DSP based Chromatic Dispersion Equalization and Carrier Phase Estimation in High Speed Coherent Optical Transmission Systems

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Abstract

Coherent detection employing multilevel modulation formats has become one of the most promising technologies for next generation high speed transmission systems due to the high power and spectral efficiencies. Using the powerful digital signal processing (DSP), coherent optical receivers allow the significant equalization of chromatic dispersion (CD), polarization mode dispersion (PMD), phase noise (PN) and nonlinear effects in the electrical domain. Recently, the realizations of these DSP algorithms for mitigating the channel distortions in the coherent transmission systems are the most attractive investigations.

The CD equalization can be performed by the digital filters developed in the time and the frequency domain, which can suppress the fiber dispersion effectively. The PMD compensation is usually performed in the time domain with the adaptive least mean square (LMS) and constant modulus algorithms (CMA) equalization. Feed-forward and feed-back carrier phase estimation (CPE) algorithms are employed to mitigate the phase noise (PN) from the transmitter (TX) and the local oscillator (LO) lasers. The fiber nonlinearities are compensated by using the digital backward propagation methods based on solving the nonlinear Schrödinger (NLS) equation and the Manakov equation.

In this dissertation, we present a comparative analysis of three digital filters for chromatic dispersion compensation, a comparative evaluation of different carrier phase estimation methods considering digital equalization enhanced phase noise (EEPN) and a brief discussion for PMD adaptive equalization. To implement these investigations, a 112-Gbit/s non-return-to-zero polarization division multiplexed quadrature phase shift keying (NRZ-PDM-QPSK) coherent transmission system with post-compensation of dispersion is realized in the VPI simulation platform. In the coherent transmission system, these CD equalizers have been compared by evaluating their applicability for different fiber lengths, their usability for dispersion perturbations and their computational complexity. The carrier phase estimation using the one-tap normalized LMS (NLMS) filter, the differential detection, the block-average (BA) algorithm and the Viterbi-Viterbi (VV) algorithm is evaluated, and the analytical predictions are compared to the numerical simulations. Meanwhile, the phase noise mitigation using the radio frequency (RF) pilot tone is also investigated in a 56-Gbit/s NRZ single polarization QPSK (NRZ-SP-QPSK) coherent transmission system with post-compensation of chromatic dispersion. Besides, a 56-Gbit/s NRZ-SP-QPSK coherent transmission system with CD pre-distortion is also implemented to analyze the influence of equalization enhanced phase noise in more detail.
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Chapter 1

Introduction

The performance of high speed optical fiber transmission systems is severely affected by fiber attenuation, chromatic dispersion (CD), polarization mode dispersion (PMD), phase noise (PN) and nonlinear effects. Coherent optical detection allows the significant equalization of the transmission system impairments in the electrical domain, and has become one of the most promising techniques for the next generation communication networks. With the full optical wave information, the fiber dispersion, the carrier phase noise and the fiber nonlinear effects can be well compensated by the powerful digital signal processing (DSP).

In this chapter, we will give a short description for the structure of the dissertation. Meanwhile, we will present an overview for the history and the state-of-the-art of the optical fiber communication systems and the coherent transmission technologies. We will also make a discussion about the attractive techniques for mitigating the system distortions in the development of the high speed coherent communication systems. Furthermore, we will make a summary of our research work in the high speed coherent transmission systems with the post- and the pre-compensation of chromatic dispersion.

1.1 Structure of thesis

In this dissertation, we present a detailed study in the DSP algorithms for mitigating the system impairments in the high speed coherent optical transmission systems. Our research mainly focuses on the chromatic dispersion compensation and the carrier phase estimation (CPE) in the coherent transmission systems with post- and pre-equalization of dispersion. We perform a comparative analysis on different CD equalization algorithms, and also describe a comparison on different carrier phase estimation methods considering the equalization enhanced phase noise (EEP).

In chapter 1, we give a brief introduction on the history and the development of the optical communication systems and the coherent transmission technologies. The state-of-the-art of the coherent systems and the attractive techniques in coherent detection are discussed in this part. Meanwhile, the DSP algorithms for chromatic dispersion compensation and phase noise mitigation are also briefly introduced.

In chapter 2, the influence of the fiber impairments and the system distortions on the high speed coherent communication systems is described in detail. The impacts of the fiber attenuation, the chromatic dispersion, the polarization mode dispersion, the phase noise and the nonlinear effects on the transmission systems are analyzed and discussed respectively. This gives a brief overview of the basic knowledge for our research.
In chapter 3, we present three implementations of high speed coherent transmission systems, involving a 112-Gbit/s non-return-to-zero polarization division multiplexed quadrature phase shift keying (NRZ-PDM-QPSK) coherent system with the post-compensation of chromatic dispersion, a 56-Gbit/s NRZ single polarization QPSK (NRZ-SP-QPSK) post-compensated coherent system with radio frequency (RF) pilot tone for phase noise mitigation, and a 56-Gbit/s NRZ-SP-QPSK coherent system with the pre-distortion of chromatic dispersion. They are all realized in the VPI platform. Meanwhile, we describe a mathematical analysis for the standard QPSK transmitter, the fiber channel and the coherent receiver based on the 112-Gbit/s NRZ-PDM-QPSK coherent system with CD post-compensation, which is also suitable for other QPSK transmission systems.

In chapter 4, the theoretical basic for our research work is described by analyzing the corresponding DSP algorithms in detail. The mitigation of the chromatic dispersion using the least mean square (LMS), the fiber dispersion finite impulse response (FD-FIR), and the blind look-up (BLU) filters are analyzed comparatively. The compensation of the carrier phase noise using the one-tap normalized LMS (NLMS), the differential detection, the block-average (BA) and the Viterbi-Viterbi (VV) methods are also significantly elaborated. Meanwhile, the adaptive equalization of the polarization mode dispersion using the LMS and the constant modulus algorithm (CMA) filters is also discussed briefly.

In chapter 5, we present the numerical simulation results in the 112-Gbit/s NRZ-PDM-QPSK coherent optical transmission system with post-compensation of dispersion employing the above DSP algorithms for chromatic dispersion compensation and carrier phase estimation. Meanwhile, the phase noise mitigation using the radio frequency (RF) pilot tone in the 56-Gbit/s NRZ-SP-QPSK coherent system with post-compensation of CD, and the carrier phase noise compensation in the 56-Gbit/s NRZ-SP-QPSK pre-distorted transmission system are also performed for the detailed analysis of the equalization enhanced phase noise (EEPN).

In chapter 6, we make a summary about our research work in this dissertation, and present a brief overview of our publications. Moreover, we also give some suggestion and plans for our future investigations in coherent transmission technologies.

1.2 Historical background

In this section, we will give a brief introduction for the history and the evolution of the optical fiber communications and the coherent transmission techniques during the past fifty years. Thus we could have a good background and understanding for the development of the optical fiber networks and the coherent optical communication systems.

1.2.1 History of optical fiber communication systems

Although A. G. Bell and his assistant C. S. Tainter have created a primitive precursor (photophone) to fiber-optic communications during the 1880s [1,2], the modern
optical fiber communication systems were born in the 1960s due to the inventions of the lasers and the applications of the glass fibers [1-4].

In 1960, the invention of the laser offered a coherent optical source for the optical fiber communication systems [3]. In 1966, C. K. Kao and G. Hockham, from standard telecommunication laboratory (STL) in England, proposed that optical fiber might be a suitable transmission medium if its attenuation could potentially be removed to an acceptable value [4]. At the time of this proposal, the loss in optical fibers is around 1000 dB/km [5,6].

The breakthrough of optical fiber communications occurred in 1970, when the optical fiber was successfully developed by Corning Glass Works, with the attenuation of 20 dB/km [7,8]. Meanwhile, the GaAs semiconductor lasers were invented, which were suitable for emitting light into fiber optic cables for long distances transmission [9,10].

The first generation of the mature lightwave systems was developed in 1975, which operated near 0.8 µm and used GaAs semiconductor lasers [7]. This first-generation system can reach a bit rate of 45-Mbit/s and a repeater spacing of up to 10 km [9].

The second generation of the fiber-optic communication systems was developed in the early 1980s, which operated near 1.3 µm and used InGaAsP semiconductor lasers [11,12]. The fiber loss in the 1.3 µm region is below 1 dB/km. The transmission speed (< 100-Mbit/s) in the initial communication systems were limited by the dispersion in the multi-mode fibers [13]. The development of the single-mode fibers overcame this problem in 1981 [14]. By 1987, the second-generation optical communication systems reached the bit rate of up to 1.7-Gbit/s and the repeater spacing of up to 50 km [5].

The third generation of fiber-optic communication systems was developed in the 1980s, which operated near 1.55 µm and used the improved InGaAsP semiconductor lasers [12]. The fiber loss in the 1.55 µm region is about 0.2 dB/km. The dispersion problem was solved by using the dispersion-shifted fibers or by limiting the laser spectrum to a single longitudinal mode [1,6]. The third-generation systems reached the bit rate of 2.5-Gbit/s and the repeater spacing of 100 km [11,12].

The fourth generation of optical fiber communication systems was developed in the recent 20 years, which employed the optical amplifiers to reduce the number of repeaters and used the wavelength-division multiplexing (WDM) techniques to increase the channel capacity [1,5,12]. These applications resulted in a revolution on the increment of the capacities in the commercial communication systems. Moreover, the polarization division multiplexing (PDM) and the space division multiplexing (SDM) techniques have also been investigated in the recent reports [15,16]. By using these multiplexing techniques (WDM&PDM&SDM), the bit-rate of up to 109-Tbit/s has been achieved over a single 16.8 km fiber until 2011 [17].

1.2.2 History of coherent optical communications

Due to the high sensitivity of the receivers, coherent optical transmission systems
were investigated extensively in the eighties of last century [18,19]. However, the development of the coherent technologies has been delayed for nearly 20 years after that period [20-22]. Until 2005, the coherent transmission techniques attracted the interests of investigation again [23], since the efficient modulation formats such as $m$-ary phase shift keying (PSK) and quadrature amplitude modulation (QAM) were implemented by employing the digital coherent receivers. Meanwhile, full access to the optical wave information offers the possibility of electrical compensation for transmission impairments as powerful as traditional optical compensation techniques. Due to the two main merits, the reborn coherent detections brought us the enormous potential for higher transmission speed and spectral efficiency in the present optical fiber communication systems [20-22].

1. Coherent optical communications in last century

With an additional local oscillator (LO) source, the sensitivity of the coherent receiver was achieved to the limitation of the shot-noise. Furthermore, compared to the traditional intensity modulation direct detection (IMDD) system, the phase detection system can also improve the receiver sensitivity because the distance between symbols is extended by the use of the signal phasors on the complex plane [19]. The multi-level modulation formats such as quadrature phase-shift keying (QPSK) and QAM can be applied into optical fiber transmission systems by using the phase modulation modules, which can include more information bits in one transmitted symbols than before.

However, the advantages of the traditional coherent optical receivers grew fainter due to the invention of erbium-doped fiber amplifiers (EDFAs) [21]. The sensitivity of coherent receivers limited by the shot-noise become less significant, because the signal-to-noise ratio (SNR) in the WDM transmission channel using EDFAs is determined by the accumulated amplified spontaneous emission (ASE) noise, which is smaller than the shot noise [20-22]. Moreover, some technical difficulties in the realization of coherent optical receivers have also prevented the development of coherent detection. For example, the coherent optical receivers are rather difficult to implement due to the high complexity and cost in stable locking of the rapid carrier phase drift. At the same time, the EDFA-based fiber communication systems employing WDM techniques played the dominant roles in the optical transmission techniques during the nineties in the last century.

2. Rebirth of coherent optical communications

Recently, there has been a renewed interest in coherent optical communication systems, due to the increment of the transmission capacity in the WDM systems [20]. With the demand of the ever-increasing bandwidth, the multi-level modulation formats based on the coherent detection need to be employed in the transmission systems to improve the spectral efficiency [24].

The first revival of the investigations in coherent optical communications comes from the differential QPSK (DQPSK) transmission experiment with the optical in-phase &
quadrature (IQ) modulation and the optical delay detection [22,25]. We can duplicate the bit rate with keeping the same symbol rate because the optical signal can carry two or more bits in one transmitted symbol. The next step of coherent technologies rebirth arises from the high-speed digital signal processing [22,23]. With the rapid development of high-speed integrated circuits, treating the electrical signal in a digital signal processing core and retrieving the IQ components from the optical carrier become feasible. Using a phase-diversity homodyne receiver (intradyne receiver) followed by the DSP circuit, the demodulation of the 10-Gsymbol/s QPSK signal with the offline digital signal processing has been realized [26,27]. Meanwhile, more advanced and powerful DSP circuits are developed, and this can provide us with more efficient methods for carrier phase estimation to substitute the optical phase-locked loop (PLL) [20-22].

3. State-of-the-art of coherent transmission technologies

The main benefit of the digital coherent receivers is the digital signal processing function [20-22]. The demodulation process is entirely linear in the coherent receivers, and all information of the transmitted optical signal including the state of polarization (SOP) is preserved. The signal processing techniques such as tight spectral filtering [23], chromatic dispersion compensation [24-29], polarization mode dispersion compensation and phase noise mitigation can be performed in the electrical domain after the coherent detection [30,31].

Once in the conventional coherent receiver, the polarization management turned out to be one of the main obstacles for the practical implementation [22,28]. For WDM systems, each channel requires a dedicated dynamic polarization controller, and this severely limits the practicality of the coherent receivers. In the digital coherent receivers, the polarization control can be solved by using the electrical adaptive polarization alignment, which is realized with much lower complexity and cost [29].

The next issue is the possibility and the applicability of the coherent transmission systems for any type of multi-level modulation formats. Besides the QPSK modulation format, the 8-PSK and the 16-QAM formats are also examined at 10-Gsymbol/s in the coherent systems [22,30-33]. Note that polarization multiplexing can always double the bit rate as mentioned before.

The most important technical issue is the real-time operation of the digital coherent receivers, which depends on the computing speed of the analog-to-digital convertors (ADCs) and the DSP components. Now the novel components come out, for example the 64-GSample/s 16-bit ADCs have been demonstrated by Opnext Inc. company, and the module of gate CMOS-ASIC with 4 integrated 8-bit ADCs over 55-GSample/s is also developed by Fujitsu company [34-37].

1.3 Structure and development of coherent lightwave systems

In this section, we will give a brief overview on the development and the structure of the coherent optical transmission systems. Employing a local oscillator (LO) in the
receiver end, the transmitted information can be demodulated by using the homodyne or the heterodyne detection techniques in the coherent lightwave communication systems [1,2]. Compared to traditional intensity modulation direct detection (IMDD), coherent detection can potentially improve the OSNR sensitivity of the receiver up to 20 dB, and the high level modulation formats can also be deployed in such systems to increase the spectral and the power efficiencies [1,38].

![Figure 1.1: Scheme of coherent lightwave communication system.](image)

The typical block diagram of the coherent optical transmission system is shown in Figure 1.1. The transmitted optical signal is combined coherently with the continuous wave (CW) from the narrow-linewidth LO laser, so that the detected optical intensity in the photodiode (PD) ends can be increased and the phase information of the optical signal can be obtained [1,39].

The structures of coherent detection can be divided into two types, the homodyne detection and the heterodyne detection. In the homodyne detection the intermediate frequency (IF) - which denotes the frequency difference between the optical carrier and the LO - is zero, while the heterodyne detection has a non-zero intermediate frequency [38-40]. The homodyne detection can improve the sensitivity of the receiver by a very large factor, and it can also demodulate the information loaded in the phase or the frequency of the optical carriers [1,40]. However, the homodyne detection needs a complicated optical phase-locked loop to synchronize the phase of the carrier and LO laser, and it also has a rigid requirement on the linewidths of the transmitter (TX) and the LO lasers. The heterodyne detection can also demodulate the information loaded in the amplitude, the phase, or the frequency of the optical carriers, while it has a lower optical signal-to-noise ratio (OSNR) improvement compared to the homodyne detection [1,38-40]. However, the heterodyne detection does not need the complicated optical phase-locked loop, and it has a moderate requirement on the linewidths of the TX and the LO lasers [1,2,41]. Therefore, the heterodyne detection was popular in the early stage of the coherent lightwave systems [38-42].

In the coherent detection, the information can be loaded in the amplitude, the phase, or the frequency of the optical carriers. The modulation formats, including the amplitude shift keying (ASK), the PSK, the frequency shift keying (FSK) can be deployed in the coherent transmission systems [1,40]. Both the homodyne detection and the heterodyne detection can be employed for demodulating all these modulation formats. Among these combinations, the PSK modulation with homodyne detection has the best theoretical performance of bit-error-rate (BER) versus OSNR, while the
stringent requirements on the optical PLL and the lasers linewidths limit the development of such transmission systems until the application of digital signal processing [1,38-43]. On the contrary, the continuous-phase FSK (CPFSK) and the differential PSK (DPSK) modulations with heterodyne detection - which have the moderate worse theoretical behaviors on BER versus OSNR - were extensively investigated due to their feasible implementations in the early 1990s, where the requirements on the optical PLL and the lasers linewidths were relaxed [40-42].

![Diagram of coherent communication system with digital signal processing.](image)

**Figure 1.2:** Scheme of coherent communication system with digital signal processing.

The development of the coherent transmission systems has stopped for more than 10 years due to the invention of EDFAs [20,21]. The coherent transmission techniques attracted the interests of investigation again around 2005, when a new stage of the coherent lightwave systems comes out by combining the digital signal processing techniques [44,45]. This type of coherent lightwave system is called as digital coherent communication system. In the digital coherent transmission systems, the electrical signals output from the photodiodes are sampled and transformed into the discrete signals by the high speed ADC components, which can be further processed by the DSP algorithms. The structure of the digital coherent detection system is illustrated in Figure 1.2.

![Diagram of DSP in digital coherent receiver.](image)

**Figure 1.3:** Typical scheme of DSP in digital coherent receiver.

The phase locking and the polarization adjustment were the main obstacles in the traditional coherent lightwave systems, while they can be solved by the carrier phase estimation and the polarization equalization respectively in the digital coherent optical transmission systems [45]. Besides, the chromatic dispersion and the nonlinear effects can also be mitigated by using the digital signal processing techniques [44-46]. The typical structure of the DSP compensating modules in the digital coherent receiver is
shown in Figure 1.3.

As we know before, the homodyne detection has the best sensitivity in the coherent detection [1], and it is now popularly used in the digital coherent optical transmission systems, where the phase fluctuations can be tracked by the DSP based carrier phase estimation algorithms [30-33]. To increase the system capacity and the spectral efficiency, high-level PSK and QAM modulation formats have been applied in the modern digital coherent communication systems, where the 8-PSK and the 256-QAM modulations have been realized recently [20,46]. Meanwhile, the multi-carrier coherent optical transmission system such as the orthogonal frequency division multiplexing (OFDM) coherent system has also been investigated to improve the capacity of the Ethernet from 2006 [47].

1.4 Brief summary of research field

The performance of high speed optical fiber transmission systems is severely affected by chromatic dispersion, polarization mode dispersion, phase noise and nonlinear effects [48-51], which can be well compensated in the coherent detection systems by employing the DSP circuit and the corresponding algorithms. Here we give a short overview about the development and the current status of the recently reported investigations related with our research work.

1. Chromatic dispersion equalization

Coherent optical receivers employing digital filters allow the significant equalization of chromatic dispersion in the electrical domain, instead of the compensation by dispersion compensating fibers (DCFs) or dispersion compensating modules (DCMs) in the optical domain [29,52-57]. This could save the costs and raise the nonlinear tolerance of the communication systems. Several digital filters have been applied to compensate the CD in the time and the frequency domain [29,55-59]. H. Bülow and A. Färbert et al. have reported their CD equalization work using the maximum likelihood sequence estimation (MLSE) method [52,55], which was the first DSP equalizer proposed. The MLSE electronic equalizer is implemented by using the Viterbi algorithm [55], where one is looking for the most likely bit sequence formed by a series of distorted signals. The MLSE is not tailored to a specific distortion but is optimum for any kind of optically distorted signal detected by the photodiode (PD), provided the inter-symbol interference (ISI) does not exceed the equalized symbols with a certain sampling period. S. J. Savory used a time-domain FD-FIR filter to compensate the CD in the 1000 km and the 4000 km transmission fibers without using dispersion compensation fibers [29,57]. The realization of the FD-FIR filter arises from the digitalization of the inverse function of the time-domain impulse response for the fiber channel [29]. The time window of the FD-FIR filter can be truncated by using the Nyquist frequency, which is determined to avoid the aliasing phenomenon in the digital systems. M. Kuschnerov and F. N. Hauske et al. have used the frequency domain equalizers (FDEs) to compensate the CD in coherent communication systems [58,59], which are considered as the most efficient digital equalizers for chromatic
dispersion compensation. The implementation of the frequency domain equalizers comes from the inverse impulse response of the fiber channel in the frequency domain. One of the most popular realizations of the FDEs is the blind look-up digital filter [58], which will be discussed later in our thesis. It has been demonstrated in the investigation that the FDEs are more efficient than the FD-FIR filter and the adaptive digital filters in the time domain, when the accumulated fiber chromatic dispersion is larger than 3000 ps/nm in the coherent transmission systems [60,61].

2. Carrier phase estimation

Phase noise from the lasers is also a significant impairment in the coherent optical transmission systems. The traditional method of demodulating the coherent optical signals is to use an optical or electrical PLL to synchronize the frequency and phase of the local oscillator (LO) laser with the transmitter (TX) laser [28]. Advances in high-speed very large-scale integration (VLSI) technology promise to change the paradigm of coherent optical receivers [62,63]. The frequency deviation (up to around 2 GHz) and the phase mismatch between the TX and the LO lasers can be tracked by the DSP algorithms and compensated in the feed-forward and the feed-back architectures [62-70]. Recent reports have demonstrated that the feed-forward carrier recovery schemes can be more tolerant to the laser phase noise than the PLL-based receivers. Several feed-forward and feed-back carrier phase estimation (CPE) algorithms have been validated as effective methods for mitigating the phase fluctuation from the laser sources [62-70]. The feed-forward carrier phase estimation algorithms mainly arise from some basic principles. One is based on the maximum-likelihood detection to estimate the transmitted sequence. The receiver consists of a soft-decision phase estimation stage and a hard-decision estimation for the carrier phase and the transmitted symbols [62,63]. The other popularly used algorithms, such as the block-average (BA) and the Viterbi-Viterbi (VV) methods, are called the N-power carrier phase estimation methods [64-69]. By applying these algorithms, the common phase value is evaluated for a block of signal samples and subtracted from the received signal prior to making a decision on the data extracted from the signal. The carrier phase is estimated by raising the signal amplitude to the power of N in order to get rid of the phase modulation in the encoded data and by averaging contributions from a block of the signal samples. The feed-back carrier phase recovery is to employ a one-tap normalized LMS (NLMS) filter to implement the decision-directed phase estimation in the coherent optical transmission systems, where a parameter of step size can influence the performance of the NLMS filter [70].

However, the analysis of the phase noise in the transmitter and the local oscillator lasers is often lumped together in these algorithms, and the influence of the large chromatic dispersion on the phase noise in the coherent systems is not considered [62-70]. Related work has been developed to deliberate the interplay between the digital chromatic dispersion equalization and the laser phase noise [71-79]. W. Shieh, K. P. Ho and A. P. T. Lau et al. have provided the theoretical assessment to evaluate the equalization enhanced phase noise (EEPEN) from the interaction between the LO phase fluctuation and the fiber dispersion, and they have also studied the EEPEN
induced time jitter in the coherent transmission systems [71-74]. C. Xie has investigated the impacts of chromatic dispersion on both the LO phase noise to amplitude noise conversion and the fiber nonlinear effects [75,76]. I. Fatadin and S. J. Savory have also studied the influence of the equalization enhanced phase noise in QPSK, 16-level quadrature amplitude modulation (16-QAM) and 64-QAM coherent transmission systems by employing the time-domain CD equalization [29,77]. Meanwhile, the effects of EEPN have also been investigated in the orthogonal frequency division multiplexing (OFDM) transmission systems [79]. Due to the existence of EEPN, the requirement of laser linewidth can not be generally relaxed for the coherent transmission systems with higher symbol rate. It would be interesting to investigate the performance of the equalization enhanced phase noise in the coherent optical communication systems employing different digital chromatic dispersion compensation methods. The behaviors of the different carrier phase estimation algorithms in the coherent transmission systems considering the impacts of the EEPN will be also discussed in this dissertation. Meanwhile, a method for extracting an RF pilot signal in the coherent receiver is also investigated to mitigate the equalization enhanced phase noise in the transmission systems.

The above analysis of the carrier phase estimation considering the equalization enhanced phase noise is performed in the coherent transmission systems with the post-compensation of chromatic dispersion. To make a more detailed analysis of the equalization enhanced phase noise, we also present an investigation of carrier phase estimation in the coherent transmission system with the pre-distortion of chromatic dispersion.
Chapter 2

Channel impairments in transmission systems

Fiber attenuation, chromatic dispersion, polarization mode dispersion, phase noise and fiber nonlinearities are the important distortions that affect the performance of optical transmission systems significantly. We will present a brief introduction in this part for the above five types of impairments in the optical fiber communication systems.

2.1 Fiber attenuation

Fiber loss reduces the optical signal power reaching the receiver, and it limits the transmission distance of the lightwave communication systems. The output power of an optical signal propagating in the silica fiber can be expressed as [1,80],

\[ P_{\text{out}} = P_{\text{in}} \cdot \exp(-\alpha \cdot z) \]  \hspace{1cm} (2.1)

where \( P_{\text{out}} \) is the output power, \( P_{\text{in}} \) is the launched power of the signal, \( \alpha \) is the attenuation coefficient of the fiber, and \( z \) is the transmission distance in the silica fiber. The attenuation coefficient \( \alpha \) (1/m) is usually described in the unit of dB/km, and we have the following relationship [80,81],

\[ \alpha \, \text{(dB/km)} = 4.343 \alpha \, \text{(1/m)} \]  \hspace{1cm} (2.2)

The fiber attenuation mainly originates from two physical phenomena in the silica fiber: material absorption and Rayleigh scattering, and it varies with the transmitted optical signal wavelength [1,82]. There are three typical wavelength windows used for the optical fiber communication systems. The first window is near 850 nm with the attenuation around 2 dB/km, the second window is near 1300 nm with the attenuation around 0.5 dB/km, and the third window is near 1550 nm with the attenuation around 0.2 dB/km [1,80].

The fiber loss in the telecommunication systems is usually compensated by the optical amplifiers such as Raman amplifiers and erbium-doped fiber amplifiers (EDFAs). The optical amplifiers will introduce the additional amplified spontaneous emission (ASE) noise (the dominant amplitude noise) in the communication systems, which has a significant influence on the optical signal-to-noise ratio (OSNR) [82,83].

2.2 Chromatic dispersion

The chromatic dispersion of an optical medium is the phenomenon that the phase velocity and the group velocity of the light depend on the optical frequency. Group delay is defined as the first derivative of the optical phase with respect to the optical frequency, and chromatic dispersion is defined as the second derivative of the optical
phase with respect to the optical frequency. Chromatic dispersion consists of the waveguide dispersion and the material dispersion [1,80-82]. The material dispersion occurs due to the changes in the refractive index of the medium with the changes in optical wavelength, which originates from the electromagnetic absorption. The waveguide dispersion occurs when the speed of an optical wave in the waveguide depends on its frequency for geometric reasons, independent of any frequency dependence of the materials. For a fiber, the waveguide dispersion arises from the dependence on the fiber parameters such as the core radius and the index difference. The common evaluation of the chromatic dispersion (dispersion parameter D) is calculated by the time delay between the unitary wavelength difference after propagating through the unitary fiber length. The unit of D is normally expressed in ps/nm/km.

![Figure 2.1: Typical wavelength dependence of dispersion parameters in normal single-mode fibers.](image)

The chromatic dispersion for different wavelength in the standard single mode fiber (SSMF) is illustrated in Figure 2.1 [84]. The example shows the characters of the single-mode fibers have zero dispersion at the wavelength of 1310 nm. We can also find that the chromatic dispersion value is around 16 ps/nm/km at 1550 nm, which is the operation wavelength for practical optical fiber transmission systems. Chromatic dispersion remains constant over the bandwidth of a transmission channel for long distance of fiber. In traditional optical fiber communication systems, the chromatic dispersion is usually compensated by the dispersion compensation fibers (DCF). In coherent transmission systems, the chromatic dispersion can be equalized by using a digital filter, which will be discussed in Chapter 4.

### 2.3 Polarization mode dispersion

Polarization mode dispersion is a phenomenon of modal dispersion that two orthogonal polarizations of light propagate at different speeds due to the random imperfections and asymmetries in the waveguide, which cause a random spreading of
the optical pulse. The ideal optical fiber core has a perfectly circular cross-section, where the two orthogonal fundamental modes travel at the same speed. However, in a realistic fiber, the random imperfections such as the circular asymmetries, can arouse the two polarizations to propagate at different speeds. The symmetry-breaking random imperfections consist of the geometric asymmetry (slightly elliptical cores) and the stress-induced material birefringences [80-82,85].

In the existence of PMD, the two polarization modes of the optical signal will separate slowly. Corresponding to the stochastic imperfections, the pulse spreading effects is also random. Due to the characteristic of random variation, the evaluation of the polarization mode dispersion is calculated by the mean polarization-dependent time-differential, which is called the differential group delay (DGD), proportional to the square root of propagation distance. The unit of the polarization mode dispersion is in $\text{ps}/\sqrt{\text{km}}$ [85,86]. In practical single mode fibers, the value of PMD is from $0.1\text{ps}/\sqrt{\text{km}}$ to $1\text{ps}/\sqrt{\text{km}}$. The pulse spreading effects in the optical fiber is shown in Figure 2.2 [87].

![Figure 2.2: The mode spreading due to the PMD in the optical fibers.](image)

The method for PMD compensation is to employ a polarization controller to compensate the differential group delay occurring in optical fibers. The PMD effects are random and time-dependent, therefore, an active feed-back device over time is required. Such systems are therefore expensive and complex. In the digital coherent receivers employing DSP, the PMD can be compensated by the adaptive filters.

### 2.4 Laser phase noise

One of the important sources for receiver sensitivity degradation in the coherent lightwave systems is the phase noise associated with the transmitter laser and the local oscillator laser [1]. Laser phase noise can be approximately regarded as a Wiener process caused by laser spontaneous emission, which can be modeled as the expression [62,88,89]:

$$\phi(t) = \int_{-\infty}^{t} \delta\omega(\tau) d\tau$$

(2.3)

where $\phi(t)$ is the instantaneous optical phase, and $\delta\omega(\tau)$ is the frequency noise with
zero mean and autocorrelation \( \rho = 2\pi \Delta \nu \delta(\tau) \). It has been demonstrated that the laser output has a Lorentzian spectrum with a 3-dB linewidth of \( \Delta \nu \) \[62\].

![Figure 2.3: The phase noise influence in QPSK coherent transmission system, (a) without phase noise, (b) with phase noise laser linewidth TX=LO=150 kHz.](image)

Phase noise is a significant impairment in the coherent transmission systems, since it impacts the optical carrier synchronization between the TX laser and the LO laser. In the non-coherent detection system (such as IMDD system), the carrier phase is not so important because the receiver only measures the power of the optical signal. In the coherent systems, the information is encoded into the variation of the carrier phase, therefore, the phase fluctuation over a symbol period has the significant influence on the signal demodulation in the receiver, as shown in Figure 2.3. The transmission fiber length is 500 km, the signal symbol rate is 28-Gsymbol/s, and the OSNR is 14.8 dB. We can find in Figure 2.3 that the QPSK constellation is distorted obviously by the phase noise. In the traditional coherent systems, the carrier phase noise is compensated by using the optical PLL in the receiver to track the phase changing with the time, and it is rather difficult to realize the corresponding control circuits. In the modern digital coherent detection systems, the carrier phase noise can be well mitigated by using the DSP algorithms such as the feed-forward and the feed-back carrier phase estimation, which are relatively easy to implement.

### 2.5 Nonlinear effects

The transmission nonlinear impairments in the long-distance high bit-rate optical fiber communication systems mainly include the fiber Kerr nonlinearities, the self-phase modulation (SPM), the cross-phase modulation (XPM) and the four-wave mixing (FWM) \[90,91\]. The Kerr effect refers to the refractive index change of a material due to the influence of the strong electric field \[90\]. The self-phase modulation leads to the phase shifting of the pulse due to the strong electric field itself \[1,90\]. The cross-phase modulation is the effect that the phase of the signal in one wavelength channel is influenced by the signals in other wavelength channels \[90\]. The four-wave mixing indicates that the interactions among three wavelengths will produce one extra
wavelength in the optical signal [90,91].

The signals transmitted through the optical fibers in presence of attenuation, chromatic dispersion and nonlinear effects follow the nonlinear Schrödinger equation (NLSE) [90],

$$\frac{\partial E(z,t)}{\partial z} + j \frac{\beta_2}{2} \frac{\partial^2 E(z,t)}{\partial t^2} + \frac{\alpha}{2} E(z,t) = j \gamma |E(z,t)|^2 E(z,t)$$

(2.4)

where $E(z,t)$ is the electric field of the optical signal, $\alpha$ is the attenuation coefficient, $\beta_2$ is the chromatic dispersion parameter, $\gamma$ is the nonlinear coefficient, and $z$ and $t$ are the propagation direction and time, respectively. The nonlinear parameter $\gamma$ scales inversely on the effective core area of the transmission fiber.

For WDM transmission systems, fiber nonlinear effects mainly consist of two aspects: inter-channel interference and intra-channel interference. Inter-channel nonlinear effects refer to the interference between different wavelength channels, which include the cross-phase modulation and the four-wave mixing. Intra-channel nonlinear effects indicate the interference between different modules in the same wavelength channel, which include the self-phase modulation, the intra-channel XPM and the intra-channel FWM. Inter-channel nonlinearities are dominant for lower bit-rate transmission systems, and intra-channel nonlinearities are dominant for higher bit-rate transmission systems.

The fiber nonlinear effects are difficult to compensate in traditional high speed IMDD transmission systems. In digital coherent systems, the nonlinear effects can be mitigated by using the backward propagation methods based on solving the nonlinear Schrödinger equation and the Manakov equation [92,93].
Chapter 3

High speed coherent transmission systems

In this chapter, we give a detailed analysis for three implementations of the high speed NRZ-QPSK coherent optical transmission systems realized in the VPI platform. The three setups include the 112-Gbit/s NRZ-PDM-QPSK coherent transmission system with post-compensation of dispersion (no RF pilot tone for phase noise correction), the 56-Gbit/s NRZ-SP-QPSK transmission system with CD post-compensation (using RF pilot tone for phase noise correction), and the dispersion pre-distorted 56-Gbit/s NRZ-SP-QPSK coherent transmission system (no RF pilot tone for phase noise correction). Meanwhile, we describe a mathematical analysis for the standard QPSK transmitter, the fiber channel, the coherent receiver based on the 112-Gbit/s NRZ-PDM-QPSK post-compensated coherent system. The analysis of these modules is also suitable for other transmission systems.

3.1 The 112-Gbit/s PDM-QPSK post-compensated system

Here we will make a description of the 112-Gbit/s NRZ-PDM-QPSK coherent system with post-compensation of chromatic dispersion (no RF pilot tone for phase noise correction). Based on this setup, we will also discuss about the mathematical modules of the standard QPSK transmitter, the fiber channel, the coherent receiver in coherent transmission systems.

3.1.1 The 112-Gbit/s NRZ-PDM-QPSK post-compensated system

As illustrated in Figure 3.1, the setup of the 112-Gbit/s NRZ-PDM-QPSK coherent transmission system with post-compensation of chromatic dispersion (no RF pilot tone for phase noise correction) is implemented in the VPI simulation platform [94]. The data output from the four 28-Gbit/s pseudo random bit sequence (PRBS) generators are modulated into two orthogonally polarized NRZ-QPSK optical signals by the two Mach-Zehnder modulators. Then the orthogonally polarized signals are integrated into one fiber channel by a polarization beam combiner (PBC) to form the 112-Gbit/s NRZ-PDM-QPSK optical signal. Using a local oscillator in the coherent receiver, the received optical signals are mixed with the LO laser to be transformed into four electrical signals by the photodiodes (PDs). Here we employ the balanced photodiode detection, which can achieve larger optical power dynamic range and higher noise sensitivity than the single detection [95]. Then the signals are digitalized by the 8-bit analog-to-digital convertors (ADCs) at twice the symbol rate [96]. The optimum bandwidth of the ADCs is half of the symbol rate [97]. The sampled signals are processed by a series of digital equalizers, and the bit-error-rate (BER) is then estimated from the data sequence of $2^{17}$ bits. The central wavelength of the TX laser and the LO laser are both 1553.6 nm. The standard single mode fibers with the CD coefficient equal to 16 ps/nm/km are employed in all the simulation work. Here we
mainly concentrate our work on the CD compensation and the carrier phase noise mitigation methods in DSP techniques, and so we neglected the influences of fiber attenuation, polarization mode dispersion and nonlinear effects [90-96].

3.1.2 Theory of QPSK transmission modules

The mathematical expressions for analyzing the modules in the NRZ-QPSK coherent transmission system with post-compensation implementation (no RF pilot tone for phase noise correction) are presented in the following descriptions.

1. Optical QPSK transmitter (Mach-Zehnder modulator)

The main structure of the QPSK transmitter is realized by using a nested Mach-Zehnder modulator [94,98-100]. The PRBS output sample pass into the non-return-to-zero signal generator to form the modulation wave, where the output bit sequence in one period could be expressed as

\[ x_I(t) = p \quad \text{and} \quad p = 0.1 \]  
\[ x_Q(t) = q \quad \text{and} \quad q = 0.1 \]

The electric field transfer function \( h_{MZM}(t) \) of the Mach-Zehnder modulator is given as the following equation,
\[ h_{\text{MZM}}(t) = \alpha_{\text{MZM}} \cdot \exp(p \cdot j\pi) \]  

(3.3)

where \( \alpha_{\text{MZM}} \) is the attenuation of the Mach-Zehnder modulator.

The I-channel electric field output from the Mach-Zehnder modulator neglecting the attenuation of the Mach-Zehnder modulator is

\[ E_i(t) = E_{\text{cw}}(t) \cdot h_{\text{MZM}}(t) \]
\[ = E_0 \exp[j(\omega_{\text{carrier}} t + \phi_{\text{carrier}} + p \cdot \pi)] \quad p = 0, 1 \]  

(3.4)

\[ E_{\text{MW}}(t) = E_0 \exp[j(\omega_{\text{carrier}} t + \phi_{\text{carrier}})] \]  

(3.5)

where \( \omega_{\text{carrier}} \) and \( \phi_{\text{carrier}} \) are the angle frequency and the phase of the optical carrier, respectively.

According to the same principle, we could obtain the Q-channel electric field as,

\[ E_Q(t) = E_{\text{cw}}(t) \cdot h_{\text{MZM}}(t) \cdot \exp\left(\frac{j}{2}\pi\right) \]
\[ = E_0 \exp\left[j\left(\omega_{\text{carrier}} t + \phi_{\text{carrier}} + q \cdot \pi + \frac{\pi}{2}\right)\right] \quad q = 0, 1 \]  

(3.6)

\[ E_{\text{QPSK}}(t) = E_i(t) \pm E_Q(t) \]  

(3.7)

2. Fiber propagation

The generalized nonlinear Schrödinger equation is used to describe the in-band effects for the fiber transmission \([90]\), which is expressed as,

\[ \frac{\partial E(z,t)}{\partial z} = \left[ \hat{D} + \hat{N} \right] \cdot E(z,t) \]  

(3.8)

where \( E(z,t) \) denotes the slowly-varying complex-envelope of the electric field of the light wave, \( |E(z,t)|^2 \) characterizes its power, \( \hat{D} \) is the dispersion operator, and \( \hat{N} \) is the nonlinearity operator.

\[ \hat{D} = j\frac{\beta_2}{2} \frac{\partial^2}{\partial t^2} + \frac{\beta_3}{6} \frac{\partial^3}{\partial t^3} - \frac{\alpha}{2} \]  

(3.9)

where \( \beta_2 \ (s^2/m) \) describes the first order group-velocity dispersion, \( \beta_3 \ (s^3/m) \) is the second order GVD slope, and \( \alpha \ (1/m) \) is the attenuation constant of the transmission fiber.

Nonlinear operator (with no Raman effect) is simply given by

\[ \hat{N} = -j\gamma |E(z,t)|^2 \]  

(3.10)
\[ \gamma = \frac{2n_2 f_{\text{ref}}}{c A_{\text{eff}}} \]  

(3.11)

where \( \gamma \) depends on the nonlinear index \( n_2 \), the effective core area \( A_{\text{eff}} \), as well as the reference frequency of optical carrier wave \( f_{\text{ref}} \) and the velocity of light in vacuum \( c \).

The propagation of the optical signal in the fiber can be calculated by the split-step Fourier method [90]. Assuming a propagation of optical signals in +z direction and an asymmetrical split-step algorithm, the mathematical formalism of the procedure can be described as the following description,

\[ E(z_0 + \Delta z, t) = \left[ \exp \left( \Delta z \hat{N} \right) E(z_0, t) \right] \exp \left( \Delta z \hat{D} \right) \]  

(3.12)

3. Coherent Receiver

In the coherent receiver, the 2×4 90 degree hybrid structure is adopted to demodulate the received optical signal, which consists of four 3-dB 2×2 fiber couplers and a phase delay components of \( \pi/2 \) phase shift in one branch [101-104].

Assuming the electric field of the received optical signal is \( E_R(t) \), and the electric field of the local oscillator laser is \( E_{LO}(t) \), which is expressed as

\[ E_{LO}(t) = E_{LO} \cdot \exp[j(\omega_{LO} t + \phi_{LO})] \]  

(3.13)

where \( \omega_{LO} \) and \( \phi_{LO} \) the angle frequency and the initial phase of the local oscillator laser.

The output electric field components of the coherent receiver are calculated as follows,

\[ \begin{bmatrix} E_0 \\ E_{\pi/2} \\ E_{x} \\ E_{3\pi/2} \end{bmatrix} = \frac{1}{2} \begin{bmatrix} 1 & 1 & 1 & 1 \\ 1 & j & -1 & -1 \\ 1 & -j & 1 & 1 \\ 1 & -j & 1 & 1 \end{bmatrix} \begin{bmatrix} E_R + E_{LO} \\ E_R + E_{LO} \cdot \exp(j \cdot \pi/2) \\ E_R - E_{LO} \\ E_R - E_{LO} \cdot \exp(j \cdot \pi/2) \end{bmatrix} \]  

(3.14)

where \( E_0, E_{\pi/2}, E_x, E_{3\pi/2} \) represent the four electric fields output from the 90 degree hybrid coherent receiver, respectively.

Due to the asymmetry of the 3-dB 2×2 fiber coupler, the two lower outputs are 90 degree phase shifted relative to the two upper outputs. With the additional 90 degree phase shift introduced, the output electric fields are revised as
\[
\begin{bmatrix}
E_0 \\
E_{5/2} \\
E_2 \\
E_{35/2}
\end{bmatrix}
= \frac{1}{2}
\begin{bmatrix}
(-j) \cdot E_R + (-j) \cdot E_{LO} \\
(-j) \cdot E_R + E_{LO} \cdot \exp\left( j \cdot \frac{\pi}{2} \right) \\
E_R - E_{LO} \\
(-j) \cdot E_R - (j) E_{LO} \cdot \exp\left( j \cdot \frac{\pi}{2} \right)
\end{bmatrix}
= \frac{1}{2}
\begin{bmatrix}
-jE_R - jE_{LO} \\
-E_R + E_{LO} \cdot \exp\left( j \cdot \frac{\pi}{2} \right) \\
E_R - E_{LO} \\
-jE_R - jE_{LO} \cdot \exp\left( j \cdot \frac{\pi}{2} \right)
\end{bmatrix}
\tag{3.15}
\]

3.2 The 56-Gbit/s SP-QPSK post-compensated system with RF tone

The complete NRZ-SP-QPSK coherent system including an optical RF signal tone or an RX extracted RF pilot tone for eliminating the phase noise is schematically shown in Figure 3.2. We consider an FDE filter for the chromatic dispersion compensation and carrier phase extraction using the one-tap NLMS method [58,70]. The FDE filter is selected as commonly used in most practical system demonstrations at this time. In the simulations, we consider the SP-QPSK transmission system with a symbol rate of 28-Gsymbol/s, and the system capacity is 56-Gbit/s. The orthogonal polarization state is used to either transmit an optical RF carrier or left empty in the case of an RX based RF pilot tone extraction. We note that it is straightforward to double the system capacity using the RX generated RF pilot tone by using also the orthogonal polarization state for QPSK signal transmission whereas this is more complicated using the optically transmitted RF pilot tone [105-107]. We utilize the VPI software tool for the system simulations, and we evaluate the bit-error-rate versus optical signal-to-noise ratio (OSNR) [94].

![Diagram of the 56-Gbit/s NRZ-SP-QPSK coherent system](image)

Figure 3.2: The 56-Gbit/s NRZ-SP-QPSK coherent system using an optical RF pilot tone (red system parts) or using an RX extracted RF pilot tone (green system parts). N(t): additive noise, PBS: polarizing beam splitter, RF: radio frequency.

The RX based pilot tone extraction is shown in generic form in Figure 3.3. It consists...
of three stages: 1) extraction of the modulated signal phase after compensation for chromatic dispersion, 2) high pass filtering (HPF) to remove as much as possible the phase noise, 3) creation of the RF tone.

The principle of phase noise cancellation by using an RF pilot tone is very simple. Let the detected coherent signal field be represented as:

$$E_i(t) = A(t) \cdot \exp(j(\phi(t) + m(t)))$$ (3.16)

where $m(t)$ represents the phase modulation, which is one of \{\pi/4, 3\pi/4, 5\pi/4, 7\pi/4\} for the QPSK modulation, $A(t)$ is the modulated (real-valued) amplitude, and $\phi(t)$ is the phase noise. The RF pilot tone in the ideal case (i.e. generated optically) is:

$$E_{RF} (t) = B \cdot \exp(j\phi(t))$$ (3.17)

where $B$ is an arbitrary constant amplitude, and the conjugated signal operation that eliminates the phase noise is given - to within the arbitrary amplitude constant, $B$ - as:

$$E_i(t) \cdot E_{RF}^\ast(t) = B \cdot A(t) \cdot \exp(jm(t))$$ (3.18)

![Figure 3.3: Structure of RF pilot tone extraction and the complex conjugation operation in the QPSK coherent receiver. HPF: high pass filter.](image)

It is well known that the leading order laser phase noise is modeled as a Brownian motion i.e. it has a Gaussian probability density function with a white frequency noise power spectral density [108]. This leads to a phase noise spectral density (in the case of no signal phase modulation) which is proportional to $f^{-2}$ (the inverse frequency squared). In the case of signal modulation, the power spectral behavior taking into account phase noise as well as the phase modulation is more complicated, but it may still be anticipated that a major part of the phase noise is situated near DC. On the other hand, the signal modulation spectrum is concentrated around the $1/T_s$ frequency ($T_s$ is the symbol time) and extending towards DC for long identical symbol modulation sequences. Thus, it appears that filtering the received modulation signal phase by a digital high pass filter will potentially take away major parts of the phase noise (and slightly distort the modulation signal). The remaining phase noise and distorted phase modulation is denoted as $\overline{m(t)}$. As a result, an RX extracted RF pilot tone can now be generated as (see Figure 3.3):
This signal can be used for phase noise compensation - equivalent to the ideal RF pilot tone in Eq. (3.17) - by generating the signal:

\[
E_{\phi}(t) = B \cdot \exp\left(j\phi(t) + m(t) - \bar{m}(t)\right)
\]  

(3.19)

So far, we have not considered that the modulated signal, as well as the RF pilot tone, is influenced by the additive noise e.g. from amplifiers in the transmission path. Taking into account the additive noise and using the modulated signal in Eq. (3.16), one can specify the bit-error-rate for the considered QPSK system in a situation without correcting the phase noise. With Equation (3.18) or Equation (3.20), the BER is specified using optically generated or RX extracted RF pilot tones to correct (as much as possible) the phase noise influence. Implementing the HPF filtering and generation of an RF carrier is equivalent to filtering away the phase noise by a mirror low pass filter (LPF).

We note that the optical or the electronically generated RF pilot carriers can also be employed to mitigate the phase noise influence in \(m\)-level PSK \((m\text{-PSK})\) and \(m\)-level QAM \((m\text{-QAM})\) coherent transmission systems.

### 3.3 The pre-distorted 56-Gbit/s SP-QPSK coherent system

The scheme for the 56-Gbit/s NRZ-SP-QPSK coherent optical transmission system employing the pre-compensation of chromatic dispersion is presented in Figure 3.4, and the detailed transmitter implementation for the CD pre-distortion is shown in Figure 3.5.

![Figure 3.4: Scheme of 56-Gbit/s SP-QPSK pre-distorted transmission system. PD: photo diode, N(t): additive noise.](image-url)
The main difference of the dispersion pre-compensated systems from the classical post-compensated coherent systems lies in the pre-distorted QPSK transmitter. Therefore, it is significant to comment on Figure 3.5 in some detail, which shows how the pre-distorted QPSK modulation is generated in the electrical domain. The general specification of the analogue modulated QPSK signal is denoted as $A(t) \exp(j\varphi(t))$, where $A(t)$ denotes the amplitude part of the modulation and $\varphi(t)$ denotes the phase part. In the case of QPSK modulation we have $A(t)=1$. The CD equalization is realized by using a digital filter with the inverse transfer function of the fiber dispersion \[29,58,109-112\]. The chromatic dispersion equalization followed by the digital-to-analogue conversion (DAC) generates the signal $A'(t) \exp(j\varphi'(t))$ which is used to drive the amplitude modulator (AM) and the phase modulator (PM). This moves the QPSK modulated signal into the optical carrier wave.

Figure 3.5: Pre-distorted QPSK transmitter in the 56-Gbit/s NRZ-SP-PSK coherent system.

Compared to the post-compensation case in Figure 3.1, it is observed that the CD equalization in the transmitter is performed in the electrical domain prior to the optical signal generation. This means that the dispersion of the CD equalizing filter does not interact with the TX laser phase noise. Because of this the net-dispersion for the EEPN generating parts of the system is the fiber for the TX laser whereas there is no dispersive system parts which interacts with the LO laser. This means that the resulting EEPN originates from the TX laser in the dispersion pre-distorted coherent transmission systems, which will be investigated in our simulation work \[113\].
Chapter 4

Digital signal processing algorithms for coherent systems

In this chapter, the chromatic dispersion mitigation and the carrier phase noise compensation are implemented and analyzed with the corresponding DSP algorithms. The adaptive PMD equalization using the least mean square (LMS) and the constant modulus algorithm (CMA) filters is also briefly discussed.

4.1 Chromatic dispersion compensation

The popular digital filters involving the time-domain LMS adaptive filter and fiber dispersion finite impulse response (FD-FIR) filter, as well as the frequency-domain filters are investigated for CD compensation. As an example, the characters of these filters are analyzed comparatively in the 112-Gbit/s NRZ-PDM-QPSK coherent transmission system with post-compensation of dispersion. We note that the FD-FIR filter and the frequency-domain filters can also be used for the pre-distorted coherent systems in the same way.

4.1.1 Time domain equalizers

1. The LMS adaptive filter

The LMS filter employs an iterative algorithm that incorporates successive corrections to weights vector in the negative direction of the gradient vector which eventually leads to a minimum mean square error [114]. The principle of LMS filter is given by the following equations:

\[ y(n) = \overrightarrow{w}(n) \overrightarrow{x}(n) \]  
\[ \overrightarrow{w}(n+1) = \overrightarrow{w}(n) + \mu \overrightarrow{x}(n)e^*(n) \]  
\[ e(n) = d(n) - y(n) \]

where \( \overrightarrow{x}(n) \) is the digitalized complex magnitude vector of the received signal, \( y(n) \) is the complex magnitude of the equalized output signal, \( n \) represents the number of sample sequence, \( \overrightarrow{w}(n) \) is the complex tap weights vector, \( \overrightarrow{w}^H(n) \) is the Hermitian transform of \( \overrightarrow{w}(n) \), \( d(n) \) is the desired symbol, which corresponds to one case of the vector \([1+i \ 1-i \ -1+i \ -1-i]\) for the QPSK coherent transmission system, \( e(n) \) represents the estimation error between the output signal and the desired symbol, \( e^*(n) \) is the conjugation of \( e(n) \), and \( \mu \) is a key real coefficient called step size. In order to guarantee the convergence of tap weights vector \( \overrightarrow{w}(n) \), the step size \( \mu \) needs to satisfy the condition of \( 0 < \mu < 1/\lambda_{\text{max}} \), where \( \lambda_{\text{max}} \) is the largest eigenvalue.
of the correlation matrix $R = \sum_{t} x(t)x^H(t)$. The LMS dispersion filter is used in the “decision-directed” mode in our work.

The tap weights in LMS adaptive equalizer for 20 km fiber CD compensation is shown in Figure 4.1. The convergence for 9 tap weights in the LMS filter with step size equal to 0.1 is shown in Figure 4.1(a), and we can find that the tap weights obtain their convergence after about 5000 iterations. The magnitudes of converged tap weights are shown in Figure 4.1(b), and it can be found that the central tap weights take more dominant roles than the high-order tap weights.

Figure 4.1: Taps weights of LMS filter. (a) Tap weights magnitudes convergence. (b) Converged tap weights magnitudes distribution.
2. The FD-FIR filter

Compared with the iteratively updated LMS filter, the tap weights in FD-FIR filter have a relatively simple specification [29,57], the tap weight in FD-FIR filter is given by the following equations:

\[
\alpha_k = \frac{j e^{T^2}}{D \lambda^2 z} \exp \left( -j \frac{\pi e^{T^2}}{D \lambda^2 z} k^2 \right) , \quad -\left\lfloor \frac{N}{2} \right\rfloor \leq k \leq \left\lfloor \frac{N}{2} \right\rfloor 
\] 

\[
N^A = 2 \times \left\lfloor \frac{D \lambda^2 z}{2 \pi e^{T^2}} \right\rfloor + 1
\] 

(4.4)

(4.5)

where \( D \) is the fiber chromatic dispersion coefficient, \( \lambda \) is the central wavelength of the transmitted optical wave, \( z \) is the fiber length in the transmission channel, \( T \) is the sampling period, \( N^A \) is the required maximum tap number for compensating the fiber dispersion, and \( \lfloor x \rfloor \) denotes the nearest integer less than \( x \).

Figure 4.2: Tap weights of FD- FIR filter.

The tap weights of FD-FIR filter according to Equation (4.4) for 20 km fiber (\( D=16 \) ps/nm/km) are shown in Figure 4.2 [115,116]. For a fixed fiber dispersion, the magnitudes of tap weights in the FD-FIR filter are constant, whereas the real and the imaginary parts vary symmetrically.

4.1.2 Frequency domain equalizers

The frequency domain equalizers (FDEs) have become the most attractive digital filters for channel equalization in the coherent transmission systems due to the low computational complexity for large dispersion and the wide applicability for different fiber distance [29,57-61]. The fast Fourier transform (FFT) convolution algorithms
involving the overlap-save (OLS) and the overlap-add zero-padding (OLA-ZP) methods are traditionally used for the equalization in the wireless communication systems [117-120]. In our research work, the OLS-FDE and the OLA-ZP-FDEs are applied to compensate the CD in the high speed coherent optical transmission systems [58,59,121,122].

1. Overlap-save method

The schematic of the FDE with overlap-save method is illustrated in Figure 4.3 [117,118,121,122]. The received signals are divided into several blocks with a certain overlap, where the block length is called the FFT-size. The sequence in each block is transformed into the frequency domain data by the FFT operation, and afterwards multiplied by the transfer function of the FDE. Next, the data sequences are transformed into the time domain signals by the inverse FFT (IFFT) operation. Finally, the processed data blocks are combined together, and the bilateral overlap samples are symmetrically discarded. One of the most popular OLS-FDEs for chromatic dispersion equalization is the blind look-up filter [58,59].

![Figure 4.3: FDE with OLS method. The parts with slants are to be discarded.](image)

2. Overlap-add method

The structure of the FDE with overlap-add one-side zero-padding (OLA-OSZP) method is shown in Figure 4.4 [117-120]. The received data are divided into small blocks without any overlap, and then the data in each block are appended with zeros at one side. To be consistent with the OLS method, the total length of data block and zero padding is called the FFT-size, while the length of zero padding is called the overlap. The zero-padded sequence is transformed by the FFT operation, and multiplied by the transfer function of the FDE. Afterwards, the data are transformed by the IFFT operation. Finally the processed data sequences are combined by overlapping and adding. Note that half of the data stream in the first block is discarded.
Figure 4.4: FDE with OLA-OSZP method. The gray parts mean the appended zeros, and the parts with slants are to be discarded.

The schematic of the FDE with overlap-add both-side zero-padding (OLA-BSZP) method is illustrated in Figure 4.5 [117-120]. The received data are also divided into several blocks without any overlap, and then the data in each block are appended with equivalent zeros at both sides. The total length of data block and zero padding is called the FFT-size, and the length of the whole zero padding is called the overlap. The zero-padded sequence is transformed by the FFT operation, and multiplied by the transfer function of the FDE, and then transformed by the IFFT operation. The processed data blocks are also combined together by overlapping and adding. Note that half of the data stream in the first block is discarded.

Figure 4.5: FDE with OLA-BSZP method. The gray parts mean the appended zeros, and the parts with slants are to be discarded.
4.2 Polarization mode dispersion and polarization rotation equalization

Due to the random character of the polarization mode dispersion and the polarization rotation (PR), the compensation of the PMD and the polarization rotation is usually realized by the adaptive algorithms such as the LMS and the CMA filters.

4.2.1 LMS adaptive PMD equalization

The influence of PMD and polarization fluctuation can be compensated adaptively by the decision-directed LMS filter [54,114], which is expressed as the following equations:

\[
\begin{align*}
\begin{bmatrix}
\vec{x}_{\text{out}}(n) \\
\vec{y}_{\text{out}}(n)
\end{bmatrix} &=
\begin{bmatrix}
\vec{H} & \vec{H} \\
\vec{H} & \vec{H}
\end{bmatrix}
\begin{bmatrix}
\vec{x}_{\text{in}}(n) \\
\vec{y}_{\text{in}}(n)
\end{bmatrix}
\end{align*}
\]

\[\begin{align*}
\vec{w}_{x}(n+1) &= \vec{w}_{x}(n) + \mu_{\rho} \vec{e}_{x}(n) \vec{x}_{\text{in}}(n) \\
\vec{w}_{y}(n+1) &= \vec{w}_{y}(n) + \mu_{\rho} \vec{e}_{y}(n) \vec{y}_{\text{in}}(n)
\end{align*} \tag{4.7}\]

\[\begin{align*}
\vec{e}_{x}(n) &= d_{x}(n) - \vec{x}_{\text{out}}(n) \\
\vec{e}_{y}(n) &= d_{y}(n) - \vec{y}_{\text{out}}(n)
\end{align*} \tag{4.8}\]

where \(\vec{x}_{\text{in}}(n)\) and \(\vec{y}_{\text{in}}(n)\) are the complex magnitude vectors of the input signals, \(\vec{x}_{\text{out}}(n)\) and \(\vec{y}_{\text{out}}(n)\) are the complex magnitudes of the equalized output signals respectively, \(\vec{w}_{x}(n), \vec{w}_{y}(n)\) and \(\vec{w}_{x}(n)\) are the complex tap weights vectors, \(d_{x}(n)\) and \(d_{y}(n)\) are the desired symbols, \(\vec{e}_{x}(n)\) and \(\vec{e}_{y}(n)\) represent the estimation errors between the output signals and the desired symbols respectively, and \(\mu_{\rho}\) is the step size parameter. The polarization diversity equalizer can be implemented subsequent to the CD compensation.

4.2.2 CMA adaptive PMD equalization

The CMA filter can also be employed for the adaptive compensation for the influence of the PMD and the polarization fluctuation [123,124], of which the tap weights can be expressed as:

\[
\begin{align*}
\begin{bmatrix}
\vec{x}_{\text{out}}(n) \\
\vec{y}_{\text{out}}(n)
\end{bmatrix} &=
\begin{bmatrix}
\vec{H} & \vec{H} \\
\vec{H} & \vec{H}
\end{bmatrix}
\begin{bmatrix}
\vec{v}_{x}(n) \\
\vec{v}_{y}(n)
\end{bmatrix}
\begin{bmatrix}
\vec{x}_{\text{in}}(n) \\
\vec{y}_{\text{in}}(n)
\end{bmatrix}
\end{align*}
\]

\[\begin{align*}
\vec{v}_{x}(n+1) &= \vec{v}_{x}(n) + \mu_{\rho} \vec{e}_{x}(n) \vec{x}_{\text{in}}(n) \\
\vec{v}_{y}(n+1) &= \vec{v}_{y}(n) + \mu_{\rho} \vec{e}_{y}(n) \vec{y}_{\text{in}}(n)
\end{align*} \tag{4.9}\]
We can find that the CMA algorithm is based on the principle of minimizing the modulus variation of the output signal to update its weight vector.

4.3 Carrier phase recovery

In this section, we will present an analysis on different carrier phase estimation algorithms, involving the one-tap NLMS, the different detection, the block-average (BA) and the Viterbi-Viterbi (VV) methods in the coherent optical transmission systems considering the equalization enhanced phase noise (EEPN).

4.3.1 Principle of equalization enhanced phase noise

The scheme of the coherent optical communication system with digital CD equalization and carrier phase estimation is depicted in Figure 4.6. The transmitter laser phase noise passes through both transmission fibers and the digital CD equalization module, and so the net dispersion experienced by the transmitter PN is close to zero. However, the local oscillator phase noise only goes through the digital CD equalization module, which is heavily dispersed in a transmission system without dispersion compensation fibers. Therefore, the LO phase noise will significantly influence the performance of the high speed coherent systems with only digital CD post-compensation [71,72]. We note that the EEPN does not exist in a transmission system with entire optical dispersion compensation for instance using DCFs [71-75].

![Figure 4.6: Scheme of equalization enhanced phase noise in coherent transmission system.](image)

Figure 4.6: Scheme of equalization enhanced phase noise in coherent transmission system. \( \Phi_{TX} \): phase fluctuation of the TX laser, \( \Phi_{LO} \): phase fluctuation of the LO laser, \( N(t) \): additive white Gaussian noise.

Theoretical analysis demonstrates that the EEPN scales linearly with the accumulated chromatic dispersion and the linewidth of LO laser [71-79], and the variance of the additional noise due to the EEPN can be expressed as:

\[
\begin{align*}
\eta_x(n) &= 1 - |x_{\text{out}}(n)|^2 \\
\eta_y(n) &= 1 - |y_{\text{out}}(n)|^2
\end{align*}
\]
where $\lambda$ is the central wavelength of the transmitted optical carrier wave, $c$ is the light speed in vacuum, $D$ is the chromatic dispersion coefficient of the transmission fiber, $L$ is the transmission fiber length, $\Delta f_{LO}$ is the 3-dB linewidth of the LO laser, and $T_S$ is the symbol period of the transmission system.

It is worth noting that the theoretical evaluation of the enhanced LO phase noise is only appropriate for the FD-FIR and the BLU dispersion equalization, which represent the inverse function of the fiber transmission channel without involving the phase noise mitigation [71].

Considering the effect of EEPN, the total phase noise variance in the coherent transmission system can be expressed as:

$$\sigma^2_{EEPN} = \frac{\pi \lambda^2}{2c} \cdot \frac{D \cdot L \cdot \Delta f_{LO}}{T_S}$$

where $\sigma^2_{EEPN}$ represents the total phase noise variance, $\sigma^2_{TX}$ and $\sigma^2_{LO}$ are the intrinsic phase noise variance of the TX and the LO lasers respectively, $\Delta f_{TX}$ is the 3-dB linewidth of the TX laser, and $\rho$ is the correlation coefficient between the EEPN and the intrinsic LO phase noise. We note that the approximation in Equation (4.13) is valid when the transmission length for the normal single mode fiber exceeds the order of 80 km [125].

Corresponding to the definition of the intrinsic phase noise from TX and LO lasers, we employ an effective linewidth $\Delta f_{Eff}$ to describe the total phase noise in the coherent system with EEPN [125,126], which can be defined as the following expression:

$$\Delta f_{Eff} = \frac{\sigma^2_{TX} + \sigma^2_{LO} + \sigma^2_{EEPN} + 2\rho \cdot \sigma_{LO} \cdot \sigma_{EEPN}}{2\pi T_S}$$

4.3.2 The normalized LMS filter for phase estimation

The one-tap NLMS filter can be employed effectively for carrier phase estimation [70,114], of which the tap weight is expressed as
\[ w_{NLMS}(n+1) = w_{NLMS}(n) + \frac{\mu_{NLMS}}{|x_{PN}(n)|} x_{PN}^*(n) e_{NLMS}(n) \]  
\[ e_{NLMS}(n) = d_{PE}(n) - w_{NLMS}(n) \cdot x_{PN}(n) \]

where \( w_{NLMS}(n) \) is the complex tap weight, \( x_{PN}(n) \) is the complex magnitude of the input signal, \( n \) represents the number of the symbol sequence, \( d_{PE}(n) \) is the desired symbol, \( e_{NLMS}(n) \) is the estimation error between the output signal and the desired symbols, and \( \mu_{NLMS} \) is the step size parameter.

The phase estimation using the one-tap NLMS filter resembles the performance of the ideal differential detection [66,67,70,127], of which the BER floor in the \( m \)-level PSK (\( m \)-PSK) coherent transmission systems can be approximately described by the following analytical expression,

\[ BER_{floor}^{NLMS} = \frac{1}{\log_2 m} \text{erfc}\left(\frac{\pi}{m\sqrt{2\sigma}}\right) \]

where \( \sigma^2 \) represents the total phase noise variance in the coherent transmission system.

### 4.3.3 Differential detection for phase estimation

It has been reported that the symbol delay detection can also be used for carrier phase estimation [66,67]. The coherent system can be operated in differential demodulation mode when the differential encoded data is recovered by a simple “delay and multiply algorithm” in the electrical domain. In such a case the encoded data is recovered from the received signal based on the phase difference between two consecutive symbols, i.e. the value of the complex decision variable \( \Psi = Z_k Z_{k+1}^* \exp\{i\pi/4\} \), where \( Z_k \) and \( Z_{k+1} \) are the consecutive k-th and (k+1)-th received symbols. The BER floor of the differential phase receiver can be evaluated using the principle of conditional probability [127]. For the \( m \)-PSK coherent systems, the BER floor in differential detection is expressed as the following equation,

\[ BER_{floor}^{DQPSK} = \frac{1}{\log_2 m} \text{erfc}\left(\frac{\pi}{m\sqrt{2\sigma}}\right) \]

where \( \sigma^2 \) represents the total phase noise variance in the coherent transmission system.

### 4.3.4 The block-average carrier phase estimation

The block-average method computes the \( m \)-th power of the symbols in each processing unit to cancel the phase modulation, and the calculated phase are summed and
averaged over the entire block (the length of the entire block is called block size). Then the phase is divided by \( m \), and the result leads to a phase estimate for the entire block \([65]\). For the \( m \)-PSK transmission system, the estimated carrier phase for each process unit using the BA method can be expressed as:

\[
\hat{\Phi}_{BA}(n) = \frac{1}{m} \arg \left\{ \sum_{k=\lceil (M-1)N_b \rceil}^{M N_b} x^m(k) \right\}
\]

\[
M = \left\lfloor \frac{n}{N_b} \right\rfloor
\]

where \( N_b \) is the block size in the BA method, and \( \lceil x \rceil \) represents the nearest integer larger than \( x \).

Using a Taylor series expansion, the BER floor in the block-average carrier phase estimation for the \( m \)-PSK transmission system can be approximately expressed as follows - see e.g. \([67,127]\):

\[
BER_{floor}^{BA} = \frac{1}{N_b \cdot \log_2 m} \sum_{k=1}^{N_b} \text{erfc} \left( \frac{\pi}{m \sqrt{2 \sigma_{BA,k}}} \right)
\]

\[
\sigma_{BA,k}^2 = \frac{\sigma^2}{6N_b} \left[ 2(k-1)^3 + 3(k-1)^2 + 2(N_b - k)^3 + 3(N_b - k)^2 + N_b - 1 \right], \quad k=1, \ldots, N_b.
\]

where \( \sigma^2 \) represents the total phase noise variance in the coherent transmission system.

**4.3.5 The Viterbi-Viterbi carrier phase estimation**

The Viterbi-Viterbi method also operates the symbols in each process unit into the \( m \)-th power to cancel the phase modulation. Meanwhile, the calculated phase are also summed and averaged over the entire block (the length of the block is also called block size). However, the difference with regard to the BA method is in the final step, where the extracted phase in the VV method is only concerned as the phase estimation for the central symbol in each block \([68,69]\). The extracted carrier phase in the \( m \)-PSK coherent transmission system using the Viterbi-Viterbi method can be expressed as:

\[
\hat{\Phi}_{VV}(n) = \frac{1}{m} \arg \left\{ \sum_{k=\lceil (N_v-1)/2 \rceil}^{(N_v-1)/2} x^m(n+k) \right\}, \quad N_v=1,3,5,7\ldots
\]

where \( N_v \) is the block size in the VV method.

The phase estimation in the \( m \)-PSK coherent system using the Viterbi-Viterbi algorithm can also be analyzed by employing the Taylor expansion \([67,127,128]\), and the BER floor can be described by the following approximate expression:
We can find that the carrier phase estimate error in Equation (4.27) for the Viterbi-Viterbi method corresponds to the smallest phase estimate error (phase error in the central symbol) in Equation (4.24) for the block-average method. Thus the Viterbi-Viterbi method will work better than the block-average method. Meanwhile, according to Equation (4.19), Equation (4.20) and Equation (4.26), the VV method will also show a better behavior than the one-tap NLMS and the differential detection methods in theory, when the block size is less than 12. However, it requires more computational complexity to update the process unit for the phase estimation of each symbol.

We note that the one-tap NLMS algorithm can also be employed for the \( m \)-QAM coherent transmission systems, while the block-average and the Viterbi-Viterbi methods can not be easily used for the classical \( m \)-QAM coherent systems except the circular constellation \( m \)-QAM systems.
Chapter 5

Simulation results in coherent transmission systems

In this part, the numerical simulations are carried out in the 112-Gbit/s PDM-QPSK coherent transmission system with post-compensation of CD to validate the effects of the digital chromatic dispersion compensation filters and the carrier phase estimation algorithms. Meanwhile, we also present the simulation work for the phase noise mitigation using the RF pilot tone in the 56-Gbit/s SP-QPSK coherent system with post-compensation of dispersion, and the impacts of the equalization enhanced phase noise in the 56-Gbit/s SP-QPSK coherent system with dispersion pre-distortion.

5.1 Performance of CD equalization and carrier phase estimation

We give a comparative analysis on the performance of different digital chromatic dispersion compensation filters and the behaviors of different carrier phase estimation algorithms in the 112-Gbit/s NRZ-PDM-QPSK coherent transmission system with post-compensation of CD, where the EEPN is considered.

5.1.1 CD compensation

The CD compensation results using three digital filters are illustrated in Figure 5.1. Figure 5.1(a) indicates the CD equalization with 9 taps for 20 km fiber and 243 taps for 600 km fiber using the LMS and the FD-FIR filters, as well as 16 FFT-size (8 overlap) for 20 km fiber and 512 FFT-size (256 overlap) for 600 km fiber using the BLU filter.
Figure 5.1: CD compensation using three digital filters neglecting fiber loss. (a) BER with OSNR. (b) BER with fiber length at OSNR 14.8 dB.

Obviously, the FD-FIR filter is not able to compensate the CD in 20 km fiber entirely. About 3 dB optical signal-to-noise ratio (OSNR) penalty from the back-to-back result at BER equal to $10^{-3}$ can be observed. Then we investigate the CD compensation for different fiber lengths using the three filters, which are shown in Figure 5.1(b). It can be found that the LMS filter and the BLU filter show the same acceptable performance for different fiber lengths, while the FD-FIR filter will not behave satisfactorily until the fiber length exceeds 320 km.
The CD equalization for 20 km and 600 km fibers using the LMS and the FD-FIR filters with different number of taps is shown in Figure 5.2. Due to the optimum characteristic of the LMS algorithm, the LMS filter has a slight improvement with the increment of tap number. However, the performance of the FD-FIR filter will degrade, when the tap number increases and exceeds the required tap number in Equation (4.5). It is because the redundant taps will lead to the pass-band of the filter exceeding the Nyquist frequency, which will further result in the aliasing phenomenon. We also find in Figure 5.2(a) that the FD-FIR filter does not achieve a satisfactory CD equalization performance for 20 km fiber even by using any other tap number.

From the above description, the FD-FIR filter does not achieve an acceptable CD equalization performance for short distance fibers, but it can work well for long fibers. When we use a series of delayed taps to approximate the filter time window $T^A$, the digitalized discrete time window $T^A = N^A \cdot T$ could not attain exactly the same value as the continuous time window $T^A$, which is illustrated in Figure 5.3.

The malfunction of FD-FIR filter for short fibers arises from this reason, and now we provide a more detailed explanation. We calculate the relative error $p$ of time window to evaluate the precision of time window approximation, which is given by
\[ p = \left( T_\text{W}^\Lambda - T_\text{N}^\Lambda \right) / T_\text{W}^\Lambda \] (5.1)

According to previous discussion, a short fiber will have a relative small time window to keep the signal bandwidth to be lower than Nyquist frequency to avoid the aliasing phenomenon. However, such a small time window is not easy to be digitalized accurately with a fixed sampling period \( T \). In order to broaden the time window, we need to raise the Nyquist frequency correspondingly. The Nyquist frequency is defined as half of the sampling frequency of the system, and this means we need to increase the sampling rate in the ADC modules. With sampling rate being increased, the Nyquist frequency are also raised, meanwhile, the sample period \( T \) is reduced, which allows the broadened continuous time window to be digitalized more precisely.

The relative errors of time window for different fiber length with different sampling rate are shown in Table 5.1, where the positive time error means the aliasing occurring. We could find that the relative error of time window for 20 km is reduced obviously with the sampling rate changing from 2 samples per symbol (Sa/Sy) to 8 Sa/Sy. Furthermore, the time error for 20 km fiber with 8 Sa/Sy is equal to the time error for 320 km fiber with 2 Sa/Sy, which is the acceptable fiber length limitation shown in Figure 5.1(b). So we consider this method could have a significant improving effect on the FD-FIR filter equalization performance for short fibers.

<table>
<thead>
<tr>
<th>Fiber length (km)</th>
<th>20</th>
<th>20</th>
<th>20</th>
<th>20</th>
<th>320</th>
</tr>
</thead>
<tbody>
<tr>
<td>Taps number</td>
<td>7</td>
<td>9*</td>
<td>33*</td>
<td>129*</td>
<td>129*</td>
</tr>
<tr>
<td>Sampling rate (Sa/Sy)</td>
<td>2</td>
<td>2</td>
<td>4</td>
<td>8</td>
<td>2</td>
</tr>
<tr>
<td>( T ) (ps)</td>
<td>17.9</td>
<td>17.9</td>
<td>8.9</td>
<td>4.5</td>
<td>17.9</td>
</tr>
<tr>
<td>( T_\text{W}^\Lambda ) (ps)</td>
<td>144.2</td>
<td>144.2</td>
<td>288.4</td>
<td>576.7</td>
<td>2306.8</td>
</tr>
<tr>
<td>( T_\text{N}^\Lambda ) (ps)</td>
<td>125</td>
<td>160.7</td>
<td>294.6</td>
<td>575.9</td>
<td>2303.6</td>
</tr>
<tr>
<td>( (T_\text{N}^\Lambda - T_\text{W}^\Lambda) / T_\text{W}^\Lambda ) (%)</td>
<td>-13.3</td>
<td>11.46</td>
<td>2.18</td>
<td>-0.14</td>
<td>-0.14</td>
</tr>
</tbody>
</table>

* means the limitation of the required tap number calculated in Equation (4.5)

The improved method for 20 km fiber CD compensation using the FD-FIR filter with different sampling rate is shown in Figure 5.4. We find in Figure 5.4(a) that the CD equalization performance shows an obvious improvement with the increment of the sampling rate, and the FD-FIR filter can equalize the CD in 20 km fiber entirely with 8 sampling points per symbol. The performance of the BER with normalized time window \( (T_\text{N}^\Lambda / T_\text{W}^\Lambda) \) using FD-FIR filter is shown in Figure 5.4(b), where a significant improvement can also be found. Meanwhile, we find that the FD-FIR filter shows the best behaviors when the value of \( T_\text{N}^\Lambda / T_\text{W}^\Lambda \) is around 1.0, which is consistent with our preceding analysis.

Although this improved method increases the necessary tap number in the FD-FIR filter and the required sampling rate in the coherent transmission systems, which may not be suitable for very long distance fibers and high speed communication systems, we could improve the FD-FIR filter to compensate the CD in short distance fibers significantly by increasing the ADC sampling rate. Meanwhile, we could also put an
adaptive post-filter after the FD-FIR filter to compensate the penalty in CD equalization for short distance fibers. However, here we mainly concentrate on analyzing and comparing the inherent characteristics of the three digital filters in CD compensation. Therefore, we hope to find the reason and improvement method in terms of the intrinsic properties of the FD-FIR filter. The fiber lengths are usually no less than hundreds of kilometers in practical transmission systems, therefore, the FD-FIR filter can be applied reasonably for the CD equalization in the systems with 2 Sa/Sy ADC sampling rate.

Figure 5.4: CD compensation using FD-FIR filter with different sampling rate. (a) BER with OSNR, (b) BER with normalized time window at OSNR 14.8 dB.
The CD compensation results using different frequency domain equalization methods are illustrated in Figure 5.5. The results refer to the CD equalization with 16 FFT-size for 20 km fiber and 512 FFT-size for 600 km fiber using OLS (BLU), OLA-BSZP and OLA-OSZP methods. The overlap size (or ZP) is all designated as half of the FFT-size. We can see that both of the OLA-ZP methods can provide the same acceptable performance as the OLS (BLU) method.

![CD compensation results using OLS and OLA-ZP methods.](image1)

Figure 5.5: CD compensation results using OLS and OLA-ZP methods.

![CD compensation using OLS and OLA-ZP methods with different FFT-sizes at OSNR 14.8 dB. The overlap is half of the FFT-size.](image2)

Figure 5.6: CD compensation using OLS and OLA-ZP methods with different FFT-sizes at OSNR 14.8 dB. The overlap is half of the FFT-size.

Figure 5.6 and Figure 5.7 show the performance of CD compensation for 20 km and 40 km fibers using OLS (BLU) and OLA-ZP methods with different FFT-sizes and
overlaps (or ZP), respectively. From Figure 5.6 we can see that for a certain fiber length, the three FDEs can show stable and converged acceptable performance with the increment of the FFT-size. The critical FFT-size values (16 FFT-size for 20 km fiber and 32 FFT-size for 40 km fiber), actually indicate the required minimum overlap (or ZP) value which are 8 overlap (or ZP) samples for 20 km fiber and 16 overlap (or ZP) samples for 40 km fiber. The similar performance demonstrates that for a fixed overlap (or ZP) value, the maximum compensable dispersion in the OLS (BLU) method is the same in the OLA-ZP methods.

![Figure 5.7: CD compensation for 4000 km fiber using OLS (BLU) and OLA-ZP methods with different overlaps at OSNR 14.8 dB. The FFT-size is 4096.](image)

We have demonstrated that the overlap (or ZP) is the pivotal parameter in the FDE, and the FFT-size is not necessarily designated as double of the overlap (or ZP). Figure 5.7 illustrates that with a fixed FFT-size (4096 samples) the three FDEs are still able to work well for 4000 km fiber, provided the overlap (or ZP) is larger than 1152 samples (1152=4096×9/32), which indicates the required minimum overlap (or ZP) for 4000 km fiber.

5.1.2 Carrier phase estimation

1. Carrier phase estimation with three CD equalization methods

Figure 5.8 shows the BER performance of the transmission system with different fiber length employing the optical and the digital dispersion compensation by further using a one-tap NLMS filter for phase noise compensation. Again the results are obtained under different combination of the TX laser and LO laser linewidths with the same summation. We can see clearly that influenced by the EEPN, the performance of the FD-FIR and the BLU equalization reveals obvious fiber length dependence with the increment of LO laser linewidth. The OSNR penalty in phase noise compensation scales with the LO phase fluctuation and the accumulated dispersion. This is in
agreement with previous studies [71-77]. On the other hand, the dispersion equalization using the LMS filter shows almost the same behavior in the three cases. That is because the chromatic dispersion interplays with the phase noise of both TX and LO lasers simultaneously in the adaptive equalization. Moreover, Figure 5.8 also shows the LMS filter is less tolerant against the phase fluctuation than the other dispersion compensation methods when the one-tap NLMS carrier phase noise compensator is employed.

We find that the EEPN has a significant impact on the long-haul high speed QPSK coherent system with electronic dispersion equalization, and it will induce more distortions on the high level modulation systems, such as the \(m\)-PSK and \(m\)-QAM transmission systems.
Figure 5.8: The one-tap NLMS phase estimation for different fiber length with inline DCF and digital CD compensation. (a) TX=4 MHz, LO=0 Hz, (b) TX=LO=2 MHz, (c) TX=0 Hz, LO=4 MHz.

2. Evaluation of BER floor in the one-tap NLMS phase estimation with EEPN

The performance of CPE using the one-tap NLMS filter with the FD-FIR dispersion equalization is compared with the theoretical evaluation in Equation (4.19), as shown in Figure 5.9. Figure 5.9(a) illustrates the numerical results for different combination of TX and LO lasers linewidths with the same summation. With the increment of OSNR value, the numerical simulation reveals the BER floor influenced by the phase noise, which achieves a good agreement with the theoretical evaluation.
Figure 5.9: BER performance in NLMS-CPE for 2000 km fiber with FD-FIR dispersion
equalization, T: theory, S: simulation. (a) different combination of TX and LO lasers
linewidth with the same summation, (b) only TX laser phase noise.

Figure 5.9(b) denotes the results with only the analysis of TX laser phase noise, where
a slight deviation is found between the simulation results and the theoretical analysis.
It arises from the approximation in the analytical evaluation of the one-tap NLMS
phase estimator in Equation (4.19). It has been validated in our simulation work that
the phase estimation with the BLU dispersion equalization performs closely the same
behavior as the FD-FIR equalization.

3. Evaluation of BER floor in differential phase estimation with EEPN

The BER performance of the DQPSK coherent transmission system with the FD-FIR
dispersion equalization is illustrated in Figure 5.10. Figure 5.10(a) shows the
simulation results for different combination of the TX and the LO lasers linewidths
with the same summation, and Figure 5.10(b) denotes the performance of the
differential demodulation system with only the TX laser phase noise. It is found that
the BER behavior in the DQPSK coherent system can achieve a good agreement with
the theoretical evaluation in Equation (4.20) for both Figure 5.10(a) and Figure
5.10(b). The consistence between simulation and theory in DQPSK demodulation is
better than the one-tap NLMS phase estimation in the case of only TX laser phase
noise.
Correlation between the intrinsic LO phase noise and the EEPN

In the above description, we have not discussed in detail the correlation among the intrinsic TX laser, LO laser phase noise and the EEPN. Obviously, the TX laser phase noise is independent from the LO laser phase noise and the EEPN. Here we mainly investigate the correlation between the intrinsic LO laser phase noise and the EEPN. The total phase noise variance in the coherent optical transmission system can be expressed as

Figure 5.10: BER performance in DQPSK system for 2000 km fiber with FD-FIR dispersion equalization, T: theory, S: simulation. (a) different combination of TX and LO lasers linewidth with the same summation, (b) only TX laser phase noise.

4. Correlation between the intrinsic LO phase noise and the EEPN
\[ \sigma^2 = \sigma^2_{\text{TX}} + \sigma^2_{\text{LO}} + \sigma^2_{\text{EEP}} + 2\rho \cdot \sigma_{\text{LO}} \sigma_{\text{EEP}} \]

(5.2)

where \( \rho \) is the correlation coefficient between the intrinsic LO laser phase noise and the EEPN, and we have the absolute value \( |\rho| \leq 1 \).

Figure 5.11: Phase noise correlation in SP-DQPSK system with BLU dispersion equalization, T: theory, S: simulation. (a) BER performance in different combination of EEPN and LO phase noise with the same summation, (b) correlation coefficient for different fiber length.

We have implemented the numerical simulation in a 28-Gsymbol/s SP-DQPSK system for different combination of the intrinsic LO laser phase noise and the EEPN
with the same summation, which is illustrated in Figure 5.11(a). It can be found that
the BER floor does not show tremendous variation due to the correlation between the
LO laser phase noise and the EEPN. The BER floor reaches the lowest value at
\( \sigma_{\text{EEP}}^2 = 0.5 \sigma_{\text{LO}}^2 \), which corresponds to the maximum value of the term \( |2\rho \cdot \sigma_{\text{LO}} \sigma_{\text{EEP}}| \).
The cases for \( \sigma_{\text{EEP}}^2 \gg \sigma_{\text{LO}}^2 \) and \( \sigma_{\text{EEP}}^2 = 0 \) correspond to the mutual term
\( |2\rho \cdot \sigma_{\text{LO}} \sigma_{\text{EEP}}| = 0 \). From Figure 5.11(a) we can find that \( \rho \) is usually a negative
value.

The EEPN arises from the electronic dispersion compensation where the phase of the
equalized symbol fluctuates during the time window of the digital filter, while the
intrinsic LO laser phase fluctuation comes from the integration during the consecutive
symbol period. Therefore, we could give an approximate theoretical evaluation of the
correlation coefficient as

\[
|\rho| = \frac{T_s}{N \cdot T} \quad (5.3)
\]

where \( N \) is the required tap number (or the necessary overlap) in the chromatic
dispersion compensation filter, and \( T \) is the sampling period in the transmission
system. The tap number (or the necessary overlap) can be calculated by the fiber
dispersion to be compensated [29,58], and the sampling period \( T = T_s/2 \) when the
sampling rate in the ADC modules is selected as twice the symbol rate.

The absolute value of the correlation coefficient \( |\rho| \) for different fiber length is
illustrated in Figure 5.11(b), in which we can see that the correlation coefficient \( |\rho| \)
determined from the numerical simulation achieves a good agreement with the
theoretical approximation. With the increment of fiber length, the magnitude of
correlation coefficient \( |\rho| \) approaches zero rapidly. Consequently, we can neglect the
correlation term in Equation (5.2) when the fiber length is over 80 km. Therefore, the
assumption in Equation (4.13) and Equation (4.16) is available when the fiber length
exceeds 80 km in the practical optical communication systems.

5. Evaluation of BER floor in BA carrier phase estimation with EEPN

The performance of the BA carrier phase extraction method with different block size
in the 112-Gbit/s NRZ-PDM-QPSK coherent system with post-compensation of CD is
illustrated in Figure 5.12. The transmission system is with 2000 km optical fiber. Here
we use the FDE with 2048 fast Fourier transform (FFT) size and 1024 overlap for the
CD compensation. We can find that for a small block size \( (N_b=3) \), the theoretical BER
floors are much lower than the simulation results. Because for the minimum block
size \( (N_b=1) \) in the BA algorithm, the theoretical prediction of the BER floor in
Equation (4.23) is zero. This indicates that Equation (4.23) - which is based on a
leading order Taylor expansion of the phase noise influence - is not accurate. A higher
order Taylor approximation is required. Therefore, for a relative small block size
\( (N_b=3) \), the theoretical BER floors will be below the simulation results. For a large
block size \( (N_b=11) \), the theoretical BER floor predictions are above the simulation
results. Because the theoretical BER floor in the BA method is derived based on the Taylor expansion [67,127], which will not work well when both the block size ($N_b=11$) and the phase noise variance are large. A similar phenomenon has also been found in previous report [67]. Therefore, for an eclectic block size ($N_b=5$), the simulation results can make a good agreement with the theoretical predictions.
Figure 5.12: CPE using BA method with different block size $N_b$. (a) $N_b=3$, (b) $N_b=5$, (c) $N_b=11$.

The theoretical BER floor is $4.7 \times 10^{-6}$ for the case of “TX=LO=5 MHz’ in (a).

6. Evaluation of BER floor in VV carrier phase estimation with EEPN

Figure 5.13 shows the performance of the VV carrier phase extraction method with different block size in the 112-Gbit/s NRZ-PDM-QPSK coherent system with post-compensation of dispersion, where the transmission fiber is also 2000 km. We also employ the FDE with 2048 FFT-size and 1024 overlap for the CD compensation.
Figure 5.13: CPE using VV method with different block size \( N_v \). (a) \( N_v=5 \), (b) \( N_v=11 \), (c) \( N_v=15 \). The theoretical BER floor is \( 1.5 \times 10^{-7} \) for the case of “TX=LO=5 MHz” in (a).

A tendency similar to the BA method can be found in the VV phase estimation algorithm. For a small block size \( \left( N_v=5 \right) \), the theoretical BER floors are lower than the simulation results. Because the minimum block size \( \left( N_v=1 \right) \) in the VV method also corresponds to a zero BER floor in Equation (4.26). This indicates that Equation (4.26) - based on a leading order Taylor expansion of the phase noise - is not accurate either. A higher order Taylor approximation is required. Therefore, for a relative small block size \( \left( N_v=5 \right) \), the theoretical BER floors will be below the simulation results. For a large block size \( \left( N_v=15 \right) \), the theoretical predictions of the BER floor are above the simulation results. This is because a large block size \( \left( N_v=15 \right) \) and a large phase noise
variance could not achieve a good approximation in the Taylor expansion for the theoretical calculation of BER floor in Equation (4.26) [67,127,128]. Similar to the BA method, the simulation results could agree well with the theoretical predictions for an eclectic block size ($N_v=11$). We can find that the eclectic block size (simulation results matching the theoretical predictions) in the BA method is smaller than in the VV method.

As discussed in the above, the analytical evaluation of the BER floors in the BA and the VV carrier phase extraction methods is derived based on the Taylor expansions. However, Taylor expansion methods - to the leading order in the phase noise – are not accurate for the large phase noise and block size values. In most practical cases, the BA and the VV algorithms allow the large phase noise and the block size parameters, and the Taylor expansion based analytical predictions are not appropriate. Therefore, a proper analysis for the practical transmission system should be described based upon the simulation results.

7. Comparison of the NLMS, the BA, and the VV carrier phase estimation

The performance of different carrier phase estimation algorithms (the NLMS, the BA, and the VV methods) with different block size is illustrated in Figure 5.14, where the transmission fiber length is 2000 km, and the TX and the LO linewidths are both 5 MHz. The FDE with 2048 FFT-size and 1024 overlap is employed for the CD compensation. The step size in the one-tap NLMS algorithm is optimized. For the one-tap NLMS method, the theoretical BER floor always makes a good agreement with the simulation result, if the step size is optimized.
Figure 5.14: Performance of three CPE methods with different block size. (a) block size is 1, (b) block size is 5, (c) block size is 11. The theoretical BER floor is $1.5 \times 10^{-7}$ for the VV method in (b).

When the block size is one, the BA algorithm shows exactly the same behavior with the VV algorithm, which is a little better than the one-tap NLMS method. Meanwhile, as described in the above section, the theoretical prediction of the BER floors matches the simulation results, only if the block size is 5 for the BA method and the block size is 11 for the VV method. Furthermore, we can find that the VV method does not make such an improvement than the BA method, even it sacrifices more complexity.

Figure 5.15 indicates the tolerable total effective linewidth for the three carrier phase
extraction algorithms (the NLMS, the BA, and the VV methods) with different block size in the 112-Gbit/s NRZ-PDM-QPSK transmission system with post-compensation of dispersion. Figure 5.15(a) is the theoretical evaluation, and Figure 5.15(b) is the numerical simulation result.

Figure 5.15: Maximum tolerable effective linewidth for different BER floors \((10^{-2}, 10^{-3}, 10^{-4})\) in the three methods versus the block size. (a) theoretical predictions, (b) simulation results.

According to the theoretical prediction in Figure 5.15(a), we can find that the block size has a significant influence on the performance of the BA and the VV methods. The BA and the VV methods degrade dramatically with the increment of the block size. The BA method is much better (allowing larger effective linewidth) than the
NLMS method with the block size less than 5, and the VV method is much better than the NLMS method with the block size less than 13. Meanwhile, the VV method gives a much better behavior than the BA method. However, the simulation results in Figure 5.15(b) demonstrate that the BA and the VV methods have a weaker dependence on the block size compared to the theoretical analysis. On the one hand, the BA method behaves a little better (allowing larger effective linewidth) than the NLMS method when the block size is less than 11, and the VV method works slightly better than the NLMS method when the block size is less than 21. On the other hand, the Viterbi-Viterbi method does not show a considerable improvement compared to the block-average method, even if it sacrifices more computational complexity.

The weak dependence on the block size in the BA and the VV algorithms implies that the additive noise in the transmission channel of the practical coherent systems can be accommodated quite well, since this requires a large block size to mitigate the additive Gaussian noise. Meanwhile, the NLMS method can also show a good performance with the additive noise in the transmission channel, if the step size is optimized [70,125].

It is worth noting that the one-tap NLMS algorithm can also be employed for the $n$-QAM coherent transmission systems, while the block-average and the Viterbi-Viterbi methods can not be easily used for the classical $n$-QAM coherent systems except the circular constellation $n$-QAM systems.

5.2 Phase noise mitigation using RF pilot tone

We investigate the phase noise mitigation using the RF pilot tone in the 56-Gbit/s SP-QPSK coherent transmission system with post-compensation of dispersion.

![Figure 5.16: BER performance for SP-QPSK coherent system using RF pilot tone. The transmission distance is 10 km, and the TX and the LO lasers linewidths are 85 MHz.](image)

Figure 5.16: BER performance for SP-QPSK coherent system using RF pilot tone. The transmission distance is 10 km, and the TX and the LO lasers linewidths are 85 MHz.
The performance in the case of a transmission distance of 10 km for normal transmission fiber with dispersion coefficient of $D=16 \text{ ps/nm/km}$ is illustrated in Figure 5.16. In the case considered here the equalization enhanced phase noise (EEPN) is negligible when we consider a TX and LO linewidth of 85 MHz which exemplifies the use of distributed feedback (DFB) laser diodes of poor quality. From the figure it appears that without phase noise compensation the phase noise generates an error-rate floor around the $10^{-3}$ level and that this is removed well below $10^{-4}$ by the use of an optical RF carrier. This result is in good qualitative agreement with previous report where a 10-Gbit/s 16-QAM system with an optical RF pilot tone was considered [105]. The use of an RX generated RF pilot tone is slightly less efficient but does for a short modulation (PRBS) sequence of $2^7-1$ move the error rate floor below $10^{-4}$. It should be noted that we are using a standard 5-th order Butterworth high-pass filter (HPF) in our simulations and that it may be possible to improve the result by considering a more carefully designed HPF for the purpose of the most efficient removal of the phase noise.

Figure 5.17: Performance of SP-QPSK coherent system with large EEPN using RF pilot tone.

The transmission distance is 2000 km, the TX and the LO linewidths are both 5 MHz.

In Figure 5.17 we consider a situation with a TX linewidth of 5 MHz, an LO linewidth of 5 MHz and a transmission distance of 2000 km. In this case the total phase noise in the RX is dominated by EEPN and it amounts to a total effective linewidth of 216 MHz. Figure 5.17 shows that the phase noise reduction using the RF pilot tone is not very efficient when the phase noise is dominated by EEPN. The reduction is less than one order of magnitude even using optical RF tone generation, and the RX generated pilot tone gives slightly poorer performance than the optically transmitted one. This behavior is attributed to the well known fact that the EEPN results in a complex combination of pure phase noise, amplitude noise and time jitter, and neither the amplitude noise nor the time jitter is compensated using the RF pilot tone.
The optically or electronically generated RF pilot tone can also be used to eliminate the phase noise influence in high constellation transmission systems, such as \(m\)-PSK and \(m\)-QAM coherent systems, by complex conjugation prior to the signal detection and error-rate specification.

### 5.3 EEPN effects in pre-distorted transmission system

In this section we will give the simulation results for the carrier phase estimation in the 56-Gbit/s SP-QPSK coherent system with dispersion pre-distortion. The CD equalization is performed by the FD-FIR filtering with a number of active taps which are adjusted according to the amount of chromatic dispersion [29]. The transmission fiber is a normal single-mode fiber with dispersion coefficient of \(D=16\) ps/nm/km. The one-tap NLMS filter is used in the RX for the carrier phase estimation.

![Figure 5.18: BER for SP-QPSK coherent system using pre-compensation of CD. The PRBS length is \(2^{16}-1\). S: simulation results, T: BER floor using Equation (4.19).](image)

Figure 5.18 shows the BER versus the OSNR for 2000 km transmission distance for this pre-distorted system, where the EEPN is dominating. From the figure it appears clearly that when the TX laser linewidth is zero, the BER-floor is well below \(10^{-3}\). When we have the TX-linewidth of 10 MHz, the BER-floor is around \(10^{-2}\). For the linewidth of 5 MHz in both lasers we have a BER floor of around \(10^{-3}\). Compared to previous post-compensation systems, the results are very similar except that in the pre-distortion implementation the EEPN results from the TX laser linewidth. We also observe a slightly poorer agreement between the theoretical BER floor predicted by Equation (4.19) and the simulation results for the pre-compensation case. This is tentatively attributed to the fact that in the post-compensation case the EEPN is generated by the digital CD equalizer in the RX, and this is the basis for the specification of the resulting phase noise variance in Equation (4.13) [71,125].
However, in the pre-compensation case the EEPN is generated in the analogue domain (by the transmission fiber dispersion) which is a somewhat different physical phenomenon.

It is obvious that the electrical CD equalization is the origin of the EEPN influence for both pre- and post-compensation implementations. This is a major difficulty in the practical implementation of the long-range high-capacity coherent optical systems with high level constellation (\(m\)-PSK and \(m\)-QAM systems). It is worth noting that this difficulty can be avoided by using pure optical CD equalization i.e. using dispersion compensating fibers (DCFs). The use of DCFs will completely eliminate the EEPN generation, and it seems to be an obvious choice for the practical implementation of advanced long-haul coherent optical transmission systems.
Chapter 6

Conclusions

6.1 Summary of the dissertation work

In this dissertation, we present a comparative analysis of different digital filters for chromatic dispersion compensation and a comparative evaluation of different carrier phase estimation methods considering equalization enhanced phase noise. These investigations are performed in the 112-Gbit/s NRZ-PDM-QPSK transmission system with post-compensation of dispersion, the 56-Gbit/s SP-QPSK post-compensated coherent system with RF pilot tone, and the 56-Gbit/s SP-QPSK coherent system with dispersion pre-distortion.

In CD equalization, the LMS adaptive filter shows the best performance in terms of safety and stability. However, it requires slow iteration for guaranteed convergence, and also the tap weights update increases the computational complexity. The FD-FIR filter affords the simplest analytical tap weights specification with respect to equalizer specification. However, it does not show acceptable performance for short distance fibers. The blind look-up filter will be faster and much more computationally efficient from the aspect of speed and efficiency, especially for large fiber dispersion. However, its performance will degrade dramatically if the overlap in the equalizer does not reach the required minimum overlap size.

In the investigation of the EEPN in carrier phase recovery, the carrier phase estimation is implemented by using the one-tap normalized LMS filter, the differential detection, the block average algorithm and the Viterbi-Viterbi algorithm. In the FD-FIR and the BLU dispersion equalization, the BER floors of the four carrier phase estimation methods with EEPN are analytically evaluated, and the theoretical predictions are compared to numerical simulations. We find that the theoretical BER floors in the one-tap NLMS algorithm and the differential detection can always make good agreements with the simulation results. However, it is not the case for the block-average and the Viterbi-Viterbi methods, where the theoretical BER floors only agree with the simulation results when a certain eclectic block size is employed. For the design of practical transmission systems, simulations should be applied for the evaluation of the carrier phase estimation methods. The one-tap NLMS method can show an acceptable behavior compared to the other approaches, while the optimization for the step size needs a complicated empirical selection. The differential detection has nearly the same behavior with the one-tap NLMS filter, except that it has about 2.5 dB OSNR penalty compared to the methods without using differential delay. The block-average method is easy and efficient to implement, but it will behave unsatisfactory when a large block size is used. The Viterbi-Viterbi method can show a slight improvement compared to the block-average method, while it sacrifices more computational complexity than the other methods.
Meanwhile, we also use the RF pilot tone to eliminate the phase noise influence in the QPSK coherent system with post-compensation of dispersion. It is found that the electronically generated RF carrier provides slightly less efficient phase noise mitigation than the optically transmitted one, but still it improves the phase noise tolerance by about one order of magnitude when a short modulation sequence is used. It is also found that equalization enhanced phase noise which appears as correlated pure phase noise, amplitude noise, and time jitter in the received signal, cannot be efficiently mitigated by the use of an (optically or electrically generated) RF pilot tone.

Furthermore, we have presented a study to specify the influence of equalization enhanced phase noise for pre- and post-compensation of chromatic dispersion in the QPSK coherent systems. Our results show that the LO phase noise determines the EEPN influence in the post-compensation implementations, whereas the TX laser phase noise determines the EEPN influence in the pre-compensation implementations. It is to be emphasized that the use of chromatic dispersion compensation in the optical domain, such as the use of DCFs, can eliminate the EEPN entirely. Thus, this seems a good option for the long-haul transmission systems operating at high constellations in the future.

6.2 Summary of the appended papers

**Paper I**

We present a novel investigation on the enhancement of phase noise in coherent optical transmission system due to electronic chromatic dispersion compensation. Two types of equalizers, including the time domain FD-FIR filter and the frequency domain BLU filter are applied to mitigate the CD in the 112-Gbit/s PDM-QPSK transmission system. The BER floor in phase estimation using the optimized one-tap NLMS filter, and considering the EEPN is evaluated analytically including the correlation effects. The numerical simulations are implemented and compared with the performance of differential QPSK demodulation system.

**Paper II**

A comparative analysis of three popular digital filters for chromatic dispersion compensation involving a time-domain least mean square adaptive filter, a time-domain fiber dispersion finite impulse response filter and a frequency-domain blind look-up filter, are applied to equalize the CD in a 112-Gbit/s NRZ-PDM-QPSK coherent transmission system in this paper. The characteristics of these filters are compared by evaluating their applicability for different fiber lengths, their usability for dispersion perturbations, and their computational complexity.

**Paper III**

In this paper, an adaptive finite impulse response filter employing normalized LMS algorithm is developed for compensating the CD in a 112-Gbit/s PDM-QPSK coherent communication system. The principle of the adaptive normalized LMS
algorithm for signal equalization is analyzed theoretically, and at the meanwhile, the taps number and the tap weights in the adaptive FIR filter for compensating a certain fiber dispersion are also investigated by numerical simulation. The CD compensation performance of the adaptive filter is analyzed by evaluating the behavior of the BER versus the OSNR, and the results are compared with other present digital filters.

**Paper IV**

The frequency domain equalizers employing two types of overlap-add zero-padding methods are applied to compensate the chromatic dispersion in the 112-Gbit/s NRZ-PDM-QPSK coherent transmission system. Simulation results demonstrate that the OLA-ZP methods can achieve the same acceptable performance as the overlap-save method. The required minimum overlap (or zero-padding) in the FDE is derived, and the optimum fast Fourier transform length to minimize the computational complexity is also analyzed.

**Paper V**

In this paper, a novel method for extracting an RF pilot carrier signal in the coherent receiver is presented. The RF carrier is used to mitigate the phase noise influence in n-level PSK and QAM systems. The performance is compared to the use of an (ideal) optically transmitted RF pilot tone. The electronically generated RF carrier provides less efficient phase noise mitigation than the optical RF. However, the electronically generated RF carrier still improves the phase noise tolerance by about one order of magnitude in bit-error-rate compared to using no RF pilot tone. It is also found that equalization enhanced phase noise - which appears as correlated pure phase noise, amplitude noise and time jitter - cannot be efficiently mitigated by the use of an (optically or electrically generated) RF pilot tone.

**Paper VI**

The RF carrier can be used to mitigate the phase noise impact in n-level PSK and QAM systems. The systems performance is influenced by the use of an RF pilot carrier to accomplish phase noise compensation through complex multiplication in combination with discrete CD compensation filters. We perform a detailed study comparing two filters for the CD compensation namely the fixed FDE and the adaptive LMS filter. The study provides important novel physical insight into the EEPN influence on the system BER versus OSNR performance. Important results of the analysis are that the FDE position relative to the RF carrier phase noise compensation module provides a possibility for choosing whether the EEPN from the TX or the LO laser influences the system quality. The LMS filter works very inefficiently when placed prior to the RF phase noise compensation stage of the RX whereas it works much more efficiently and gives almost the same performance as the FDE when placed after the RF phase noise compensation stage.

**Paper VII**

The analytical model for the phase noise influence in differential n-level phase shift
keying (n-PSK) systems and 2n-level quadrature amplitude modulated (2n-QAM) systems employing electronic dispersion equalization and quadruple carrier phase extraction is presented. The model includes the dispersion equalization enhanced local oscillator phase noise influence. Numerical results for phase noise error-rate floors are given for dual polarization DQPSK, D16PSK and D64PSK system configurations with basic baud-rate of 25 GS/s. The transmission distance in excess of 1000 km requires local oscillator lasers with sub MHz linewidth.

**Paper VIII**

In this paper we present a comparative study in order to specify the influence of EEPN for pre- and post-compensation of chromatic dispersion in high capacity and high constellation systems. Our results show that the local oscillator phase noise determines the EEPN influence in post-compensation implementations whereas the transmitter laser determines the EEPN in pre-compensation implementations. As a result of significance for the implementation of practical longer-range systems it is to be emphasized that the use of chromatic dispersion equalization in the optical domain – e.g. by the use of dispersion compensation fibers – eliminates the EEPN entirely. Thus, this seems a good option for such systems operating at high constellations in the future.

**Paper IX**

In this paper, we demonstrate the chromatic dispersion equalization employing a time-domain FIR filter in a 112-Gbit/s PDM-QPSK coherent communication system. The required tap number of the filter is analyzed from anti-aliasing and pulse broadening. The dynamic range of the filter is evaluated by using different number of taps. We find that the time domain FIR filter does not work well for the coherent systems with short transmission distance fibers. However, this penalty can be compensated by using a post-added few-tap LMS adaptive filter or by increasing the sampling rate in the ADCs modules.

**Paper X**

In this paper, we investigate the phase noise elimination employing an optical pilot carrier in the high speed coherent transmission system considering the EEPN. The numerical simulations are performed in a 28-Gsymbol/s QPSK coherent system with a polarization multiplexed pilot carrier. The carrier phase estimation is implemented by the one-tap NLMS filter and the differential phase detection, respectively. Simulation results demonstrate that the application of the optical pilot carrier is very effective for the intrinsic laser phase noise cancellation, while is less efficient for the EEPN mitigation.

### 6.3 Suggestions for our future work

In the CD compensation investigation, to demonstrate the fundamental features of the time-domain adaptive and fixed filters as well as the frequency-domain equalizers, different digital filters are applied to compensate the CD in the 112-Gbit/s
NRZ-PDM-QPSK coherent optical transmission system with post-compensation of dispersion. Besides the chromatic dispersion equalization, our analysis does not take into account the influences of PMD and fiber nonlinearities, only the carrier phase estimation is investigated. Future efforts should incorporate comparing the CD equalization performance of these methods in the return-to-zero (RZ) and the NRZ polarization division multiplexed QPSK coherent transmission systems, as well as compensating the PMD and the fiber nonlinearities in such coherent systems. Furthermore, these CD compensation methods should be studied for the utilization in the QAM coherent systems.

In the carrier phase recovery investigation, three electronic CD equalizers are applied to compensate the dispersion in the coherent optical transmission system to investigate the impact of the dispersion equalization enhanced phase noise. The carrier phase estimation is implemented by using the one-tap normalized LMS filter, the differential phase detection, the block-average method and the Viterbi-Viterbi method. The analytical predictions are compared to the simulation results. Further investigation will involve the evaluation of the BER floor in phase estimation with LMS adaptive CD equalization, which is rather complicated due to the equal enhancement of both the TX and the LO lasers phase noise. Moreover, the entire mitigation of the EEPN in coherent transmission systems will be also studied in the future investigations.
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Acronyms

ADCs: analog-to-digital convertors
AM: amplitude modulator/modulation
ASE: amplified spontaneous emission
ASK: amplitude shift keying
BER: bit-error-rate
BLU: blind look-up
CD: chromatic dispersion
CMA: constant modulus algorithms
CPE: carrier phase estimation
CPFSK: continuous-phase frequency shift keying
CW: continuous wave
DCFs: dispersion compensating fibers
DCMs: dispersion compensating modules
DGD: differential group delay
DPSK: differential phase shift keying
DQPSK: differential quadrature phase shift keying
DSP: digital signal processing
EDFAs: erbium-doped fiber amplifiers
EEPN: equalization enhanced phase noise
FDEs: frequency domain equalizers
FD-FIR: fiber dispersion finite impulse response
FFT: fast Fourier transform
FSK: frequency shift keying
FWM: four-wave mixing
GVD: group velocity dispersion
HPF: high pass filter/filtering
IF: intermediate frequency
IFFT: inverse fast Fourier transform
IMDD: intensity modulation direct detection
IQ: in-phase and quadrature
ISI: inter-symbol interference
LMS: least mean square
LO: local oscillator
MLSE: maximum likelihood sequence estimation
NLMS: normalized least mean square
NLSE: nonlinear Schrödinger equation
NRZ: non-return-to-zero
OFDM: orthogonal frequency division multiplexing
OLA: overlap-add
OLA-BSZP: overlap-add both-side zero-padding
OLA-OSZP: overlap-add one-side zero-padding
OLA-ZP: overlap-add zero-padding
OLS: overlap-save
OSNR: optical signal-to-noise ratio
PBC: polarization beam combiner
PD: photodiode
PDM: polarization division multiplexed/multiplexing
PLL: phase-locked loop
PM: phase modulator/modulation
PMD: polarization mode dispersion
PN: phase noise
PRBS: pseudo random bit sequence
PSK: phase-shift keying
QAM: quadrature amplitude modulation
QPSK: quadrature phase shift keying
RF: radio frequency
RZ: return-to-zero
SDM: space division multiplexed/multiplexing
SNR: signal-to-noise ratio
SOP: state of polarization
SP: single polarization
SPM: self-phase modulation
SSMF: standard single mode fiber
TX: transmitter
VLSI: very large-scale integration
WDM: wavelength-division multiplexed/multiplexing
XPM: cross-phase modulation