA adaptive equalizer

Our CMA adaptive equalizer still follows the same receiver scheme described in figure 4.4. As we mentioned in section 4.5.1, a phase noise compensator is no longer an optional device in this scheme. Because CMA adaptive equalizer does not keep track on current receiving bit and do not know the detail of default constellation point, it can only restore the signal towards a QPSK scheme with a random rotation remained. A phase noise compensator must be employed after CMA adaptive equalizer to remove this random rotation (figure 4.6).



(a) (b)

(c) (d)

Figure 4.6 Constellation map of received signal at different CMA receiver stage.

Figure 4.6 (a) shows the signal constellation map after digital filter (before any compensation). At this stage, the received signal is totally unreadable. Figure 4.6 (b) shows the signal constellation map after fixed CD compensation. At this stage, major chromatic dispersion effect has been removed and the signal becomes grouped and somewhat distinguishable. Figure 4.6 (c) shows the signal constellation map after adaptive CMA equalizer. The most linear impairments have been compensated, but the signal is rotated randomly from default constellation point. Figure 4.6 (d) shows the signal constellation map after phase noise compensator. The signal constellation map is correctly rotated back to default constellation point now, and the signal is ready for detection. Simulation parameters for figure 4.6: 30Gbit/s each channel, PM-QPSK (4 channels), launch power = -1dbm, transmitter and receiver laser line-width 1Mhz, ASE nsp=1.5, CD $\beta_2 = 21e-27ps^2/km$, nonlinearity $\gamma = 1.1\,W^{-1}/km$, PMD D_p = 1 $ps/\sqrt{km}$, lose $\alpha = 0.2dB/km$, 96k bits, $\mu = 0.2$, adaptive equalizer tap length =10, fiber length L = 10*115km. Figure 4.6 only shows the constellation maps for X polarization.

4.6 Least Mean Square Adaptive Equalizer

Comparing to constant modulus algorithm (CMA), least mean square (LMS) algorithms has many advantages including better stability, better performance, and faster initialization speed. On the other hand, CMA is capable of blind detection and able to self-recover in any case owing to its blind detection capability. There is no way to say one is better than the other. It is just the matter of which one is best fit your need. We implemented both solutions for our program.

As we discussed in section 4.4, a butterfly formation equalizer is used in our PM-QPSK system. According to figure 4.3, we have

$$X_{pol\_o}[n] = X_{pol}^T[n] * W_{11}[n] + Y_{pol}^T[n] * W_{21}[n] \quad (4.4)$$

$$Y_{pol\_o}[n] = X_{pol}^T[n] * W_{12}[n] + Y_{pol}^T[n] * W_{22}[n]$$

$$Let\ \ \widetilde{W}[n] = \begin{bmatrix} W_{11}[n] & W_{12}[n] \\ W_{21}[n] & W_{22}[n] \end{bmatrix}$$

$$I[n] = \begin{bmatrix} X_{pol}^T[n] \\ Y_{pol}^T[n] \end{bmatrix}$$

$$O[n] = \begin{bmatrix} X_{pol\_o}[n] \\ Y_{pol\_o}[n] \end{bmatrix}$$

$$Then\ O[n] = \widetilde{W}^T[n] I[n] \quad (4.5)$$

Where W_{11}, W_{12}, W_{21}, and W_{22} are the tap weighs of adaptive equalizer, X[n] and Y[n] is the input signal from x and y polarization, X_{out}[n] and Y_{out}[n] is the compensated signal output from equalizer. These formulas are true for both LMS and CMA adaptive equalizer.

Specifically, LMS update its tap weigh according to equation 4.6 and 4.7.

$$Assume \qquad d[n] = \begin{bmatrix} {X_{ref}}^T[n] \\ {Y_{ref}}^T[n] \end{bmatrix}$$

$$Thus \quad e[n] = d[n] - O[n] \quad (4.6)$$

$$and \quad \widetilde{W}[n+1] = \widetilde{W}[n] + \mu I^*[n] e^T[n] \quad (4.7)$$

where d[n] is the reference signal, e[n] is the error vector.

A major disadvantage of LMS algorithm is that it requires a reference signal (desired response) to compare with when updating its tap weigh. How should this reference signal be generated is important to LMS algorithm. In our program, we employ two decision devices for two purposes. The first one is the dedicated soft decision module which will detect the estimated output signal of next bit, and then generate the desired response d[n+1] based on the detection result. The adaptive equalizer will use this d [n+1] to calculate new e[n+1] and $\widetilde{W}[n+1]$, and use the updated tap weigh to equalize I [n+1]. The result of them will now send to the second decision module (hard decision) for final output. Similar to phase noise compensator, the separation of soft and hard decision is not mandatory, and does not improve system performance dramatically.

4.6.1 Performance and optimization of LMS adaptive equalizer

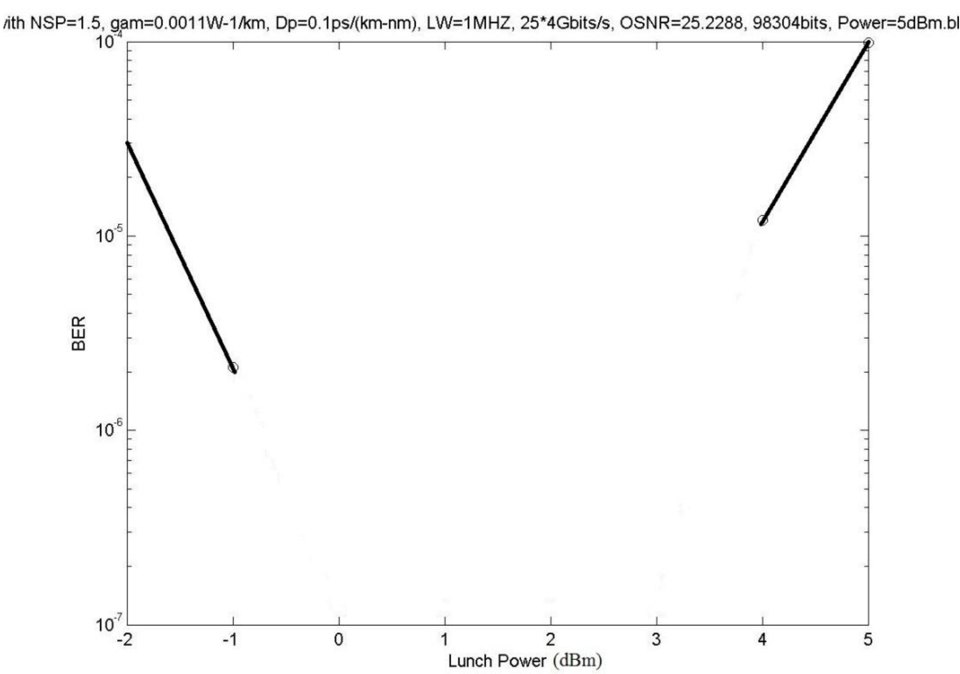


Figure 4.7 BER vs. launch power optimization. Simulation parameters: 25Gbit/s each

channel, PM-QPSK (4 channels), launch power from -2dbm to 5dbm, transmitter and receiver laser line-width 1Mhz, ASE nsp=1.5, CD $\beta_2 = 21e - 27ps^2/km$, nonlinearity $\gamma = 1.1\,W^{-1}/km$, PMD D_p = 0.1 $ps/\sqrt{km}$, lose $\alpha = 0.2dB/km$, 96k bits, $\mu = 0.2$, adaptive equalizer tap length =10, fiber length L = 10*115km.

The center region in figure 4.7 (launch power from 0dbm to 3 dBm) is blank due to the limitation of low simulation length which limits the minimum BER our simulation program could obtain. In that power range, our simulations yield zero error and thus the optimal launch power lies in that range. Because the BER is extremely low when operating near optimal power range, our following simulation will be performed under lower launch power.

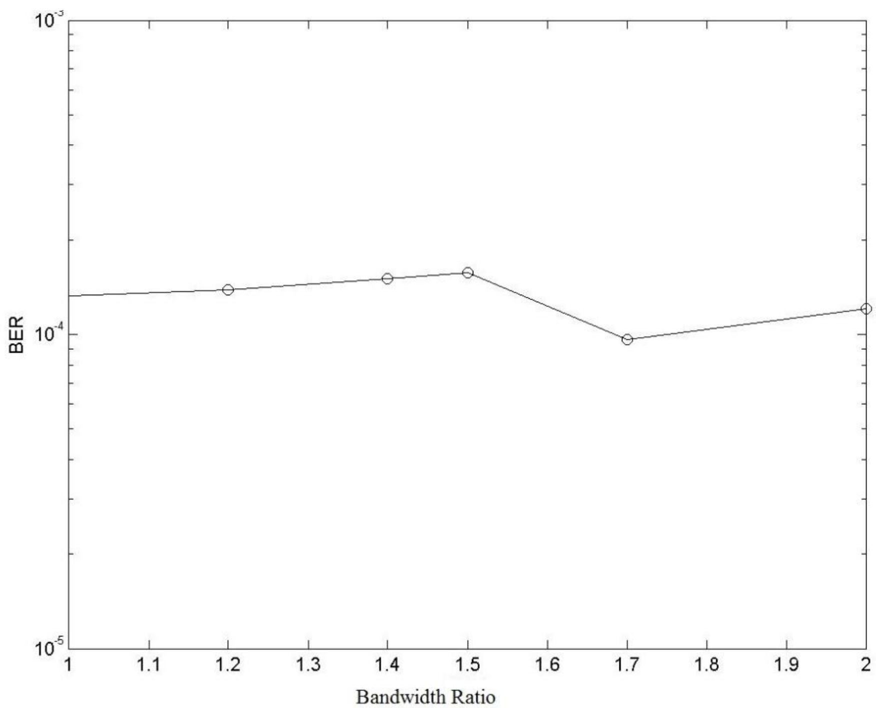


Figure 4.8 BER vs. filter bandwidth ratio optimization. Simulation parameters: 25Gbit/s each channel, PM-QPSK (4 channels), base bandwidth = 50Ghz, applied bandwidth = base bandwidth * bandwidth ratio, launch power from -3dbm, transmitter and receiver laser line-width 1Mhz, ASE nsp=1.5, CD $\beta_2 = 21 \times 10^{-27} ps^2/km$, nonlinearity $\gamma = 1.1\, W^{-1}/km$, PMD D_p = 0.1 $ps/\sqrt{km}$, lose $\alpha = 0.2dB/km$, 96k bits, $\mu = 0.2$, adaptive equalizer tap length =10, fiber length L = 10*115km.

In this simulation, we notice that the filter bandwidth nearly does not affect out system performance. In this case, we decided to set our bandwidth ratio to be 2 for future simulation.

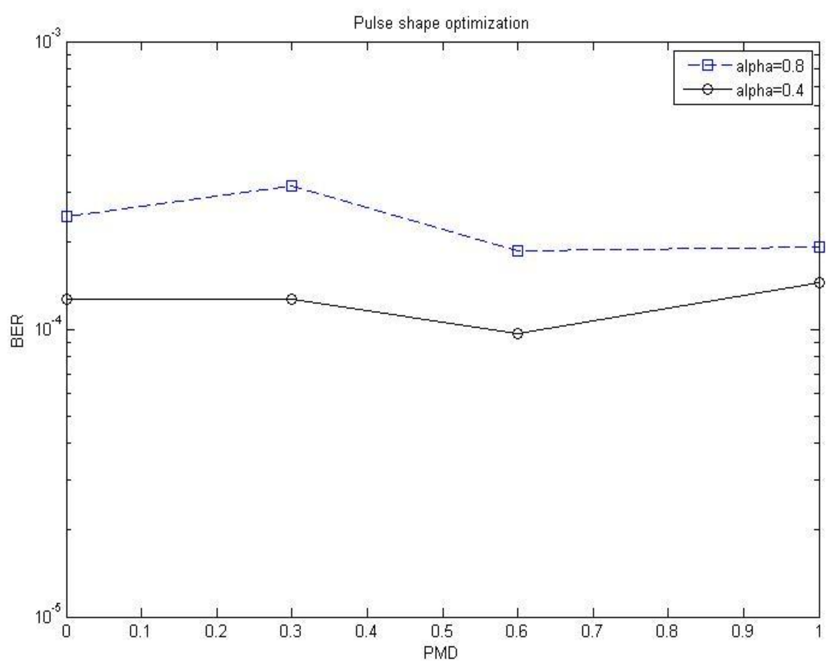
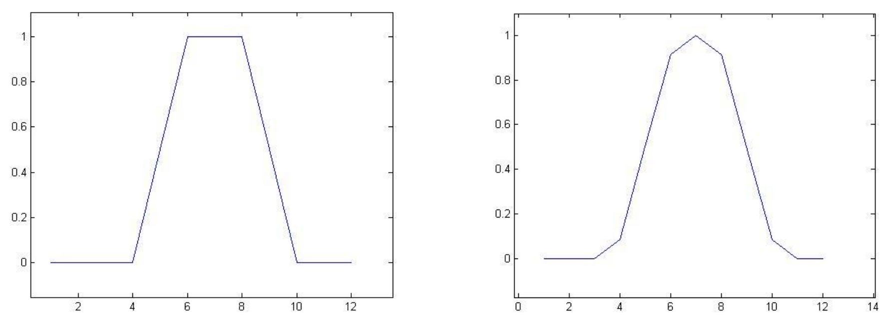


Figure 4.9 Pulse shape optimization. Simulation parameters: 25Gbit/s each channel,

PM-QPSK (4 channels), raised cosine pulse, applied bandwidth = base bandwidth * bandwidth ratio, launch power from -3dbm, transmitter and receiver laser line-width 1Mhz, ASE nsp=1.5, CD $\beta_2 = 21e - 27ps^2/km$, nonlinearity $\gamma = 1.1\,W^{-1}/km$, PMD D_p = 0 ~ 1 $ps/\sqrt{km}$, lose $\alpha = 0.2dB/km$, 96k bits, $\mu = 0.2$, adaptive equalizer tap length =10, fiber length L = 10*115km.



(a) (b)

Figure 4.10 Illustration of two pulse shapes. (a) alpha = 0.4, shaper (b) alpha = 0.8 smoother.

In this simulation, we not only found out that pulse shape (a) shown in figure 4.10 yields better performance, but also found out its interference with PMD. In our earlier simulations, our simulation results always showed low PMD penalties. The lower the PMD distortion effect the system has, the lower overall performance they system will yield. This strange effect lead to this optimization which we found out that although smoother pulse shape will reduce the signal bandwidth and should improve the performance, it brings in another bigger problem that it lead to more ISI effect when PMD exists. Thus, we choose figure 4.10 (a) pulse scheme for our simulations.

4.7 Constant Modulus Adaptive Equalizer

The major difference between CMA and LMS algorithm is how it updates the adaptive equalizer tap weight. Due to the unique ability of CMA, CMA adaptive equalizer does not need a desired signal (reference signal) to update itself which cause CMA equalizer a blind-detection capable equalizer. The basic idea of CMA comes from the nature of PSK system. All PSK constellation points have same modulus (distance from origin) but different phase. Thus, CMA will calculate the cost (difference between compensated signal modulus and standard modulus) of the current compensated signal, and then update the tap weight to adapt the distortion effect. The blind-detection ability comes from the fact that the CMA cost estimation does not rely on any signal approximation but only the constant modulus, which is constant and specified in system preset. CMA also natively only support PSK modulation scheme, but later advancement also enable it to be used in some other modulation schemes like ASK by using complex cost function. However, using PSK scheme can achieve the best CMA convergent rate and stability.

$$cost[n] = \begin{bmatrix} \frac{(1-|X_{out}[n]|)^2}{M_{constant}} \\ \frac{(1-|Y_{out}[n]|)^2}{M_{constant}} \end{bmatrix} \tag{4.8}$$

$$W_{CMA}[n] = W_{CMA}[n-1] + \mu * \begin{bmatrix} X[n] \\ Y[n] \end{bmatrix}^* * cost[n]^T \tag{4.9}$$

where cost[n] means the cost function of current sample, $M_{constant}$ means the standard modulus of the system, $[\ ]^*$ means complex conjugate, $[\ ]^T$ means transpose.

With the same system configuration, the optimization result of CMA equalizer is generally the same as LMS. The detailed result will not present here for this reason.

4.8 Performance comparison of LMS and CMA

The following figure 4.11 shows a general system performance of LMS and CMA adaptive equalizer. Generally, CMA has worse system performance than LMS algorithm due to the fact that it does not keep track of current detection result even after system converged. The blind-detection nature limits its compensation ability. However, the LMS algorithm shows a nearly perfect system performance. Its performance almost overlapped with ASE only case, meaning the LMS equalizer is almost able to compensate all linear impairment effects. Another important advantage of LMS which is not shown in figure 4.11 is that the system stability. Once LMS equalizer have been optimized and converged, it will almost never diverge again, and its maximum convergent time is limited and can be calculated. However, the CMA doesn’t have such characteristic. The convergent time of

CMA generally decreased as the step size increased, but its stability also decreased meanwhile.

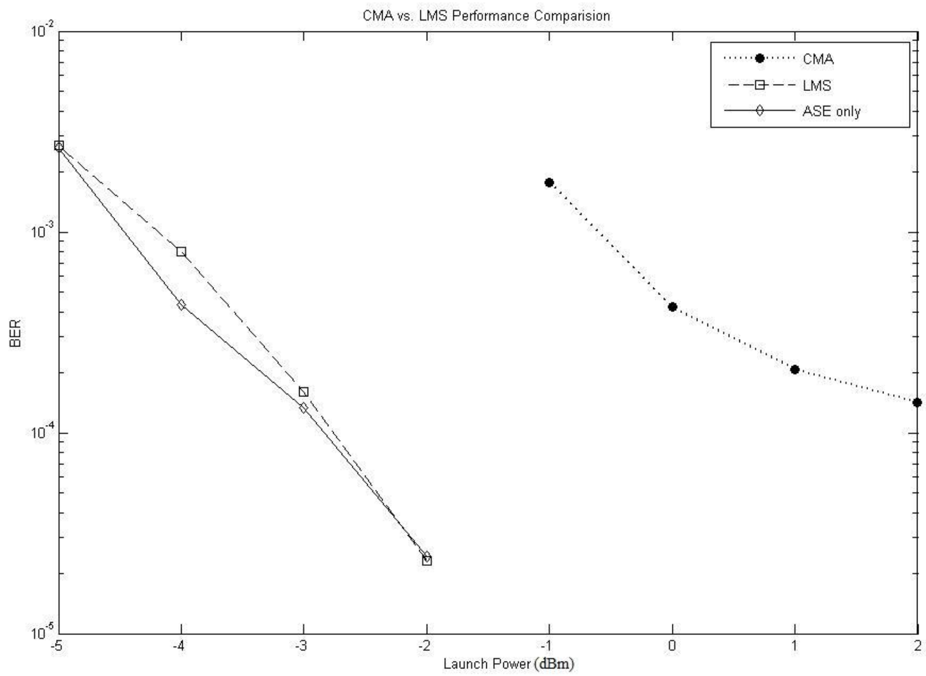


Figure 4.11 Performance comparison of CMA and LMS adaptive equalizer. Simulation parameters: 96k bits simulation length, 4 samples/symbol before ADC, 2 samples/symbol after ADC, adaptive equalizer training length = 15k bits, 30G symbol/s per polarization PM-QPSK system (120G bit/s total), laser line-width = 1MHz, number of WDM channels = 1, 1550nm system, ASE nsp=1.5, CD $\beta_2 = 21e - 27$, nonlinearity $\gamma = 1.1\,W^{-1}/km$, PMD D_p = 1 ps/(km-nm), lose $\alpha = 0.2dB/km$, $\mu = 0.2$, adaptive equalizer tap length =10, fiber length L = 10*115km. Note: CMA performance penalty also comes from higher power nonlinearity effect which is not handled in receiver.

Chapter 5

Conclusion and Future Work

The laser frequency drift (laser phase noise), chromatic dispersion, and polarization mode dispersion of the optical fiber could lead to performance degradation. In this thesis, various electrical equalization techniques to mitigate these impairments are discussed.

In chapter 3, various techniques to compensate the laser phase noise are presented. We discussed the M-th power method, unwarping algorithm, and update step size control technique used in phase noise compensation module. Performance comparison of different system configurations is also given in the end. We concluded that the block phase noise compensation scheme with a block size of 10 and phase update limit enforced is the best of all. We used this scheme in our following simulations.

In chapter 4, we discussed the implementation of both LMS and CMA adaptive equalizer solution, and the corresponding phase noise compensator for both equalizers. Although wavelength-division multiplexing (WDM) is not discussed in this thesis, such system configuration is also supported in this tool. With this simulation tool, we are able to compare the performance of many different system configurations, and examine the impact of a specify system parameter change. Also, we are able to prove the performance of DSP compensation device in varies situations.

Our results show that the DSP compensation scheme in coherent polarization multiplexing QPSK is very effective in compensating linear distortion effects. The LMS adaptive equalizer scheme has almost perfect compensating capability, while the CMA adaptive equalizer scheme has bind-detection ability. Although the current version simulation tool does not have the ability to use LMS and CMA adaptive equalizer together, it still would be a very good future work direction. Combining the LMS and CMA algorithm together, by using CMA in training stage and LMS in receiving stage, might combine the advantage of both algorithms together. It will give such DSP receiver not only the bind initialization and the capability to automatically recover from errors from CMA, but also the near perfect compensation performance and stability from LMS.

Finally, there are still many directions we can improve our equalization techniques such as experimental validation, additional function and mode support, and/or improved simulation speed.

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