HIGH-SPEED ELECTRONIC SIGNAL PROCESSING FOR PRE-COMPENSATION IN OPTICAL COMMUNICATIONS

by

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Abstract

Narrowband optical filtering and chromatic dispersion are two important issues that affect optical fiber transmission performance. Recent technological developments in high-speed digital signal processors, digital-to-analog converters and analog-to-digital converters have enabled the implementation of electronic signal processing (ESP) in optical transmission systems leading to adaptive and cost efficient integrated solutions. This thesis focuses on applying ESP at the transmitter to pre-compensate for narrowband optical filtering and chromatic dispersion.

A novel electronic pre-compensation approach was proposed to deal with narrowband optical filtering. The effectiveness was demonstrated by a straight-line experiment and a recirculating loop experiment for 10 Gb/s non-return-to-zero on-off-keying (NRZ-OOK). Moreover, the work was extended to NRZ differential-phase-shift-keying as well as 20 Gb/s NRZ differential-quadrature-phase-shift-keying. Experimental results demonstrate that electronic pre-compensation effectively reduces the degradation in system performance induced by narrowband optical filtering.

Electronic dispersion pre-compensation was investigated using a semiconductor InP Mach-Zehnder modulator (MZM) for the NRZ-OOK modulation format at 10.709 Gb/s aiming at providing a cost efficient implementation for core and metro transmission networks. A brute-force method was developed to determine the requisite drive
voltages due to the nonlinear voltage dependent attenuation and phase constants of the InP MZM. The transmission results for the recirculating loop and straight-line experiments demonstrate that an InP MZM provides comparable dispersion pre-compensation performance with a conventional LiNbO$_3$ MZM. Use of the NRZ-OOK modulation format and InP MZM provides a simple and cost-efficient solution for core and metro transmission network.

Dispersion pre-compensation was also performed for a 85.672 Gb/s polarization multiplexed 16-ary quadrature amplitude modulation (PM-16QAM) modulation format with digital coherent detection and offline digital signal processing. The transmitter was characterized to ensure the quality of the 16QAM signal generation. Simulation results indicate the impact of the modulator bias voltage error on system performance. Recirculating loop experimental results demonstrate that the performance of dispersion pre-compensation is comparable with dispersion post-compensation, thus providing the possibility to combine dispersion pre- and post-compensation for PM-16QAM coherent transmission for further performance improvement.
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Gratefully appreciate to Corning, Inc. for offering their recirculating loop and digital coherent receiver which contribute the completion of this thesis. Dr. John D. Downie is particularly acknowledged.

I am also appreciate my family in which education is always regarded as the most important thing and this encourages me to make the decision to study for the PhD. My parents and sister are always my most beloved person.

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# Acronyms

<table>
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<tr>
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<th>Description</th>
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<tbody>
<tr>
<td>ADC</td>
<td>analog-to-digital converter</td>
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<tr>
<td>AOM</td>
<td>acousto-optic modulator</td>
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<tr>
<td>AOWG</td>
<td>arbitrary optical waveform generator</td>
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<tr>
<td>APSK</td>
<td>amplitude phase shift keying</td>
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<tr>
<td>ASE</td>
<td>amplified spontaneous emission</td>
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<tr>
<td>ASIC</td>
<td>application-specific integrated circuit</td>
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<tr>
<td>BER</td>
<td>bit-error-ratio</td>
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<td>BPF</td>
<td>band pass filter</td>
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<td>CPE</td>
<td>carrier phase estimation</td>
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<tr>
<td>CW</td>
<td>continuous wave</td>
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<tr>
<td>DAC</td>
<td>digital-to-analog converter</td>
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<td>DC</td>
<td>direct current</td>
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<tr>
<td>DCF</td>
<td>dispersion compensating fiber</td>
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<td>DD-MZM</td>
<td>dual-drive Mach-Zehnder modulator</td>
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<tr>
<td>DFB</td>
<td>distributed feedback</td>
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<td>DFE</td>
<td>decision-feedback equalization</td>
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<tr>
<td>DP-MZM</td>
<td>dual-parallel Mach-Zehnder modulator</td>
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<tr>
<td>DPSK</td>
<td>differential-phase-shift-keying</td>
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<td>Abbreviation</td>
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<tr>
<td>DQPSK</td>
<td>differential-quadrature-phase-shift-keying</td>
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<tr>
<td>DSP</td>
<td>digital signal processing</td>
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<tr>
<td>ECL</td>
<td>external cavity laser</td>
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<td>ECP</td>
<td>eye-closure penalty</td>
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<td>EDC</td>
<td>electronic dispersion compensation</td>
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<td>EDFA</td>
<td>erbium doped fiber amplifier</td>
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<td>ESP</td>
<td>electronic signal processing</td>
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<td>EVM</td>
<td>error vector magnitude</td>
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<td>FDE</td>
<td>frequency-domain equalization</td>
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<td>FPGA</td>
<td>field programmable gate array</td>
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<td>GVD</td>
<td>group velocity dispersion</td>
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<td>IMDD</td>
<td>intensity modulation and direct detection</td>
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<td>ISI</td>
<td>intersymbol interference</td>
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<td>local oscillator</td>
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<td>maximum likelihood sequence estimation</td>
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<td>MZDI</td>
<td>Mach-Zehnder delay line interferometer</td>
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<td>MZM</td>
<td>Mach-Zehnder modulator</td>
</tr>
<tr>
<td>MZI</td>
<td>Mach-Zehnder interferometer</td>
</tr>
<tr>
<td>NRZ</td>
<td>non-return-to-zero</td>
</tr>
<tr>
<td>NZDSF</td>
<td>non-zero dispersion shifted fiber</td>
</tr>
<tr>
<td>OOK</td>
<td>on-off-keying</td>
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xx
<table>
<thead>
<tr>
<th>Abbreviation</th>
<th>Description</th>
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<tbody>
<tr>
<td>OSA</td>
<td>optical spectrum analyzer</td>
</tr>
<tr>
<td>OSNR</td>
<td>optical signal-to-noise ratio</td>
</tr>
<tr>
<td>PBS</td>
<td>polarization beam splitter</td>
</tr>
<tr>
<td>PBC</td>
<td>polarization beam combiner</td>
</tr>
<tr>
<td>PDL</td>
<td>polarization-dependent loss</td>
</tr>
<tr>
<td>PLL</td>
<td>phase-locked loop</td>
</tr>
<tr>
<td>PM</td>
<td>polarization multiplexed</td>
</tr>
<tr>
<td>PMD</td>
<td>polarization mode dispersion</td>
</tr>
<tr>
<td>PSK</td>
<td>phase-shift-keying</td>
</tr>
<tr>
<td>QAM</td>
<td>quadrature amplitude modulation</td>
</tr>
<tr>
<td>RZ</td>
<td>return-to-zero</td>
</tr>
<tr>
<td>ROADM</td>
<td>reconfigurable optical add-drop multiplexer</td>
</tr>
<tr>
<td>SE</td>
<td>spectral efficiency</td>
</tr>
<tr>
<td>SNR</td>
<td>signal-to-noise ratio</td>
</tr>
<tr>
<td>SSMF</td>
<td>standard single-mode fiber</td>
</tr>
<tr>
<td>SPM</td>
<td>self-phase modulation</td>
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<tr>
<td>TDE</td>
<td>time-domain equalization</td>
</tr>
<tr>
<td>VOA</td>
<td>variable optical attenuator</td>
</tr>
<tr>
<td>VNA</td>
<td>vector network analyzer</td>
</tr>
<tr>
<td>WDM</td>
<td>wavelength division multiplexed</td>
</tr>
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</table>
Chapter 1

Introduction

1.1 Electronic Signal Processing Technology

The potential of utilizing electronic signal processing (ESP) in optical fiber communication systems to mitigate the effect of chromatic dispersion and polarization mode dispersion (PMD) was first demonstrated in 1990 [1]. In a computer simulated 8 Gb/s system, different ESP techniques such as linear equalization via a tapped delay line, nonlinear cancelations using variable threshold detection, coding, and multilevel signaling were investigated. Although the simulation results showed great performance improvement, ESP was not implemented in optical systems until the late 1990s due to the lack of high-speed electronics.

However, over the past 10 years, ESP technology has undergone substantial technological developments and attracted research efforts aimed at increasing the per-channel bit rate and extending the transmission distance. Additional ESP techniques have appeared such as electronic equalizers and electronic pre-compensation. Moreover, the realization of high-speed digital signal processors, digital-to-analog converters (DACs) and analog-to-digital converters (ADCs) has made ESP feasible for
current optical fiber communication systems at bit rates of 10 Gb/s and above, as shown in Table 1.1. ESP has been implemented in the transmitter [3] and/or the receiver [5], [10] to mitigate signal distortions that occur during fiber transmission.

<table>
<thead>
<tr>
<th>Year</th>
<th>ESP technology development</th>
</tr>
</thead>
<tbody>
<tr>
<td>2004</td>
<td>25 GSa/s 3 bit ADC in 130 nm CMOS application-specific integrated circuit (ASIC) [2]</td>
</tr>
<tr>
<td>2005</td>
<td>22 GSa/s 6 bit DAC in 130 nm BiCMOS ASIC [3]</td>
</tr>
<tr>
<td>2006</td>
<td>22 GSa/s 5 bit ADC in 130 nm BiCMOS [4]</td>
</tr>
<tr>
<td>2007</td>
<td>20 GSa/s 4 bit ADC in 90 nm CMOS ASIC [5]</td>
</tr>
<tr>
<td>2008</td>
<td>20 GSa/s 8 bit DAC in 250 nm BiCMOS [6]</td>
</tr>
<tr>
<td>2009</td>
<td>35 GSa/s 4 bit flash ADC-DAC in 180 nm BiCMOS [7]</td>
</tr>
<tr>
<td>2010</td>
<td>40 GSa/s 4 bit ADC in 130 nm BiCMOS [8]</td>
</tr>
<tr>
<td>2011</td>
<td>56 GSa/s 6 bit DAC in 65 nm CMOS [9]</td>
</tr>
</tbody>
</table>

1.2 ESP in Direct Detection Receiver

In an intensity modulation and direct detection (IMDD) system, binary information is translated into optical intensity, that is, 1’s are represented by high optical intensity and 0’s by low intensity. At the receiver, the incoming optical signal is converted to an electrical signal that is proportional to the power of the incoming signal. Thus, high electrical output is for 1’s and low for 0’s. Several ESP techniques, such as feed-forward equalization (FFE), decision-feedback equalization (DFE) and maximum likelihood sequence estimation (MLSE), have been used with direct detection to deal with intersymbol interference (ISI) induced by chromatic dispersion and PMD.
In IMDD systems, as electronic analog signal processing structures, FFE and DFE have been realized to mitigate chromatic dispersion and PMD for bit rates of 10 Gb/s [11], [12] and 40 Gb/s [13]. Figure 1.1 shows the structure of a 5 - tap feed-forward equalizer. Its output is given by [12]:

\[ y(t) = \sum_{k=0}^{N-1} c_k \cdot x(t - k \cdot \Delta t) \]  \hspace{1cm} (1.1)

where \( c_k \) are the variable tap weights. \( x(t) \) is the input signal at time \( t \). \( \Delta t \) is the tap spacing. \( N \) is the number of the equalizer taps. Since the output signal is a linear combination of delayed versions of the input signal, the feed-forward equalizer is referred to as a linear equalizer. In order to avoid noise enhancement and also ensure stable adaptation of the tap weights, the tap spacing is chosen from between 0.3 - 0.6 of the bit period [14]. Typically, half the bit period is used (\( \Delta t = 50 \) ps for a 10 Gb/s system). The number of feed-forward equalizer taps should not be too large as more taps will increase the implementation complexity and power consumption. Moreover, a large number of taps may induce adaptation-related slow convergence and instability [14].

Figure 1.1: A 5-tap feed-forward equalizer.
The decision feedback equalizer is a nonlinear equalizer which provides better performance than a feed-forward equalizer when coping with severe signal distortion [14]. It consists of a feed-forward filter followed by a feedback filter, as shown in Figure 1.2. The feed-forward and feedback filters are conceptually similar to the feed-forward equalizer. The input signal is passed through a feed-forward filter. The decisions of the past symbols are fed back into a feedback filter whose output is then subtracted from the output of the feed-forward filter. Therefore, the feed-forward filter is used to deal with part of the ISI while the remaining ISI is removed by the feedback filter.

The analog implementation of the feed-forward and decision feedback equalizers provides a significant power consumption advantage with a possible sacrifice in performance compared to a digital implementation. However, since the feed-forward and decision feedback equalizers are both used following square law detection, the phase information of the received optical signal is not preserved thus leading to limited performance.

![Decision feedback equalizer](image)

Figure 1.2: Decision feedback equalizer.

Maximum likelihood sequence estimation (MLSE) is an electronic digital signal processing technique which is implemented following an ADC. Different from the FFE and DFE, MLSE determines the most likely bit sequence that was transmitted by comparing the received waveform with all possible transmitted bit sequences. The
MLSE equalizer is usually implemented with a $N$ state Viterbi detector, where $N = 2^M - 1$ and $M$ is the sequence length being considered. The performance using 16 state MLSE and 4 bit ADC in a chromatic dispersion limited 10 Gb/s standard single-mode fiber (SSMF) transmission system has been reported [15]. The transmission of a non-return-to-zero on-off-keying (NRZ-OOK) signal over 210 km of uncompensated SSMF was achieved with an optical signal-to-noise ratio (OSNR) penalty of only 2.6 dB at BER of $10^{-3}$. When the number of states is increased to 4096 and the ADC resolution is 5 bits, a 10 Gb/s NRZ-OOK signal was transmitted over 1040 km of uncompensated SSMF [16]. However, the drawback of the MLSE equalizer is that the computational complexity increases exponentially with the number of states. Moreover, the MLSE equalizer becomes impractical for a very large number of states even for offline signal processing and is therefore impractical for implementation [17].

Today, the MLSE equalizer in 10 Gb/s applications is limited to 16 states. In 2008, a 16 state MLSE digital equalizer chip set was accomplished in which the 3 bit ADC was implemented in a BiCMOS chip and the digital Viterbi detector was achieved in a CMOS chip [18]. A 4 state MLSE at 42.7 Gb/s would be feasible in today’s integrated circuit technology [19].

### 1.3 ESP in Transmitter

ESP can also be implemented at the transmitter to pre-compensate for chromatic dispersion. The basis of the dispersion pre-compensation technique is to generate a pre-compensated optical signal with controlled amplitude and phase at the transmitter. During transmission, the fiber chromatic dispersion reverses the effect of the pre-compensation so that a desired optical signal is received. For example, let $E(t)$,
with a Fourier transform of $E(f)$, be the desired optical signal at the receiver. Chromatic dispersion is modeled by the frequency response $H(f)$. The pre-compensated optical signal at the transmitter is therefore calculated by

$$E_{tr}(t) = \mathcal{F}^{-1}(E(f)H^{-1}(f))$$  \hspace{1cm} (1.2)

After passing through the fiber, the received optical signal $E'(t)$ has a spectrum $E'(f)$ given by

$$E'(f) = E(f)H^{-1}(f) \cdot H(f) = E(f)$$  \hspace{1cm} (1.3)

Therefore, the received optical signal $E'(t)$ corresponds to the desired optical signal $E(t)$. The pre-compensation concept is illustrated in Figure 1.3.

![Figure 1.3: Electronic dispersion pre-compensation concept.](image)

The input data is filtered by two linear finite impulse response (FIR) filters implemented in a digital signal processing (DSP) circuit. The FIR filter tap coefficients are determined by the in-phase and quadrature part of the pre-compensated optical signal. After the DACs and amplifiers, the analog voltages drive the modulator to generate the required pre-compensated optical signal $E_{tr}(t)$ which is launched into the fiber. After fiber transmission, the desired optical signal $E(t)$ is obtained at the receiver.
In 2005, the first proof-of-principle implementation using pulse pattern generators, high-speed data multiplexers, wide-band RF combiners, phase shifters and attenuators was demonstrated [20]. It allowed for transmitting a return-to-zero differential-phase-shift-keying (RZ-DPSK) signal through 5120 km of SSMF with 82.4 ns/nm uncompensated chromatic dispersion. In the same year, an ASIC with 6 bit DACs enabled the transmission of an NRZ-DPSK signal over 3840 km of SSMF with complete electronic dispersion compensation (EDC) at 10 Gb/s [21]. In 2010, electronic pre-compensation of chromatic dispersion for 43 Gb/s NRZ differential-quadrature-phase-shift-keying (NRZ-DQPSK) signal transmission over 135 km SSMF was demonstrated employing a test chip with 6 bit DACs [22].

In addition to chromatic dispersion, the pre-compensation technique is also able to deal with self-phase modulation (SPM). A technique of using a dual-drive Mach-Zehnder modulator (DD-MZM) driven by a look-up table based nonlinear digital filter implemented in a DSP circuit was proposed in 2005 [23]. A drawback of this method is the memory size scales exponentially with the amount of dispersion. A promising approach to simultaneously compensate for both chromatic dispersion and SPM is to combine linear FIR filters and nonlinear filters [24]. The linear FIR filters carry out chromatic dispersion compensation and the nonlinear filters deal with SPM. It is demonstrated that, at 10 Gb/s, electronic nonlinear pre-compensation can mitigate the SPM degradation on NRZ-DPSK signals across 320 km and 1280 km with average fiber launch powers of 9 dBm and 2.6 dBm, respectively.

However, the pre-compensation technique is not suitable for PMD which is a fast varying signal distortion since this would require adaptive feedback from the receiver to the transmitter. The update frequency has to be less than the reciprocal of the round transmission time and this limits the capacity of PMD pre-compensation.
1.4 ESP in Digital Coherent Receiver

Coherent receivers extract the amplitude, phase and polarization information of the signal by mixing the received optical signal with a local oscillator (LO) signal. In the 1980s, the coherent detection technique was investigated extensively [25], [26] since it provides several important advantages compared to direct detection receivers [27]. Firstly, the ability to detect multilevel modulation formats such as quadrature-phase-shift-keying (QPSK) allows a more efficient use of the fiber bandwidth by increasing the bit rate for the same symbol rate. Secondly, since the LO signal has a much larger power relative to the received signal power, shot-noise limited receiver sensitivity can be approached.

However, the development of coherent detection techniques was not pursued further until recently. There are two main reasons which contributed to this. In the case of homodyne detection, the phase and frequency of the LO signal has to be matched exactly to the incoming signal. This can be realized by an optical phase-locked loop (PLL) which is complicated in practical use. Comparatively, heterodyne detection requires a simpler receiver without the need for phase and frequency matching, but it results in a 3 dB signal-to-noise ratio (SNR) penalty [28]. Another important reason was the advent of the erbium doped fiber amplifier (EDFA) which makes the shot-noise limited coherent receiver sensitivity less significant since the SNR of the signal in an amplified transmission link is determined by the amplified spontaneous emission (ASE) of EDFAs. Moreover, the deployment of EDFAs in wavelength division multiplexed (WDM) systems revolutionized the increase of transmission capacity in the early 1990s and thus changed in the direction of optical communications.
As the transmission capacity increased in WDM systems, coherent detection began to attract substantial interest during recent years. One reason is the demand for multi-level modulation formats which carry more than one bit per symbol. Therefore the bit rate is increased while keeping the same symbol rate or the spectral width is reduced while maintaining the bit rate. Any multi-level modulation format can be demodulated by coherent detection which allows complete access to the amplitude and phase of the optical field. Another reason is due to the availability of high-speed digital circuits which enable the signal processing to be performed in a DSP circuit. Real-time signal processing is necessary in a commercial product since the payload data in a high-speed optical signal transmission system is always continuous in real-time. However, the real-time hardware implementation is limited by the availability of fast enough circuit technology to achieve the ADC having high resolution and high sampling rate. Moreover, the DSP unit with high integrability and lower power requirement is required. A milestone in the implementation of a digital coherent receiver was the real-time operation with a 90 nm CMOS ASIC at 40 Gb/s accomplished by Nortel Networks (now Ciena Corporation) in 2008 [5]. A year later, Ciena Corporation announced the 100 Gb/s ASIC PM-QPSK architecture by dual sub-carrier implementation where each carrier delivers 50 Gb/s and single carrier implementation [29]. In 2011, AT&T labs published their real-time single-carrier coherent 100 Gb/s PM-QPSK transmission system tested in a field environment [30].

In contrast to a real-time implementation, offline DSP allows for the development, verification and optimization of signal processing algorithms. With the same set of data, a comparison of different DSP algorithms can be made. Experiments utilizing coherent detection and offline DSP allow a full understanding of digital coherent transmission systems before hardware implementation. Therefore, the offline
DSP is especially important in university research environments. Figure 1.4 shows a polarization and phase diversity digital coherent receiver front-end with subsequent offline DSP. The input signal and a LO signal are first split into two polarizations by two polarization beam splitters (PBSs) and combined in 90° hybrids followed by four pairs of balanced detectors. The four output signals are sampled using ADCs with a certain sampling rate and resolution and then offline processed in Matlab. The offline DSP includes chromatic dispersion compensation, clock recovery, polarization recovery, carrier frequency and phase estimation, symbol recovery of the transmitted bit sequences and bit-error-ratio (BER) counting.

![Figure 1.4: Digital coherent receiver front-end with subsequent offline DSP.](image)

Recent achievements of digital coherent detection with offline DSP includes transmission of 32 Tb/s PM 8-ary quadrature amplitude modulation (PM-8QAM) over 580 km [31], 112 Gb/s PM-16QAM [32], 10×224 Gb/s WDM transmission of PM-16QAM over 1200 km [33], and 69.1 Tb/s C- and extended L-band transmission of PM-16QAM over 240 km [34].


1.5 Comparison of ESP Techniques

Several ESP technologies have been briefly outlined in the previous sections. In Table 1.2, a comparison is made for the benefits and shortcomings that each ESP technology provides.

Table 1.2: Comparison of ESP techniques.

<table>
<thead>
<tr>
<th></th>
<th>Benefits</th>
<th>Shortcomings</th>
</tr>
</thead>
<tbody>
<tr>
<td>FFE &amp; DFE in direct-detection receiver</td>
<td>Compensate for chromatic dispersion and PMD; low power consumption</td>
<td>Limited performance due to loss of phase information after direct detection</td>
</tr>
<tr>
<td>MLSE in direct-detection receiver</td>
<td>Compensate for chromatic dispersion, PMD and SPM</td>
<td>Computational complexity</td>
</tr>
<tr>
<td>ESP in transmitter</td>
<td>Compensate for chromatic dispersion and SPM</td>
<td>Can not compensate PMD</td>
</tr>
<tr>
<td>ESP in digital coherent receiver</td>
<td>Compensate for chromatic dispersion and PMD; achieve polarization alignment and allows multilevel signal transmission to increase capacity</td>
<td>Power consumption and heat dissipation</td>
</tr>
</tbody>
</table>

1.6 Techniques for Overcoming Narrowband Optical Filtering and Chromatic Dispersion

Optical fibers are widely utilized as the communication channel in lightwave systems due to several reasons. The first is optical fibers have low loss, as small as 0.16 dB/km at around 1550 nm. The second reason is the large usable bandwidth. The transmission capacity of a system is proportional to the carrier frequency of transmitted signals. The optical carrier frequency is in the range of $10^{14}$ Hz which enables optical
fibers to have enormous transmission bandwidth and support high data rates. Moreover, their small size, light weight, low cost and reliability make optical fibers widely and effectively used. However, in practice signal propagation through an optical fiber is limited by various factors including:

- Fiber loss limits the maximum transmission distance by reducing the signal power reaching the receiver. While EDFAs can compensate for the fiber loss, the induced ASE noise will accumulate and eventually degrade the signal quality.

- Narrowband optical filtering degrades the signal quality due to the reduced effective bandwidth caused by the concatenation of reconfigurable optical add-drop multiplexers (ROADMs) in the transmission link.

- Fiber chromatic dispersion and PMD limit the performance by broadening optical pulses as they propagate along the fiber.

- Nonlinear effects limit the transmission distance by inducing signal distortion.

In this thesis, only narrowband optical filtering and chromatic dispersion are considered. Both of them can be compensated either optically or electronically.

1.6.1 Narrowband Optical Filtering and Compensation Techniques

Due to the use of ROADM with flexible add-drop or routing capabilities in WDM networks, the performance of such networks can be affected by the overall frequency response of the cascaded optical filters used for multiplexing and demultiplexing. Since the overall frequency response is the product of the frequency responses for the individual filters in the link between the transmitter and receiver, the effective
bandwidth is reduced as the number of cascaded filters is increased, as shown in Figure 1.5. Therefore, the reduction in the effective bandwidth has to be considered to ensure satisfactory system performance.

![Figure 1.5: Reduction of the effective filter bandwidth.](image)

The narrowband optical filtering was addressed optically by using a semiconductor optical amplifier based 3R regenerator at 40 Gb/s for 100 GHz channel spacing with a carrier suppressed return-to-zero signal and for 200 GHz channel spacing with a RZ signal [35]. It was also mitigated by using MLSE at 10 Gb/s [36], [37].

### 1.6.2 Chromatic Dispersion and Compensation Techniques

Chromatic dispersion in an optical fiber leads to the pulse broadening since different frequencies traveling at different velocities along the fiber. It results in ISI and therefore the transmission distance is restricted without dispersion compensation. Chromatic dispersion consists of material dispersion and waveguide dispersion. Material dispersion is due to the frequency dependence of a fiber’s refractive index. Waveguide dispersion is because the propagation constant is nonlinearly dependent on frequency.

The impact of the chromatic dispersion is described by the frequency dependent
propagation constant \( \beta(\omega) \). It can be approximated using a Taylor series expansion up to third order around the carrier frequency \( \omega_0 \) as

\[
\beta(\omega) = \beta_0 + \beta_1(\omega - \omega_0) + \frac{\beta_2}{2}(\omega - \omega_0)^2 + \frac{\beta_3}{6}(\omega - \omega_0)^3 \tag{1.4}
\]

where \( \beta_1 \) is the inverse group velocity defined as \( \beta_1 = 1/v_g \) and \( \beta_n \) is the \( n^{th} \) derivative of \( \beta(\omega) \) with respect to angular frequency,

\[
\beta_n = \frac{d^n\beta(\omega)}{d\omega^n}igr|_{\omega=\omega_0} \tag{1.5}
\]

The second order term \( \beta_2 \) is the group velocity dispersion (GVD) coefficient in \([\text{ps}^2/\text{km}]\). The third order term \( \beta_3 \) refers to the GVD slope in \([\text{ps}^3/\text{km}]\). \( \beta_2 \) and \( \beta_3 \) are related to the dispersion \( (D) \) and dispersion slope \( (S) \) which are defined theoretically as

\[
D = -\frac{2\pi c}{\lambda^2} \beta_2 \tag{1.6}
\]

\[
S = \frac{4\pi c}{\lambda^3} \left( \beta_2 + \frac{\pi c}{\lambda} \beta_3 \right) \tag{1.7}
\]

where \( \lambda \) is the carrier wavelength. \( D \) is often expressed in units of \([\text{ps/\text{nm/km}}]\) and \( S \) in units of \([\text{ps/\text{nm}^2/\text{km}}]\) [28].

There are several optical compensation methods to deal with chromatic dispersion. For example, dispersion compensating fiber (DCF) [38], [39] is currently widely used in commercial systems but it is expensive with large size, high loss and lack of adaptability. Fiber Bragg gratings [40], Gires-Tournois etalons [41] and optical finite impulse response filters [42] have also been implemented. However, they provide limited tunable dispersion compensation and/or high insertion loss. Different
modulation formats such as duobinary or DQPSK are used due to their narrower signal spectrum, thus leading to a higher tolerance to chromatic dispersion. Optical phase conjugation is an alternative solution [43], [44]. Although these are effective, the increased cost and complexity can not be ignored.

In addition to optical dispersion compensation techniques, chromatic dispersion can also be coped with using EDC techniques which become increasingly attractive due to their low cost and adaptability. As mentioned in previous sections, FFE, DFE, MLSE, dispersion pre-compensation and digital filters in coherent receivers [10], [45] are effective methods to compensate for chromatic dispersion.

1.7 Thesis Organization

This thesis focuses on utilizing the electronic pre-compensation technique to deal with the signal distortion induced by narrowband optical filtering and chromatic dispersion. In links employing direct detection receiver, the advantage of carrying out the compensation at the transmitter is because both the amplitude and phase of the signal can be fully controlled. Thus the phase loss during the direct detection process is avoided. Moreover, the pre-compensation technique provides a possibility to combine the pre- and post-compensation to further improve the compensation effectiveness.

Chapter 2 describes the arbitrary optical waveform generator (AOWG) including its structure, operation principle as well as two commonly used LiNbO$_3$ MZMs. One is the DD-MZM and the other is the dual-parallel Mach-Zehnder modulator (DP-MZM). The difference between these two modulators and the back-calculation equation for each modulator are presented. Three different modulation formats including NRZ-OOK, NRZ-DPSK and NRZ-DQPSK will be used in following chapters. Therefore,
the signal generation and detection methods for each modulation format are reviewed.

Chapter 3 presents simulation and experimental results using the AOWG to pre-compensate for narrowband optical filtering for 10 Gb/s NRZ-OOK and NRZ-DPSK as well as 20 Gb/s NRZ-DQPSK modulation formats. Two types of LiNbO$_3$ MZMs (DD-MZM and DP-MZM) are considered. The effectiveness of pre-compensation in terms of eye closure penalty (ECP) for different overall filter bandwidths is investigated by simulation. An experiment is performed in which a single narrow filter with a 3 dB bandwidth of 9.13 GHz is used to emulate the resultant narrowband optical filtering caused by the concatenation of ROADMs. Moreover, recirculating loop experiments in which an optical filter with a 3 dB bandwidth of 28 GHz is used to represent the ROADM optical filter are implemented.

Chapter 4 investigates electronic dispersion pre-compensation using a semiconductor InP MZM and the NRZ-OOK modulation format at 10.709 Gb/s. Different from widely used LiNbO$_3$ MZMs, the InP MZM has nonlinear voltage dependent attenuation and phase constants. This requires a distinct method to obtain the required drive voltages. For the purpose of pre-compensation comparison, a recirculating loop experiment with non-zero dispersion shifted fiber (NZDSF) and a straight-line experiment with SSMF spans are performed using both an InP MZM and a LiNbO$_3$ MZM.

Chapter 5 focuses on the electronic dispersion pre-compensation for 85.672 Gb/s PM-16QAM with coherent detection and offline DSP. In order to improve the quality of the signal generation, especially for 16QAM, characterization of the AOWG is performed. The effect of modulator bias voltage error on the system performance is investigated by simulation in terms of BER and error vector magnitude (EVM).
A recirculating loop experiment is done to investigate the pre-compensation performance which is then compared with dispersion post-compensation. In addition, the optimal signal laser power for dispersion pre-compensation and post-compensation is discussed.

Chapter 6 concludes the thesis contributions and proposes future work.
Chapter 2

Background Review

2.1 Arbitrary Optical Waveform Generator

The AOWG is a powerful tool which can generate complex optical signals used for electronic pre-compensation. A simplified schematic of the AOWG is shown in Figure 2.1 [21]. It consists of an ASIC and amplifiers as well as a continuous wave (CW) laser and an optical modulator. The ASIC allows for the generation of arbitrary electrical signals with a sampling rate of 21.418 GSa/s and 6 bit DAC resolution per sample. This ASIC is provided by Ciena Corporation.

Figure 2.1: Arbitrary optical waveform generator.
In order to generate a desired optical signal with specified amplitude and phase at the output of modulator, two required electrical drive voltages are back-calculated based on an inverse modulator transfer function. After considering the bandwidth limitations of the DACs, amplifiers and modulator, the voltage waveforms are sampled, quantized and loaded into external DSP memory. This procedure is depicted in Figure 2.2. The digital signals stored in the external DSP memory are converted to analog signals by DACs to drive the modulator through individual amplifiers.

Figure 2.2: Procedure of backcalculation using AOWG.
The block diagram shown in Figure 2.3 describes how the AOWG works in a practical implementation [46]. The input data at 10.709 Gb/s are upsampled to 21.418 GSa/s with two samples per symbol and then applied to a linear filter to compensate for linear transmission impairments, such as chromatic dispersion. The nonlinear filter is responsible for dealing with nonlinear SPM effects. Adaptive control can be realized through a communication path with receiver feedback which provides the necessary information, such as BER or Q-factor to the DSP controller. Provided that the transmission impairments are not fast-varying effects, which is generally true for chromatic dispersion and optical filtering, the filter settings do not require rapid adaptation and the feedback delay does not affect the system performance.

![Block diagram of AOWG transmitter in practical implementation.](image)

Except for dispersion and nonlinear SPM pre-compensation applications, the AOWG transmitter can also be utilized as a signal generator according to different requirements, such as synthesizing an optical side-band modulation for fiber to the premises [47], generating single side-band signals for measuring the amplitude and phase responses of passive optical components [48], sub-picosecond optical pulse generation [49], and generating an optical signal with known sinusoidal jitter to measure the jitter transfer function of a distributed feedback (DFB) self-pulsing laser [50].
2.2 Mach-Zehnder Modulators

A LiNbO$_3$ MZM works based on the electro-optic effect [28]. Variation of the applied drive voltages cause the refractive index of the waveguide in the modulator arms to change. Without applying drive voltages, the optical fields experience identical phase shifts in the two arms leading to constructive interference. When electrical drive voltages are applied, the voltage-induced index change will cause different phase shifts in the two arms. This destroys the constructive interference and the transmitted light intensity is reduced. If the phase difference between the signals in the two arms equals $\pi$, destructive interference occurs and no light is transmitted. Two modulators widely used in the AOWG are the DD-MZM and DP-MZM.

2.2.1 Dual-Drive Mach-Zehnder Modulator

A schematic of a DD-MZM is shown in Figure 2.4. The light from a laser diode $E_{in}(t)$ is divided by a 3 dB coupler into two equal parts that propagate in the lower and upper arms of the DD-MZM. Two electrical drive voltages controlled independently are applied to the modulator.

![Figure 2.4: Dual-drive Mach-Zehnder modulator.](image-url)
The relationship between optical input and output is defined as [46]

\[
E_{\text{out}}(t) = \frac{E_{\text{in}}(t)}{2} \left[ \exp \left( j \frac{\pi}{V_{\pi}} (v_{r_f,1}(t) + V_{dc,1}) \right) + \exp \left( j \frac{\pi}{V_{\pi}} (v_{r_f,2}(t) + V_{dc,2}) \right) \right] \quad (2.1)
\]

where \( V_{\pi} \) is the parameter to indicate the drive voltage required to change the output light intensity from its maximum to a minimum level. \( v_{r_f,1}(t) \) and \( v_{r_f,2}(t) \) are the electrical drive voltages applied to two arms. \( V_{dc,1} \) and \( V_{dc,2} \) are the corresponding direct current (DC) bias voltages. When the modulator is operated in a differential mode in which \( V_{dc,1} \) is the reverse of \( V_{dc,2} \) for any bias voltage value, Equation (2.1) reduces to

\[
E_{\text{out}}(t) = E_{\text{in}}(t) \cos \left[ \frac{\pi}{2V_{\pi}} (v_{r_f,1}(t) - v_{r_f,2}(t) + 2V_{dc,1}) \right] \exp \left[ j \frac{\pi}{2V_{\pi}} (v_{r_f,1}(t) + v_{r_f,2}(t)) \right] \quad (2.2)
\]

The input optical signal is from a CW laser, therefore it can be regarded as a constant level (i.e., \( E_{\text{in}}(t) = \text{constant} \)). Here, let \( E_{\text{in}}(t) = 1 \) to simplify the calculation. The complex envelope of the optical signal generated at the output of the modulator is represented as

\[
E_{\text{out}}(t) = |E_{\text{out}}(t)| \exp \left[ j \angle E_{\text{out}}(t) \right] \quad (2.3)
\]

With Equations (2.2) and (2.3), we thus have

\[
|E_{\text{out}}(t)| = \cos \left[ \frac{\pi}{2V_{\pi}} (v_{r_f,1}(t) - v_{r_f,2}(t) + 2V_{dc,1}) \right] \quad (2.4)
\]

\[
\angle E_{\text{out}}(t) = \frac{\pi}{2V_{\pi}} (v_{r_f,1}(t) + v_{r_f,2}(t)) \quad (2.5)
\]
Solving Equation (2.4) and (2.5), the required electrical drive voltages can be obtained

\[ v_{rf,1}(t) = \frac{V\pi}{\pi} \left[ \angle E_{out}(t) + \arccos(|E_{out}(t)|) \right] - V_{dc} \] (2.6)

\[ v_{rf,2}(t) = \frac{V\pi}{\pi} \left[ \angle E_{out}(t) + \arccos(|E_{out}(t)|) \right] + V_{dc} \] (2.7)

where \( V_{dc} = V_{dc,1} = -V_{dc,2} \) represent the differential DC bias voltages applied to the DD-MZM.

### 2.2.2 Access Field Limitations of DD-MZM

According to Equation (2.4), without applying electrical drive voltages, the output power is given by

\[ P_{out} = \cos^2 \left( \frac{\pi}{V\pi} V_{dc} \right) \] (2.8)

The relationship of the output power and DC bias voltage is plotted in Figure 2.5. Along the curve, points A, B and C refer to different bias voltages of 0, \( V/3 \) and \( V/2 \), respectively.

With infinite drive voltages \( v_{rf,1}(t) \) and \( v_{rf,2}(t) \), the modulator is able to access the entire complex plane with desired magnitude and phase. However, in practice, the drive voltages are finite and bandwidth limited. The maximum peak-to-peak voltage \( (V_{pp}) \) swing and bias voltage confine the area in the complex plane which can be accurately reproduced by the modulator. When the modulator is provided with drive voltages of \( V_\pi \) peak-to-peak on each of two arms, the accessible area on the complex plane is determined by the bias voltage, as illustrated in Figure 2.6 [46]. When the bias voltage is set at 0, \( V/3 \) and \( V/2 \), the accessible area corresponds to Figure 2.6 (a), (b) and (c), respectively.
CHAPTER 2. BACKGROUND REVIEW

Figure 2.5: Power transfer function of a DD-MZM.

Figure 2.6: Accessible optical field of a DD-MZM at different bias voltages with $V_\pi$ peak-to-peak swing.
2.2.3 Dual-Parallel Mach-Zehnder Modulator

Figure 2.7 shows a schematic of a DP-MZM. Two inner MZMs are embedded in a main Mach-Zehnder interferometer (MZI). Following the output of the lower MZM is a 90° phase shifter. Due to the consideration of simplicity, the two inner MZMs are both single-drive instead of dual-drive modulators. The DP-MZM allows accessing the entire complex plane when the electrical drive voltages are limited to have a $V_{pp}$ of $V_r$. This provides considerable benefits for arbitrary optical waveform synthesis.

![Figure 2.7: Dual-parallel Mach-Zehnder modulator.](image)

The output signal from a DP-MZM can be represented as [51], [52]

$$E_{out}(t) = \frac{E_{in}(t)}{2} \left[ \frac{1}{2} \left( e^{j \frac{\pi}{V_r} (v_{rf,1}(t)+V_{dc,1})} + e^{-j \frac{\pi}{V_r} (v_{rf,1}(t)+V_{dc,1})} \right) + ... ight]$$

$$= \frac{E_{in}(t)}{2} \left[ \cos \left( \frac{\pi}{V_r} (v_{rf,1}(t) + V_{dc,1}) \right) + ... \right]$$

$$\cos \left( \frac{\pi}{V_r} (v_{rf,2}(t) + V_{dc,2}) \right) e^{j \frac{\pi}{V_r} V_{dc,3}} \right] \quad (2.9)$$

where $v_{rf,1}(t)$ and $V_{dc,1}$ are the electrical drive voltage and DC bias voltage which are
applied to the upper inner MZM. \( v_{rf,2}(t) \) and \( V_{dc,2} \) are the electrical drive voltage and DC bias voltage which go to the lower inner MZM. \( V_{dc,3} \), the DC bias voltage applied to the 90° phase shifter, is equal to \( V_\pi/2 \). The input optical signal of the DP-MZM is from a CW laser. Let \( E_{in}(t) = 2 \) for simplification. Then Equation (2.9) becomes

\[
E_{out}(t) = \cos \left( \frac{\pi}{V_\pi} \left( v_{rf,1}(t) + V_{dc,1} \right) \right) + j \cos \left( \frac{\pi}{V_\pi} \left( v_{rf,2}(t) + V_{dc,2} \right) \right) \tag{2.10}
\]

The required electrical drive voltages are calculated as follows:

\[
v_{rf,1}(t) = \frac{V_\pi}{\pi} \arccos \left[ |E_{out}(t)| \cos \left( \angle E_{out}(t) \right) \right] - V_{dc,1} \tag{2.11}
\]

\[
v_{rf,2}(t) = \frac{V_\pi}{\pi} \arccos \left[ |E_{out}(t)| \sin \left( \angle E_{out}(t) \right) \right] - V_{dc,2} \tag{2.12}
\]

When \( V_{dc,1} = V_{dc,2} = V_\pi/2 \), Equation (2.10) is simplified as

\[
E_{out}(t) = \sin \left( \frac{\pi}{V_\pi} v_{rf,1}(t) \right) + j \sin \left( \frac{\pi}{V_\pi} v_{rf,2}(t) \right) \tag{2.13}
\]

Inversion of Equation (2.13) yields the required electrical drive voltages

\[
v_{rf,1}(t) = \frac{V_\pi}{\pi} \arcsin \left[ |E_{out}(t)| \cos \left( \angle E_{out}(t) \right) \right] \tag{2.14}
\]

\[
v_{rf,2}(t) = \frac{V_\pi}{\pi} \arcsin \left[ |E_{out}(t)| \sin \left( \angle E_{out}(t) \right) \right] \tag{2.15}
\]

Assuming small electrical drive voltages, the output signal can be approximated as

\[
E_{out}(t) = \frac{\pi}{V_\pi} v_{rf,1}(t) + j \frac{\pi}{V_\pi} v_{rf,2}(t) \tag{2.16}
\]

The output signal becomes a linear combination of two electrical drive voltages. Therefore, driven with small signals, the DP-MZM operates in a linear region.
2.3 Optical Waveform Generation

The complex-valued electric field (baseband equivalent) of an optical signal can be written as

\[ E(t) = \sum_{n=-\infty}^{\infty} a_n d(t - nT) \]  

(2.17)

where the sequence \( \{a_n\} \) takes the value of \( \{0, 1\}, \{+1, -1\} \) and \( \{\pm 1 \pm j\} \) to represent the OOK, DPSK and DQPSK modulation formats, respectively. Here \( j = \sqrt{-1} \), \( T \) is the bit period and \( d(t) \) is the symbol pulse shape. Assuming \( E(t) \) is the desired optical signal to be received at the receiver, the required optical signal to be transmitted is calculated as follows

\[ E_{tr}(t) = F^{-1}\left(\frac{F(E(t))}{H(f)}\right) \]  

(2.18)

where \( F \) denotes the Fourier transform, and \( F^{-1} \) denotes the inverse Fourier transform. The transfer function of the pre-compensated channel effect \( H(f) \) can be known either analytically or from measurement. For example, the transfer function for fiber chromatic dispersion can be theoretically expressed as

\[ H_{\text{fiber}}(f) = \exp\left(-j \frac{DL\lambda^2 \pi f^2}{c}\right) \]  

(2.19)

\( \lambda \) is the carrier wavelength, \( c \) is the speed of light, \( D \) is the dispersion coefficient and \( L \) is the length of the fiber. The total accumulated dispersion can be represented by \( DL \). If pre-compensating for both chromatic dispersion and narrowband optical filtering simultaneously, the required transmitted optical signal is given by

\[ E_{tr}(t) = F^{-1}\left(\frac{F(E(t))}{H_{\text{fiber}}(f)H_{\text{filter}}(f)}\right) \]  

(2.20)

As seen from Equation (2.19), the chromatic dispersion only affects the signal...
phase, not the amplitude. The fiber phase responses at 10 km and 100 km are plotted in Figure 2.8. The influence of the phase response increases as the distance increases.

![Figure 2.8: Phase response of fiber dispersion.](image)

**2.3.1 Raised-Cosine Pulse Shape**

The pulse shape $d(t)$ in Equation (2.17) is chosen to be raised-cosine. There are two major advantages of using such a pulse shape. One is that this pulse shape is strictly band-limited and the bandwidth can be easily controlled through adjusting the roll-off factor. Another advantage is that there is no ISI between neighboring pulses due to zero crossings at sampling times other than the pulse center. An ideal raised-cosine pulse is defined as

$$d(t) = \frac{\sin\left(\frac{t}{T}\right) \cos\left(\frac{\pi \beta t}{T}\right)}{\frac{t}{T} - \frac{\beta t^2}{2}}$$  \hspace{1cm} (2.21)
where \( \beta \) is the roll-off factor which governs the bandwidth and rate at which the pulse decays as \( t \) approaches \( \pm \infty \).

The shape of the raised-cosine signal is shown in Figure 2.9. As can be seen, the ripple level in the time domain increases as \( \beta \) decreases. A value of \( \beta = 1 \) offers the fastest decay rate. The name “raised-cosine” refers to the pulse spectrum in the frequency domain, not its time domain shape. Its frequency domain description is given in Equation (2.22) and frequency spectra are shown in Figure 2.10 [53]

\[
H_{rc}(f) = \begin{cases} 
T & \text{for } 0 \leq |f| \leq \frac{1-\beta}{2T} \\
\frac{T}{2} \left[ 1 + \cos \left( \frac{\pi T}{\beta} (|f| - \frac{1-\beta}{2T}) \right) \right] & \text{for } \frac{1-\beta}{2T} \leq |f| \leq \frac{1+\beta}{2T} \\
0 & \text{for } \frac{1+\beta}{2T} \leq |f| 
\end{cases} \tag{2.22}
\]

Figure 2.9: Raised-cosine pulses with different roll-off factors in time domain.
2.3.2 NRZ-OOK Modulation Format

An NRZ-OOK signal is obtained by switching ON and OFF the amplitude of an optical carrier signal. It is a simple modulation format in terms of the structure of the transmitter and receiver. The most commonly used generation scheme for an NRZ-OOK signal is to externally modulate the laser signal using an intensity MZM, as depicted in Figure 2.11 (a). The modulator is biased at the quadrature point and the electrical drive voltage swings from minimum to maximum transmittance with a peak-to-peak value of $V_p$. Figure 2.11 (b) shows a measured optical eye diagram for a 10.709 Gb/s NRZ-OOK signal which is generated from AOWG with a raised-cosine pulse shape ($\beta = 0.7$). The constellation diagram of an optical NRZ-OOK signal is given in Figure 2.11 (c).

The binary NRZ-OOK signal can be detected using direct detection with a simple photodetector. An example of a pre-amplified optical NRZ-OOK signal direct detection receiver front-end is given in Figure 2.12.
Figure 2.11: Optical NRZ-OOK signal. (a) generation scheme, (b) measured 10.709 Gb/s optical eye diagram of a NRZ-OOK signal. $\beta = 0.7$ for the raised-cosine pulse shape; time base is 30 ps/div, (c) NRZ-OOK signal constellation diagram.

Figure 2.12: Direct detection receiver front-end for a pre-amplified optical NRZ-OOK signal. OBPF: optical band-pass filter.
2.3.3 NRZ-DPSK Modulation Format

The DPSK signal format is a phase-shift-keying (PSK) format which carries the information by modulating the optical phase. The information is differentially encoded by using the phase difference between two neighboring bits. For example, no phase change results in a bit “1” and phase change of $\pi$ results in a bit “0”. A DPSK differential encoder is used at the transmitter, as follows:

$$b_k = a_k \oplus b_{k-1}$$  \hspace{1cm} (2.23)

where $a_k \in \{0, 1\}$ are the original binary bits, $b_k \in \{0, 1\}$ are the encoded binary bits, and $\oplus$ is the logic XOR. The encoded $b_k$ is then converted into a bipolar form $c_k \in \{-1, 1\}$ for the electrical drive voltage.

The most common generation scheme for an NRZ-DPSK signal is to use an external MZM which is biased at null and the electrical drive voltage has a peak-to-peak value of $2V_\pi$, as illustrated in Figure 2.13 (a). This modulator therefore switches between two peaks with a phase jump of $\pi$. Ideally, the NRZ-DPSK signal should only change the phase while keeping the amplitude constant. However, intensity dips are observed in the measured 10.709 Gb/s optical NRZ-DPSK eye diagram which is synthesized from the AOWG with a raised-cosine pulse shape ($\beta = 1$), as shown in Figure 2.13 (b). The non-constant intensity is due to the fact that the signal intensity returns to zero when the phase switches between 0 and $\pi$. The width of intensity dips depends on the rise and fall times of electrical drive voltages. The constellation diagram of an optical NRZ-DPSK signal is illustrated in Figure 2.13 (c). Compared to the NRZ-OOK signal, the most important benefit of the NRZ-DPSK signal is the 3 dB lower OSNR requirement to reach a given BER [54]. Moreover, the peak
power is 3 dB lower for NRZ-DPSK than for OOK given the same average optical power and the optical power is more evenly distributed in every bit slot. These make NRZ-DPSK more tolerant to nonlinear effects [54].

Figure 2.13: Optical NRZ-DPSK signal. (a) generation scheme, (b) measured 10.709 Gb/s optical eye diagram of a NRZ-DPSK signal. \( \beta = 1 \) for the raised-cosine pulse shape; time base is 30 ps/div, (c) NRZ-DPSK signal constellation diagram.

A differential detection receiver front-end of a pre-amplified optical NRZ-DPSK signal is shown in Figure 2.14. The differential phase modulation is converted into

Figure 2.14: Differential detection receiver front-end for a pre-amplified optical NRZ-DPSK signal.
amplitude modulation by a Mach-Zehnder delay line interferometer (MZDI) with a one bit delay in one arm. Given the input optical field to the MZDI $E(t)$, the two output fields at the second coupler are:

\begin{align*}
E_1(t) &= E(t - T) - E(t) \\
E_2(t) &= jE(t - T) + jE(t)
\end{align*}

$E_1(t)$ is the destructive port at which destructive interference occurs when there is no phase change between two consecutive bits and constructive interference occurs when the phase of between subsequent bits shifts by $\pi$. $E_2(t)$ is the constructive port which yields inverted pattern. Either $E_1(t)$ or $E_2(t)$ allows full detection of an NRZ-DPSK signal. However, the 3 dB lower OSNR improvement is only available using balanced detection from which the photocurrent is obtained:

\[ I_{\text{bal}} = |E_1(t)|^2 - |E_2(t)|^2 \]  

The resultant photocurrent only has two levels without a zero rail. The high and low level photocurrent corresponds to 1’s and 0’s in the original binary bits.

### 2.3.4 NRZ-DQPSK Modulation Format

DQPSK is a four level phase modulation format with the information encoded on four phases ($\pi/4, 3\pi/4, 5\pi/4, 7\pi/4$). Each symbol represents two bits. Therefore, for the same bit rate, the spectral occupancy is reduced by half compared to OOK and DPSK and results in an increased tolerance to chromatic dispersion and PMD. Alternatively, the bit rate can be doubled for the same spectral occupancy as OOK.
and DPSK. The DQPSK encoder is given as follows [55]:

\[
I_k = U_k I_{k-1} - Q_{k-1} + U_k I_{k-1} Q_{k-1} + V_k I_{k-1} Q_{k-1} \quad (2.27)
\]

\[
Q_k = U_k I_{k-1} Q_{k-1} + U_k I_{k-1} Q_{k-1} + V_k I_{k-1} Q_{k-1} + V_k I_{k-1} Q_{k-1} \quad (2.28)
\]

where \( U_k \) and \( V_k \) ∈ \{0, 1\} are the original transmitted binary bits for the in-phase and quadrature components. \( I_k, Q_k \in \{0, 1\} \) are encoded binary bits which are then converted into a bipolar form taking values of \{-1, 1\}. The DQPSK symbol is represented by \(\{I_k + jQ_k\}\).

In order to generate an optical NRZ-DQPSK signal, a DP-MZM is used as shown in Figure 2.15 (a). The two MZMs are both, in effect, NRZ-DPSK modulated with independent electrical drive voltages. The NRZ-DPSK signal generated in the upper arm is combined with a 90° phase shifted NRZ-DPSK signal in the lower arm. A measured 21.418 Gb/s optical NRZ-DQPSK optical eye diagram with a raised-cosine pulse shape (\(\beta = 1\)) is given in Figure 2.15 (b) from which double intensity dips are observed. Figure 2.15 (c) shows the constellation diagram.

A NRZ-DQPSK signal can be decoded using two separate MZDIs to convert phase to amplitude information, as depicted in Figure 2.16 [54]. The differential phase between the interferometer arms is set to \(\pi/4\) and \(-\pi/4\) for each of two transmitted sequences. Given the optical signal at the input to MZDI \(E(t)\), the output fields at the second coupler for the U channel are:

\[
E_1(t) = E(t - T) - E(t)e^{j\frac{\pi}{4}} \quad (2.29)
\]

\[
E_2(t) = jE(t - T) + jE(t)e^{j\frac{\pi}{4}} \quad (2.30)
\]
Figure 2.15: Optical NRZ-DQPSK signal. (a) generation scheme, (b) measured 21.418 Gb/s optical eye diagram of an NRZ-DQPSK signal. $\beta = 1$ for the raised-cosine pulse shape; time base is 30 ps/div, (c) NRZ-DQPSK signal constellation diagram.

Figure 2.16: An optical differential detection pre-amplified receiver front-end for an NRZ-DQPSK signal.
The photocurrent after the balanced detection is

\[
I_U = |E_1(t)|^2 - |E_2(t)|^2 \quad (2.31)
\]

\(I_U\) has only two levels. High and low levels corresponds to 1’s and 0’s in the original binary transmitted bit for in-phase component \(U_k\). Similarly,

\[
E_3(t) = E(t - T) - E(t)e^{-j\frac{\pi}{4}} \quad (2.32)
\]

\[
E_4(t) = jE(t - T) + jE(t)e^{-j\frac{\pi}{4}} \quad (2.33)
\]

The photocurrent after the balanced detection for the V channel is

\[
I_V = |E_3(t)|^2 - |E_4(t)|^2 \quad (2.34)
\]

High and low levels of \(I_V\) corresponds to 1’s and 0’s in the original binary transmitted bit for quadrature component \(V_k\).

## 2.4 Summary

This chapter described the generation of an optical signal with specified amplitude and phase using an AOWG and the back-calculation procedure using two widely used \(\text{LiNbO}_3\) MZMs, DD-MZM and DP-MZM. The signal generation and detection methods for three different modulation formats including NRZ-OOK, NRZ-DPSK and NRZ-DQPSK have been briefly reviewed.
Chapter 3

Electronic Pre-Compensation of Narrowband Optical Filtering

3.1 Introduction

The high capacity WDM optical network is recognized as an attractive technology to increase system transmission capacity. A typical network with add-drop or routing capabilities includes a significant number of ROADMs. An optical signal will pass through many such network elements before reaching its destination at the receiver. In such networks, performance degradation occurs primarily due to the narrowband optical filtering which can be traced to the optical filters used to (de)multiplex the optical channels. The shape of the optical filter response determines its performance characteristics. The realistic transfer function of the optical filter deviates from a brick-wall shape and the effective spectral transfer function of cascaded filters is the multiplication of the transfer functions for the individual filters. Therefore, the effective filter bandwidth gets narrower with an increasing number of filters that the
signal traverses. The decrease of filter bandwidth as a function of number of cascaded filters is plotted in Figure 3.1. If a second-order gaussian filter is used, the 3 dB bandwidths of 1, 20 and 140 cascaded filters are 30 GHz, 14.5 GHz and 8.9 GHz, respectively assuming all filters are central wavelength aligned. However, if a third-order gaussian filter with a flattened passband is used, the 3 dB bandwidths of 1, 20 and 140 cascaded filters are 30 GHz, 18.2 GHz and 13.1 GHz, respectively. Therefore, the third-order gaussian filter is less susceptible to concatenation due to its flat passband. However, when there are too many cascaded filters, the bandwidth narrowing will lead to distortion of the time domain signal causing the eye-closure penalty (ECP). Therefore, it is necessary to find a solution to overcome the resultant degradation in the signal quality.

Figure 3.1: Simulated dependence of filter 3 dB bandwidth on the number of cascaded optical filters.

The work presented in this chapter focuses on using a high-speed DSP technique to reduce the signal distortion induced by the narrowband optical filtering for three different modulation formats, NRZ-OOK, NRZ-DPSK and NRZ-DQPSK signals using a
DD-MZM and a DP-MZM as appropriate. For the simulation results, an indication of the impact that pre-compensation has on system performance is obtained in terms of the ECP for different overall filter bandwidths. For the experimental results, a single optical filter with a 3 dB bandwidth of 9.13 GHz is used to emulate the equivalent narrowband optical filtering caused by the concatenation of filters. The simultaneous pre-compensation for narrowband optical filtering and fiber chromatic dispersion is also considered. Moreover, recirculating loop experiments are performed in which an optical filter with a 3 dB bandwidth of 28 GHz is used.

Since all experiments presented in this chapter are performed at 10 Gb/s which does not include a 7% forward error correction (FEC) overhead, the pre-compensation is characterized in terms of the penalty in OSNR to achieve a BER of $10^{-9}$ which is standard performance criteria for optical fiber transmission systems. However, if the 7% is included to implement FEC, it leads to a bit rate of 10.709 Gb/s instead of 10 Gb/s. Then a signal having a BER of $10^{-3}$ is able to be corrected to a quasi error free (BER better than $10^{-12}$) [56].

3.2 Pre-Compensation for Single Narrow Filter

3.2.1 Optical Single Narrow Filter Characterization

An optical spectrum analyzer (OSA) which provides the filtered output signal was used as an optical filter with different bandwidths to represent the overall response of cascaded filters. Its filter response was characterized by using a Luna optical vector analyzer. When measuring the optical signal spectral density, the resolution bandwidth of the OSA defines the frequency resolution. A smaller resolution bandwidth means a more detailed characterization of the optical signal.
However, when the OSA is working in a filter mode, the resolution bandwidth determines the bandwidth of the optical filter. By adjusting the resolution bandwidth of the OSA to be 0.01 nm, 0.05 nm, 0.07 nm, 0.1 nm, 0.2 nm and 0.5 nm, an optical filter with a 3 dB bandwidth of 6.54 GHz, 7.93 GHz, 9.13 GHz, 15.95 GHz, 24.85 GHz and 61.79 GHz was obtained, respectively. The measured filter responses were used in the simulations in order to examine the signal distortion and pre-compensation performance as a function of the filter bandwidth. The normalized transfer function is shown in Figure 3.2 for the case of a 3 dB bandwidth of 9.13 GHz. Since the phase response is linear over the passband, the signal distortion induced by the filter is determined exclusively by the magnitude response.

![Figure 3.2: Magnitude and phase response of the optical filter with a 3 dB bandwidth of 9.13 GHz.](image-url)
3.2.2 Experimental Setup

The experimental setup is illustrated in Figure 3.3. The carrier wavelength was 1547.72 nm. An optical signal with a raised-cosine pulse shape was synthesized by the AOWG and then sent to the optical filter and 76 km of SSMF. The electrical drive voltages were determined for a $2^{12}$ deBruijn bit sequence. The bit sequence length can be extended up to $2^{14}$. A broadband noise source was used to vary the OSNR of the signal detected by a pre-amplified receiver. Different detectors were used appropriately according to the modulation formats. For the measurements of the BER, an optimum decision threshold level was used.

![Experimental setup for single narrowband optical filtering](image)

**Figure 3.3:** Experimental setup for single narrowband optical filtering. OBPF: optical band-pass filter.

Experimental results include pre-compensation of the narrowband optical filtering for the 10 Gb/s NRZ-OOK modulation format with DD-MZM and DP-MZM, pre-compensation of both the narrowband optical filtering and chromatic dispersion...
for the 10 Gb/s NRZ-DPSK modulation format with DP-MZM and 20 Gb/s NRZ-DQPSK modulation format with DP-MZM.

### 3.2.3 NRZ-OOK Modulation with DD-MZM and DP-MZM

For a DD-MZM, the signal is restricted to taking values in a portion of the complex plane for electrical drive voltages $V_{rf,1}$ and $V_{rf,2}$ with peak-to-peak values of $V_\pi$. For example, consider the waveform synthesis for the pre-compensation of a 10 Gb/s NRZ-OOK signal for a filter bandwidth of 9.13 GHz. In Figure 3.4 (a), the gray shaded region represents the accessible portion of the complex plane when the modulator is biased at $\pm V_\pi/3$. The black trace represents the trajectory of the ideal pre-compensated signal for narrowband optical filtering. Since the ideal pre-compensated signal extends outside the accessible region in the vicinity of the origin, it cannot be realized with a DD-MZM. This can be remedied by changing the extinction ratio of the optical field: $\{a_n\}$, which is used in Equation (2.17), takes values of 0.2 and 1 instead of 0 and 1. Figure 3.4 (b) illustrates that the pre-compensated signal now lies within the accessible region.

If the 3 dB bandwidth of the optical filter is reduced to 7.93 GHz, the data sequence $\{a_n\}$ has to take values of 0.3 and 1 in order to fit the pre-compensated signal into the accessible region. If the 3 dB bandwidth of the optical filter is further reduced to 6.54 GHz, the data sequence $\{a_n\}$ will take the values of 0.4 and 1. Therefore, the extinction ratio is sacrificed. Figure 3.5 shows how the extinction ratio [28]

$$r_{ex} = \frac{P_1}{P_o} \quad (3.1)$$

varies with filter bandwidth when a DD-MZM is used to synthesize the pre-compensated
Figure 3.4: Trajectory of the pre-compensated NRZ-OOK signal for a filter bandwidth of 9.13 GHz (black trace) and the accessible portion of the complex plane (gray region) using a DD-MZM. (a) $\{a_n\} = 0$ or 1, (b) $\{a_n\} = 0.2$ or 1.
optical signal for narrowband optical filtering. $P_0$ ($P_1$) is the power at the center of the bit period for the 0 bit (1 bit, respectively). The required extinction ratio decreases with a decrease in the filter bandwidth. For a DP-MZM, the decrease in the extinction ratio can be avoided since the restriction on the accessible portion of the complex plane is removed, as shown in Figure 3.6.

\begin{figure}
\centering
\includegraphics[width=\textwidth]{extinction_ratio.png}
\caption{Simulated extinction ratio as a function of the filter bandwidth for an NRZ-OOK signal using a DD-MZM.}
\end{figure}

The simulated pre-compensation performance for narrowband optical filtering is characterized in terms of the eye closure $P_{1,\text{min}} - P_{0,\text{max}}$, where $P_{1,\text{min}}$ and $P_{0,\text{max}}$ are the minimum power for 1 bits and maximum power for 0 bits, respectively, at the pulse center [57]. The ECP in decibels is

$$\text{ECP(dB)} = 10 \times \log_{10}\left(\frac{EC_0}{P_0}\right) - 10 \times \log_{10}\left(\frac{EC}{P}\right)$$  \hspace{1cm} (3.2)
Figure 3.6: Trajectory of the pre-compensated NRZ-OOK signal for a filter bandwidth of 9.13 GHz (black trace) and the accessible portion of the complex plane (gray region) using a DP-MZM.
where $EC_0$ and $\overline{P}_0$ are the eye closure and average power, respectively, for the back-to-back transmission. $EC$ and $\overline{P}$ are the eye closure and average power, respectively, for the transmission with or without filter pre-compensation. The ECP as a function of optical filter bandwidth is shown in Figure 3.7 for signals generated by the DD-MZM and DP-MZM. Narrowband filtering degrades the transmission performance and, as expected, the pre-compensation performance is better for the DP-MZM than for the DD-MZM. For a filter bandwidth of 9.13 GHz, pre-compensation improves the ECP by 1.2 dB for the DD-MZM. However, the penalty is not completely eliminated. This is partly due to the quantization noise which makes the actual synthesized waveform deviate from its ideal. Moreover, this deviation is accentuated by the non-flat response of the DD-MZM. The quantization noise can not be avoided but it has been demonstrated that 6 bit DAC resolution results negligible penalty [58]. Higher DAC resolution does not provide obvious improvement. However, the non-flat frequency response of the DACs, driver amplifiers and optical modulator can be measured and pre-emphasized in the back-calculation. On the other hand, for the DP-MZM case, the ECP was improved by 1.43 dB for a filter bandwidth of 9.13 GHz. For a filter bandwidth larger than 20 GHz, no significant penalty occurs.

Figure 3.8 shows simulated and measured optical eye diagrams for a 10 Gb/s NRZ-OOK signal, DD-MZM, and optical filter bandwidth of 9.13 GHz: the transmitted signal without pre-compensation, the filtered signal without pre-compensation, the pre-compensated signal for narrowband optical filtering, and the filtered signal with pre-compensation. The roll-off factor of the raised-cosine pulse is 0.7. The simulated and measured results are in good agreement. The eye diagrams for the DP-MZM (not shown) are similar to those of the DD-MZM except for an increase in the extinction ratio.
Figure 3.7: Dependence of the simulated ECP on the filter bandwidth for a 10 Gb/s NRZ-OOK signal using a DD-MZM and a DP-MZM.
Figure 3.8: Simulated and measured optical eye diagrams for 10 Gb/s NRZ-OOK signals using a DD-MZM. $\beta = 0.7$ for the raised-cosine pulse shape. (a) back-to-back, (b) without filter pre-compensation, (c) pre-compensated signal for a filter bandwidth of 9.13 GHz, and (d) with filter pre-compensation. For measured eye diagrams, time base is 30 ps/div.
Measured optical spectra for the NRZ-OOK signal are shown in Figure 3.9. For the back-to-back signal, the normalized power of the spectral component at the optical carrier is 0 dBm, but the power of spectral components at ±10 GHz away from the carrier drop by over 30 dB. The narrowing of the filtered spectrum without pre-compensation is clearly evident, as is the broadening of the spectrum for the pre-compensated signal. In a WDM system with a channel spacing of 50 GHz, the extent of the spectral broadening is not significant enough to cause crosstalk for neighboring channels. However, from Figure 3.9, it is seen that in order not to induce crosstalk, the channel spacing has be larger than 30 GHz. The spectra of the NRZ-OOK signals generated with the DP-MZM (not shown) are similar to those for the DD-MZM.

Figure 3.9: Measured optical spectra for 10 Gb/s NRZ-OOK signals using a DD-MZM with a resolution bandwidth of 0.01 nm.
Figure 3.10 plots the dependence of the measured BER on the OSNR (0.1 nm resolution bandwidth) of 10 Gb/s NRZ-OOK signals, with and without 9.13 GHz filter pre-compensation. For the DD-MZM (Figure 3.10 (a)), an OSNR penalty of 5.8 dB (at a BER of $10^{-9}$) is induced by the narrowband optical filtering. The penalty is reduced to 1.8 dB with filter pre-compensation. The extent of improvement is limited by the requirement that the pre-compensated signal has to lie within the limited accessible area of the complex plane which in turn results in a decreased extinction ratio. This limitation is removed by using the DP-MZM. As shown in Figure 3.10 (b), for the DP-MZM, the narrowband optical filtering causes an OSNR penalty of 5.5 dB. The penalty is reduced to only 0.9 dB with filter pre-compensation since the required portion of the complex plane can be accessed (Figure 3.6). However, due to the quantization noise and transmitter imperfections such as non-ideal frequency responses of DACs, driver amplifiers and the DP-MZM, perfect pre-compensation is not achieved. The pre-compensation performance can be improved by measuring realistic frequency responses of the DACs, driver amplifiers and DP-MZM and using the measured data to pre-emphasize the back-calculated drive voltages.

### 3.2.4 NRZ-DPSK Modulation with DP-MZM

For NRZ-DPSK signal generation, the DD-MZM needs to be biased at $\pm V_\pi/2$ and drive voltages swing with peak-to-peak voltage of $2V_\pi$. The roll-off factor of the raised-cosine pulse is 1. The accessible region of the complex plane with a peak-to-peak drive voltage of $V_\pi$ is shown by the gray shaded region in Figure 3.11 (a). The trajectory of the ideal pre-compensated signal for a filter bandwidth of 9.13 GHz, which is also shown in the figure, cannot be synthesized since it takes values in the inaccessible region in the vicinity of the origin. The pre-compensated signal can be
Figure 3.10: Measured dependence of the BER on the OSNR for 10 Gb/s NRZ-OOK signals with (a) a DD-MZM, (b) a DP-MZM.
CHAPTER 3. PRE-COMPENSATION FOR OPTICAL FILTERING

synthesized with a DP-MZM, as shown in Figure 3.11 (b). Simulated and measured optical eye diagrams are presented in Figure 3.12 for 10 Gb/s NRZ-DPSK signals with a DP-MZM. Corresponding electrical eye diagrams are shown in Figure 3.13. It can be observed that for both optical and electrical eye diagrams, simulation and measurement results are in good agreement. After passing through the filter, the signal distortion can be clearly seen but it is eliminated by the filter pre-compensation.

The simulated ECP is plotted in Figure 3.14. The narrowband optical filtering pre-compensation improves the system performance significantly. For a filter bandwidth of 9.13 GHz, the penalty is reduced from 1.19 dB to 0.11 dB.

Measured optical spectra for the 10 Gb/s NRZ-DPSK signals are shown in Figure 3.15. The filter pre-compensation emphasizes the high frequency components of the signal so that the effects of the narrowband optical filtering are alleviated.

The dependence of the measured BER on the OSNR (0.1 nm resolution bandwidth) of the NRZ-DPSK signal is illustrated in Figure 3.16. Without filter pre-compensation, the OSNR penalty (at a BER of $10^{-9}$) is 3.5 dB. With the filter pre-compensation, the OSNR penalty is only 0.6 dB. This demonstrates that the NRZ-DPSK signal is more tolerant to narrowband optical filtering than the NRZ-OOK signal. Figure 3.16 also illustrates results for the simultaneous pre-compensation for narrowband optical filtering and fiber chromatic dispersion (1.3 nm/nm). In this case the OSNR penalty is 0.9 dB. As mentioned above, the deviation from the ideal is due to the quantization noise and then accentuated by the non-ideal and/or nonlinear responses of the DACs, driver amplifier and the modulator. The pre-compensation performance can be improved by measuring the realistic frequency responses of the DACs, driver amplifier and the modulator and pre-emphasizing in the back-calculation.
Figure 3.11: Trajectory of the pre-compensated NRZ-DPSK signal for a filter bandwidth of 9.13 GHz (black trace) and the accessible portion of the complex plane (gray region). (a) a DD-MZM, (b) a DP-MZM.
Figure 3.12: Simulated and measured optical eye diagrams for 10 Gb/s NRZ-DPSK signals using a DP-MZM. $\beta = 1$ for the raised-cosine pulse shape. (a) back-to-back, (b) without filter pre-compensation, (c) pre-compensated signal for a filter bandwidth of 9.13 GHz, (d) with filter pre-compensation. For measured eye diagrams, time base is 30 ps/div.
Figure 3.13: Simulated and measured electrical eye diagrams for 10 Gb/s NRZ-DPSK signals using a DP-MZM. (a) back-to-back, (b) without filter pre-compensation, (c) with filter pre-compensation. For measured eye diagrams, time base is 30 ps/div.
Figure 3.14: Dependence of the simulated ECP on the filter bandwidth for a 10 Gb/s NRZ-DPSK signal using a DP-MZM.
Figure 3.15: Optical spectra for 10 Gb/s NRZ-DPSK signals using a DP-MZM with a resolution bandwidth of 0.01 nm.
Figure 3.16: Measured dependence of the BER on the OSNR for 10 Gb/s NRZ-DPSK signals with a DP-MZM.
Simulated and measured optical eye diagrams for the filter and dispersion pre-compensated NRZ-DPSK signal are illustrated in Figure 3.17 (a). Figure 3.17 (b) shows measured optical and electrical eye diagrams with both narrowband optical filtering and dispersion pre-compensation. This demonstrates the possibility of simultaneous pre-compensation for filter and dispersion.

Figure 3.17: Eye diagrams for 10 Gb/s NRZ-DPSK signals using a DP-MZM. (a) (left) simulated pre-compensation optical eye diagram for narrow filter and 1.3 ns/nm of fiber chromatic dispersion, (right) measured corresponding pre-compensated optical eye diagram, (b) (left) measured optical signal with pre-compensation for narrow filter and 1.3 ns/nm of fiber chromatic dispersion, (right) measured corresponding electrical eye diagram. For measured eye diagrams, time base is 30 ps/div.
3.2.5 NRZ-DQPSK Modulation with DP-MZM

For an NRZ-DQPSK signal, each symbol represents two bits. For a 20 Gb/s NRZ-DQPSK signal generated with the DP-MZM, the trajectory of the ideal pre-distorted signal for a filter bandwidth of 9.13 GHz is shown in Figure 3.18. As for the NRZ-DPSK signals in the previous section, the filter pre-compensation can only be achieved with the DP-MZM.

![Figure 3.18: Trajectory of the pre-compensated NRZ-DQPSK signal for a filter bandwidth of 9.13 GHz (black trace) and the accessible portion of the complex plane (gray region) using a DP-MZM.](image)

Simulated and measured optical eye diagrams are shown in Figure 3.19 and electrical eye diagrams are shown in Figure 3.20. The roll-off factor of the raised-cosine pulse shape is 1. The three distinct levels that occur during bit transitions are characteristic of an NRZ-DQPSK signal.
Figure 3.19: Simulated and measured optical eye diagrams for 20 Gb/s NRZ-DQPSK signals using a DP-MZM. $\beta = 1$ for the raised-cosine pulse shape. (a) back-to-back, (b) without filter pre-compensation, (c) pre-compensated signal for a filter bandwidth of 9.13 GHz, and (d) with filter pre-compensation. For measured eye diagrams, time base is 30 ps/div.
Figure 3.20: Simulated and measured electrical eye diagrams for 20 Gb/s NRZ-DQPSK signals using the DP-MZM. (a) back-to-back, (b) without filter pre-compensation, and (c) with filter pre-compensation. For measured eye diagrams, time base is 30 ps/div.
Figure 3.21 illustrates the dependence of the ECP on the filter bandwidth. The filter pre-compensation reduces the penalty from 2.42 dB to only 0.12 dB. The measured optical spectra are shown in Figure 3.22 which looks the same as those of NRZ-DPSK signal (Figure 3.15) due to the same symbol rate.

![Figure 3.21: Dependence of the simulated ECP on the filter bandwidth for a 20 Gb/s NRZ-DQPSK signal using a DP-MZM.](image)

As mentioned in Chapter 2, in the differential receiver, when the phase difference between two interferometer arms is set to $\pi/4$, in-phase channel (denoted U) of the NRZ-DQPSK signal is detected. When the phase difference is $-\pi/4$, quadrature channel is detected (denoted V). The measured BER performance for the in-phase and quadrature channels is shown in Figure 3.23. The 20 Gb/s NRZ-DQPSK signal is severely distorted by the narrowband optical filtering. This leads to a floor in the BER of $10^{-5}$. Compared to NRZ-DPSK, the poor performance is not predicted by
Figure 3.22: Optical spectra for 20 Gb/s NRZ-DQPSK signals using a DP-MZM with a resolution bandwidth of 0.01 nm.
the simulation results. In the simulation, an ideal DAC model is used assuming 6 bit resolution without noise and timing jitter. Moreover, the digital-to-analog conversion is carried out by interpft which is a built-in Matlab function. However, in practice, a DAC can not be perfect and the digital-to-analog conversion is based on a zero-order hold model followed by a low-pass filter. The effective number of bits of a real DAC is smaller than the number of bits and this induces some distortion to the signal [59]. Therefore, in the experiment, the addition of noise (amplitude and phase) and timing jitter could play an important role to affect the NRZ-DQPSK signal which has a strong pattern dependence. With filter pre-compensation, the OSNR penalty is 2.1 dB (at a BER of $10^{-9}$). Figure 3.23 also illustrates results for the simultaneous pre-compensation for narrowband optical filtering and fiber chromatic dispersion (1.3 ns/nm). In this case the ONSR penalty is 3.6 dB. Except for the reasons mentioned above, the penalty may also come from the disturbance of the other tributary due to non-strictly orthogonal in-phase and quadrature components.

Figure 3.24 (a) shows the simulated and measured pre-compensated optical eye diagrams for narrowband optical filtering and 1.3 ns/nm of fiber chromatic dispersion. Figure 3.24 (b) shows the measured optical and electrical eye diagrams with both narrowband optical filtering and dispersion pre-compensation and thus demonstrates the capacity of simultaneous pre-compensation for narrowband filtering and dispersion.
Figure 3.23: Measured dependence of the BER on the OSNR for 20 Gb/s NRZ-DQPSK signals with a DP-MZM.
Figure 3.24: Eye diagrams for 10 Gb/s NRZ-DQPSK signals using a DP-MZM. (a) (left) simulated pre-compensated optical eye diagram, (right) measured pre-compensated optical eye diagram, (b) (left) measured corresponding pre-compensated optical eye diagram, (right) measured electrical signal with pre-compensation for narrowband optical filtering and 1.3 ns/nm of fiber chromatic dispersion. For measured eye diagrams, time base is 30 ps/div.
3.3 Pre-Compensation for Cascaded Optical Filtering

An effective way to further test the performance in the lab is to perform a recirculating loop experiment to fully evaluate the influence of cascaded optical filtering with and without the use of pre-compensation. The performance is evaluated in terms of the required OSNR (measured with 0.1 nm resolution bandwidth) at a BER of $10^{-9}$.

3.3.1 Recirculating Loop Setup

The recirculating loop setup is illustrated in Figure 3.25. Two acousto-optic modulators (AOMs) operate as loop switches. First, AOM1 is open while AOM2 is closed to let the data signal fill the loop. After the loop is filled with the data signal, AOM1 closes and AOM2 opens allowing the data signal to circulate inside the loop which consisted of 80 km NZDSF, DCF, a JDSU TB9 filter with a 3 dB bandwidth of 28 GHz. There were two spans of NZDSF. The first fiber span was 50.558 km (attenuation of 0.2 dB/km and dispersion of 4.6 ps/nm/km at 1550 nm). The second fiber span was 29.862 km (attenuation of 0.2 dB/km and dispersion of 4.6 ps/nm/km at 1550 nm). The DCF had attenuation of 2.95 dB and dispersion of -363.07 ps/nm at 1550 nm. The residual dispersion per loop was -0.89 ps/nm. After each loop, a portion of the data signal was coupled to a pre-amplified receiver for time-gated BER measurement. No polarization controller or polarization scrambler was used inside the loop. However, in order to make the recirculating loop perform more like a straight-line system, polarization controllers or a polarization scrambler are suggested to be used inside the loop to break the periodic polarization-dependent loss (PDL) and polarization-dependent gain (PDG) effects brought by EDFAs as well as the PDL of the optical filters [60], [61].
It has been demonstrated in section 3.2.3 that the per-compensation performance using a DD-MZM is worse than using a DP-MZM. However, in order to compare with the post-compensation using adaptive electronic equalizer which was performed in Athens Information Technology Centre, a DD-MZM was used in the AOWG for the NRZ-OOK signal. A DP-MZM was used for the NRZ-DPSK and NRZ-DQPSK signals.

Figure 3.25: Experimental loop setup for cascaded optical filtering. AOM: acousto-optic modulator, OBPF: optical band-pass filter.

Figure 3.26 shows the measured response of the JDSU optical filter using Luna optical vector analyzer. This JDSU filter has a 3 dB bandwidth of 28 GHz and a flat phase response across the passband. The input power to the loop was 0 dBm. The bit pattern was a deBruijn sequence of length $2^{14}$. 
3.3.2 Results for NRZ-OOK Modulation

For the NRZ-OOK signal, the dependence of the required OSNR at BER $= 10^{-9}$ on the number of loops without and with filter pre-compensation is illustrated in Figure 3.27. Without the use of filter pre-compensation, the maximum number of cascaded filters under the requirement of BER $= 10^{-9}$ is 10. When the number of loops is less than 5, the signal distortion induced by the cascaded filters is negligible. Therefore, the filter pre-compensation is performed when the number of loops is larger than 5 loops. For 10 cascaded filters, the required OSNR is reduced by 6.65 dB with the filter pre-compensation. The advantage of the filter pre-compensation is further emphasized in extending the transmission to 14 loops with a required OSNR of 27.1 dB.
Figure 3.27: Required OSNR at BER = $10^{-9}$ versus number of loops without and with filter pre-compensation for NRZ-OOK signal.
Measured optical eye diagrams are displayed in Figure 3.28 from which the mitigation of the filter induced signal distortion can be clearly seen. The influence of noise accumulation and residual dispersion is observed for longer transmission distances. In addition, measurements were done using the filter pre-compensation for 12 filters for the cases of 5 to 14 filters. The results indicate that a single filter pre-compensation performs well with relatively small penalty.

Figure 3.28: Measured optical eye diagrams for NRZ-OOK signal. (a) 10 loops without filter pre-compensation, (b) 10 loops with filter pre-compensation, (c) 12 loops with filter pre-compensation, (d) 14 loops with filter pre-compensation. Time base is 30 ps/div.
3.3.3 Results for NRZ-DPSK Modulation

For the 10 Gb/s NRZ-DPSK signal, Figure 3.29 shows the dependence of the required OSNR for $\text{BER} = 10^{-9}$ on the number of loops without and with filter pre-compensation. When the number of loops is less than 8, the signal distortion induced by the cascaded filters is not obvious and the compensation is not necessary. Without filter pre-compensation, the maximum number of cascaded filters for $\text{BER} = 10^{-9}$ is 14. The required OSNR is reduced by 8.3 dB for 14 cascaded filters when pre-compensation is implemented. Moreover, with the filter pre-compensation optimized for the specific number of loops, the transmission is extended to 16 loops with a required OSNR of 27.7 dB. Also included in Figure 3.29 are the results obtained by using the filter pre-compensation for 12 filters for the cases of 8 to 16 filters. The results indicate that a single pre-compensation performs well for a different number of filters with small penalties.

The effectiveness of pre-compensation can be limited by the bandwidth restriction of the required pre-compensated signal. The spectrum of the generated optical signal is limited within $\pm 10$ GHz from the carrier. As the optical filtering pre-compensation broadens the spectrum by emphasizing high frequency components of the signal, when the spectrum of the required pre-compensated signal extends beyond 20 GHz, the signal cannot be accurately generated and thus degrades the pre-compensation performance. The electrical eye diagrams shown in Figure 3.30 reveal considerable distortion after the signal propagates through 12 and 14 filters. With the filter pre-compensation, open eye diagrams are obtained.
Figure 3.29: Required OSNR at BER = $10^{-9}$ versus number of loops without and with filter pre-compensation for NRZ-DPSK signal.
Figure 3.30: Measured electrical eye diagrams for the NRZ-DPSK signal. (a) 12 loops without filter pre-compensation, (b) 12 loops with filter pre-compensation, (c) 14 loops without filter pre-compensation, (d) 14 loops with filter pre-compensation. Time base is 30 ps/div.
3.3.4 Results for NRZ-DQPSK Modulation

Figure 3.31 plots the results for the 20 Gb/s NRZ-DQPSK in-phase component without and with filter pre-compensation. Without the use of filter pre-compensation, the maximum number of cascaded filters is only 3 with a required OSNR of 27.1 dB to achieve BER = $10^{-9}$. The filter pre-compensation improves the performance by reducing the required OSNR by 6.2 dB for 3 cascaded filters and further extends the transmission up to 8 filters with a required OSNR of 24.8 dB. However, when the single pre-compensation for 5 filters is used for 3 to 8 filters, a high OSNR of 26.1 dB was needed to achieve BER = $10^{-9}$ for 6 filters. Compared to NRZ-DPSK, the reduced tolerance to filtering largely comes from the addition of noise, on both the amplitude and phase, and timing jitter of the DQPSK signal which has a stronger pattern dependence.

Measured electrical eye diagrams are shown in Figure 3.32 for the in-phase component. Obvious signal distortion after 3 loops transmission without filter pre-compensation is observed from Figure 3.32 (a) but clear and open eye diagrams are obtained with filter pre-compensation for 3, 5, and 8 loops as shown in Figure 3.32 (b). Similar performance was obtained for the quadrature component.
Figure 3.31: Required OSNR at BER = $10^{-9}$ versus number of loops without and with filter pre-compensation for NRZ-DQPSK in-phase component.
Figure 3.32: Measured electrical eye diagrams for the NRZ-DQPSK in-phase component. (a) 3 loops without filter pre-compensation, (b) 3 loops with filter pre-compensation, (c) 5 loops with filter pre-compensation, (d) 8 loops with filter pre-compensation. Time base is 30 ps/div.
3.4 Summary

In this chapter, the pre-compensation for narrowband optical filtering has been investigated by using a single narrow filter with a 3 dB bandwidth of 9.13 GHz and recirculating loop experiment for the 10 Gb/s NRZ-OOK and NRZ-DPSK and 20 Gb/s NRZ-DQPSK modulation formats.

For NRZ-OOK, the DP-MZM offers improved performance compared to the DD-MZM as the required pre-compensated optical signal can be generated more accurately. For NRZ-DPSK and NRZ-DQPSK, the DP-MZM has to be used to generate the pre-compensated signal since it accommodates the limitations imposed by the DD-MZM. The simultaneous pre-compensation of narrowband optical filtering and fiber chromatic dispersion (1.3 ns/nm) has also been achieved for NRZ-DPSK and NRZ-DQPSK formats. Recirculating loop experiments were performed for each of three modulation formats. Both the straight-line and recirculating loop results demonstrate that electronic pre-compensation is a promising approach for coping with the narrowband optical filtering that occurs in networks.
Chapter 4

Electronic Dispersion Pre-Compensation Using InP Mach-Zehnder Modulator

4.1 Introduction

The realization of suitable DSP and DACs has led to electronic dispersion pre-compensation technology for 10.709 Gbit/s optical fiber transmission systems. Utilizing the NRZ-DPSK modulation format and a LiNbO$_3$ DP-MZM, an ASIC with 21.418 GSa/s DACs has enabled transmission over 3840 km of SSMF without optical dispersion compensation [21].

Although this technique reduces the system cost by elimination of optical dispersion compensators and supporting EDFAs from the link, the combination of an NRZ-DPSK modulation format and LiNbO$_3$ DP-MZM is not a low-cost solution. The modulation and demodulation schemes for a DPSK signal need a complex optical transmitter and receiver. Moreover, the LiNbO$_3$ DP-MZM is also an expensive modulator compared to LiNbO$_3$ DD-MZM or semiconductor modulators. However,
in core and metro networks which have shorter reach requirements, low cost and small size are crucial factors.

This chapter investigates a cost effective approach for core and metro networks using electronic dispersion pre-compensation with the NRZ-OOK modulation format and an InP DD-MZM. Experimental results are presented for an assessment of electronic dispersion pre-compensation using InP and LiNbO$_3$ DD-MZMs for 10.709 Gb/s NRZ-OOK transmission. For the rest of the chapter, the words InP MZM and LiNbO$_3$ MZM refer to dual-drive InP and LiNbO$_3$ modulators unless otherwise specified.

4.2 Difference Between InP and LiNbO$_3$ MZMs

Compared to widely available LiNbO$_3$ MZMs, InP MZMs are attractive because of the smaller required drive voltage, reduced size, increased stability, and monolithic integrability with other opto-electronic devices such as lasers and optical amplifiers [63]. In particular, the size of InP MZMs makes them suitable for small industry standard packages. InP MZMs differ from LiNbO$_3$ MZMs in several important ways.

- Each arm of a LiNbO$_3$ MZM can be considered a pure phase modulator. An applied drive voltage changes the phase constant of the optical signal, but there is no appreciable change in the attenuation constant. InP MZMs exhibit a change in both the phase and attenuation constants.

- The dependence of the phase change on applied drive voltage is linear for a LiNbO$_3$ MZM whereas it is nonlinear for an InP MZM.

- The dependence of the phase change on applied voltage is independent of the wavelength of the optical signal for a LiNbO$_3$ MZM whereas the dependence
of the phase and amplitude change on applied voltage is wavelength dependent for an InP MZM.

4.3 InP Modulator Characterization

The InP MZM used in this chapter is a $\pi$-shift modulator designed to achieve near zero-chirp across the C-band. The $\pi$ phase shift is induced by making one arm half a wavelength longer than the other arm. The InP MZM is in the OFF state with equal voltages applied to the two arms. This results in the same optical power for each arm and thus a high extinction ratio [62]. The extinction ratio is theoretically infinity. When the InP MZM is turned ON when two applied voltages are unequal. The phase increase in the deeply biased arm is larger than phase decrease in the other arm. To enable the near zero-chirp operation, the InP multiple quantum well MZM is optimized so that the residual nonlinearity of the phase-versus-voltage characteristic is reduced and independent DC control electrodes are utilized [63].

The output signal from the modulator is given by [64]

$$E(V_1, V_2) = E_0 \left( \sqrt{S_i S_o} \exp \left( -\frac{\alpha(V_1)L}{2} + j\Delta\phi(V_1)L - \pi \right) + \right.$$ \[4.1\]

$$\sqrt{(1-S_i)(1-S_o)} \exp \left( -\frac{\alpha(V_2)L}{2} + j\Delta\phi(V_2)L \right) \right)$$

where $S_i$ and $S_o$ are the input and output power splitting ratios which are equal to 0.5 and 0.63, respectively. $V_1$ and $V_2$ are the voltages applied to two arms of the modulator, $\alpha(V)$ and $\Delta\phi(V)$ are the voltage dependent attenuation and phase constants, and $L$ is the interaction length for the arms of the modulator. $\alpha(V)L$ and $\Delta\phi(V)L$ are obtained based on experimental measurements and non-linear fit
functions. The following expressions and data were provided by Santur Corporation with parameters given in Tables 4.1 and 4.2, respectively.

\[
\alpha(V)L = \frac{A + |V|^C}{B} \quad (4.2)
\]

\[
\Delta\phi(V)L = p_1V + p_2V^2 \quad (4.3)
\]

Table 4.1: Parameters for attenuation constants.

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<th>A</th>
<th>B</th>
<th>C</th>
</tr>
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<td>1565.5 nm</td>
<td>2.0000</td>
<td>10000.000</td>
<td>3.0500</td>
</tr>
<tr>
<td>1549.32 nm</td>
<td>16.4825</td>
<td>3000.0354</td>
<td>3.1643</td>
</tr>
<tr>
<td>1530.33 nm</td>
<td>5.0032</td>
<td>400.0005</td>
<td>2.7634</td>
</tr>
</tbody>
</table>

Table 4.2: Parameters for phase constants.

<table>
<thead>
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<th>Wavelength</th>
<th>( p_1 )</th>
<th>( p_2 )</th>
</tr>
</thead>
<tbody>
<tr>
<td>1565.5 nm</td>
<td>0.243315</td>
<td>0.093853</td>
</tr>
<tr>
<td>1549.32 nm</td>
<td>0.284984</td>
<td>0.114342</td>
</tr>
<tr>
<td>1530.33 nm</td>
<td>0.395947</td>
<td>0.143285</td>
</tr>
</tbody>
</table>

The transmission, defined as \( T = 1 - \exp(-\alpha(V)L) \), as a function of the applied voltage for different operating wavelengths is plotted in Figures 4.1. It is found that there is more attenuation at the same applied voltage for shorter wavelength. Figure 4.2 plots the voltage dependent phase for different wavelength. The required applied voltages to achieve \( \pi \) phase shift are -3.5 V, -4.2 V and -4.62 V for 1530.33 nm, 1549.32 nm and 1565.50 nm, respectively.

A contour plot of normalized transmission at 1549.32 nm is given in Figure 4.3. High and low transmission refer to the ON and OFF states of the modulator, respectively. Since the InP MZM is a \( \pi \)-shift modulator, when the voltages applied to two arms are equal, the InP MZM is in the OFF state. An unbalanced splitting ratio and
Figure 4.1: Wavelength dependence of the transmission on applied voltage for an InP MZM.

Figure 4.2: Wavelength dependence of the phase shift on applied voltage for an InP MZM.
nonlinear voltage dependent phase characteristic can be observed from the contour plot which is asymmetric and has curved lines. Since an InP MZM is a semiconductor type of reverse biased modulator, the applied voltages are within the range of \([-6 \text{V}, 0]\). Figure 4.4 shows the contour plot of normalized transmission of a LiNbO$_3$ MZM which is strictly symmetric and has straight lines. Different from InP MZM, the LiNbO$_3$ MZM is in the ON state with two equally applied voltages. The LiNbO$_3$ MZM can be either positively or negatively biased over a range of typically \([-10 \text{V}, 10 \text{V}]\).

![Contour plot of normalized transmission for InP MZM](image)

Figure 4.3: Calculated dependence of the normalized transmission on the voltages applied to both arms of an InP MZM.

### 4.4 Optical Signal Generation using InP MZM

In order to generate an optical signal with specified amplitude and phase at the output of the modulator, the two requisite drive voltages, $V_1$ and $V_2$ have to be determined. For the LiNbO$_3$ MZM, $V_1$ and $V_2$ can be back-calculated through an
analytical expression, as described in Equation (2.8). However, for the InP MZM, it is not possible to perform the back-calculation analytically because the attenuation and phase constants are measurement based nonlinear fit functions of the applied voltage. The properties of the InP MZM alter the procedure for synthesizing the requisite drive voltages to obtain a specified optical field.

In this case, a brute-force search method was used to obtain the required drive voltages for a desired time-varying optical field to be generated from the InP MZM. The real and imaginary parts of the desired optical field are represented by $E_{r,tar}$ and $E_{i,tar}$. $V_1$ and $V_2$ vary within the range of $[-6 \, V, \, 0]$. A combination of any value of $V_1$ and $V_2$ within that range create a $E_{r,cal}$ and $E_{i,cal}$, but that leads to an infinite number of $V_1$ and $V_2$.

In order to avoid infinite number of solutions, let $V_1$ and $V_2$ vary across the range
of [-6V, 0] in N uniform steps. For each sample time, only one unique pair of $E_{r,tar}$ and $E_{i,tar}$ is available. The $N \times N$ combination of $V_1$ and $V_2$ create $N \times N$ pairs of $E_{r,cal}$ and $E_{i,cal}$. For the first sample time, $N \times N$ pairs of $E_{r,cal}$ and $E_{i,cal}$ and the unique $E_{r,tar}$ and $E_{i,tar}$ are substituted into Equation (4.4). The combination of $V_1$ and $V_2$ which minimizes the cost function is regarded as most appropriate voltage value of the first sample time.

$$C(V_1, V_2) = \sqrt{(E_{r,cal} - E_{r,tar})^2 + (E_{i,cal} - E_{i,tar})^2} \quad (4.4)$$

Similarly, the voltage search is performed for the next sample time and repeated until the last sample time. By doing this, the required drive voltages as a function of time are obtained for a specific desired optical field. Considering the resolution of DACs in the AOWG is 6 bit with 64 quantization levels, 64 values for $V_1$ and 64 values of $V_2$ are good numbers with a step of 0.0923V for each. A coarser voltage step induces more error and a finer step will not improve accuracy while increasing the computation complexity.

As an example, for the first sample time, when $V_1$ and $V_2$ vary in the range of [-6V, 0] with 0.0923V per step, the dependence of the cost function in Equation (4.4) on the drive voltages is given in Figure 4.5. It can be found that for the first sample time, the minimum cost function is 0.005 when $V_1(1) = -3.28V$ and $V_2(1) = -2.89V$.

The brute-force search method is simple to implement but is slow and computationally intensive. A more elegant approach would be a useful advance in future work. For example, optimization algorithms such as brute-force Monte Carlo method [65], [66], the stimulated annealing method [67] and genetic algorithm method [68] could be promising solutions. The process of optimization involves finding the minimum of an objective function defined in Equation (4.4).
Figure 4.5: Calculated dependence of cost function on the voltages applied to both arms of InP MZM for the first sample time.

Figure 4.6 shows the simulated optical eye diagram, two calculated required electrical drive voltages using the brute-force search method, measured optical spectrum and chirp for a 10.709 Gb/s NRZ-OOK signal. The roll-off factor for the raised-cosine pulse shape equals 1. The two drive voltages have unequal peak-to-peak values and there is residual positive chirp with 1 GHz peak-to-peak. This is due to the unbalanced power splitting ratio and nonlinear phase shift of the InP MZM. Comparatively, balanced drive voltages and near-zero chirp are obtained using a LiNbO$_3$ MZM, as shown in Figure 4.7.
Figure 4.6: Signals generated from an InP MZM. (a) Simulated optical eye diagram for NRZ-OOK signal with raised-cosine pulse shape ($\beta = 1$), (b) required two drive voltages, (c) measured optical spectrum (resolution bandwidth = 0.01 nm), (d) measured normalized power and chirp of the NRZ-OOK signal.
Figure 4.7: Signals generated from a LiNbO$_3$ MZM. (a) Simulated optical eye diagram for NRZ-OOK signal with raised-cosine pulse shape ($\beta = 1$), (b) required two drive voltages, (c) measured optical spectrum (resolution bandwidth = 0.01 nm), (d) measured normalized power and chirp of the NRZ-OOK signal.
4.5 Recirculating Loop Experiment with NZDSF

A recirculating loop experiment was done to assess the performance of the dispersion pre-compensation using the InP and LiNbO$_3$ MZMs, as shown in Figure 4.8. A 10.709 Gb/s optical signal was generated from the AOWG and applied to a recirculating loop through AOM1 which operates as the loop switch. After the prescribed number of loops, measurements were performed and the loop was emptied by AOM2, the second loop switch. The loop consisted of two spans of NZDSF, an EDFA and an optical band-pass filter. The first fiber span was 50.609 km (attenuation of 0.203 dB/km and dispersion of 4.3 ps/nm/km at 1550 nm). The second fiber span was 34.624 km (attenuation of 0.201 dB/km and dispersion of 4.5 ps/nm/km at 1550 nm). The PDL and PDG of EDFA as well as the PDL of optical filters sitting inside the loop may induce periodic polarization effects during a recirculating loop [60], [61]. Polarization controllers or polarization scramblers are suggested to be used in future loop experiments in order to more closely mimic a straight-line system.

Two measurements were performed: without and with the electronic dispersion pre-compensation. The received optical signal was degraded by a broadband noise source placed before the optical pre-amplifier. Since the experiment was performed at 10.709 Gb/s which includes 7% FEC overhead, the transmission performance was characterized in terms of OSNR penalty at a BER of $10^{-3}$ assuming an effective FEC is installed.

Figure 4.9 shows the dependence of the required OSNR (0.1 nm resolution bandwidth) for BER = $10^{-3}$ without and with dispersion pre-compensation. For the InP MZM, without pre-compensation, the maximum transmission is 1513.8 ps/nm (336 km, 4 loops). For the same amount of dispersion, the required OSNR is reduced by 7.1 dB by using pre-compensation. Moreover, the dispersion pre-compensation
further extends the maximum transmission to 4920 ps/nm (1092 km, 13 loops) with a required OSNR of 19.1 dB.

For comparison, the recirculating loop experiment was repeated using a LiNbO$_3$ MZM with results given in Figure 4.9. Without pre-compensation, the performance is similar to that of the InP MZM. The maximum transmission is 1520 ps/nm (336 km, 4 loops). With pre-compensation, the maximum transmission is 5071.2 ps/nm (1260 km, 15 loops) with a required OSNR of 18.1 dB.

The measured optical eye diagrams using the InP MZM are displayed in Figure 4.10 from which the mitigation of the dispersion induced signal distortion can be clearly seen. The influence of noise accumulation is observed for longer transmission distances. Similarly, measured optical eye diagrams using the LiNbO$_3$ MZM are displayed in Figure 4.11.
CHAPTER 4. DISPERSION PRE-COMPENSATION USING INP MZM

Figure 4.9: Required OSNR versus dispersion without and with dispersion pre-compensation for 10.709 Gb/s NRZ-OOK signal using an InP and a LiNbO$_3$ MZM.
Figure 4.10: Measured optical eye diagrams using an InP MZM. (a) back-to-back, (b) 4 loops without dispersion compensation, (c) 4 loops pre-compensated signal, (d) 4 loops with dispersion pre-compensation, (e) 10 loops pre-compensated signal, (f) 10 loops with dispersion pre-compensation. Time base is 30 ps/div.
Figure 4.11: Measured optical eye diagrams using a LiNbO$_3$ MZM. (a) back-to-back, (b) 4 loops without dispersion compensation, (c) 4 loops pre-compensated signal, (d) 4 loops with dispersion pre-compensation, (e) 10 loops pre-compensated signal, (f) 10 loops with dispersion pre-compensation. Time base is 30 ps/div.
4.6 Straight-Line Experiment with SSMF

A straight-line experiment with amplified SSMF spans was also performed. For the InP MZM, the dependence of the BER on the OSNR (0.1 nm resolution bandwidth) for 76 km, 152 km and 225 km transmission is shown in Figure 4.12.

![Graph showing BER vs OSNR for different SSMF lengths](image)

Figure 4.12: Measured dependence of the BER on the OSNR for 10.709 Gb/s NRZ-OOK signals with an InP MZM for different SSMF lengths.

Measured optical eye diagrams for back-to-back, 76 km, 152 km and 225 km with dispersion pre-compensation are displayed in Figure 4.13 from which it is found that the pre-compensation leads to clear and open optical eye diagrams up to 225 km SSMF transmission.
Figure 4.13: Measured optical eye diagrams with InP MZM in straight-line experiment. (a) back-to-back, (b) 76 km with pre-compensation, (c) 152 km with pre-compensation, (d) 225 km with pre-compensation.
For the LiNbO$_3$ MZM, Figure 4.14 plots the dependence of the measured BER on the OSNR (0.1 nm resolution bandwidth) for 76 km, 152 km and 225 km transmission. Measured optical eye diagrams are presented in Figure 4.15 which demonstrates the capability of the dispersion pre-compensation for up to 225 km SSMF transmission.

Figure 4.14: Measured dependence of the BER on the OSNR for 10.709 Gb/s NRZ-OOK signals with a LiNbO$_3$ MZM for different SSMF lengths.
Figure 4.15: Measured optical eye diagrams with a LiNbO$_3$ MZM in straight-line experiment. (a) back-to-back, (b) 76 km with pre-compensation, (c) 152 km with pre-compensation, (d) 225 km with pre-compensation.
Table 4.3 illustrates the OSNR penalties for a BER of $10^{-3}$. The penalties are relative to the back-to-back performance. The penalties with pre-compensation are attributed to the limitations from using a dual-drive MZM, quantization noise and the non-ideal modulator frequency response which is not considered while calculating the required drive voltages. The pre-compensated performance is comparable for 76 km and 152 km between the InP MZM and LiNbO$_3$ MZM. However, there is 1.65 dB difference in OSNR penalty at BER of $10^{-3}$ for 225 km. This can be attributed to the inaccurate model in Equation (4.1) in which the power splitting ratio, the dependence of the attenuation and phase constants on the applied drive voltage are all based on measured data. Slight deviation makes it hard to accurately generate the required drive voltages.

Table 4.3: Comparison of straight-line OSNR penalty at BER $10^{-3}$.

<table>
<thead>
<tr>
<th></th>
<th>Without pre-compensation</th>
<th>With pre-compensation</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>76 km</td>
<td>76 km</td>
</tr>
<tr>
<td>InP MZM</td>
<td>3.79 dB</td>
<td>1.07 dB</td>
</tr>
<tr>
<td>LiNbO$_3$ MZM</td>
<td>3.75 dB</td>
<td>1.05 dB</td>
</tr>
</tbody>
</table>

4.7 Summary

In this chapter, an experimental investigation of the performance of dispersion pre-compensation using InP and LiNbO$_3$ dual-drive MZMs for the NRZ-OOK modulation format has been presented. A recirculating loop experiment with NZDSF and a straight-line experiment with amplified SSMF spans were performed. It was demonstrated that the combination of the NRZ-OOK modulation format and a dual-drive InP MZM provides a compact and cost effective solution for transmission systems with moderate reach requirements.
Chapter 5

Electronic Dispersion Pre-Compensation on PM-16 Quadrature Amplitude Modulation

5.1 Introduction

In a WDM system, the growth of optical fiber transmission capacity can be achieved by increasing the number of channels simultaneously transmitted along the fiber or by increasing the bit rate per channel. One approach to increase the aggregate bit rate at the same optical amplification bandwidth is to use multi-level modulation formats which encode multiple bits on one symbol and polarization multiplexing. This promises the enhancement of spectral efficiency (SE) in a unit of b/s/Hz which is defined as

\[
SE = \frac{C}{\Delta f}
\]  

(5.1)

where \( C \) is the bit rate per channel and \( \Delta f \) is the allocated channel bandwidth. Recent achievements in high SE WDM transmission systems are listed in Table 5.1.
Table 5.1: Recent achievements in high SE WDM transmission.

<table>
<thead>
<tr>
<th>Modulation formats</th>
<th>SE</th>
<th>Bit rate</th>
<th>WDM grid</th>
</tr>
</thead>
<tbody>
<tr>
<td>PM-RZ-QPSK [69]</td>
<td>4.1 b/s/Hz</td>
<td>112 Gb/s</td>
<td>25 GHz</td>
</tr>
<tr>
<td>PM-RZ-8PSK [70]</td>
<td>4.2 b/s/Hz</td>
<td>114 Gb/s</td>
<td>25 GHz</td>
</tr>
<tr>
<td>PM-RZ-8QAM [31]</td>
<td>4.2 b/s/Hz</td>
<td>114 Gb/s</td>
<td>25 GHz</td>
</tr>
<tr>
<td>PM-36QAM [71]</td>
<td>8 b/s/Hz</td>
<td>107 Gb/s</td>
<td>12.5 GHz</td>
</tr>
<tr>
<td>PM-64QAM [72]</td>
<td>9 b/s/Hz</td>
<td>120 Gb/s</td>
<td>12.5 GHz</td>
</tr>
</tbody>
</table>

The modulation formats listed in Table 5.1 transmit data by modulating either the phase only or both the amplitude and phase of the signal. Therefore, at the end of transmission, a coherent receiver is used to fully access the information of the signal. In a coherent transmission system, the chromatic dispersion is usually post-compensated using a time-domain digital FIR filter [73] or a frequency-domain equalizer [74] in the coherent receiver.

In this chapter, the transmission performance of dispersion pre-compensation for 85.672 Gb/s PM-16QAM signals is investigated for the first time using digital coherent detection and offline DSP. In a laboratory test environment, the recirculating loop coherent transmission would require offline DSP unless the real-time implementation hardware designed specifically to receive the burst-mode signal is available. The AOWG with a DP-MZM connected is characterized to ensure the quality of the 16QAM and pre-compensated 16QAM signal generation at the transmitter. For the simulation results, the impact that the three bias voltages of the DP-MZM have on the dispersion pre- and post-compensation has been studied in terms of BER and EVM. For the experimental results, Corning Vascade EX2000 fiber which is designed to have ultra low attenuation with large effective area is used. The transmission up to 2400 km EX2000 fiber with dispersion pre- and post-compensation is compared.
Also, the optimized power of external cavity laser (ECL), used as the signal laser source, for dispersion pre- and post-compensation is investigated.

5.2 16QAM and Constellation Diagrams

A 16QAM signal has 4 bits per symbol. While higher order QAM modulation formats can carry more bits per symbol, if the average power of the signals remains the same, the constellation symbols must be closer together. Therefore, the transmission becomes more susceptible to noise and an increase in OSNR is required for the same BER.

With respect to the constellation distributions, there are three types of 16QAM: star 16QAM [75], [76], 16-ary amplitude phase shift keying (16APSK) [77] and square 16QAM, as shown in Figure 5.1. It can be observed that by distributing the constellation symbols in a square grid with equal horizontal and vertical spacing, the distance between adjacent symbols in square 16QAM is greater than that in star 16QAM and 16APSK. The constellation symbols are more distinct and therefore decision errors can be reduced. Recently, square 16QAM has been more extensively investigated. In this chapter, all simulation and experimental work are based on square 16QAM.
Figure 5.1: Constellation diagrams (a) star 16QAM, (b) 16APSK, (c) square 16QAM.
5.3 AOWG Characterization

16QAM is a modulation format where both the in-phase and quadrature components are four-level modulated. The imperfections and limitations of the transmitter are potentially more significant as the order of the modulation format increases. Therefore, it is very important to take into account the properties of the transmitter to improve the quality of the signal generation at the transmitter.

In this section, a characterization of the AOWG is described. One of the major limitations of the AOWG is the non-flat frequency response of the DACs, amplifiers and optical modulator over the modulated signal bandwidth. Magnitude ripple and a non-linear phase response can lead to signal distortion. Another limitation is due to the limited peak-to-peak voltage \( V_{pp} \) at the amplifier output.

5.3.1 DACs and Amplifiers Characterization

The DACs and amplifiers used in the AOWG both have finite bandwidths. According to the product datasheet, the 3 dB signal bandwidth of the amplifier is 8 GHz. Due to unavailability of datasheet of the DAC and the DAC and following amplifier are mounted on a circuit board, it is not possible to obtain the signal at the DAC output. However, the actual overall frequency response of the DAC and amplifier can be measured.

The magnitude response was acquired using the measurement setup shown in Figure 5.2 which however can not measure the phase response. The optical modulator was removed from the AOWG. The amplifier output ports for path a and b were connected to a RF spectrum analyzer (RFSA). Quantized values of sinusoidal signals were synthesized offline using a computer and then loaded into the external memory of the AOWG. The signal power for each path was measured on the RFSA. The
start and end frequencies were 0.6 MHz and 10.709 GHz, respectively. A total of 64 sinusoidal signals were generated with a frequency step of 167 MHz. The gain of the amplifier was small to ensure it was working in a linear region.

Figure 5.2: Setup for measuring overall magnitude response of the DAC and amplifier.

The measured magnitude responses of path a and path b at DC are assumed to have the same value at 0.6 MHz. All data are then normalized to the value at 0.6 MHz. Figure 5.3 plots the normalized measured results as well as the sinc function. Magnitude ripples can be observed in the measured data. The inverse of the measured magnitude response is used in the back-calculation to pre-equalize the magnitude rolloff effect of the DAC. The phase response of the DAC is not measured due to the difficulty for the measurement and thus is not taken into account in the back-calculation. However, according to the design specification of the AOWG, the phase response of the DAC and amplifier is quite linear.

5.3.2 Dual-Parallel MZM Characterization

In previous chapters, the optical modulators were assumed to have flat magnitude responses and linear phase responses which, however, is not the case. The DP-MZM used for 16QAM signal generation has a non-flat magnitude response and non-linear phase response over a wide frequency range. The DP-MZM was characterized at each
of the two RF ports using an electrical vector network analyzer (VNA) and a calibrated, unamplified photodiode module using the setup shown in Figure 5.4. A laser source was required to provide an optical input to the DP-MZM and the frequency response was obtained from the modulated optical signal. Since the VNA was only capable of generating and measuring electrical signals, a characterized photodiode was needed to convert the modulated optical signal back to an electrical signal which was measured by the VNA. Before the measurement, a calibration is needed to remove systematic error and improve the accuracy and repeatability. The VNA and photodiode module are both from Anritsu Corporation and work from 40 MHz to 65 GHz.

The magnitude and phase responses are all normalized to the values at 40 MHz. The normalized magnitude and phase responses for both RF ports are given in Figure 5.5 and Figure 5.6, respectively. The 3 dB bandwidth of the DP-MZM is about
CHAPTER 5. DISPERSION PRE-COMPENSATION ON PM-16QAM

Figure 5.4: Setup for measuring the frequency response of the DP-MZM.

15 GHz. The phases fluctuate within ±12° of 0°. Some magnitude and phase ripples which go up with frequency can be observed from measured data. Although the systematic errors such as imperfections in the VNA can be removed through calibration, drift occurring after the calibration, random errors induced by instrument noise and the package imperfections of the DP-MZM can all contribute to the ripples. These ripples are not deterministic and vary from trace to trace. For the purpose of calculation, the values of magnitude and phase responses at DC are required and they are assumed to have the same values at 40 MHz.

The measured frequency responses of the DACs, amplifiers and DP-MZM are first interpolated. Then, using the inverse of the interpolated data, a pre-equalization is done in the back-calculation to overcome the signal distortion induced by the bandwidth limitations of the DACs, amplifiers and DP-MZM. For example, the two required drive voltages become as follows

\[
v_{\text{rf}, A}(t) = \mathcal{F}^{-1} \left( \mathcal{F} \left( v_{\text{rf}, 1}(t) \right) \cdot H_{\text{path}, A}^{-1} \cdot H_{\text{MZM}, A}^{-1} \right) \tag{5.2}
\]

\[
v_{\text{rf}, B}(t) = \mathcal{F}^{-1} \left( \mathcal{F} \left( v_{\text{rf}, 2}(t) \right) \cdot H_{\text{path}, B}^{-1} \cdot H_{\text{MZM}, B}^{-1} \right) \tag{5.3}
\]
where $v_{rf,1}(t)$ and $v_{rf,2}(t)$ are back-calculated required drive voltages according to Equation (2.13) and (2.14), $H^{-1}_{\text{path},A}$ and $H^{-1}_{\text{path},B}$ are the inverse of measured magnitude responses of DAC and amplifier for each RF path. $H^{-1}_{\text{MZM},A}$ and $H^{-1}_{\text{MZM},B}$ are inverse of measured frequency responses of DP-MZM for each RF port.

![Figure 5.5: Measured magnitude response for RF1 and RF2 ports of DP-MZM.](image)

**5.3.3 Limited Peak-to-Peak Voltage of Amplifier Output**

For the 16QAM signal generation, each inner MZM of the DP-MZM is biased at null and driven by a four-level voltage ideally with a $V_{pp}$ of $2V_\pi$. However, the DP-MZM has a large $V_\pi$ of 6.4 V. For the amplifiers in the AOWG, the maximum $V_{pp}$ of the output electrical drive voltage is 7.34 V which is only 1.15$V_\pi$. This can be solved by using a DP-MZM which has smaller $V_\pi$ or changing to an amplifier with large enough output. However, with the current DP-MZM and amplifier used in the AOWG, a
method has to be found to avoid signal distortion induced by the limited $V_{pp}$ at the amplifier output. As discussed in Chapter 2, by biasing two inner MZMs at null, the output signal from a DP-MZM is represented as

$$E_{\text{out}}(t) = \sin\left(\frac{\pi}{V_{\pi}}v_{rf,1}(t)\right) + j\sin\left(\frac{\pi}{V_{\pi}}v_{rf,2}(t)\right) \quad (5.4)$$

The two required drive voltages are back-calculated as follows assuming they have a $V_{pp}$ of $2V_{\pi}$,

$$v_{rf,1}(t) = \frac{V_{\pi}}{\pi} \arcsin\left[|E_{\text{out}}(t)| \cos(\angle E_{\text{out}}(t))\right] \quad (5.5)$$

$$v_{rf,2}(t) = \frac{V_{\pi}}{\pi} \arcsin\left[|E_{\text{out}}(t)| \sin(\angle E_{\text{out}}(t))\right] \quad (5.6)$$
One drive voltage is shown in Figure 5.7. It is found that the four-level drive voltage is not equally spaced ($D_1 = D_3 > D_2$).

![Figure 5.7: Calculated unequally spaced ($D_1 = D_3 > D_2$) four-level drive voltage assuming amplifier output has a $V_{pp}$ of $2V_\pi$.](image)

If the amplifier output $V_{pp}$ reaches $2V_\pi$, due to the nonlinear modulator electrical/optical conversion, the unequally spaced four-level drive voltages lead to a uniform constellation diagram, as shown in Figure 5.8 (a). However, if the amplifier output $V_{pp}$ is limited to $1.15V_\pi$, which drives the modulator in a semi-linear region, the unequally spaced electrical drive voltages result in a non-uniform constellation diagram in which both the amplitude and phase deviate from the ideal constellation, as presented in Figure 5.8 (b). Moreover, the limited amplifier output leads to a reduced average signal power.

In order to compensate for the limited $V_{pp}$ at the amplifier output to obtain a uniform constellation diagram, a scale factor ($S$) should be used to make the modulator operate in a semi-linear region. Different $V_{pp}$ of drive voltages leads to different $V_{pp}$ of the optical field at the output of the modulator. The scale factor can be defined as
the voltage ratio, $V_{pp}/2V_\pi$, which is equal to 0.57. The voltage ratio was used for all simulation and experimental results in this chapter. However, to be more accurate, the scale factor can be defined as the ratio of $V_{pp}$ of the optical field. According to Equation (5.4), when drive voltages have a $V_{pp}$ of $2V_\pi$, both in-phase and quadrature components of the optical field have $V_{pp}$ of 2. However, when the drive voltages have only a $V_{pp}$ of $1.15V_\pi$, the $V_{pp}$ of in-phase and quadrature components of the optical field becomes 1.4. Therefore, the scale factor is calculated as $1.4/2$ which is 0.7. The scale factor of either 0.57 or 0.7 won’t lead to a big difference in the experimental results.

Considering the effect of limited $V_{pp}$ at the amplifier output, Equation (5.4), the output signal from the DP-MZM, becomes

$$E_{out}(t) \times S = \sin \left( \frac{\pi}{V_\pi} v_{rf,1}(t) \right) + j \sin \left( \frac{\pi}{V_\pi} v_{rf,2}(t) \right)$$

(5.7)
The corresponding required electrical drive voltages are back-calculated by

\[ v_{r,1}(t) = \frac{V}{\pi} \arcsin \left( |E_{\text{out}}(t)| \cos \left( \angle E_{\text{out}}(t) \right) \right) \]  \hspace{1cm} (5.8)

\[ v_{r,2}(t) = \frac{V}{\pi} \arcsin \left( |E_{\text{out}}(t)| \sin \left( \angle E_{\text{out}}(t) \right) \right) \]  \hspace{1cm} (5.9)

The obtained four-level drive voltages with \( S = 0.57 \) are nearly equally spaced. One of them is plotted in Figure 5.9. Under the condition that the modulator works in a semi-linear region with the amplifier output limited to a \( V_{\text{pp}} \) of 1.15\( V_\pi \), a uniform constellation diagram with reduced average power, presented in Figure 5.10, is obtained.

![Figure 5.9: Calculated nearly equally spaced \((D_1 \simeq D_2 \simeq D_3)\) four-level drive voltage assuming 1.15\( V_\pi \) \( V_{\text{pp}} \).](image)

With consideration of the actual response of the DAC, amplifier and DP-MZM as well as the limited amplifier output \( V_{\text{pp}} \), a simulated and a measured optical eye diagram with a raised-cosine pulse shape \( (\beta = 1) \) for a 42.836 Gb/s 16QAM signal are shown in Figure 5.11 (a) and (b), respectively. After considering the frequency
response of the DACs, driver amplifiers and DP-MZM in both simulation and measurement, the measured optical eye diagrams still deviates from the simulated eye diagram due to the implementation. Two back-calculated, nearly equally spaced, four-level electrical drive voltages are given in Figure 5.11 (c). Figure 5.11 (d) shows the measured carrier suppressed optical spectrum in which first two nulls are at around ±10 GHz from optical carrier.

5.4 Simulation Investigation of Optical Modulator Bias Voltage Error

For the DP-MZM, the required bias voltage for a certain operating point drifts over time. In the MZM's electro-optic lithium-niobate material, the combination of pyroelectric, photorefractive and photoconductive effects causes a specific bias voltage to drift due to environmental perturbation [78]. For example, a bias voltage set at quadrature point on the modulator transfer function curve at one time may yield a different transmission point on the curve at a later time. This will degrade the quality of the generated signal and hence the transmission performance.
Figure 5.11: A 42.836 Gb/s 16QAM signal. (a) Simulated optical eye diagram with raised-cosine pulse shape (\(\beta = 1\)), (b) measured optical eye diagram; time base is 30 ps/div, (c) back-calculated drive voltages, (d) measured optical spectrum with a resolution bandwidth 0.01 nm.
In this section, the effect of bias voltage error on a 85.672 Gb/s PM-16QAM signal with dispersion pre- or post-compensation in a coherent transmission system is numerically investigated using VPIphotonics transmission maker software. The performance degradation is evaluated by the BER and EVM.

5.4.1 Simulation Setup

The VPI simulation setup of a 85.672 Gb/s PM-16QAM coherent optical transmission system is illustrated in Figure 5.12. The transmitter was executed with co-simulation in Matlab files which generate 42.836 Gb/s 16QAM signals with or without dispersion pre-compensation. A certain amount of bias voltage error for each of three bias voltages of a DP-MZM was intentionally induced. The transmitted signal was launched to a PBS. The signals in the two orthogonal polarizations were decorrelated by 10 symbols using a passive delay-line and are then combined through a polarization beam combiner (PBC). The optical fiber had a total length of 1200 km with 100 km per span. The attenuation and dispersion of the fiber were 0.162 dB/km and 19.36 ps/nm/km at a wavelength of 1550 nm, respectively. The fiber has a PMD parameter of 0.05 ps/√km and the nonlinearity coefficient was assumed to be zero. EDFAs were used to compensate the fiber span loss. The output signal power of each EDFA was 0 dBm and the noise figure for each EDFA was 5 dB.

After noise loading, the signal was sent to a coherent receiver front end which is assumed to be ideal. Using an optical LO, the received optical signal was converted into in-phase and quadrature components of two orthogonal polarizations by two 90° hybrids, four balanced detectors and ADCs. The data were captured after electrical filters and processed offline in Matlab. The linewidth was zero for the signal laser but was 200 kHz for the LO. The power and frequency differences between the signal
and LO were 20 dB and 100 MHz, respectively. The simulation was performed at an OSNR of 22 dB which was measured by the OSNR monitor. Table 5.2 summarizes all parameters used for the simulation setup.

![Simulation setup of a 85.672 Gb/s PM-16QAM coherent transmission system. PD: photodiode.](image)

**Figure 5.12**

---

### 5.4.2 Offline Digital Signal Processing

The offline DSP used to recover the PM-16QAM signal includes

- Differential encoding and decoding.

- Chromatic dispersion post-compensation.

- Adaptive equalization for compensation of residual chromatic dispersion and polarization separation.
Table 5.2: Parameters for simulation setup.

<table>
<thead>
<tr>
<th>Components</th>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Delay</td>
<td>Length</td>
<td>933.794 ps</td>
</tr>
<tr>
<td>EDFA</td>
<td>Output power</td>
<td>0 dBm</td>
</tr>
<tr>
<td></td>
<td>Noise figure</td>
<td>5 dB</td>
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<tr>
<td>Fiber</td>
<td>Length/span</td>
<td>100 km</td>
</tr>
<tr>
<td></td>
<td>Attenuation</td>
<td>0.162 dB/km</td>
</tr>
<tr>
<td></td>
<td>Dispersion</td>
<td>19.36 ps/nm/km at 1550 nm</td>
</tr>
<tr>
<td></td>
<td>PMD parameter</td>
<td>0.05 ps/√km</td>
</tr>
<tr>
<td>OBPF</td>
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<td></td>
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<td>50 GHz</td>
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<td>Noise power density</td>
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</tr>
<tr>
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<td>Linewidth</td>
<td>200 kHz</td>
</tr>
<tr>
<td></td>
<td>Frequency offset</td>
<td>100 MHz</td>
</tr>
<tr>
<td>Photodiodes</td>
<td>Responsivity</td>
<td>1 A/W</td>
</tr>
<tr>
<td></td>
<td>Dark current</td>
<td>0 A</td>
</tr>
<tr>
<td></td>
<td>Thermal noise</td>
<td>$10^{-12}$ A/√Hz</td>
</tr>
<tr>
<td></td>
<td>Shot noise</td>
<td>Included</td>
</tr>
<tr>
<td>Electrical filters</td>
<td>Filter type</td>
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<tr>
<td></td>
<td>Transfer function</td>
<td>Bessel</td>
</tr>
<tr>
<td></td>
<td>Bandwidth</td>
<td>17.1344 GHz</td>
</tr>
<tr>
<td></td>
<td>Filter order</td>
<td>5</td>
</tr>
</tbody>
</table>

- Second-order loop filter for carrier phase and frequency offset estimation.
- Bit error counting and EVM calculation.

This section briefly presents each of these DSP algorithms which were implemented in Matlab.
Differential Encoding and Decoding

Due to the symmetric nature of the 16QAM signal, the constellation pattern of the 16QAM signal remains unchanged with a rotation of $m \cdot \pi/2$ radians where $m$ is an integer. Therefore, a four-fold phase ambiguity arises at the receiver which is able to recognize the constellation pattern of the signal but however can not distinguish a phase rotation of $m \cdot \pi/2$. This may cause catastrophic failure even in simulation [79]. This may be avoided by using differential encoding at the transmitter. Alternatively, a short known reference pattern can be used to test regularly at the receiver to determine the correct phase rotation of the 16QAM constellation and therefore the cycle slip induced long error bursts can be avoided [32]. In this chapter, angle differential encoding is used to resolve the phase ambiguity. However, since the differential encoding can not preserve the Gray code rule, it leads to a worse BER compared to the nondifferential encoded case [80].

For a 16QAM signal, every 4 bits are mapped to one of the $2^4$ constellation symbols. The first two bits are represented in terms of a differential angle $\Delta \theta_1$ and the last two bits correspond to another differential angle $\Delta \theta_2$ according to the mapping rule in Table 5.3 [80]:

<table>
<thead>
<tr>
<th>$B_0$</th>
<th>$B_1$</th>
<th>$\Delta \theta$</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>0</td>
<td>1</td>
<td>$\pi/2$</td>
</tr>
<tr>
<td>1</td>
<td>1</td>
<td>$\pi$</td>
</tr>
<tr>
<td>1</td>
<td>0</td>
<td>$3\pi/2$</td>
</tr>
</tbody>
</table>

The 16 QAM symbols $s_k$ are represented as

$$s_k = a_k + b_k$$
and the differential encoding scheme is given by the following recursive formulae [80]:

\[ a_k = a_{k-1} \cdot e^{j \Delta \theta_{1,k}} \] (5.11)
\[ b_k = b_{k-1} \cdot e^{j \Delta \theta_{2,k}} \] (5.12)

with initial values of \( a_1 \) and \( b_1 \) are \( 2 + j2 \) and \( 1 + j1 \), respectively. The differential decoding scheme is described as follows [80]:

\[ c_k = 2 \sqrt{2} \cdot \left[ \text{sign}(\text{real}(r_k)) + j \text{sign}(\text{imag}(r_k)) \right] \] (5.13)
\[ d_k = \sqrt{2} \cdot \text{sign}(\text{real}(r_k - c_k)) + j \text{sign}(\text{imag}(r_k - c_k)) \] (5.14)

where real and imag denote the real and imaginary part of the received symbol \( r_k \) and sign represents the sign function. From \( c_k \) and \( d_k \) and their previous values, the differential angle and the corresponding dibit can be obtained according to the following rule [80]:

\[ c_k \times c_{k-1}^* = \begin{cases} 
16 & \Delta \theta_1 = 0 \\
16j & \Delta \theta_1 = \pi/2 \\
-16 & \Delta \theta_1 = \pi \\
-16j & \Delta \theta_1 = 3\pi/2 
\end{cases} \] (5.15)

\[ d_k \times d_{k-1}^* = \begin{cases} 
4 & \Delta \theta_2 = 0 \\
4j & \Delta \theta_2 = \pi/2 \\
-4 & \Delta \theta_2 = \pi \\
-4j & \Delta \theta_2 = 3\pi/2 
\end{cases} \] (5.16)
Chromatic Dispersion Compensation

The chromatic dispersion post-compensation can be achieved using time-domain equalization (TDE) [73] or frequency-domain equalization (FDE) [74]. In the simulation, TDE implemented with a digital FIR filter was used. The fiber transfer function with chromatic dispersion is modeled as

\[
H_{\text{fiber}}(f) = \exp\left(-j \frac{L D \lambda^2 \pi f^2}{c}\right)
\]  

(5.17)

\(L\) is the length of the fiber, \(D\) is the dispersion coefficient, \(\lambda\) is the carrier wavelength, and \(c\) is the speed of light. Its impulse response is obtained by taking the inverse Fourier transform, resulting in [81]

\[
g(t) = \sqrt{\frac{c}{jD\lambda^2 z}} \exp\left(j \frac{\pi c}{D\lambda^2 z} t^2\right)
\]  

(5.18)

Considering a digital FIR filter with an odd number of taps, the tap weights are determined by [73]

\[
a_k = \sqrt{\frac{j c}{D\lambda^2 z}} \exp\left(-j \frac{\pi c T^2}{D\lambda^2 z} k^2\right) - \left\lfloor \frac{N}{2} \right\rfloor \leq k \leq \left\lceil \frac{N}{2} \right\rceil
\]  

(5.19)

where \(\lfloor x \rfloor\) is the integer part of \(x\) rounded towards minus infinity and \(N\) is the total number of taps defined by [73]

\[
N = 2 \times \left\lceil \frac{|D| \lambda^2 L}{2c T^2} \right\rceil + 1
\]  

(5.20)
Adaptive Equalization

Adaptive equalization is used for the polarization separation and compensation for residual chromatic dispersion. An adaptive equalizer consists of four FIR filters arranged in a butterfly structure, as shown in Figure 5.13 [82].

![Figure 5.13: Adaptive equalizer.](image)

The output of the equalizer is given by

\[ X_{out} = h_{xx} \cdot x_{in} + h_{xy} \cdot y_{in} \]  \hspace{1cm} (5.21)

\[ Y_{out} = h_{yx} \cdot x_{in} + h_{yy} \cdot y_{in} \]  \hspace{1cm} (5.22)

where \( h_{xx}, h_{xy}, h_{yx} \) and \( h_{yy} \) are adaptive filters with a small number of taps for each. \( x_{in} \) and \( X_{out} \) are the equalizer input and output for the x-polarization signal. \( y_{in} \) and \( Y_{out} \) are the equalizer input and output for the y-polarization signal. The tap weights
are adapted according to following rule [73]:

\[ h_{xx} \rightarrow h_{xx} + \mu \varepsilon_x X_{out} \cdot x^*_\text{in} \]  
\[ h_{xy} \rightarrow h_{xy} + \mu \varepsilon_x X_{out} \cdot y^*_\text{in} \]  
\[ h_{yx} \rightarrow h_{yx} + \mu \varepsilon_y Y_{out} \cdot x^*_\text{in} \]  
\[ h_{yy} \rightarrow h_{yy} + \mu \varepsilon_y Y_{out} \cdot y^*_\text{in} \]  

where \( \mu \) is the step size parameter. \( x^*_\text{in} \) and \( y^*_\text{in} \) denote the complex conjugate of \( x_\text{in} \) and \( y_\text{in} \). \( \varepsilon_x \) and \( \varepsilon_y \) are the error signals defined as the error between the equalizer output and the nearest constellation radius \( R \) [82]

\[ \varepsilon_x = R - |X_{out}|^2 \]  
\[ \varepsilon_y = R - |Y_{out}|^2 \]  

For 16QAM, the constellation points define three distinct radii.

**Carrier Phase and Frequency Offset Estimation**

A second-order loop filter, shown in Figure 5.14 [83], performs carrier phase and frequency offset estimation simultaneously. The governing equations of the tracking loop are given as [83]

\[ \hat{\theta}(k + 1) = \hat{\theta}(k) + \zeta(k) \]  
\[ \zeta(k) = \zeta(k - 1) + \gamma(1 + \rho)e(k) - \gamma e(k - 1) \]
where $\gamma$ and $\rho$ are the loop parameters. The decided symbols $\hat{a}_k$ and the received symbols $\tilde{x}_k$ are both sent to the phase error detector to calculate the phase error [83]

$$e(k) = \text{Imag}(\hat{a}_k^* \cdot \tilde{x}_k e^{-j\hat{\theta}(k)})$$  \hspace{1cm} (5.31)

After obtaining a new estimate for the carrier phase, the received signal is de-rotated by the phase correcting term. The optimum values of two parameters $\gamma$ and $\rho$ need to be chosen to achieve the best performance.

**Bit Error Counting and EVM Calculation**

After carrier phase estimation, the received symbols are decoded into binary bits. Bit error counting is done using the Monte Carlo method for each polarization and the final BER is calculated as the average of the results for the two polarizations.

The EVM is defined as

$$EVM = \sqrt{\frac{1}{N} \sum_{k=1}^{N} [(I_k - \tilde{I}_k)^2 + (Q_k - \tilde{Q}_k)^2]}$$  \hspace{1cm} (5.32)

where $N$ is the total number of symbols. $\tilde{I}_k$ and $\tilde{Q}_k$ are in-phase and quadrature components of received symbols. $I_k$ and $Q_k$ are in-phase and quadrature components
of ideal constellation symbols. The final EVM is also calculated as the average EVM for the two polarizations.

5.4.3 Simulation Results

Figure 5.15 shows the simulated BER versus the bias voltage error which is defined as the difference between the actual bias voltage and the optimal value. When varying one of three bias voltages, the other two are set to their optimum values. For the DP-MZM, $V_1$ and $V_2$ represent the bias voltages applied to upper and lower inner MZMs and $V_3$ is the bias voltage for the phase shifter. The $V_\pi$ of the DP-MZM is 6.4 V. It is observed that at a BER of $10^{-2}$, for pre-compensation, the tolerance of bias voltage error for $V_1$ and $V_2$ is within $\pm 0.064$ V which is 0.01 $V_\pi$. For post-compensation, the tolerance of bias voltage error of $V_1$ and $V_2$ is over the range of $\pm 0.192$ V ($0.03 V_\pi$). However, the bias voltage error of $V_3$ can go up to $\pm 0.64$ V ($0.1 V_\pi$) for both pre- and post-compensation. Therefore, it is concluded that $V_1$ and $V_2$ are more critical than $V_3$ to ensure the transmission performance, especially for the pre-compensation case.

Figure 5.16 demonstrates the calculated dependence of EVM as a function of bias voltage error. As for the BER results, it leads to a conclusion that $V_1$ and $V_2$ influence the overall transmission performance more than $V_3$ and are more critical for pre-compensation than post-compensation.

Figure 5.17 shows constellation diagrams of PM-16QAM for the $X$ polarization when one of the three bias voltage errors is 0.064 V. It is found that when the bias voltage error for either $V_1$ or $V_2$ is 0.064 V, more severe signal distortions occur for pre-compensation (Figure 5.17 (a) and (b)) than post-compensation (Figure 5.17 (c) and (d)). Compared to $V_1$ and $V_2$, the same amount of bias voltage error in $V_3$ induces much smaller distortion.
Figure 5.15: Dependence of BER on the DP-MZM bias voltage error for dispersion pre- and post-compensation. $V_{\pi}$ of the DP-MZM is 6.4 V.
Figure 5.16: Dependence EVM on the DP-MZM bias voltage error for dispersion pre- and post-compensation. $V_x$ of the DP-MZM is 6.4 V.

**Parameters of Offline DSP Algorithms for Simulation Data**

Parameters used for the offline DSP algorithms are listed in Table 5.4. When using the second-order loop for carrier phase and frequency offset estimation, an optimization of $\gamma$ and $\rho$ is required. Using the simulated data for pre-compensation at 1200 km without inducing bias voltage error to $V_1$, $V_2$ and $V_3$ as an example, the optimization maps for $X$ and $Y$ polarization are shown in Figure 5.18 (a) and (b) respectively. There are 20 values of $\gamma$ and $\rho$ equally spaced within $[0.001, 0.1]$ with a step of 0.005. It is found that for both polarizations, the lowest $\log_{10}(\text{BER})$ occurs more than once meaning that the pair of $\gamma$ and $\rho$ which produces the best performance is not unique. For example, for $X$ polarization, the lowest $\log_{10}(\text{BER})$ is given by $\gamma = 0.056$ and $\rho = 0.016$ or $\gamma = 0.01$ and $\rho = 0.021$, etc. For $Y$ polarization, $\gamma = 0.071$ and $\rho = 0.031$ or $\gamma = 0.076$ and $\rho = 0.016$ provide lowest $\log_{10}(\text{BER})$ result with the same number
Figure 5.17: Constellation diagrams of 85.672 Gb/s PM-16QAM signal in X polarization with bias voltage error of 0.064V in (a) $V_1$ for pre-compensation, (b) $V_2$ for pre-compensation, (c) $V_1$ for post-compensation, (d) $V_2$ for post-compensation, (e) $V_3$ for pre-compensation, (f) $V_3$ for post-compensation. Red dots represent the ideal constellation symbols.
of errors. In this chapter, the optimum values of $\gamma$ and $\rho$ are always chosen to provide the best BER result for both simulation and experimental results.

Table 5.4: Parameters for offline DSP algorithms used for simulation data.

<table>
<thead>
<tr>
<th>Algorithms</th>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Dispersion post-compensation</td>
<td>Method</td>
<td>TDE</td>
</tr>
<tr>
<td></td>
<td>Number of FIR taps</td>
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</tr>
<tr>
<td>Adaptive equalizer</td>
<td>Method</td>
<td>RDE</td>
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<td></td>
<td>Convergence parameter</td>
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<td></td>
<td>Number of taps</td>
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</tr>
<tr>
<td></td>
<td>Number of iterations</td>
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</tr>
<tr>
<td>Second-order loop filter</td>
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<td>optimized</td>
</tr>
<tr>
<td></td>
<td>$\rho$</td>
<td>optimized</td>
</tr>
</tbody>
</table>

5.5 Recirculating Loop Experiment

A recirculating loop experiment was performed in order to evaluate the electronic pre-compensation performance on the 85.672 Gb/s PM-16QAM. The experiment was conducted by John Cartledge and John Downie at Corning Corporation.

Figure 5.19 shows the experimental configuration which was originally designed for thirty-seven channel transmission of 112 Gb/s PM-QPSK signals with 50 GHz spacing. Thirty-seven DFB lasers were modulated by a DP-MZM driven by I and Q data with $2^{15} - 1$ pseudo random bit sequence at 28 Gb/s for each and then polarization multiplexed to 112 Gb/s PM-QPSK signals. A 42.836 Gb/s 16QAM signal was generated from the AOWG using a DP-MZM driven by a 10.709 Gb/s $2^{16}$ deBruijn bit sequence and optically polarization multiplexed to produce the 85.672 Gb/s PM-16QAM signal. The 85.672 Gb/s PM-16QAM was multiplexed with the
Figure 5.18: Optimization map for the dependence of $\log_{10}(BER)$ on the parameters of second-order loop filter. (a) $X$ polarization, (b) $Y$ polarization.
thirty-seven 112 Gb/s PM-QPSK signals in the recirculating loop experiment. The PM-16QAM signal and nearest PM-QPSK signals were spaced by 100 GHz.

The loop consisted of 3 spans of Corning Vascade EX2000 fiber with 100 km for each span. Each fiber span was counter propagated with Raman pump signals at 1427 nm, 1443 nm and 1462 nm. The fiber had a dispersion of 19.359 ps/nm/km at 1550 nm and the average fiber attenuation was 0.162 dB/km. The gain of each Raman amplifier was set to fully compensate for the span loss. A loop synchronous polarization scrambler (LSPS) was used inside the loop to mitigate any periodic loop polarization effects. The Finisar Waveshaper filter was configured to flatten the signal spectrum for all channels. No optical dispersion compensation was used in the transmission system. All dispersion was either electronically pre-compensated in the transmitter or post-compensated in the digital coherent receiver.

At the end of the transmission, 50% of the optical signal was tapped off and sent to an EDFA followed by a VOA. After the channel selection filter (0.4 nm bandwidth), the signal was amplified and detected in a polarization- and phase-diverse digital coherent receiver combined with a free-running LO laser with nominal linewidth of 100 kHz, the same type of ECL used for the AOWG. Four electrical signal outputs from the balanced photodetectors were sampled and digitized at 50 GSa/s using a real-time oscilloscope with 20 GHz electrical bandwidth. Captured data consisted of 50,000 sample values corresponding to 10709 symbols. Data were stored at different moments and the BER was calculated as the average of the results for all data sets.
All experimental results shown in this chapter are based on offline processing. In order to process the experimental data, several additional DSP algorithms were required compared to the simulation in section 5.4.2:

- Quadrature imbalance compensation to correct for optical front-end errors, such as sampling skew between in-phase and quadrature components or phase errors in 90° hybrids [84].

- Retiming and resampling using digital square and filter clock recovery [85].

- Frequency offset estimation (FOE) [86].

- Correlation between transmitted and received bit sequences and bit error counting.
Although the second-order loop filter is able to deal with carrier phase and frequency offset estimation simultaneously, a separate FOE was used to increase the range of the frequency offset estimation. These DSP algorithms in Matlab implementation were provided by John Cartledge. A block diagram describing all offline DSP algorithms used for the experimental results is given in Figure 5.20. Corresponding parameters are listed in Table 5.5.

Figure 5.20: Block diagram of offline DSP algorithms used within digital coherent receiver.
Table 5.5: Parameters for offline DSP algorithms used for experimental data.

<table>
<thead>
<tr>
<th>Algorithms</th>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
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<tr>
<td>Dispersion post-compensation</td>
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<td>TDE</td>
</tr>
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<td></td>
<td>Number of taps</td>
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</tr>
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<td>Adaptive equalizer</td>
<td>Method</td>
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<td>Convergence parameter</td>
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<td></td>
<td>Number of iterations</td>
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<tr>
<td>Second-order loop filter</td>
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</tr>
<tr>
<td></td>
<td>$\rho$</td>
<td>optimized</td>
</tr>
</tbody>
</table>

### 5.6 Experimental Results

The back-to-back transmission performance was first investigated. The average output power of the ECL used for the AOWG was 7 dBm. Figure 5.21 plots the dependence of the averaged BER for 5 measured data sets versus OSNR (0.1 nm resolution bandwidth). An OSNR of 19.2 dB is required to achieve a BER of $10^{-3}$. Figure 5.22 shows the corresponding constellation diagrams obtained using one data set for the X and Y polarization signals at an OSNR of 23 dB.

The second investigation was the dispersion compensation performance. Figure 5.23 plots the averaged BER for 5 measured data sets as a function of transmission distance for dispersion pre- and post-compensation. The OSNR varied from 19.85 dB to 22.56 dB. The ECL with 7 dBm output power was used for post-compensation. The pre-compensated signal is no longer a four-level signal and only occasional peaks reach the highest and lowest voltage levels. This reduces the average signal power at the modulator output. Therefore, for the pre-compensation, the ECL output power was increased to 13 dBm, the maximum output power, to compensate for the power
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Figure 5.21: Dependence of BER versus OSNR of 85.672 Gb/s PM-16QAM for back-to-back transmission.

Figure 5.22: Constellation diagrams of 85.672 Gb/s PM-16QAM for back-to-back transmission at an OSNR of 24 dB, (a) X polarization, (b) Y polarization. Red dots represent the ideal constellation symbols.
reduction. The OSNR remained the same and varied from 19.75 dB to 22.68 dB since no additional optical amplifiers were used. With 7 dBm ECL output power for post-compensation and 13 dBm for pre-compensation, the fiber launch power was -7.5 dBm.

Theoretically, dispersion pre- and post-compensation should generate the same BER results. However, the experimental results show that the post-compensation performance is marginally better than pre-compensation. The performance difference might be attributed to the experimental error in BER results as well as the modulator bias voltage error. For all three biases ($V_1$, $V_2$ and $V_3$) of the DP-MZM, BER is more sensitive to $V_1$ and $V_2$ for pre-compensation than post-compensation.

Also included in Figure 5.23 is the dependence of the BER on transmission distance for post-compensation when the power of the ECL was increased to 8.5 dBm which lead the fiber launch power to be -6 dBm. The increase in the ECL power improved the OSNR which varied from 21.40 dB to 23.15 dB and therefore resulted in a lower BER. This leads to the third investigation of the impact of the ECL power on the performance of pre- and post-compensation.

The power reduction induced by the pre-compensation will affect the launch power into the fiber span and therefore the BER results. The dependence of averaged BER for 2 measured data sets on the ECL output power is presented in Figure 5.24. For post-compensation, a BER reduction is observed when the ECL power is increased from 7 dBm to 8 dBm since the increase in ECL power leads to an increase in fiber launch power and improves the OSNR. However, increasing the ECL power further results in a degradation of the BER performance due to SPM induced by the increased fiber launch power. Therefore, it is concluded that 8 dBm is an optimal value of ECL power for post-compensation. However, pre-compensation requires a higher ECL
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Figure 5.23: Dependence of BER versus transmission distance of 85.672 Gb/s PM-16QAM for dispersion pre- and post-compensation.

power. As can be seen from the Figure 5.24, the BER is decreased when the ECL output power is increased from 10 dBm to 12 dBm but it starts to degrade at 13 dBm. Since the maximum output power from ECL is limited to 13 dBm, the measurement was not continued. It is expected that the BER will get worse if the ECL power is increased further. Figure 5.25 shows the recovered constellation diagrams in X and Y polarizations for pre- and post-compensation with ECL output power of 12 dBm and 8 dBm respectively at 1500 km transmission using only one data set.

5.7 Summary

In this chapter, chromatic dispersion pre-compensation for 85.672 Gb/s PM-16QAM signals in a digital coherent transmission system has been demonstrated for the first time. Characterization of the AOWG ensures the quality of the 16QAM signal generation at the transmitter. A simulation investigation of the effect of modulator bias
voltage error was done using VPIphotonics software. Results indicate that the bias voltages for two inner MZMs of the DP-MZM have a larger impact than the bias voltage for the phase shifter.

In this chapter, dispersion pre-compensation for a 85.672 Gb/s PM-16QAM signal with coherent detection has been demonstrated and provides comparable results with dispersion post-compensation achieved with offline DSP. The implementation of dispersion pre-compensation allows to share the DSP requirements between the transmitter and receiver. The fixed FIR filter which consumes the most power [29] can be removed from the coherent receiver or the number of FIR filter taps can be reduced. Moreover, the pre-compensation enables pulse shaping at the transmitter and mitigation of nonlinear effects. Future work could consider combining the dispersion pre- and post-compensation for further performance improvement.

It has also been demonstrated that the dispersion pre-compensation requires much
Figure 5.25: Constellation diagrams of 85.672 Gb/s PM-16QAM signal at 1500 km transmission for (a) pre-compensation with a ECL output power of 12 dBm in $X$ polarization, (b) pre-compensation with a ECL output power of 12 dBm in $Y$ polarization, (c) post-compensation with a ECL output power of 8 dBm in $X$ polarization, (d) post-compensation with a ECL output power of 8 dBm in $Y$ polarization. Red dots represent the ideal constellation symbols.
higher ECL output power than post-compensation in order to obtain the same average optical power at the output of the modulator since the pre-compensation reduces the average signal power at the modulator output.
Chapter 6

Conclusions

6.1 Thesis Contributions

This thesis focuses on utilizing the electronic pre-compensation technique to deal with optical filtering and chromatic dispersion to improve transmission system performance.

Original contributions in the thesis include the following:

- Although some former investigations have been performed to deal with the narrowband optical filtering induced by multi-node signal transmission throughROADMs for intensity modulated signals, the electronic pre-compensation approach remains attractive since no change in the receiver configuration is required. Moreover, it has sufficient flexibility to enable transmission with not only intensity modulation but also phase modulation formats. The simultaneous pre-compensation for a narrow optical filter which has a 3 dB bandwidth of 9.13 GHz and fiber chromatic dispersion (1.3 ns/nm) has been accomplished
for 10 Gb/s NRZ-OOK and NRZ-DPSK as well as 20 Gb/s NRZ-DQPSK signals in a straight-line experiment. Results have been published in *Photonics Technology Letters* 2008 [87] and *Journal of Lightwave Technology* 2009 [88].

- A recirculating loop experiment with an optical filter with a 3 dB bandwidth of 28 GHz was performed to fully evaluate the pre-compensation performance under the influence of cascaded optical filtering for 10 Gb/s NRZ-OOK and NRZ-DPSK, and 20 Gb/s NRZ-DQPSK signals. Results demonstrate that the electronic pre-compensation is an effective approach for coping with narrow-band optical filtering. Results were published in *LEOS Summer Topic Meeting* 2009 [89] and *European Conference on Optical Communication* 2009 [90].

- Dispersion pre-compensation using a semiconductor InP MZM at 10.709 Gb/s has been reported for the first time. Instead of using advanced modulation formats and a conventional LiNbO$_3$ MZM, a combination of the NRZ-OOK modulation format and an InP MZM provides a cost effective approach for core and metro networks in which low cost and small size are crucial factors. With the pre-compensation using InP MZM, the maximum tolerable dispersion reaches 4920 ps/nm. Comparatively, if a conventional LiNbO$_3$ MZM is used, the pre-compensation allows the maximum allowable dispersion to be 5071 ps/nm. This demonstrates that the cost-efficient approach of using InP MZM for pre-compensation comes with only slight performance degradation. Results have been published in *Electronics Letters* 2011.

- The first demonstration of the dispersion pre-compensation on a 85.672 Gb/s PM-16QAM coherent transmission system has been achieved. The performance degradation induced by bias voltage error was investigated using VPIphotonics
CHAPTER 6. CONCLUSIONS

Simulation. Recirculating loop results show that the pre-compensation provides comparable results with post-compensation. Due to the properties of pre-compensated signals, the dispersion pre-compensation requires a much higher signal power at the modulator input compared to post-compensation in order to keep the same modulator output power. Results have been submitted to *Optical Fiber Communication Conference 2012*.

6.2 Future Work

6.2.1 Higher Bit Rate for Electronic Pre-Compensation

Optical signal transmission systems at 100-400 Gb/s per wavelength have been researched. Therefore, increasing the bit rate of pre-compensated systems is needed. This can be achieved by using multi-level formats such as 16QAM signal or lowering the sampling rate of the signal. The former has been discussed in this thesis. The latter has been demonstrated for the generation of 112 Gbits/s PM-16QAM signal using 1.5 samples per symbol with 21 GSa/s DACs [92]. Another solution is to upgrade the AOWG to a higher data rate by assembling DACs with a higher sampling rate as well as driver amplifiers and optical modulator operating at a higher frequency. DACs with sampling rate up to 25 GSa/s and 34 GSa/s are commercially available from Micram, Mitsubishi Electric Corporation has reported a 43 GSa/s DAC, and Ciena Corporation has demonstrated a 56 GSa/s DAC.

6.2.2 InP Photonic Devices at 40 Gb/s

The investigation utilizing InP photonic devices can be further extended. The availability of InP dual-parallel MZMs [93], [94] and InP-based DQPSK receivers [95] at
40 Gbit/s will benefit DQPSK transmission by reducing system level complexity with smaller footprint. Combining these with high-speed DACs will enable widespread deployment of pre-compensation for the DQPSK modulation format.

6.2.3 FPGA-Based Electronic Pre-Compensation

The electronic pre-compensation work presented in this thesis are all based on offline DSP. The required drive voltages are calculated offline and loaded into the DSP memory of the AOWG. However, this is not a real time approach. Recently, field programmable gate array (FPGA) implementations have attracted a lot of attention. Different from an ASIC, an FPGA allows the design to be translated into a configuration which is programmable and more flexible. The next stage of this work should consider using an FPGA to realize a more powerful real time implementation. The sampling rate for an FPGA implementation can go up to 28 GSa/s which allows a symbol rate of the signal to be 14 Gsymbol/s assuming 2 samples per symbol.
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