Transmission Performance of High Baud Superchannels in Optical Fiber Communications

by

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Abstract

In high baud fiber-optic communication systems, the bandwidth limitation at the transmitter and receiver, and the linear and nonlinear responses of electro-optical components, drive amplifiers and digital-to-analog converters can introduce pattern dependent distortion that caused the signal-quality degradation. Furthermore, the use of narrow subcarrier spacing leads to cross-phase modulation (XPM) distortion considered one of the main obstacles to achieving large transmission distances in superchannel systems. The primary objective of this research is to mitigate the effect of pattern dependent distortion and enhance the transmission performance of high baud superchannel systems.

Transmission of single-subcarrier 448 Gb/s dual-polarization 16-ary quadrature-amplitude modulation (DP 16-QAM) signal and a 1.206 Tb/s three-subcarrier superchannel signal using DP 16-QAM is demonstrated by applying a fixed look-up table (LUT) based on maximum-a-posteriori probability (MAP) detector at the receiver. Different decision rules for the MAP detector are considered based on multiple observations of the same symbol as the detection window advances through the received symbol sequence. Alternative approaches include a LUT based nonlinear pre-distorter (NLPD) at the transmitter, and a Volterra nonlinear equalizer (VNLE) or sparse-VNLE at the receiver. The LUT
based NLPD with iterative calculation of the pre-distortion provides the best performance in back-to-back systems and transmission systems followed by the MAP detector, VNLE and sparse-VNLE.

For a 1.206 Tb/s DP 16-QAM Nyquist-superchannel using 3, 5 and 9 subcarriers, the impact of the number of subcarriers on the relative strength of the intra- and inter-subcarrier nonlinearities and on the maximum achievable transmission distance for an information rate of 1 Tb/s is determined by simulation. The degree of correlation for the inter-subcarrier nonlinearity induced phase perturbation is assessed between a modulated center subcarrier and a continuous wave (CW) probe center subcarrier. The results provide an indication of the reliability of statistical information about the nonlinear phase perturbation that can be obtained from a probe signal either through simulation or experiment.
Acknowledgments

To my supervisor Prof. John C. Cartledge. Your exceptional supervision, technical knowledge, and enthusiasm for research have tremendously enriched me. I feel immensely fortunate to have had the opportunity to work with you. Words can not express the admiration and gratitude I feel toward you. Thank you.

To my beautiful wife, Dr. Nikoo Parvinnejad. I feel blessed to have you by my side. Thank you for your advice and devoted support. This achievement could not have been accomplished without your help.

To my parents. A thousands thanks for your endless support, guidance, care and love. Huge credit for this accomplishment goes to you.
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<th>Definition</th>
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<tr>
<td>ACF</td>
<td>auto-correlation function</td>
</tr>
<tr>
<td>ADC</td>
<td>analog-to-digital converter</td>
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<tr>
<td>ASE</td>
<td>amplified spontaneous emission</td>
</tr>
<tr>
<td>ASIC</td>
<td>application-specific integrated circuit</td>
</tr>
<tr>
<td>AWGN</td>
<td>additive white Gaussian noise</td>
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<tr>
<td>BER</td>
<td>bit error ratio</td>
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<tr>
<td>BPG</td>
<td>bit pattern generator</td>
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<td>BPS</td>
<td>blind phase search</td>
</tr>
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<td>CD</td>
<td>chromatic dispersion</td>
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<tr>
<td>CMA</td>
<td>constant modulus algorithm</td>
</tr>
<tr>
<td>CMOS</td>
<td>complementary metal-oxide-semiconductor</td>
</tr>
<tr>
<td>CPR</td>
<td>carrier phase recovery</td>
</tr>
<tr>
<td>CW</td>
<td>continuous wave</td>
</tr>
<tr>
<td>DAC</td>
<td>digital-to-analog converter</td>
</tr>
<tr>
<td>DP</td>
<td>dual polarization</td>
</tr>
<tr>
<td>DSP</td>
<td>digital signal processing</td>
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<tr>
<td>EDFA</td>
<td>erbium doped fiber amplifier</td>
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<tr>
<td>ENOB</td>
<td>effective number of bit</td>
</tr>
<tr>
<td>ETDM</td>
<td>electrical time division multiplexing</td>
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</tbody>
</table>
FDE    frequency domain equalizer
FEC    forward error correction
FFT    fast Fourier transform
FIR    finite impulse response
FOE    frequency offset estimation
FWM    four wave mixing
GE     gain equalizer
GVD    group velocity dispersion
HD     hard-decision
IFWM   intra-channel four-wave mixing
IIR    infinite impulse response
ISI    inter-symbol interference
IXPM   intra-channel cross-phase modulation
LMS    leas mean square
LO     local oscillator
LPN    linear phase noise
LSPS   loop synchronous polarization scrambler
LUT    look-up table
MAP    maximum-a-posteriori probability
MI     mutual information
MMA    multi modulus algorithm
NLPD   nonlinear pre-distortion
NLSE   nonlinear Schrödinger equation
OADM   optical add-drop multiplexer
<table>
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<tr>
<th>Abbreviation</th>
<th>Description</th>
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<tr>
<td>OFDM</td>
<td>orthogonal frequency division multiplexing</td>
</tr>
<tr>
<td>OSA</td>
<td>optical spectrum analyzer</td>
</tr>
<tr>
<td>OSNR</td>
<td>optical signal-to-noise ratio</td>
</tr>
<tr>
<td>OTDM</td>
<td>optical time division multiplexing</td>
</tr>
<tr>
<td>PBC</td>
<td>polarization beam combiner</td>
</tr>
<tr>
<td>PBS</td>
<td>polarization beam splitter</td>
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<tr>
<td>PD</td>
<td>photo diode</td>
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<tr>
<td>PDF</td>
<td>probability density function</td>
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<td>PLL</td>
<td>phase-locked loop</td>
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<tr>
<td>PMD</td>
<td>polarization mode dispersion</td>
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<tr>
<td>POF</td>
<td>programmable optical filter</td>
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<tr>
<td>PRBS</td>
<td>pseudorandom bit sequence</td>
</tr>
<tr>
<td>PS</td>
<td>polarization synthesizer</td>
</tr>
<tr>
<td>PSK</td>
<td>phase-shift-keying</td>
</tr>
<tr>
<td>QAM</td>
<td>quadrature-amplitude modulation</td>
</tr>
<tr>
<td>QPSK</td>
<td>quadrature-phase-shift-keying</td>
</tr>
<tr>
<td>RC</td>
<td>raised cosine</td>
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<tr>
<td>RDA</td>
<td>radius directed algorithm</td>
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<tr>
<td>RLS</td>
<td>recursive least square</td>
</tr>
<tr>
<td>RS</td>
<td>Reed-solomon</td>
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<tr>
<td>SD</td>
<td>soft-decision</td>
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<tr>
<td>SMF</td>
<td>single mode fiber</td>
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<tr>
<td>SNR</td>
<td>signal-to-noise ratio</td>
</tr>
<tr>
<td>SPM</td>
<td>self-phase modulation</td>
</tr>
<tr>
<td>Acronym</td>
<td>Description</td>
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<tr>
<td>TDE</td>
<td>time domain equalizer</td>
</tr>
<tr>
<td>TED</td>
<td>time error detector</td>
</tr>
<tr>
<td>TIA</td>
<td>trans-impedance amplifier</td>
</tr>
<tr>
<td>VNLE</td>
<td>Volterra nonlinear equalizer</td>
</tr>
<tr>
<td>VOA</td>
<td>variable optical attenuator</td>
</tr>
<tr>
<td>WDM</td>
<td>wavelength division multiplexing</td>
</tr>
<tr>
<td>WSS</td>
<td>wavelength selective switch</td>
</tr>
<tr>
<td>XPM</td>
<td>cross-phase modulation</td>
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<tr>
<td>XPOLM</td>
<td>cross-polarization modulation</td>
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Chapter 1

Introduction

The continuous expansion of multimedia terminals, like smart-phones, besides the introduction of new multimedia services, like high-definition streaming television, determines a constant increase in the high-speed connectivity demand [1–3]. In fact, the most important novelty in today’s telecommunication networks is the predominance of data traffic originated from multimedia terminals [1]. Some Global traffic forecasting studies indicate that data traffic will be constantly growing. The overall mobile data traffic is expected to grow to 30.6 exabytes per month by 2020, a fivefold increase over 2016 [2]. Therefore, rules-update in the telecommunication networks design is necessary to have a suitable quality of service for all end-user applications. Optical fiber communication systems have played an important role in the evolution and development of modern communication networks. Due to their low cost, low attenuation and high bandwidth, optical fibers have dominated the area of long-haul terrestrial and transoceanic transmission [3]. Nevertheless, a proper capacity upgrades plan is required for back-bone optical networks in order to avoid possible network congestion.
The achievable information capacity and transmission distance for a single optical fiber has seen a substantial increase with the deployment of wavelength division multiplexing (WDM) systems and erbium doped fiber amplifiers (EDFAs). WDM allows up to 100 optical channels in the C-band (1528.71 - 1568.36 nm) with different wavelengths to propagate simultaneously through the same fiber. Typically, each individual channel travels a different path through the network which can be added, dropped or routed in a reconfigurable manner depending on the network requirements. Long haul transmission systems are required to operate at a high bit rate per WDM channel to fully utilized the installed fiber resource. Today, 100 Gb/s single channel systems are being deployed for long haul transmission. The next generation of long-haul WDM transmission systems are expected to operate at 400 Gb/s or 1 Tb/s per WDM wavelength.

Such high data rates are difficult to achieve without finding new technological paths. Increasing the symbol rate and encoding more bits per symbol are the most common approaches to increase the data rate. However, several challenges make their practical implementation difficult. High symbol rate signals can exhibit a pattern-dependent distortion due to the bandwidth limitations and nonlinear responses of electronic amplifiers and opto-electronic devices at the transmitter and receiver. On the other hand, high spectral efficiency modulation formats require very high optical signal to noise ratios (OSNRs) which strongly increase the sensitivity to fiber nonlinear effects. Due to nonlinear effects, sending 1 Tb/s over a single- or multi-carrier with high symbol rates and high modulation formats can reach only a few hundred kilometers if only linear processing is carried out which is not the desired transmission reach in the context of long-haul transmission. As
detection of the data signal becomes quite a challenging task in the presence of these impairments, it is desired to find ways to eliminate or at least mitigate these distortions.

Coherent detection is very beneficial within the design of high-order modulation systems, because all the optical field parameters (amplitude, phase frequency and polarization) are available in the electrical domain [4]. The preservation of the temporal phase enables more effective methods for the electronic compensation of transmission impairments like fiber nonlinearities [5]. Recently, a new way for using a WDM wavelength has been proposed, called superchannel, defined as a group of modulated optical subcarriers that are treated as a single transport entity. By superchannel technique, a number of subcarriers are seamlessly aggregated to form individual superchannels routed optically through a network as a single channel. Superchannel systems combined with multi-level modulation are considered promising candidates to achieve terabit per second bit rates in terms of practical implementation and performance [4–7].

This research is motivated by extensive demand for higher spectral efficiency for long-haul optical communication links in which transition into higher baud and modulation orders requires effective mitigation of system impairments. Two main topics have been focused on to assess the transmission performance of a 448 Gb/s single-carrier using dual-polarization 16-ary quadrature-amplitude-modulation (DP 16-QAM) and 1.206 Tb/s multi-subcarrier DP 16-QAM superchannel signal. More in details, the transmission performance has been evaluated after applying a nonlinear pre-distortion (NLPD) at the transmitter, a fixed look-up table (LUT) based on a maximum-a-posterior probability (MAP) detection and Volterra
nonlinear equalizer (VNLE). Furthermore, a simulative study has been performed to investigate the impact of the number of subcarriers on the relative strength of the intra- and inter-subcarrier nonlinearities in superchannel and to indicate the reliability of statistical information about the inter-subcarrier nonlinear phase perturbation on a modulated subcarrier and a continuous wave (CW) probe.

1.1 Principle of High Baud Superchannels

A superchannel is comprised of several optical subcarriers considered as a single entity with the subcarriers being routed through optical add-drop multiplexers (OADMs) and wavelength selective switches (WSSs) together. A pictorial description of the superchannel concept is displayed in Figure 1.1 [8].

![Figure 1.1: A pictorial description of superchannel concept for a comb of superchannels separated by a guard band and individual superchannel made of 9 subcarriers.](image)

Creating a superchannel by tightly packing subcarriers is a possible approach by
applying coherent optical orthogonal frequency division multiplexing (CO-OFDM) or alternatively using symbol-rate spacing of subcarriers called Nyquist-WDM. Despite the similarity of multi-carrier signal generation between CO-OFDM and Nyquist-WDM, there are some differences with respect to their operation principles and capabilities [7]. By the Nyquist-WDM approach, the spectral utilization between each subcarrier is minimized. Nyquist-WDM can be generated either by digital filtering as a part of the digital signal processing (DSP) block at the transmitter or in the optical domain by optical filtering [7]. Nyquist-WDM is quite a promising method to generate future superchannels with bit rates beyond 400 Gb/s. Systems with a bit rate of 1 Tb/s are being investigated as the next Ethernet standards [6].

Recently there have been many demonstration of terabit per second transmission systems based on coherent detection, DSP and various advanced modulation techniques [8–23]. These papers applied the state-of-the-art technology advancements in some key areas of optical communications, such as the recent availability of high-sample-rate analog-to-digital converters (ADCs) and digital-to-analog converters (DACs) technologies to generate superchannel signal.

With a view to possible transmission product development of terabit per second superchannel systems in the near future, it is anticipated that significant efforts is required to reduce the pattern-dependent distortion, mitigate nonlinear fiber effects and simplify the associated DSP chips.

Recently, for high baud superchannel experiments steps are taken to minimize the distortion of the transmitted signal and to mitigate the distortion that does exist [10, 12–16]. Moreover, it has been shown that applying more number of subcarriers
1.2. CONTRIBUTIONS

to generate terabit per second superchannel leads to increase the inter-subcarrier nonlinearities which is a limiting factor during fiber transmission in superchannel systems [24–34].

1.2  Contributions

The following publications are the outcome of the research presented in this thesis. For all journal and conference papers, the idea was conceived by my supervisor and myself. I performed the experiment or simulation and wrote the paper. The co-authors contributed in editing the paper and discussing the idea.

1.2.1  Journal Publications


1.2.2  Conference Publications

1.2. CONTRIBUTIONS

The main contributions of this PhD research are listed as follows:


1.2.3 Original Contributions

The main contributions of this PhD research are listed as follows:
16-QAM superchannel signal. These experimental results are reported in the following publications [C5], [C4], [C3].

• We manage to provide a novel solution for different decision rules of the MAP detector. We show that the multiple observations of the same symbol as the detection window advances through the received symbol sequence can be used to update the LUT entries and improve the performance. Transmission of single-carrier 448 Gb/s DP 16-QAM electrical time-division multiplexing (ETDM) signal and a 1.206 Tb/s three-carrier superchannel ETDM signal using DP 16-QAM is demonstrated by applying novel LUT based MAP detection algorithms. Experimental results are reported in [J2] and the proposed algorithms discussed in Chapter 3.

• Using experimental results, we demonstrate that pattern-dependent distortion compensation based on NLPD at the transmitter can outperform MAP detector and time-domain VNLE at the receiver. The NLPD is superior in terms of performance and computational complexity. The proposed solutions and results are presented in [J1] and discussed in Chapter 3 for a 1.206 Tb/s DP 16-QAM superchannel signal.

• There has been a lot of research on splitting the high baud single-carrier signal into multiple low baud subcarriers and optimizing the number of subcarriers to improve the nonlinearity tolerance. In this thesis, we determine the impact of the number of subcarriers on the intra- and inter-subcarrier nonlinearities on the maximum achievable transmission distance for an information rate of 1 Tb/s. The simulation results are presented in [C2] and discussed in Chapter 4.
1.3. ORGANIZATION OF THESIS

- For a 1.206 Tb/s dual-polarization 16-QAM superchannel, we assess the degree of correlation between a modulated center subcarrier and a CW probe center subcarrier for the inter-subcarrier nonlinearity induced phase perturbation. Simulation results are published in [C1] and discussed in Chapter 4.

1.3 Organization of Thesis

An overview of the thesis is summarized in Table 1.2.

Table 1.1: Thesis overview

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<th>CHAPTER</th>
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<th>RESULTS</th>
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<td>Chapter 2</td>
<td>Literature background and relevant concepts</td>
<td>————</td>
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<tr>
<td>Chapter 3</td>
<td>Compensation schemes for transmitter and receiver based pattern-dependent distortion in high baud modulation systems(^1)</td>
<td>Experiment</td>
</tr>
<tr>
<td>Chapter 4</td>
<td>Impact of the number of subcarriers on the intra- and inter-subcarrier nonlinearities on the maximum achievable transmission distance and modeling of XPM-induced distortion for 1 Tb/s Nyqyist-superchannel(^2)</td>
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\(^1\) Chapter 3 is based on papers J1, J2, C3, C4 and C5. The IEEE owns the copyrighted material which is used with permission in this thesis.

\(^2\) Chapter 4 is based on papers C1 and C2. The IEEE owns the copyrighted material which is used with permission in this thesis.
Chapter 2

Literature Background and Relevant Concepts

2.1 Fiber Optic Impairments

In fiber optic communication systems, linear impairments are due to fiber loss, chromatic dispersion (CD) and polarization mode dispersion (PMD). Optical power-loss due to light propagation inside a fiber results from absorption and scattering and it can be easily compensated by optical amplifiers. CD and PMD are the main linear impairments for optical communication systems.

Nonlinear distortion is an important source of signal degradation in optical communication systems using advance modulation formats and a high baud, because of the limit placed on the maximum power that can be launched into a fiber and thus the achievable optical signal to noise ratio (OSNR) at the receiver. From a system requirements point of view, this limitation restricts the available OSNR to a lower value as opposed to an ideal case without nonlinear effects. This yields a
limitation in system margin and maximum achievable propagation distance. Nonlinear impairments result from the power dependent refractive index, also referred to as the optical Kerr effect [35]. The optical Kerr effect can be classified in terms of intra-channel and inter-channel nonlinearities. The former defines the nonlinear effects in a single wavelength channel and the latter describes the nonlinear interactions between neighbouring wavelength channels. The intra-channel effects are categorized into self-phase modulation (SPM), intra-channel cross-phase modulation (IXPM), and intra-channel four-wave mixing (IFWM). Moreover, the inter-channel effects are classified into cross phase-modulation (XPM), and four-wave mixing (FWM). Additionally, in dual polarization systems, Kerr nonlinearity couples the polarizations of the same or neighbouring channels together denoted as cross-polarization modulation (XPolM).

2.1.1 Attenuation

Light traveling in an optical fiber loses power over distance which is mainly caused by material absorption and Rayleigh scattering [35]. The former is affected by the level of impurity in the fiber. The latter is caused by local fluctuation of the refractive index. Attenuation in fiber optic systems is shown with $\alpha$ measured in dB/km. If $P$ (W) is the launch power at the input of a fiber, the power of the optical signal $P(z)$ decreases exponentially during the propagation

$$P(z) = P \exp(-\alpha z) \quad (2.1)$$
2.1. FIBER OPTIC IMPAIRMENTS

In the wavelength range around 1550 nm corresponding to the C-band, the attenuation is around 0.2 dB/km which is the minimum loss and one reason why WDM channels are distributed around this wavelength. The total attenuation of a 75 km fiber is 15 dB. Therefore, re-amplification of the signal is necessary.

2.1.2 Chromatic Dispersion

Group velocity dispersion (GVD) or chromatic dispersion is caused by a combination of waveguide and material dispersion. Waveguide dispersion is the result of wavelength-dependence of the propagation constant of the optical waveguide. It is important in single-mode waveguides. The larger the wavelength, the more the fundamental mode will spread from the core into the cladding. Waveguide dispersion can be controlled by careful design [35]. Material refractive index of the fiber varies with wavelength and therefore causes the group velocity to vary. It is classified as material dispersion. Indeed, the spectral components of the modulated signals travel at a different speeds in the fiber which causes some wavelengths to arrive before others and therefore make the signal pulse to broaden. As a result short pulses become longer which leads to significant inter-symbol interference (ISI) and degradation of the system performance.

Mathematically, the effects of fiber dispersion are accounted for expanding the mode-propagation constant $\beta(\omega)$ in a Taylor series around the carrier frequency $\omega_0$ at which the pulse spectrum is centered [35]

$$\beta(\omega) = \beta_0 + \beta_1(\omega - \omega_0) + \frac{1}{2}\beta_2(\omega - \omega_0)^2 + \frac{1}{6}\beta_3(\omega - \omega_0)^3 + \cdots$$  \hspace{1cm} (2.2)
The constant $\beta_0 = \beta(\omega_0)$ and

$$
\beta_m = \left( \frac{d^m \beta(\omega)}{d\omega^m} \right)_{\omega = \omega_0} \quad (m = 1, 2, \cdots) \quad (2.3)
$$

$\beta_1 = 1/\vartheta_g$ defines the group velocity $\vartheta_g$ of the signal. Here, terms with $m \geq 3$ can be neglected assuming the so called slowly varying envelope approximation of the electrical field, since the optical signal bandwidth is considerably less than the carrier frequency $\omega_0$ [35]. Of particular interest is $\beta_2$, which is known as the group velocity dispersion (GVD) and is responsible for pulse broadening. Conventionally, the dispersion parameter $D$ is defined in terms of $\beta_2$ as [35]

$$
D = -\frac{2\pi c}{\lambda^2} \beta_2 \quad (2.4)
$$

where $c$ is the speed of light in vacuum and $\lambda$ is the carrier wavelength. From 2.4, it can be understood that $D$ has the opposite sign of $\beta_2$. If $\beta_2 > 0$ (or $D < 0$), an optical signal exhibits normal dispersion. In the normal dispersion regime high-frequency components of optical signal travel slower than low-frequency components. On the contrary, when $\beta_2 < 0$ (or $D > 0$), an optical signal exhibits anomalous dispersion. In the anomalous dispersion regime high-frequency components of optical signal travel faster than low-frequency components [35]. The anomalous dispersion regime has a considerable interest for the study of nonlinear effects, because it is in this regime that optical fibers support solitons through a balance between the dispersive and nonlinear effects. The impact of GVD can be conventionally described using the dispersion length, $L_D = \frac{T_0^2}{\beta_2}$ where $T_0$ is the temporal pulse width [35]. These lengths provide a scale over which the dispersive effect becomes significant for pulse
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Evolution along a fiber. The GVD plays an important role in signal quality over fiber transmission. The interaction between dispersion and nonlinearity is an important issue in lightwave system design.

2.1.3 Kerr Effect

The refractive index of silica is power dependent and it increases with intensity. The anharmonic oscillations of the electrons in response to a high intensity electromagnetic field is the physical reason for this effect. By including the power dependency of the refractive index, the nonlinear refractive index is defined as [35]

\[
n(\omega, P) = n_0(\omega) + n_2 \frac{P}{A_{eff}}
\]  

where \( n_0 \) is the linear refractive index, \( n_2 \) is the Kerr coefficient with typical value of \( 2 \times 3.4 \times 10^{-20} \text{ (m}^2/\text{W}) \), \( P \) is the optical power, and \( A_{eff} \) is the effective core area. In spite of the intrinsically small values of the nonlinear coefficients in silica, the nonlinear effects in optical fibers can be observed at relatively low power levels. This is possible because of two important characteristics of SMF: (i) a small effective core area and (ii) extremely low loss (< 0.2 dB/km). The dependence of the refractive index on the light intensity results in the propagation constant, \( \beta \), varying due to \( \beta = \frac{2\pi n}{\lambda} \), and the propagation constant can be written as [35]

\[
\beta(\omega, P) = \beta_0(\omega) + \frac{2\pi n_2}{\lambda A_{eff}} P
\]  

(2.6)
where $\beta_0(\omega)$ is the propagation constant in the absence of nonlinear effects, and

$$\gamma = \frac{2\pi n_2}{\lambda A_{eff}}$$  \hspace{1cm} (2.7)

is known as the nonlinear coefficient. The total nonlinear phase shift due to the Kerr effect after the distance $L$ is given by

$$\phi_{NL} = \int_0^L [\beta - \beta_0] \, dz$$  \hspace{1cm} (2.8)

By substituting equation 2.6 in 2.8, and considering equation 2.1 for attenuation along the fiber, the total nonlinear phase shift is defined as [36]

$$\phi_{NL} = \gamma P \int_0^L e^{-\alpha z} \, dz = \gamma P \frac{1 - e^{-\alpha L}}{\alpha} = \frac{L_{eff}}{L_{NL}}$$  \hspace{1cm} (2.9)

where $L_{eff}$ and $L_{NL}$ are called the effective length and the nonlinear length of the fiber, respectively.

$$L_{eff} = \frac{1 - e^{-\alpha L}}{\alpha}, \quad L_{NL} = \frac{1}{\gamma P}$$  \hspace{1cm} (2.10)

Physically, the nonlinear length, $L_{NL}$, indicates the distance at which the nonlinear phase shift reaches 1 radian, and it provides a length scale over which the nonlinear effects become relevant for optical fibers. It can be seen from equation 2.10 that the fiber nonlinear effect enhances when $L_{NL}$ decreases, or equivalently the power $P$ increases. There are three types of fiber nonlinearities due to the Kerr effect: (i) self phase modulation (SPM) (ii) cross phase modulation (XPM), and (iii) four wave
mixing (FWM).

SPM refers to the self-induced power-dependent phase shift experienced by an optical field during propagation in the optical fiber as shown in equation 2.9, and it is responsible for the spectral broadening of optical pulses under certain conditions. The interaction with fiber dispersion can cause temporal pulse broadening in the normal dispersion regime, or pulse compression in the anomalous dispersion regime. The overlap of adjacent pulses in the same wavelength channel, due to the broadening of the pulses as a result of fiber dispersion, leads to the nonlinear interaction between the pulses known as intra-channel cross-phase modulation (IXPM). The process of IXPM is the change of the phase of one pulse due to the variation of the power of other pulses that are overlapping in time. The shift in the phase of pulses in the channel of interest results in a change in its propagation velocity and consequently timing jitter [37]. The effect of IXPM becomes small if the overlapping pulse power is low due to a large amount of broadening. However, in this case, another nonlinear effect called intra-channel four-wave mixing (IFWM) occurs, which causes amplitude jitter. As a result of IFWM, when a pulse has large broadening due to dispersion, echo pulses in the time domain appear in the neighbouring slots and generate ghost pulses. In fact, IFWM causes the transference of energy among the interacting pulses. There are various approaches to compensate the effect of timing jitter (due to the IXPM), and amplitude jitter (due to IFWM). It was analytically shown in [38] that complete cancellation of the amplitude and timing jitter can be achieved by dividing dispersion compensation equally between the input and output of a link. This means that a symmetric dispersion map should apply about the center of the link. By means of data dependent coding, a scheme for mitigating IFWM and amplitude jitter was
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proposed in [39].

In addition to the intra-channel nonlinear effects, inter-channel nonlinear effects in the WDM systems are important limiting factors. Using the WDM technique, when two or more wavelength channels are transmitted simultaneously inside the fiber, the power of the neighbouring channels causes a nonlinear phase shift for the channel of interest. This effect is known as cross-phase modulation (XPM). The XPM effects are quite important for WDM lightwave systems since the phase of each optical channel is affected by both the average powers and bit patterns of all other channels. In WDM systems with \( N \) channels, the nonlinear phase shift in the \( k \)-th channel can be written as [36]

\[
\phi_{NL} = \gamma L_{eff} P^{(k)} + 2 \sum_{h=1,h\neq k}^{N} \gamma L_{eff} P^{(h)} \tag{2.11}
\]

where \( P^{(k)} \) denotes the peak power in the \( k \)-th channel. The first term is the SPM and the second term denotes the contribution of XPM. In deriving equation 2.11, \( P^{(k)} \) was assumed to be constant. In particular, time dependence of \( P^{(k)} \) makes \( \phi_{NL} \) vary with time. In fact, the nonlinear optical phase shift changes with time in exactly the same fashion as the optical pulse due to SPM. It can be seen from equation 2.11 that the XPM induced phase shift is twice that SPM when the optical powers of the channels are equal. XPM causes asymmetric spectral broadening of optical pulses, timing jitter and amplitude distortion in the time domain.

In WDM transmission, the nonlinear interaction between light at different frequencies results in the generation of signals at new frequencies termed FWM. When optical fields with \( \omega_i, \omega_j \) and \( \omega_k \) carrier frequencies co-propagate in the fiber, several frequencies \( \omega_h = \omega_i \pm \omega_j \pm \omega_k \) corresponding to different combinations of
these frequencies can be generated [35]. However, not all these combinations have the phase matching requirement. In other words, significant FWM occurs only when the phase matching requirement is satisfied.

In dual polarization systems, the temporal and spatial evolution of a pulse as it propagates through the dispersive fiber medium can be described by the Manakov equation [36]

\[
\frac{\partial A_p(z,t)}{\partial z} + \frac{\alpha}{2} A_p(z,t) + \beta_1 \frac{\partial A_p(z,t)}{\partial t} + j \frac{\beta_2}{2} \frac{\partial^2 A_p(z,t)}{\partial t^2} = j \gamma \left( |A_p(z,t)|^2 + \frac{2}{3} |A_q(z,t)|^2 \right) A_p(z,t)
\]

(2.12)

where \( p = x, y \) and \( q = x, y \) such that \( p \neq q \), \( A_x \) and \( A_y \) are the slowly varying envelopes of the electrical field in two orthogonal polarization tributaries.

In order to emphasize different nonlinear effects, a system with two channels, namely, \( A_1 \exp(j(\omega_1 t - \beta(\omega_1)z)) \) and \( A_2 \exp(j(\omega_2 t - \beta(\omega_2)z)) \), is considered. Assume the two channels have the same state of polarization (SOP). In this case, equation 2.12 can be expanded [36]

\[
\frac{\partial A_{1p}}{\partial z} + \frac{\alpha_1}{2} A_{1p} + j \beta_{21} \frac{\partial^2 A_{1p}}{\partial t^2} = j \gamma_1 \left( |A_{1p}|^2 + \frac{2}{3} |A_{1q}|^2 \right) A_{1p} \\
+ (2|A_{2p}|^2 + \frac{2}{3} |A_{2q}|^2) A_{1p} + A_{2p} A_{1q}^* A_{1q}
\]

(2.13)

\[
\frac{\partial A_{2p}}{\partial z} + \frac{\alpha_2}{2} A_{2p} + \frac{\alpha_2}{2} \frac{\partial A_{2p}}{\partial T} + j \beta_{22} \frac{\partial^2 A_{2p}}{\partial T^2} = j \gamma_2 \left( |A_{2p}|^2 + \frac{2}{3} |A_{2q}|^2 \right) A_{2p} \\
+ (2|A_{1p}|^2 + \frac{2}{3} |A_{1q}|^2) A_{2p} + A_{1p} A_{1q}^* A_{2q}
\]

(2.14)
where

\[ T = t - \frac{z}{\vartheta_{g1}} \quad \text{and} \quad d = \frac{\vartheta_{g1} - \vartheta_{g2}}{\vartheta_{g1}\vartheta_{g2}} \]  

(2.15)

Time \( T \) is the reference time frame based on propagation normalized by \( \vartheta_{g1} \). The parameter \( d \) is a measure of the group-velocity mismatch between the two channels. The first terms of the right-hand side of equations 2.13 and 2.14 represent SPM. In addition to the XPM contribution, it can also be inferred from the second term that the cross-channel interaction in the case of co-polarized channels is twice that in the case of orthogonally polarized channels. XPolM is also embedded in the second term. Finally, the last term expresses FWM.

The system tolerance to fiber nonlinear effects depends on many elements of the system configuration. In addition to the link engineering, factors such as the modulation format, symbol rate, and channel spacing are of major importance. Being the main obstacle towards capacity upgrade, mitigation techniques for fiber nonlinear effects are of major interest. The main challenge for compensation of inter-channel impairments is the accessibility to each optical field amongst WDM coherent systems. Due to current limitations of application-specific integrated circuit (ASIC) technology, sharing field information of multiple channels with precise timing between several coherent receivers is almost impossible. So far, no commercial product has been released which truly compensates inter-channel nonlinear impairments. However, with the introduction of elastic networks that create the generation and transmission of densely packed subcarriers in the form of superchannels, the destructive effects of neighbouring channels would be more severe compared to conventional WDM systems [40].
2.2 Relevant Concepts

In this section, relevant concepts appearing throughout the thesis are described.

2.2.1 System Configuration

Figure 2.1 shows a canonical model of a digital coherent optical system. At the transmitter, a digital signal processor converts input symbols into four digitized waveforms corresponding to the in-phase (I) and quadrature (Q) components of the two transmitted polarizations. These are converted to the analog domain by digital-to-analog converters (DACs). The DAC output signals are applied to two optical transmitters including drive amplifiers and an IQ-modulator for performing electrical-to-optical up-conversion. The two polarizations are combined orthogonally by a polarization beam combiner (PBC) and then the dual-polarized modulated signal is transmitted over a channel consisting of multiple spans of SMF with amplifiers.

At the receiver, a coherent optical front-end consists of polarization-beam splitters (PBS) in order to split the received signal as well as the local oscillator (LO) light into two orthogonal states of polarization, and two 90° optical hybrids and eight photodiodes (PD) to provide balanced detection of the superposition of the received signal and LO. Four linear trans-impedance amplifiers (TIA) deliver the required signals to analog-to-digital converters (ADCs) and digital signal processing (DSP) for further processing.
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2.2.2 Raised Cosine Pulse Shaping

Throughout this thesis, raised cosine (RC) pulse shaping and filtering is used in both experiment and simulation setups. The RC pulse shape offers the advantage of being strictly band-limited, with flexible bandwidth control through adjusting the roll-off factor. Furthermore, RC pulses do not experience any inter-symbol-interference due to the presence of zero crossings at the center of neighbouring symbols. An RC pulse shape is mathematically described in the time domain

\[
h(t) = \text{sinc}(t/T) \frac{\cos(\frac{\pi \beta t}{T})}{1 - \frac{4\beta^2 t^2}{T^2}} \tag{2.16}\]
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where $T$ is the symbol period and $\beta$ is the roll-off factor. In the frequency domain the RC pulse shape is described by

$$H(f) = \begin{cases} T, & |f| \leq \frac{1-\beta}{2T} \\ \frac{T}{2} [1 + \cos(\frac{\pi T}{\beta} [|f| - \frac{1-\beta}{2T}])], & \frac{1-\beta}{2T} < |f| \leq \frac{1+\beta}{2T} \\ 0, & |f| \geq \frac{1+\beta}{2T} \end{cases}$$ \hspace{1cm} (2.17)

The roll-off factor can vary between $\beta = 0$, known as the Nyquist pulse shape (rectangular shape), and $\beta = 1$. The flexibility to control the pulse shape is particularly important for spectrally efficient modulation formats [41]. The sampling theorem requires that the sampling rate be at least twice a signal bandwidth. For a DAC sampling rate, the achievable symbol rate can be increased by choosing a pulse shape that reduces the modulated signal bandwidth.

In practical systems, the transfer functions of physical filters are mainly established by manufacturing tolerances and parasitic behaviours of devices. Figure 2.2 shows the filtering of optical signals for generating a sinc-shape Nyquist pulse.

Generally, the Nyquist filter can be implemented optically (Figure 2.2(a)), or digitally (Figure 2.2(b)) [7]. An optical Nyquist-filter can be applied to carve an ideal rectangle spectrum out of a non-return to zero (NRZ) signal. However, the optical filters are quite complicated compared to the digital filters [7]. At symbol rates of interest in optical communications, the development of high-speed DACs and ADCs with a sufficient number of bits is fundamental to implement effective Nyquist-WDM systems. Doing filtering digitally (DSP at the transmitter) is more flexible compared to the optical domain [7]. Digital filtering requires an analog anti-aliasing filter to remove spectral image. This filter can be standard low-pass filters. Furthermore, only
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DSP offers the flexibility to vary the filter coefficients during run-time and therefore the capability to adapt to changes in the channel response. The system advantage of the DSP-based Nyquist filtering at the transmitter is evident, and it leads to the design of superchannel systems with very tight subcarrier spacing.

Figure 2.2: Filtering solution for Nyquist transmitter (a) Optical (b) Digital.
2.2.3 Digital Coherent Receiver

As the transmission capacity increased in WDM systems, coherent detection began to attract substantial interest during recent years and is now widely deployed in commercial systems. 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 by a digital signal processor. Real-time signal processing is necessary in a commercial product since the payload data in a high-speed optical signal transmission system are always continuous. However, the real-time hardware implementation is limited by the availability of circuit technology to implement ADC with high resolution and high sampling rate. Moreover, a DSP unit with high integrability and lower power requirement is required.

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, offline DSP is especially important in research environments. Table 2.1 shows a possible sequence of the processing functions applied by DSP in a coherent receiver.
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Table 2.1: The sequence of the processing functions by DSP in a coherent receiver

<table>
<thead>
<tr>
<th>Function</th>
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<tbody>
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<td>Data from ADC</td>
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<tr>
<td>Front-end compensation</td>
</tr>
<tr>
<td>Resampling</td>
</tr>
<tr>
<td>Fixed equalizer</td>
</tr>
<tr>
<td>Timing recovery</td>
</tr>
<tr>
<td>Adaptive equalizer</td>
</tr>
<tr>
<td>Frequency offset compensation</td>
</tr>
<tr>
<td>Carrier phase recovery</td>
</tr>
<tr>
<td>Measuring system performance</td>
</tr>
</tbody>
</table>

First the four digitized signals (i.e., in-phase and quadrature components for each polarization) are passed through a block for the compensation of front-end imperfections. The front-end imperfections might include timing skew between the four channels due to the difference in both optical and electrical path lengths within a coherent receiver. Other types of front-end imperfections can be the difference in output powers for four channels due to different responses of the PDs in the receiver, and quadrature imbalance due an imperfect phase shift in the optical hybrid [42]. A resampling block is used to resample from an asynchronous sampling rate of the ADC to two samples per symbol.

While compensating for the transmission impairments of an optical fiber, it is important to note the different time scales of the dynamics of these impairments. For example, if PMD varies on a millisecond time scale, the chromatic dispersion can be considered constant on that scale. Because of such a difference in the dynamics of
these effects, it is necessary to divide the equalization of the received signal into two steps. First, fixed equalization for chromatic dispersion compensation is performed on each polarization signal separately [43], and then fast adaptive equalization is carried out jointly for two polarization signals. After the fixed equalizer, timing recovery is necessary because there are two samples per symbol but after the fixed equalizer they are not necessarily in the right place within the symbol period. Timing recovery determines what the modulation frequency is and where the two samples should be located relative to the symbol period. When the timing error is known the waveform is resampled using interpolation [44].

The adaptive equalizer performs polarization demultiplexing most importantly, as well as dealing with residual dispersion and PMD. For quadrature-phase-shift-keying (QPSK) signals the adaptive equalization can use the constant modulus algorithm (CMA) which minimizes the deviation of the amplitude of the equalized signal from a desired fixed value. For a 16-ary quadrature-amplitude-modulation (16-QAM) signal, CMA alone is not sufficient due to the presence of three amplitude levels in a 16-QAM constellation, and is typically used in combination with a multi-modulus algorithm (MMA), such as a radius directed algorithm (RDA) [45–47].

Depending on the intermediate frequency ($f_{IF}$), three different coherent methods can be distinguished: (i) homodyne receiver, (ii) heterodyne receiver, and (iii) intradyne receiver. Throughout this work the intradyne receiver is used. The coherent receiver is called intradyne when

$$f_{IF} = f_s - f_{LO} \neq 0 < B_{opt} \quad (2.18)$$

where $f_s$, $f_{LO}$ and $B_{opt}$ are the carrier frequency, local oscillator frequency and optical
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bandwidth of the transmitted signal, respectively. Because of the intradyne detection, $f_s$ is not locked to $f_{LO}$. The frequency offset must be estimated and removed to prevent constellation rotation [48, 49]. Finally, the carrier phase noise is estimated and removed from the modulated signal by a carrier phase recovery algorithm [50, 51]. The system performance is measured by the bit error ratio (BER) and/or mutual information (MI) calculation.

2.2.4 Chromatic Dispersion Compensation

Chromatic dispersion in an optical fiber can be modeled as an all-pass filter. The effect of chromatic dispersion on a signal can be modeled by a complex transfer function in the frequency domain [43].

\[ H(\omega) = \exp\left(-j\frac{D\lambda^2}{4\pi c}\omega^2z\right) \]  

(2.19)

where $\lambda$ is the wavelength, $c$ is the speed of the light, $D$ is the dispersion coefficient of the fiber, $z$ is the transmission distance and $\omega$ is the angular frequency. The dispersion compensating filter can be considered as an all-pass filter with transfer function $1/H(\omega)$ which can be a non-recursive [52] or recursive [53] digital filter. In theory, an all-pass filter can only be created using a recursive filter; however, it is not possible to design such a filter with the desired phase response [53]. Dispersion can either be compensated using time domain equalization (TDE) or frequency domain equalization (FDE) [43]. The choice between TDE and FDE depends on the maximum dispersion of the channel and the required filter length [46].

For TDE of the chromatic dispersion compensation, a large fixed equalizer is
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used to approximately compensate for the fiber dispersion. By means of the Fourier transform, equation 2.19 can be inverted to give the impulse response of a dispersive fiber

\[ h(z, t) = A \exp \left( j \frac{\pi c}{D\lambda^2 z} t^2 \right) \text{ where } A = \sqrt{\frac{c}{j D\lambda^2 z}} \quad (2.20) \]

As a result, by inverting the sign of the chromatic dispersion, the impulse function of the chromatic dispersion compensating filter is defined by

\[ h_c(z, t) = A \exp (-j\phi(t)) \text{ where } \phi(t) = \frac{\pi c}{D\lambda^2 z} t^2 \quad (2.21) \]

In the sense of a digital implementation, the impulse response given by equation 2.21 has some issues which must be considered. This impulse response is not only non-causal and infinite in duration, but since it passes all frequencies, aliasing will occur. To solve this problem the impulse response is truncated to a finite duration. Therefore, if the signal is sampled above the Nyquist rate, a fractionally spaced equalizer can compensate for the chromatic dispersion. Such a filter can effectively be realized by using an finite impulse response (FIR) filter with the tap spacing equal to the sampling interval, \( T/2 \), where \( T \) denotes the symbol period. For an odd number of taps such that the total number is \( N \), the tap weights to compensate for CD accumulated through a transmission distance \( z \) are given as [43]

\[ a_n = \sqrt{\frac{j c T^2}{D\lambda^2 z}} \exp \left( -j \frac{\pi c T^2}{D\lambda^2 z} n^2 \right) - \left[ \frac{N}{2} \right] \leq n \leq \left[ \frac{N}{2} \right] \quad (2.22) \]

If the tap number is large enough then the sampled impulse response will approximate
the continuous time impulse response. It means the more dispersion along the fiber, the more broadening of a pulse and the more taps are needed. Another possibility for dispersion compensation is infinite impulse response (IIR) filtering [53]. However, the inherent feedback of IIR equalizers makes this approach virtually impossible to implement in high-speed applications with parallelized signal processing.

For a large number of $N$, FIR filtering is more efficiently implemented in the frequency domain using the overlap-and-add or overlap-and-save methods [46]. Figure 2.3 shows the principle of FDE of CD compensation for two polarizations ($x$ and $y$). First, the received signals after the resampling block ($P_x(n)$ and $P_y(n)$) are transferred to the frequency domain by the fast Fourier transform (FFT)-block, and then multiplied by the inverse transfer function of equation 2.19. Finally, the IFFT block returns the signal to the time domain.

![Figure 2.3: The principle of frequency domain equalizer of dual polarization data.](image)

It should be noted that it is difficult to precisely determine the amount of
dispersion. Therefore, even after the fixed equalizer, there is some amount of remaining dispersion. The adaptive equalizer can deal with the residual dispersion or mismatch between the estimated value of dispersion and the actual value of dispersion. In this thesis, in the case of fiber transmission, the frequency domain equalization for chromatic dispersion is used.

2.2.5 Timing Recovery

A time recovery block retimes the received signal using a time error detector (TED) and interpolation [54]. Feed-forward timing recovery uses TED algorithms to estimate the current timing error based on received samples [44, 54, 55]. The Gardner algorithm [54] can be employed as the TED algorithm because of its low complexity compared to hybrid feedback timing recovery introduced by Tanimura et al. [55]. However, the Gardner algorithm has low tolerance towards dispersion, and cannot work properly when the chromatic dispersion is large.

It should be noted that timing recovery estimates the delay for optimal detection of data. Because the delay varies very slowly, the received signal can be processed block by block. For each block it is possible to assume the delay is constant and estimate delay [44]. Since the conventional timing recovery algorithms lack high dispersion tolerance, it is commonly performed after CD compensation.

2.2.6 Adaptive Equalizer

The purpose of an adaptive equalizer is to do polarization de-multiplexing most importantly and to compensate for time varying channel impairments such as the
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state of polarization and PMD. In contrast to chromatic dispersion, the polarization
dependent effect is a time varying phenomenon. This is mainly due to the relatively
rapid variation in the polarization states. The adaptive equalizer required to
compensate for these dynamic effects may be realized using a set of four FIR
filters [45], as illustrated in Figure 2.4

![Figure 2.4: The butterfly structure of an adaptive equalizer.](image)

Therefore, a $T/2$–fractionally spaced equalizer can compensate for impairment
incurred, where $x_{in}(n)$ and $y_{in}(n)$ are the complex vectors of the input signals, $x_{out}(n)$
and $y_{out}(n)$ are the complex vectors of the equalizer output signals respectively, $h_{xx}(n)$,
h$_{xy}(n)$, $h_{yx}(n)$ and $h_{yy}(n)$ are the complex tap weights vectors. Depending on whether
a training sequence is used to estimate the channel, several estimators can be used.
In single carrier based optical transmission, blind equalizers are commonly applied in
offline processing because of their simplicity and their tracking property. The most
common adaptive blind equalizer algorithms cited in the literature are:

- The constant modulus algorithm (CMA),
2.2. RELEVANT CONCEPTS

- The radius directed algorithm (RDA) which is an adaptation of CMA to QAM modulation formats with different amplitude levels.

For CMA, the estimation of the filter coefficients is based on the constant modulus criterion [45] which minimizes the following cost function (e.g., $x$-polarization)

$$J_{CMA}(n) = E \left[ (|x_{out}(n)|^2 - R_x)^2 \right] \quad \text{where} \quad R_x = \frac{E[|x_{in}(n)|^4]}{E[|x_{in}(n)|^2]}$$

(2.23)

In the context of an adaptive algorithm, it is usual to implement the stochastic gradient descent algorithm version of the CMA to update the filter coefficients at each symbol. For example, the filter coefficient update $h_{xx}(n)$, is defined

$$h_{xx}(n+1) = h_{xx}(n) - \mu \nabla J_{CMA}(n)$$

(2.24)

Therefore,

$$h_{xx}(n+1) = h_{xx}(n) - \mu \left( \frac{e(n)}{|x_{out}(n)|^2 - R_x} \right) x_{out}(n) x_{in}^*(n)$$

(2.25)

where $\nabla J_{CMA}$ is the gradient, $\mu$ is the constant step-size parameter and $e(n)$ is the error signal.

Since a 16-QAM signal is clearly not a constant modulus format, the CMA error function, $e_x(n)$, does not tend to be zero. Therefore, the CMA may be used for pre-convergence. Another cost function is used which ensures that the error signal tends to zero. This result in an equalizer based on RDA [45].

The RDA criterion corresponds to an adaptation of CMA to 16-QAM constellations [45, 46] where the modulus of the constellation is not constant. In
RDA, the cost function (e.g., $x$-polarization) can be written as

$$J_{RDA}(n) = E \left[ (|x_{out}|^2 - R_{0x})^2 \right]$$

where

$$R_{0x} = \begin{cases} 
R_1 & \text{if } |x_{out}| < (\sqrt{R_1} + \sqrt{R_2})/2 \\
R_3 & \text{if } |x_{out}| > (\sqrt{R_2} + \sqrt{R_3})/2 \\
R_2 & \text{else} 
\end{cases}$$

where $R_1$, $R_2$ and $R_3$ are equal 2, 10 and 18 for 16-QAM signal, respectively. The stochastic gradient algorithm associated with this cost function is the same as in equations 2.24 and 2.25 where $R_x$ is replaced with $R_{0x}$.

### 2.2.7 Frequency Offset Estimation

The presence of a frequency offset between the transmitter laser and the LO can have an adverse impact on the performance of the system. Frequency offset estimation (FOE) prior to carrier phase recovery is necessary to ensure reliable performance. Numerous techniques have been reported that exhibit varying properties in terms of the accuracy and range of the frequency estimate, and the computational complexity [48,49,56].

For real-time implementation of 16-QAM coherent systems, it is important to investigate the performance of feed-forward algorithms for frequency estimation where hardware parallelization and pipelining are necessary. The QPSK partition scheme has been presented in [48] for frequency offset estimation in a 16-QAM coherent system. The 16-QAM constellation can be partitioned into three classes ($C_1$, $C_2$ and $C_3$) based on their amplitude.

In Class I ($C_1$) and class III ($C_3$), symbols lie on the inner and outer rings,
These two classes form two QPSK constellations with different amplitudes. It is therefore possible to employ these symbols for frequency estimation as described in [48] for the QPSK modulation format. It should be noted that raising the signal to the $M^{th}$-power eliminates the $M$-phase-shift-keying ($M$-PSK) modulation [56]. Therefore, for a 16-QAM signal to obtain modulation-free samples, the power of four operation is used on the partitioned Class I and Class III symbols. To minimize the impact of additive white Gaussian noise on the frequency estimation, the modulation-free samples are averaged over a block of data.

In [57] and [58] a spectral domain algorithm is applied to estimate and compensate the frequency offset. For this algorithm, first the signal is zero-padded up to certain length. Second, the fast Fourier transform (FFT) of the $4^{th}$ power signal is computed. Morelli proved that the frequency where the largest absolute value of signal is located is equal to the frequency offset. The length of zero-padding considerably affects the estimation performance. Most papers recommend the length of zero-padding in the range from 2 to 8 times the original length of signal. In [49] a FOE algorithm is proposed based on eighth-order statistics for optical QAM signals. The eighth-order algorithm consumes substantially less memory resources compared the Morelli algorithm [49].

In this thesis, the spectral domain algorithm (Morelli algorithm) is used for recovery of the carrier frequency offset.
2.2.8 Carrier Phase Recovery

The tolerance of high-order QAM modulation formats towards laser phase noise decreases because of the smaller Euclidean distance between symbols. As a result, carrier phase recovery (CPR) algorithms with better laser linewidth tolerance are very important. For coherent detection, CPR is one of the essential steps in the DSP at the receiver. However, the performance of DP 16-QAM can be affected by the sensitivity of CPR algorithms to the laser linewidth. Therefore, CPR algorithms with high performance and high linewidth tolerance are required. The process of CPR for the $M$-PSK modulation format is straightforward. As the phase angles for the constellation points are equally spaced, the modulation can be removed by taking the $M^{th}$ power of the received complex symbols. However, the phase angles for 16-QAM constellation points are not equally spaced and alternative approaches are required. Many carrier phase recovery algorithms for 16-QAM have been proposed [50, 51, 56, 59–61]. Among them, the QPSK partitioning [56] and blind phase search (BPS) [50] algorithms using the feed-forward approach are attractive because they can be implemented in real time using DSP techniques in contrast to the feedback algorithms such as the decision-directed phase-locked loop (DD-PLL). Although QPSK partitioning for 16-QAM systems introduces a stringent linewidth requirement, in comparison to the BPS algorithm, it requires significantly less computational effort. Two-stage algorithms have been introduced as a way of reducing complexity and/or improving performance [61]. In [51] a two-stage algorithm has been proposed for which the first stage utilizes either one of the simplified QPSK partitioning algorithms or the BPS algorithm to obtain a coarse estimate. The second stage utilizes a novel QPSK constellation transformation...
algorithm to obtain a fine estimate.

In this thesis, the two-stage simplified QPSK partitioning with QPSK constellation transformation algorithm and BPS algorithm are used in Chapter 3 and Chapter 4, respectively.

2.2.9 Forward Error Correction

The use of forward error correction (FEC) encoding at the transmitter and decoding at the receiver is fundamental to recent advances in optical fiber communications. Often, researchers like to reuse experimental data obtained in expensive optical transmission experiments to evaluate the performance of different FEC schemes, without needing to redo the transmission experiment and/or signal processing. Therefore, thresholds are commonly used to decide whether the BER after FEC decoding is below the required target BER, which can be in the range of $10^{-13}$ to $10^{-15}$. The most commonly used threshold in the optical communications literature is the pre-FEC BER.

Throughout this thesis, for single-carrier 448 Gb/s DP 16-QAM signal, 7% FEC coding overhead is chosen. Based on this overhead the pre-FEC BER limit is $3.8 \times 10^{-3}$ [62]. For bit rate of 1.206 Tb/s, the FEC coding overhead is 15% corresponding to pre-FEC BER limit of $1.9 \times 10^{-2}$ [63].

It is shown in the literature that the different threshold of BER can be used for back-to-back systems [62]. In this thesis, the BER threshold is $1 \times 10^{-3}$ for back-to-back systems. The theoretical curve for the BER versus OSNR is presented and the corresponding implementation penalty is calculated at BER of $1 \times 10^{-3}$. 
2.2.10 Measuring System Performance

To quantify the performance of different systems, a typical approach involves adding broadband noise from an amplified spontaneous emission (ASE) source to the optical signal prior to the receiver. This process is referred to as noise loading and facilitates measuring the BER as a function of the OSNR. Throughout this work, the BER is measured using bit error counting and the OSNR is measured using the conventional 0.1 nm noise bandwidth using an Advantest Q8384 optical spectrum analyzer (OSA). The BER values convert to corresponding signal to noise ratio (SNR) values for systems with coherent detection for a modulation order $M$ using the well-known theoretical relation between them [64]

$$BER_{M-QAM} \approx \frac{2}{\log_2 M} \cdot \left(1 - \frac{1}{\sqrt{M}}\right) \cdot \text{erfc} \left(\sqrt{\frac{3\text{SNR}}{M-1}}\right)$$

(2.27)

where $\text{SNR} = \text{OSNR} \cdot \frac{2\Delta f}{R_s}$

where $\Delta f$ is the reference bandwidth (most commonly a reference bandwidth of 0.1 nm or 12.5 GHz is used) and $R_s$ is the symbol-rate. In this thesis, the superchannel BER is defined as the average of the BERs for individual subcarriers.

Mutual information (MI) has been a key quantity in the field of information theory since Shannon’s seminal paper [65]. Recently, in optical communications, it has been used to state lower-bound estimates on channel capacity [66]. The mutual information $MI(X;Y)$ between the input and output signals of a system (denoted $X$ and $Y$, respectively) represents the amount of information about $X$ that is contained in the
observation $Y$ when $X$ is transmitted. The MI is defined \[67\]

$$MI (X; Y) = \sum_{m=1}^{M} \sum_{Y} P_X (X_m) P_{Y|X} (Y|X_m) \cdot log_2 \left[ \frac{P_{Y|X} (Y|X_m)}{P_Y (Y)} \right]$$ \hspace{1cm} (2.28)

where $P_{Y|X} (Y|X_m)$ is the conditional probability density function (PDF) of $Y$ given the $m^{th}$ input realization of $X$, $P_Y (Y)$ is the PDF of the output $Y$ and $P_X (X)$ is the PDF of the input $X$ which is taken as uniformly distributed $P_X (X_m) = 1/M$. The MI can be estimated using a histogram as a PDF estimator in the case that the optimal histogram bin widths are obtained from blind algorithms \[68–70\]

In an optical fiber link, MI is used to find the maximum potential data throughput over a fiber. Further, MI serves as a relative figure of merit to assess the impact of changing a design parameter or adding a routine to the receiver with respect to achievable rate. Choosing the correct value for the bin width is a key requirement for estimating the MI precisely.
Chapter 3

Pattern-Dependent Distortion Compensation

With the deployment of 100 Gb/s transceivers and the increasing demand for data transmission capacity, systems with bit rates of 400 Gb/s and 1 Tb/s are now under investigation [4, 6]. Possible solutions include single-carrier multi-level modulation with a high baud [9–12, 62, 71–74] for a bit rate of 400 Gb/s, and multi-carrier modulation with a low symbol rate for each subcarrier, such as Nyquist-WDM [8–23] and coherent optical orthogonal frequency division multiplexing (CO-OFDM) [75–77], for bit rates of 400 Gb/s and 1 Tb/s. Single-carrier approaches for a bit rate of 400 Gb/s are attractive because of the simple structure in the transmitter and a receiver complexity that is similar to those for multi-carrier signals [23]. For a bit rate of 400 Gb/s, single-carrier dual-polarization 16-ary quadrature-amplitude modulation (DP 16-QAM), combined with coherent detection and digital signal processing provides a high spectral efficiency and reasonable OSNR requirement. Experiments for 400 Gb/s or
1 Tbps systems have used either optical time division multiplexing (OTDM) or electrical time division multiplexing (ETDM). While OTDM overcomes the bandwidth limitations of electrical and opto-electronic components, ETDM is seen as a more practical approach [78].

For single-carrier signals, demonstrations of high symbol rate modulation include

- 56 Gbaud using DP 16-QAM (448 Gb/s) [10–12, 62];
- 64 Gbaud using DP 16-QAM (512 Gb/s) [73];
- 72 Gbaud using DP 64-QAM (864 Gb/s) [71];
- 80 Gbaud using DP 16-QAM (640 Gb/s) [9];
- 107 Gbaud using DP QPSK (428 Gb/s) [9];
- 110 Gbaud using DP QPSK (440 Gb/s) [72].

For 1 Tbps superchannel signals, recent demonstrations of high symbol rate modulation include

- DP 8-QAM and 9 subcarriers with 23 Gbaud (1.242 Tbps) [15];
- DP 16-QAM and DP 8-QAM and 8 subcarriers with of 24.8 Gbaud (1.5872 Tbps and 1.1904 Tbps, respectively) [20];
- DP 16-QAM and 5 subcarriers with 28 Gbaud (1.12 Tbps) [21];
- DP 16-QAM and 4 subcarriers with 36 Gbaud (1.152 Tbps) [23], 39 Gbaud (1.248 Tbps) [19] and 40 Gbaud (1.28 Tbps) [18];
• DP 16-QAM and 3 subcarriers with 50.25 Gbaud (1.206 Tb/s) [10,12–14,16];
  DP 32-QAM and 3 subcarriers with 46 Gbaud (1.38 Tb/s) [22];

• DP 16-QAM and 2 subcarriers with 80 Gbaud (1.28 Tb/s) [9].

The number of optical subcarriers, the bit rate per subcarrier, and the modulation format are important aspects of superchannel design.

Due to the bandwidth limitations and nonlinear responses of electronic amplifiers and opto-electronic devices at the transmitter and receiver, high symbol rate signals can exhibit a pattern-dependent distortion or nonlinear inter-symbol-interference. This distortion can adversely affect system performance [10,12,16,79].

Recently, different approaches have been proposed to mitigate pattern-dependent distortion, including digital signal processing techniques to mitigate the nonlinearities of the Mach-Zehnder modulator [80] or to improve the signal quality of digital-to-analog converters (DACs) with limited bandwidth [79]. Volterra series, which has served as a powerful tool to analyze and compensate for fiber nonlinearities [81–83], has received a lot of attention in the context of pattern-dependent distortion compensation [16,84]. Another approach is to capture the nonlinearity-degraded data signals and use this knowledge to build a look-up table (LUT) to pre-compensate or implement at the receiver to mitigate the distortion that does exist [10,12,74]. Since the transmitted symbols are equally likely, there is no difference between the maximum likelihood sequence estimation (MLSE) and the MAP detection approaches.

In this Chapter, the transmission performance of single-carrier 448 Gb/s DP 16-QAM and a 1.206 Tb/s three-carrier DP 16-QAM superchannel signal is evaluated using a maximum-a posteriori probability (MAP) detector based on a fixed LUT
at the receiver to mitigate the effects of pattern-dependent distortion due to the high symbol rate (50.25 Gbaud) [10, 13]. Alternative approaches include a LUT based nonlinear pre-distorter (NLPD) at the transmitter [12,16,84,85] and a Volterra nonlinear equalizer (VNLE) at the receiver [16].

3.1 Fixed Look-Up Table MAP Detection

MAP detection has previously been investigated in the context of mitigating the pattern-dependent distortion generated by intra-channel fiber nonlinear effects [86] and by narrow filtering [87]. In this thesis, fixed LUTs based on MAP detection are used to mitigate the pattern-dependent distortion in high baud systems. For each polarization signal, one-dimensional MAP detection for two 4-ary signals is used rather than two-dimensional MAP detection for one 16-ary signal. The LUT entries need to be updated for any change in the modulation format or symbol rate.

Figure 3.1 illustrates how the LUT entries are generated using a training sequence. The rows of the LUT contain the average received sample values for each of the possible symbol sequences of length \( N = 2n + 1 \) \( (n = 1, 2, 3, \ldots) \).
3.1. FIXED LOOK-UP TABLE MAP DETECTION

A sequence of training symbols $T$ is transmitted in a back-to-back system at the highest available OSNR. As a sliding window advances through the training sequence, the known $N$ symbol sequence forms the address of the LUT and the amplitudes of the $N$ sample values for the signal $R$ are recorded as the corresponding LUT entry. Each time a specific symbol sequence occurs, the amplitudes of the sample values are added to the corresponding LUT entry and a counter ($C$) is incremented to register the number of updates for that sequence. After the training is completed, the average sample values are determined and stored in memory. The effect of noise on the final LUT entries is reduced by the averaging. Since the MAP detection is used to mitigate transmitter- and receiver-based pattern-dependent distortion, the LUT is applied for system configurations with different fiber lengths and OSNRs. For a 16-QAM signal, the LUT has $4^N$ rows and $N$ columns.

The conventional MAP detection algorithm performs symbol detection based on determining the Euclidean distances between the received sample values for $2n + 1$
consecutive symbols and the LUT entries. The decision about the center symbol is obtained from the address for the LUT entry with the minimum Euclidean distance.

For conventional LUT-based MAP detection, the Euclidean distance calculation requires \( N \times 4^N \) real multiplications. For \( n = 2 \), Table 3.1 shows an example of a sequence of transmitted symbols, the sequence of received sample values of back-to-back system with OSNR = 28.8 dB, the contents of the 5-symbol LUT row entries selected by MAP detection, and the corresponding LUT address that yields the symbol decisions. Note that in the sliding window implementation of the Table 3.1: Transmitted, Received Sequence, Corresponding Content of 5-symbol LUT Entry and Address.

<table>
<thead>
<tr>
<th>Transmitted sequence</th>
<th>1</th>
<th>-3</th>
<th>1</th>
<th>-3</th>
<th>-1</th>
<th>3</th>
<th>-1</th>
<th>1</th>
<th>3</th>
</tr>
</thead>
<tbody>
<tr>
<td>Received sequence</td>
<td>1.39</td>
<td>-2.41</td>
<td>0.87</td>
<td>-2.92</td>
<td>-2.05</td>
<td>2.78</td>
<td>-1.02</td>
<td>0.69</td>
<td>3.06</td>
</tr>
<tr>
<td>Selected LUT entry</td>
<td>0.65</td>
<td>-2.96</td>
<td>0.60</td>
<td>-3.07</td>
<td>-1.33</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>LUT address</td>
<td>1</td>
<td>-3</td>
<td>1</td>
<td>-3</td>
<td>-1</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Selected LUT entry</td>
<td>-2.98</td>
<td>0.60</td>
<td>-3.18</td>
<td>-1.29</td>
<td>2.71</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>LUT address</td>
<td>-3</td>
<td>1</td>
<td>-3</td>
<td>-1</td>
<td>3</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Selected LUT entry</td>
<td>0.75</td>
<td>-3.20</td>
<td>-2.88</td>
<td>2.76</td>
<td>-1.06</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>LUT address</td>
<td>1</td>
<td>-3</td>
<td>-3</td>
<td>3</td>
<td>-1</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Selected LUT entry</td>
<td>-3.16</td>
<td>-2.86</td>
<td>2.77</td>
<td>-0.90</td>
<td>0.84</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>LUT address</td>
<td>-3</td>
<td>-3</td>
<td>3</td>
<td>-1</td>
<td>1</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Selected LUT entry</td>
<td>-1.12</td>
<td>2.79</td>
<td>-1.03</td>
<td>0.96</td>
<td>2.67</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>LUT address</td>
<td>-1</td>
<td>3</td>
<td>-1</td>
<td>1</td>
<td>3</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

LUT-based MAP detection, each received sample value participates in \( 2n + 1 \) consecutive detection blocks, hence receiving \( 2n + 1 \) initial decisions. In this thesis, three modified versions of the MAP detector are proposed in order to combine these initial decisions and improve the system performance. For each position of the detection window, an initial decision is made based on the minimum Euclidean
distance between the received sample values and the LUT entries.

For the minimum distance rule, the final decision is based on the detection window with the smallest Euclidean distance between the sample value for the symbol of interest and the corresponding LUT entry. Table 3.1 is used to illustrate this rule for the case of making an incorrect decision with a conventional 5-symbol MAP detector (third detection window of the Table 3.1). The LUT entry -1.33 from the first window has the smallest Euclidean distance to the received sample value -2.05. Although, the sample value for the symbol under decision is itself in the wrong decision region, it is correctly decided by the minimum distance rule.

For the majority rule, the final decision is based on the majority of the initial decisions for the symbol of interest obtained from the $2n + 1$ observations. In Table 3.1, three observations yield initial decisions of $-1$ and two observations yield initial decisions of $-3$. The final decision is $-1$.

For the averaging rule, the final decision is based on the smallest Euclidean distance between the sample value for the symbol of interest and the averages of the LUT entries for the symbol of interest which correspond to the same initial decision. In Table 3.1, the averages of the LUT entries for initial decisions of $-1$ and $-3$ are -1.25 and -2.87, respectively. The final decision for the sample value -2.05 is -1.

The conventional MAP detector uses a single observation of the symbol of interest (the detection window is centered about the symbol of interest). The three modified versions use $2n + 1$ observations of the same symbol to form a final decision. A final decision does not directly affect subsequent ones, although as a next final decision is made, the result for one detection window (left-most in Table 3.1) is removed and the result for one detection window (right-most) is added.
3.1. FIXED LOOK-UP TABLE MAP DETECTION

The complexity of MAP detection scales with the memory length $2n + 1$. As the detection window advances, the initial decisions described above are common to the four rules for making a final decision (conventional, minimum distance, majority and averaging). The complexities of the additional calculations to determine the final decisions are relatively small since they involve a few computational steps as summarized in Table 3.2.

Table 3.2: Computational steps of the three decision rules.

<table>
<thead>
<tr>
<th>Minimum Distance</th>
<th>Majority</th>
<th>Averaging</th>
</tr>
</thead>
<tbody>
<tr>
<td>$2n + 1$ initial decisions</td>
<td>Identify which of the $2n + 1$ initial decisions are the same (4 possible decisions are $\pm 1, \pm 3$)</td>
<td>Identify which of the $2n + 1$ initial decisions are the same (4 possible decisions are $\pm 1, \pm 3$)</td>
</tr>
<tr>
<td>Calculate $2n + 1$ distances between the symbol of interest and the corresponding LUT entry</td>
<td>Determine the number of initial decisions that correspond to each of the 4 possible decisions</td>
<td>For initial decisions that are the same, calculate the average of the LUT entries for the symbol of interest (from 1 to 4 averages)</td>
</tr>
<tr>
<td>Select the minimum from $2n + 1$ values</td>
<td></td>
<td>Calculate the distances between the sample value for the symbol of interest and the averages of the LUT entries (from 1 to 4 distances)</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Select the minimum from 1 to 4 values</td>
</tr>
</tbody>
</table>
3.1. EXPERIMENTAL SETUP

The first setup utilized a SHF 12103 multi-channel bit pattern generator (BPG). For the 448 Gb/s single-carrier DP 16-QAM signal and 1.206 Tb/s three-carrier DP 16-QAM superchannel signal, LUTs entries were obtained based on the received signal for a back-to-back system configuration with the highest available OSNR. Figure 3.2 shows the back-to-back experimental setup for the 448 Gb/s single-carrier DP 16-QAM and 1.206 Tb/s three-carrier DP 16-QAM superchannel transmission systems. For the single carrier system, the optical carrier was generated using an external cavity laser with a nominal linewidth of 100 kHz. Four 56 Gb/s signals modulated by a $2^{15} - 1$ pseudorandom bit sequence (PRBS) were attenuated and combined to generate two four-level signals for driving an IQ modulator (25 GHz 3-dB bandwidth). The 16-QAM signal from the IQ modulator was split and then recombined in orthogonal polarizations after delaying one of the signals to de-correlate it from the other signal. The transmitter is shown as an inset in Figure 3.2. The DP 16-QAM signal obtained by polarization multiplexing emulation was applied to an EDFA and programmable optical filter (POF) with a response set to obtain an output signal with a raised-cosine response (roll-off factor of 0.2) subject to the limitations imposed by a resolution of 1 GHz. The spectral filtering reduced the requirement on the receiver bandwidth and enhanced the spectral efficiency.

For the three-carrier system, the optical carriers with different channel spacings were generated using three external cavity lasers with nominal linewidths of 100 kHz. Four 50.25 Gb/s signals modulated by a $2^{15} - 1$ PRBS were attenuated and combined to generate two four-level drive signals that were split and applied to two IQ modulators (25 GHz 3-dB bandwidth) with relative time delays to de-correlate the
symbol patterns. The signal quality was degraded after splitting the four level signal for two modulators. This could have been avoided had a second pulse generator been available. One modulator was used for the odd channels (channels 1 and 3), and one modulator was used for the even channel (channel 2). After emulating polarization multiplexing, the even and odd channels were separately amplified and applied to POFs for spectral filtering. For the back-to-back system, a broadband source was used to add amplified spontaneous emission noise to the modulated signals to measure the dependence of the BER on OSNR.

Figure 3.2: Experimental setup for back-to-back measurement. PS: phase shifter; IQM: IQ modulator; EDFA: erbium doped fiber amplifier; POF: programmable optical filter; VOA: variable optical attenuator; BBS: broadband source; OBPF: optical bandpass filter; OSA: optical spectrum analyzer.

Figure 3.3 illustrates the transmission experimental setup. The re-circulating loop was comprised of four 75 km spans of standard SMF with a nominal dispersion
3.1. FIXED LOOK-UP TABLE MAP DETECTION

A coefficient of 17 ps/km/nm at a wavelength of 1550 nm and an attenuation of 0.19 dB/km. An EDFA, optical bandpass filter (OBPF, 3 dB bandwidth of 1.2 nm and 18 nm for the single-carrier and three -subcarrier signals, respectively), and variable optical attenuator (VOA) were used to set the input power for the re-circulating loop. The OBPF removed the out-of-band noise and avoided saturation of the EDFAs. A gain equalizer (GE) was used in the loop for the three-carrier signal. A polarization synthesizer (PS) was used in the loop to scramble the state-of-polarization of the re-circulating signals.

Figure 3.3: Experimental setup for transmission measurements. AOM: acousto-optic modulator; EDFA: erbium doped fiber amplifier; OBPF: optical bandpass filter.

The received signal was amplified and filtered (1.3 nm bandwidth) before detection by a polarization- and phase-diverse coherent receiver with 32 GHz bandwidth. The local oscillator laser had a nominal linewidth of 100 kHz. The four signals from the balanced photodetectors were digitized by 80 GSa/s ADCs using two synchronized real-time sampling oscilloscopes (33 GHz bandwidth).
The off-line signal processing included (i) compensation for quadrature imbalance [42], (ii) re-sampling to two samples per symbol using interpolation algorithm in MATLAB, (iii) frequency domain equalization for chromatic dispersion in the case of fiber transmission [43, 45], (iv) clock recovery using the digital square and filter technique [44], (v) polarization demultiplexing and compensation for residual signal distortion using 13-tap adaptive $2 \times 2$ equalizer [45], (vi) recovery of the carrier frequency offset using a spectral domain algorithm [58], (vii) recovery of the carrier phase using a sliding window two-stage simplified QPSK partitioning and QPSK constellation transformation algorithm [51], (viii) MAP detection for pattern-dependent distortion compensation and (ix) symbol decisions. The adaptive equalizer first used the constant modulus algorithm for pre-convergence and then the radius directed algorithm [45]. Rectilinear decision boundaries were used in obtaining the BER by direct bit error counting. Corresponding parameters for these DSP algorithms are listed in Table 3.3.

<table>
<thead>
<tr>
<th>Algorithms</th>
<th>Parameters</th>
<th>Values</th>
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<td>Method</td>
<td>frequency domain</td>
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<td>Adaptive equalizer</td>
<td>Method</td>
<td>CMA-RDA</td>
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<td></td>
<td>Convergence parameters</td>
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<td>Number of taps</td>
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<td>Number of iterations</td>
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<td>Carrier phase recovery</td>
<td>Block lengths</td>
<td>optimized</td>
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3.1.2 Optical Pulse Shaping

Raised-cosine optical pulse shaping was applied to narrow the modulated signal spectrum and thereby enhance the spectral efficiency and reduce the requirements on the receiver bandwidth and analog-to-digital converter (ADC) sampling rate. A programmable optical filter (POF) with a frequency resolution of 1 GHz was used to shape the modulated signal spectrum [15]. The POF response was set based on obtaining raised-cosine spectra with a specified roll-off factor and included pre-emphasis to compensate for (i) non-ideal features in the output spectra that result from the sharp roll-off of the desired response and the POF resolution, and (ii) the bandwidth of the coherent receiver (32 GHz). Modulated signal spectra before and after the POF are shown in Figure 3.4 for the 448 Gbit/s single-carrier DP 16-QAM signal with a roll-off factor of 0.2.

Figure 3.4: Optical spectrum of the 448 Gb/s DP-16QAM signal before and after the POF with roll-off factor of 0.2.
3.1.3 Results - LUT Entries

The fixed LUTs for the single-carrier and three-carrier signals are created using training sequence data sets comprised of 564,645 symbols in a back-to-back system configuration operating at the highest available OSNR (36.6 dB and 36.9 dB, respectively). For a single carrier, Figure 3.5 superimposes all the LUT entries of the center symbols for symbol sequences of length 3, 5 and 7. The range of the LUT entries increases with an increase in the length of the symbol sequence until the memory of the pattern-dependent distortion is captured. For a training sequence of a given length, there are fewer occurrences of each $2n+1$ symbol sequence as $n$ increases. As a result there is less averaging of the sample values for each specific symbol sequence.

![LUT Entries](image)

Figure 3.5: LUT entries for the center symbol for 3-, 5- and 7-symbol sequence lengths.
For LUTs with symbol sequences of length 3, 5 and 7, Figure 3.6 indicates the average of the four variances corresponding to the center symbol LUT entries (Figure 3.5) as a function of the number of data sets used for training. In the results that follow, five data sets were used for determining the LUT entries as this provides near-minimum variances.

![Figure 3.6: The average variance of LUT entries as a function of the number of data sets](image)

3.1.4 Results - 448 Gb/s Single-Carrier

Figure 3.7 shows the dependence of the BER on OSNR for the 448 Gb/s DP 16-QAM back-to-back system without and with MAP detection. The pre-FEC BER limit is $1 \times 10^{-3}$ aiming for a post-FEC BER below $10^{-15}$. Without MAP detection, a BER floor larger than pre-FEC thresholds is observed. With conventional MAP detection,
the BER improves significantly as the length of the symbol sequence \((2n + 1)\) increases from 3 to 7. The required OSNRs (0.1 nm noise bandwidth) for BER = \(1 \times 10^{-3}\) using 3-, 5- and 7-symbol MAP detection are 29.7 dB, 28.6 dB and 27.7 dB, respectively. In contrast to the theoretical limit, around 4.7 dB implementation penalty is observed at BER = \(1 \times 10^{-3}\) compared with 7-symbol MAP detection.

Figure 3.7: Dependence of the BER on OSNR for the 448 Gb/s DP 16-QAM back-to-back system without and with conventional MAP detection.

The different decision rules are compared in Figure 3.8 in terms of how the BER depends on the OSNR for the 448 Gb/s DP 16-QAM back-to-back system with 7-symbol MAP detection. The minimum distance MAP detector yields the best performance. Compared to the conventional MAP detector, for each symbol the minimum distance MAP detector requires additional processing for the decision rule
3.1. FIXED LOOK-UP TABLE MAP DETECTION

comprised of $2n + 1$ subtractions, $2n + 1$ multiplications, and a select minimum from $2n + 1$ values. All decision rules share the LUT search which yields the final decision for the conventional rule and the initial decisions for the minimum distance, majority, and averaging rules. For $\text{BER} = 1 \times 10^{-3}$, a back-to-back OSNR sensitivity of 26.9 dB is achieved. The required OSNR at pre-FEC threshold is reduced by about 0.8 dB compared to conventional MAP detection. This corresponds to an implementation penalty at $\text{BER} = 1 \times 10^{-3}$ of 3.9 dB from theory.

![Figure 3.8: Dependence of the BER on OSNR for 448 Gb/s DP 16-QAM back-to-back system with different decision rules for 7-symbol MAP detection.](image)

The dependence of the BER on launch power for transmission over 900 km of SMF for the 448 Gb/s DP 16-QAM signal is illustrated in Figure 3.9 without LUT-MAP and with 3-, 5- and 7-symbol minimum distance MAP detection. Without MAP
detection, the BER curve is above the FEC threshold of $3.8 \times 10^{-3}$.

Figure 3.9: Dependence of the BER on launch power for a fiber length of 900 km without and with 3-, 5- and 7-symbol minimum distance MAP detection.

Figure 3.10 shows the dependence of the BER on the fiber length for the 448 Gb/s DP 16-QAM signal with an optimum launch power of 1 dBm with conventional and minimum distance 7-symbol MAP detection. The improved performance of minimum distance MAP detection allows transmission over 1200 km at the FEC threshold. The 7-symbol minimum distance MAP detector increases the achievable transmission distance by about 107 km (8.9%) compared to conventional MAP detection at the FEC threshold of $3.8 \times 10^{-3}$. 
3.1.5 Results - 1.206 Tb/s Three-Carrier

The OSNR sensitivity (0.1 nm noise bandwidth) for $\text{BER} = 1 \times 10^{-3}$ as a function of the prescribed roll-off factor $\beta$ for the POF response and the spacing between the three channels is shown in Figure 3.11. These results are for the center channel (channel 2) with conventional 7-symbol MAP detection. The best back-to-back performance is obtained for $\beta = 0.6$ for all the channel spacings. For $\beta < 0.6$ the increase in the required OSNR is attributed to the limitations in achieving raised-cosine spectra with sharp roll-off using the POF and for $\beta > 0.6$ the increase is attributed to inter-channel crosstalk between the three subcarriers. A roll-off factor of 0.6 and subcarrier spacing
of 75 GHz were chosen by considering the performance and spectral efficiency. The net spectral efficiency is 4.4 b/s/Hz (bit rate of 1.005 Tb/s and bandwidth of 225 GHz). The optical spectrum is shown as an inset in Figure 3.11. The similarity of the dependencies on the prescribed roll-off factor for different subcarrier spacings is attributed to the actual response of the programmable optical filter (resolution of 1 GHz) being different from the ideal prescribed response. The discrepancy between the actual and prescribed responses depends on the sharpness as the passband response.

Figure 3.11: Dependence of the required OSNR for BER = $1 \times 10^{-3}$ on the prescribed roll-off factor for different channel spacings for the 1.206 Tb/s superchannel signal in a back-to-back system with 7-symbol conventional MAP detection.

For a back-to-back system, the dependence of the BER on OSNR is illustrated in Figure 3.12 without and with 7-symbol minimum distance MAP detection. Results
are presented for the three individual channels and the superchannel (average of the individual channels), and for comparison for the case of a single channel (channel 2). Without the MAP detector, a BER floor larger than $1 \times 10^{-3}$ is observed. The 7-symbol minimum distance MAP detector yields an OSNR sensitivity of 26.4 dB for BER = $1 \times 10^{-3}$.

Figure 3.12: Dependence of the BER on OSNR for the 1.206 Tb/s superchannel signal back-to-back system without and with 7-symbol minimum distance MAP detection.

Figure 3.13 shows the dependence of the BER on OSNR for the 1.206 Tb/s superchannel back-to-back system with 7-symbol MAP detection based on the different decision rules. The minimum distance MAP detector yields the best performance. The required OSNR for BER = $1 \times 10^{-3}$ is reduced by 1 dB using
7-symbol minimum distance MAP detection compared to conventional MAP detection.

Figure 3.13: Dependence of the BER on OSNR for the 1.206 Tb/s superchannel signal back-to-back system with different decision rules for 7-symbol MAP detection.

Figure 3.8 and Figure 3.13 indicate that it is better to use a final decision rule based on the best initial decision (i.e., minimum distance rule) rather than one based on all the initial decisions.

For 3-, 5- and 7-symbol minimum distance MAP detection, Figure 3.14 illustrates the dependence of the BER on the per-channel launch power for transmission over 1500 km of SMF. A launch power range of 2.7 dB was achieved for 7-symbol minimum distance MAP detection.
3.1. FIXED LOOK-UP TABLE MAP DETECTION

Figure 3.14: Dependence of the BER on launch power for the 1.206 Tb/s superchannel signal and fiber length of 1500 km with 3-, 5- and 7-symbol minimum distance MAP detection.

The dependence of the BER on fiber length at an optimum launch power of 2 dBm is shown in Figure 3.15 for conventional and minimum distance 7-symbol MAP detection. The improvement in the performance with minimum distance MAP detection allows transmission over 1500 km at the FEC threshold of $1.9 \times 10^{-2}$. The achievable transmission distance increases by about 127 km (7.6%) by using 7-symbol minimum distance MAP detection compared to conventional MAP detection at the FEC threshold of $1.9 \times 10^{-2}$. 
As indicated in Figure 3.14, there are two input powers which yield BER = 1.9 × 10^{-2} for a fiber length of 1500 km. In Figure 3.16, the minimum and maximum per-channel fiber launch powers that yield BER = 1.9 × 10^{-2} with 7-symbol minimum distance MAP detection are shown for different fiber lengths. For comparison, results are also shown for the case of a single channel. The high input power is limited by fiber nonlinearities and low input power is limited by optical amplifier noise. As expected, the range in input power that corresponds to BER ≤ 1.9 × 10^{-2} decreases as the fiber length increases.
3.1.6 Results - Mutual Information

In this thesis, mutual information is used to find the maximum potential data throughput. Mutual information can serve as a relative figure of merit to assess the impact of changing a design parameter with respect to the achievable rate. The mutual information $I(S_T; \tilde{S}_R)$ between the discrete transmitted symbols $S_T$ and the receiver symbol decisions $\tilde{S}_R$ represents the amount of information about $S_T$ that is contained in the decisions $\tilde{S}_R$. For a modulation format of order $M$, the mutual
information is [67]

\[
MI(S_T; S_R) = \frac{1}{M} \sum_{m=1}^{M} \sum_{n=1}^{M} P_{R|T}(\tilde{S}_{R,n}|S_{T,m}) \\
\times \log_2 \left[ \frac{P_{R|T}(\tilde{S}_{R,n}|S_{T,m})}{P_{R}(\tilde{S}_{R,n})} \right]
\] (3.1)

where \(P_{R|T}(\tilde{S}_{R,n}|S_{T,m})\) is the conditional probability of \(n^{th}\) received symbol decision of \(\tilde{S}_R\) given the \(m^{th}\) transmitted symbol of \(S_T\) (\(n^{th}\) and \(m^{th}\) are both referring to the received and transmitted constellation points, respectively). \(P_{R}(\tilde{S}_{R,n})\) is the probability of \(n^{th}\) received symbol decision of \(\tilde{S}_R\). Figure 3.17 indicates the calculation of the mutual information after the conventional DSP algorithms and LUT based MAP detector at the receiver. After carrier phase recovery, the mutual information is calculated between a discrete-transmitted symbol and continuous-received signal value, whereas after LUT based MAP detector the mutual information is calculated by equation 3.1.

The mutual information is estimated using histograms as probability density function (PDF) estimators. Choosing the correct value for the bin width is a key requirement for estimating the MI precisely. For a large number of bins (small bin width) the MI tends to be overestimated, and for a small number of bins the histogram does not approximate the true PDF well [67]. Scott derived a method for the optimal bin width by minimizing the integrated mean squared error of the histogram [68]. Shimazaki proposed a method based on the spike count static [69]. Knuth introduced a straightforward blind method by assigning a multinomial likelihood and a non-informative prior of the given data [70].
3.1. FIXED LOOK-UP TABLE MAP DETECTION

Figure 3.17: Calculation of the mutual information before and after LUT-MAP detector.

\[ M_{I}(S_{T}; S_{R}) = \frac{1}{M} \sum_{m=1}^{M} \int p_{R|T}(s_{R}|s_{T,m}) \times \log_2 \left[ \frac{p_{R|T}(s_{R}|s_{T,m})}{p_{R}(s_{R})} \right] ds_{R} \]

\[ M_{I}(\tilde{S}_{R}; \tilde{S}_{R}) = \frac{1}{M} \sum_{m=1}^{M} \sum_{n=1}^{M} p_{R|T}(\tilde{s}_{R,n}|s_{T,m}) \times \log_2 \left[ \frac{p_{R|T}(\tilde{s}_{R,n}|s_{T,m})}{p_{R}(\tilde{s}_{R,n})} \right] \]

Figure 3.18 and Figure 3.19 illustrate the two-dimensional histogram for the conditional probability of \( \tilde{S}_{R} \) given the transmitted symbol of \( S_{T} = 3 + 3i \) and the histogram for the probability of received symbols \( P_{R}(\tilde{S}_{R}) \) of back-to-back result for three-carrier setup with OSNR = 26.8 dB, respectively. The bin width was determined based on the Knuth algorithm.
3.1. FIXED LOOK-UP TABLE MAP DETECTION

Figure 3.18: Histogram of the $P_{RT}(S_R|S_T = 3 + 3i)$.

Figure 3.19: Histogram of the $P_R(S_R)$. 
3.1. FIXED LOOK-UP TABLE MAP DETECTION

The performance of a 1.206 Tb/s three-carrier DP 16-QAM superchannel signal is also evaluated in terms of the mutual information without and with the minimum distance MAP detection rule based on a fixed LUT at the receiver. The superchannel mutual information is the average of all mutual information between the transmitted and received symbol sequences for each subcarrier.

Figure 3.20 shows the dependence of the MI and BER on OSNR for the 1.206 Tb/s three-carrier DP 16-QAM superchannel in a back-to-back system without LUT-MAP detection. MI results for three different methods of choosing a histogram bin width are compared [68–70].

Figure 3.20: MI and BER versus OSNR for a back-to-back system without LUT-MAP detection.
The MI results illustrate that all three algorithms yield similar results with the approach of Knuth [70] providing slightly larger values. The optimal bin width is calculated by Knuth algorithm in the following results.

The dependence of the MI and BER on OSNR is illustrated in Figure 3.21 without and with the 7-symbol minimum distance LUT-MAP detector. MI is calculated for each polarization and then averaged. It is clear that the amount of MI increases with the LUT-MAP detector. The BER = $2.5 \times 10^{-5}$ corresponds to MI of 3.9 bits/symbol.

![Graph](image)

Figure 3.21: MI and BER versus OSNR for a back-to-back system without and with 7-symbol minimum distance LUT-MAP detection for the three-carrier 1.206 Tb/s.

Figure 3.22 illustrates the dependence of the MI and BER on launch power for 1500 km of SMF without and with 7-symbol minimum distance LUT-MAP detection.
At the optimum launch power of 2 dBm, the MI is 2.5 and 2.8 bits/symbol without and with MAP detection, respectively. Figure 3.23 indicates the dependence of the MI and BER on the fiber length for a launch power of 2 dBm without and with 7-symbol minimum distance LUT-MAP detection. With MAP detection, an increase in the BER from 0.004 to 0.02 corresponds to a decrease in the MI from 3.15 to 2.6 bits/symbol.

Figure 3.22: MI and BER versus launch power for 1.206 Tb/s superchannel and 1500 km fiber without and with 7-symbol minimum distance LUT-MAP detection.
3.2 LUT BASED NONLINEAR PRE-DISTORTION

Nonlinear pre-distortion (NLPD) can be implemented at the transmitter using a LUT that contains the required pre-distorted value for the center symbol within each possible symbol sequence of length $N = 2n + 1$ [12, 16, 84, 85]. This approach has the advantage of using known transmitted symbol sequences to always identify the correct LUT address and apply the required correction to the symbol of interest based on the LUT entries.

Figure 3.24 illustrates how the LUT entries are generated for the LUT based NLPD scheme. During a training phase, a sequence of training symbols is transmitted in a...
back-to-back system at the highest available OSNR. The amplitude pre-distortions are determined from the differences between the actual received sample values, $\tilde{R}(k)$, obtained during training and the corresponding transmitted symbol values $\tilde{T}(k) \in \{\pm 1, \pm 3\}$. To obtain the LUT entries a sliding window of length $2n + 1$ symbols is advanced through the training sequence. The known $2n+1$ sequence of symbols forms the LUT address and each time a specific symbol sequence appears, the amplitude difference for the center symbol is accumulated as the corresponding LUT entry. Counters ($C$) track the number of occurrences for each specific sequence allowing the average pre-distortions to be stored as the LUT entries upon completion of the training.

Since the pattern-dependent distortion is nonlinear and the values of adjacent symbols are changed when the sliding window advances through the training sequence, the correctness of the pre-distorted values can be improved by iterative calculations until convergence is reached [85]. In this study, three iterations are considered enough to achieve more performance improvement compared to other compensation algorithms.
In operation mode, the pre-distorted value replaces the center symbol of each $N$ sequence. Since the transmitted symbol sequence is known, the LUT address is always determined correctly. For a 16-QAM signal, the LUT has $4^N$ rows and 1 column. The NLPD does not require computational steps during operation. For each polarization signal, two one-dimensional LUTs are used to separately mitigate the amplitude distortions of the in-phase and quadrature signals.

To illustrate the LUT contents for NLPD, a 3-symbol LUT ($n = 1$) is used as an example. Figure 3.25 shows the average corrections of a LUT for each 3-symbol pattern after 3 iterations. The LUT entries are clearly pattern-dependent and range between -0.4 to 0.4 relative to the symbol values $\{\pm1, \pm3\}$.

To demonstrate the role of the symbols adjacent to the center symbol for 3-, 5- and 7-symbol sequences, the LUT pre-distortion for the center 3-symbol pattern $[1 \ 1 \ 3]$ is shown in Figure 3.26. The $4^4$ possibilities of the 7-symbol patterns $[c \ a \ -1 \ -1 \ -1 \ -3 \ b \ d]$
are numbered sequentially on the abscissa, where $a, b, c,$ and $d$ assumes all possible values $\{\pm1, \pm3\}$. The $4^2$ possibilities of the 5-symbol pattern $[a - 1 - 1 - 3 b]$ and 1 possibility for 3-symbol pattern $[-1 - 1 - 3]$ are also shown. Compared with the 5-symbol LUT patterns, the additional symbols $c$ and $d$ in the 7-symbol patterns need to be included in order to determine the amplitude of the pre-distortion accurately.

Figure 3.25: LUT entries of NLPD for 3-symbol patterns.
3.2. LUT BASED NONLINEAR PRE-DISTORTION

Figure 3.26: LUT entries for 3-symbol pattern \([-1 \ -1 \ -3]\), 5-symbol pattern \([a \ -1 \ -1 \ -3 \ b]\) and 7-symbol pattern \([c \ a \ -1 \ -1 \ -3 \ b \ d]\), respectively.

Figure 3.27: Pre-distortion distribution for the pattern \([-1 \ -1 \ -3]\) using the 3-symbol LUT.
3.3. **VOLterra NONLINEar EQUALIZATION**

Figure 3.27 indicates a histogram for the pre-distortion for the different occurrences of the symbol sequence \([-1 \ -1 \ -3]\) in the training phase. The variation about the average value of 0.4 is due to the pattern-dependent distortion. The variation is smaller for 5- and 7-symbol LUTs.

Constellation diagrams without and with 7-symbol LUT based NLPD\((n = 3)\) are shown in Figure 3.28. The 7-symbol LUT-NLPD yields a significant reduction in the error vector magnitude (EVM) from 14.6 % to 8.9 %.

![Constellation Diagrams](image)

Figure 3.28: Constellation diagrams for 402 Gb/a DP 16QAM signal obtained without LUT pre-distortion (left, EVM = 14.6 %), with 7-symbol LUT pre-distortion (right, EVM = 8.9 %).

### 3.3 Volterra Nonlinear Equalization

It has been shown that a Volterra series based nonlinear equalizer (VNLE) can compensate distortion introduced by fiber nonlinearity [81–84]. The nonlinear
3.3. VOLTERRA NONLINEAR EQUALIZATION

System identification based on Volterra models can be performed in the time domain [81, 83, 84] or frequency domain [82]. The discrete time formulation for the Volterra series expansion up to third-order for a nonlinear system with finite memory is given by [83]

$$y(n) = \sum_{k_1=0}^{N-1} h_1(k_1) x(n - k_1) + \sum_{k_1=0}^{N-1} \sum_{k_2=0}^{N-1} h_2(k_1, k_2) x(n - k_1) x(n - k_2) + \sum_{k_1=0}^{N-1} \sum_{k_2=0}^{N-1} \sum_{k_3=0}^{N-1} h_3(k_1, k_2, k_3) x(n - k_1) x(n - k_2) x(n - k_3)$$

where $x(n)$ and $y(n)$ are the input and output signals, respectively. $N = 2n + 1$ is the memory length. $h_1, h_2$ and $h_3$ are the linear, quadratic and cubic Volterra kernels, respectively. The total number of kernel coefficients for the VNLE is $L = N + N^2 + N^3$.

Notice that the Volterra series is a linear combination of Volterra kernels. As a result, when a Volterra model is used to model a nonlinear system, the Volterra kernels can be updated using the least mean square (LMS) or the recursive least square (RLS) methods. The kernel coefficients are obtained in the training phase by the RLS algorithm with highest available OSNR to compensate the pattern-dependent distortion at the receiver. One-dimensional Volterra equalizers are implemented to separately mitigate the distortion of the in-phase and quadrature signals for each polarization.

One major disadvantage of the Volterra series is its complexity. Consequently, it may not be feasible to apply a Volterra model based compensator in real-time signal processing applications. The VNLE compensation requires $L$ real multiplications.
for each evaluation of the output signal $y(n)$. The complexity of the VNLE can be reduced by only using a subset of the kernel coefficients with highest significance. The resulting Volterra model is referred to as a sparse Volterra model. The Gram-Schmidt orthogonalization technique enables a sparse-VNLE implementation with an adjustable performance-complexity trade-off [83].

3.4 Experimental Setup

The field of arbitrary electronic waveform generation has seen tremendous improvement with the development of high speed digital-to-analog converters (DACs) capable of producing over 65 GSa/s and more [88]. These technological advances came with the success of the integrated circuit industry in mastering complementary metal-oxide-semiconductor (CMOS) transistors down to 40 nm [89]. The second setup utilized a Keysight Technology M8195A 4-channel arbitrary waveform generator (AWG) to generate the desired signal. The AWG allows for the generation of arbitrary electrical signals with a sampling rate of 65 GSa/s and 8 bit DAC resolution per sample with 6.4 effective numbers of bits (ENOB) at 1 GHz which gradually decreases to 4.8 at 25 GHz [90].

The experimental setup is illustrated in Figure 3.29. Experimental results were obtained from a $3 \times 402$ Gb/s DP 16-QAM superchannel experiment with 50.25 Gbaud per subcarrier and 75 GHz subcarrier spacing. First, a $2^{15} - 1$ pseudo random bit sequence was mapped to 16-QAM symbols to generate the 402 Gb/s DP 16-QAM signals. Next, the symbol sequence was re-sampled to 1.29 sample/symbol to match the sampling rate of the DACs and root-raised cosine pulse shaping with a roll-off factor of 0.075 was performed. The AWG converted the digital waveforms to electrical
signals to drive two optical transmitters (SHF 46213D which includes drive amplifiers and an IQ modulator) with 25 GHz bandwidth IQ modulators. One modulator and polarization multiplexing emulator were dedicated for the center subcarrier, and another modulator and polarization multiplexing emulator were used to generate the two side channels. The two side subcarriers and the center channel were separately amplified and then combined.

Figure 3.29: Experimental setup. DAC: digital-to-analog converter; EDFA: erbium doped fiber amplifier; AOM: acousto-optic modulator; OBPF: optical band-pass filter; LSPS: loop synchronous polarization scrambler; GE: gain equalizer.

The superchannel signal was launched into a recirculating loop with four spans, a loop synchronous polarization scrambler (LSPS), and a gain equalizer (GE). Each span consisted of 75 km of standard SMF, an erbium doped fiber amplifier (EDFA) with a noise figure of 5 dB, and an optical band-pass filter (OBPF) with a bandwidth...
of 17.3 nm. The OBPFs prevented the EDFAs in the loop from being saturated by out-of-band amplified spontaneous emission noise. The received signal was amplified and filtered before detection by a polarization- and phase-diverse coherent receiver with 32 GHz bandwidth. The local oscillator laser had a nominal linewidth of 100 kHz. The four signals from balanced photodetectors were digitized by 80 GSa/s ADCs using two synchronized real-time sampling oscilloscopes. The sample values were match filtered and processed off-line.

DSP was performed off-line and included (i) compensation for quadrature imbalance [42], (ii) re-sampling to two samples per symbol using an interpolation algorithm, (iii) frequency domain equalization for chromatic dispersion in the case of fiber transmission [43], (iv) clock recovery using the digital square and filter technique [44], (v) polarization demultiplexing and compensation for residual signal distortion using 25-tap adaptive $2 \times 2$ equalizers with the CMA for pre-convergence and then the RDA [45], (vi) recovery of the carrier frequency offset using a spectral domain algorithm [48], (vii) recovery of the carrier phase using a sliding window two-stage simplified QPSK partitioning and QPSK constellation transformation algorithm [51]. The BER was obtained from decisions based on rectilinear decision boundaries. Corresponding parameters for these DSP algorithms were identical to those illustrated in Table 3.3

3.5 Results

For each polarization, the one-dimensional 4-ary in-phase and quadrature signals are separately processed by the VNLE, sparse-VNLE, MAP detector and NLPD to compensate the pattern-dependent distortion caused in the transmitter and receiver.
To determine the Volterra kernel coefficients and LUT entries for the NLPD and MAP detector, five sets of training sequence data (each consisting of 564,645 symbols) were used in a back-to-back system configuration with highest available OSNR (37.9 dB, 0.1 nm noise bandwidth). The five sets of kernel coefficients and LUT entries are then averaged. Kernel coefficients and LUT entries are obtained separately for the odd subcarriers and the even subcarrier and are used for all other system configurations (e.g., back-to-back systems for different OSNRs and transmission systems with different fiber lengths). Given the properties of the pattern-dependent distortion for the particular case considered here, \( N = 2n + 1 = 7 \) represents a good compromise between performance and complexity. The Volterra kernel coefficients and LUT entries depend on the modulation format and symbol rate. For the carrier phase recovery, the two-stage simplified QPSK partitioning and QPSK constellation transformation algorithm is implemented with a block-length of \( B \) for both stages [51]. For the back-to-back systems, \( B = 60 \) is the optimum value for the four compensation techniques.

A VNLE up to third-order with a filter memory of \( N = 7 \) symbols was implemented at the receiver. The role of the three kernels in compensating the pattern-dependent distortion is illustrated in Figure 3.30 in terms of the dependence of the BER (average BER for the three subcarriers) on OSNR. Without compensation, a BER floor larger than \( 10^{-3} \) is observed. Results without compensation and with a first-order and second-order VNLE are shown for comparison. The third-order VNLE with 399 kernel coefficients yields a significant improvement in the performance and outperforms first and second order VNLEs.
An orthogonal search was used to determine the most significant kernel coefficients and reduce the computational complexity of the VNLE. Figure 3.31 shows the quadratic kernel coefficients ($h_2$) for the VNLE (solid circle) and the corresponding sparse-VNLE (empty circle) for the in-phase signal. The amplitudes of the least significant kernel coefficients are set to zero and the number of coefficients is reduced to 155 and 203 coefficients for the in-phase and quadrature signals, respectively. For this case, the sparse-VNLE provides an average reduction in the number of kernel coefficients by 55% compared to the VNLE.
Figure 3.31: Quadratic kernel coefficients for the in-phase signal.

Figure 3.32 compares the back-to-back performance of the VNLE, sparse-VNLE and 7-symbol MAP detector. The required OSNR for BER $= 10^{-3}$ is reduced by about 0.9 and 1.2 dB for the MAP detector compared to the VNLE and sparse-VNLE, respectively. For the 7-symbol MAP detector, the number of real multiplications is $7 \times 4^7 = 114,688$ compared to $7 + 7^2 + 7^3 = 399$ for the VNLE. The complexity reduction for the sparse-VNLE yields a 0.3 dB penalty in required OSNR compared to the VNLE.
3.5. RESULTS

For a back-to-back system, Figure 3.33 shows the dependence of the BER on OSNR for the 7-symbol NLPD with one, two and three iterations (denoted 1i, 2i and 3i, respectively) in calculating the pre-distortion of the center symbol within each of the possible 7-symbol sequences. As expected, the BER improves as the number of iterations increases. Results for the LUT based MAP detector are shown for comparison. For the NLPD with three iterations, the required OSNR for $\text{BER} = 10^{-3}$ is 0.85 dB smaller than that for the MAP detector. Further improvement in the performance of NLPD is expected with more than three iterations.
Figure 3.33: Dependence of the BER on OSNR for a 7-symbol NLPD in a back-to-back system. Results are shown for one, two and three iterations in calculating the pre-distortions and without LUT based compensation.

For transmission over 1500 km of SMF, the dependencies of the BER on launch power are compared in Figure 3.34 for the four compensation techniques. The launch power ranges for a BER below the FEC threshold of $\text{BER} = 1.9 \times 10^{-2}$ are 2.6, 2.8 and 3.2 dB for the VNLE, MAP detector, and NLPD after 3 iterations, respectively. The launch power range is about 0.2 dB smaller for the sparse-VNLE compared to the VNLE. For launch powers of 3 and 4 dBm, the optimum CPR block-lengths $B$ are 72 and 60 for the VNLE and MAP detector, respectively. As the nonlinear effects become more dominant, a larger block-length is needed for the VNLE compared to the MAP detector.
The dependence of the BER on launch power at an optimum launch power of 2 dBm is illustrated in Figure 3.35. The VNLE, sparse-VNLE and MAP detector enable transmission over 1500 km of SMF. The BER values for 1500 km transmission are converted to corresponding OSNR values using equation 2.27. The VNLE and MAP detector provide additional OSNR margins of 0.16 and 0.47 dB relative to the sparse-VNLE for transmission over 1500 km. The improvement in performance for the LUT based NLPD with three iterations allows transmission over 1800 km of fiber for the 1.206 Tb/s DP 16-QAM superchannel signal.
3.6 Summary

In this Chapter, the performance of a single-carrier 448 Gb/s DP 16-QAM signal and a three-carrier 1.206 Tb/s DP 16-QAM superchannel signal was assessed using MAP detection to mitigate transmitter and receiver based pattern-dependent distortion. Four different decision rules were considered. For the single-carrier 448 Gb/s signal, a BER floor was not reached for the back-to-back system (at least within the constraints imposed by offline signal processing). A back-to-back OSNR sensitivity of 26.9 dB
3.6. SUMMARY

(BER = 10^{-3}) and transmission over 1200 km of SMF were achieved using 7-symbol minimum distance MAP detection. For the three-carrier 1.206 Tb/s superchannel signal, a back-to-back OSNR sensitivity of 26.4 dB (BER = 10^{-3}) and transmission over 1500 km of SMF with a spectral efficiency of 4.4 b/s/Hz were achieved. A launch power range of 2.7 dB was obtained and the achievable transmission distance increased by about 7.6% with 7-symbol minimum distance MAP detection.

The performance of the 1.206 Tb/s three-carrier DP 16-QAM superchannel signal was demonstrated by evaluating the MI for the LUT-MAP detector. The 7-symbol minimum distance LUT-MAP detector provided an increase in the MI of around 0.3 bits/symbol per polarization in the transmission systems.

Four techniques for mitigating the pattern-dependent were compared in the context of a 1.206 Tb/s DP 16-QAM superchannel signal with three subcarriers. The best performance was achieved by the NLPD followed by the MAP detector, VNLE and sparse-VNLE. The NLPD does not require online computations. At the FEC threshold of 1.9 \times 10^{-2}, the NLPD provided an increase in the transmission distance of 300 km compared to the sparse-VNLE, VNLE and MAP detector. For transmission over 1500 km of SMF, the VNLE and MAP detector exhibited additional OSNR margins of 0.16 dB and 0.47 dB relative to the sparse-VNLE. Compared to the VNLE, the sparse-VNLE exhibited a 55% reduction in number of kernel coefficients.
Chapter 4

Nonlinear Phase Noise in Terabit Nyquist-Superchannel

To answer the demand for a ceaseless increase of capacity, high spectral efficiency and high data-rate systems have been the focus of significant research. Recently, the generation of single carrier DP 32-QAM with 124 Gbaud (1.24 Tb/s) using parallel generation and a combination of four spectral sub-bands has been demonstrated [91]. Even with the most recent components, sending 1 Tb/s over a single carrier would face limiting factors such as poor sensitivity, intolerance to fiber nonlinearities, hardware bandwidth limitations, and the need for ultra-fast digital-to-analog converters (DACs) at the transmitter. The use of superchannels can overcome such limitations to achieve very high interface rates.

For Nyquist-superchannels, the nonlinear interaction between the subcarriers, called inter-subcarrier nonlinearity (i.e., XPM), during fiber transmission is a key limiting factor [24, 25]. While intra-subcarrier nonlinearity (i.e., SPM) can be compensated by DSP algorithms such as digital back-propagation at the
receiver [43, 92] or predistortion at the transmitter [93], similar approaches to compensate inter-subcarrier XPM are considered impractical.

A joint compensation of inter-subcarrier XPM was proposed in [26, 27, 94]. This approach consists of detecting the adjacent subcarriers, and removing the inter-subcarrier XPM effect from the subcarrier of interest. Due to high data transmission rates and current limitations of the application-specific integrated circuit (ASIC) technology, sharing field information of multiple subcarriers with precise timing between several coherent receivers is almost impossible. In addition, dealing with inter-subcarrier XPM distortion is critical because of the fact that the polarization states of different subcarriers evolve randomly in the presence of polarization mode dispersion.

Recently, the modeling of inter-subcarrier induced distortion in coherent systems has received extensive attention [28–34]. In [28] the impact of inter-subcarrier nonlinearity is described in terms of an XPM-induced phase noise that is modeled as a time-varying stochastic process. In [29] the nonlinear interference noise in dispersion-uncompensated transmission systems is modeled as an additive white Gaussian noise process with the variance determined by a perturbation technique. In [30] an analytical expression for the variance of a probe Gaussian pulse due to its interaction with other channels in a DWDM system is developed and validated numerically. These models are extremely useful in the quest to mitigate the effects of XPM through transmission system design and coherent receiver digital signal processing. These models allow for tracking and compensation of the inter-subcarrier XPM by employing approximate stochastic filtering methods such as the extended Kalman filter (EKF) [31] or machine learning algorithms [95].
4.1 IMPACT OF INTRA- AND INTER-SUBCARRIER NONLINEARITIES

In this Chapter, the impact of the number of subcarriers on the relative strength of the intra- and inter-subcarrier nonlinearities and on the information rate is investigated for 1.206 Tb/s superchannel systems through simulation. Moreover, the nonlinear phase perturbation induced by inter-subcarrier XPM is quantified for the center subcarrier in a Nyquist-superchannel system. The degree of correlation for the nonlinear phase perturbation is determined between a modulated subcarrier and a continuous wave (CW) probe signal.

4.1 Impact of Intra- and Inter-Subcarrier Nonlinearities

Recently, it has been shown that by using multiple low baud subcarriers instead of single carrier high baud signal, the fiber nonlinearity tolerance can be improved [94, 96–99]. The dependence of the multi-carrier performance and maximum system reach on the per carrier symbol rate has been considered based on numerical simulations [96–98] and experiments [94, 99]. The improvement in nonlinearity tolerance benefiting from subcarrier multiplexing can be explained by the walk-off between subcarriers due to chromatic dispersion [100].

For a 1.206 Tb/s DP 16-QAM superchannel, the impact of the number of subcarriers on the intra- and inter-subcarrier nonlinearities and on the maximum achievable transmission distance for an information rate of 1 Tb/s is determined. The superchannel bit rate and spectral efficiency are kept constant for 3, 5 and 9 subcarriers which takes into consideration practical constraints for such systems. The performance is quantified in terms of the BER and information rate. With an
overhead of 15% for FEC coding the maximum achievable transmission distance is determined as a function of the number of subcarriers. The superchannel information rate is the summation of products of the mutual information between the transmitted and received symbol sequences and the symbol rate for each subcarrier.

### 4.1.1 Simulation Setup

The numerical simulation was first performed with the commercial software package VPItransmissionMaker V9.3 to generate and transmit the optical signal. After coherent detection the saved data were processed by MATLAB. VPItransmissionMaker provides a fiber module called UniversalFiber. A user can activate the nonlinear effects in fiber transmission selectively. Figure 4.1 shows the parameter editor of the UniversalFiber module.

![Parameter editor of the UniversalFiber module in VPItransmissionMaker.](image)

VPItransmissionMaker provides a unique visual photonic design environment that
allows a user to create a new schematic diagram, open an existing schematic, place module icons and link them together, run the simulation, save and display results. In VPITransmissionMaker, a user’s access is limited to the parameters of each photonic and electronic module.

The numerical simulation was also performed using Optilux [101], an open source collection of tools. Optilux is implemented as a Matlab toolbox. Hence, any piece of code is completely accessible. The whole toolbox is composed of free software which a user has a complete control over simulation parameters. Optilux has a more user-friendly interface compared to VPITransmissionMaker. Working with Optilux is not complex, but instead is straightforward and a user has quick access to common features or commands. All toolboxes are well-organized making it easy to locate different tools and options.

In Optilux the NLSE for electrical filed of multi-subcarrier can be solved for a unique field or separate field. The unique field solution is the most complete one and it is more reliable compared with separate filed method [101].

For a superchannel bit rate of 1.206 Tb/s, configurations with 3, 5 and 9 subcarriers were considered, each modulated at symbol rates of 50.25 Gbaud, 30.15 Gbaud and 16.75 Gbaud, respectively. Root-raised-cosine pulse shaping with a roll off factor equal to 0.075 was applied to each subcarrier. Each superchannel configuration has a constant spectral efficiency $R/\Delta f$ (where $R$ is the bit rate and $\Delta f$ is the subcarrier frequency spacing, respectively) and a total bandwidth occupancy of 162.5 GHz as summarized in Table 4.1.
4.1. IMPACT OF INTRA- AND INTER-SUBCARRIER NONLINEARITIES

Table 4.1: Simulated system configurations.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>System 1</th>
<th>System 2</th>
<th>System 3</th>
</tr>
</thead>
<tbody>
<tr>
<td>Baud (Gbaud)</td>
<td>50.25</td>
<td>30.15</td>
<td>16.75</td>
</tr>
<tr>
<td>Number of subcarriers</td>
<td>3</td>
<td>5</td>
<td>9</td>
</tr>
<tr>
<td>Subcarrier spacing (GHz)</td>
<td>54.167</td>
<td>32.5</td>
<td>18.05</td>
</tr>
<tr>
<td>Bandwidth occupancy (GHz)</td>
<td>162.5</td>
<td>162.5</td>
<td>162.5</td>
</tr>
</tbody>
</table>

The linewidths of the transmitter lasers were 100 kHz. The laser wavelength for the center subcarrier was 1552.524 nm. A $2^{15} - 1$ de-Bruijn bit sequence was used for the bit-to-symbol mapping. For the two polarization signals, the de-Bruijn bit sequences were de-correlated by 128 bits before the bit-to-symbol mapping. For the subcarriers, the bit sequences were de-correlated by 300 bits. For simplicity the effects of polarization mode dispersion were neglected. The transmitted signals were applied to a recirculating loop. The typical system parameters and specification of components used in all simulations are given in Table 4.2.
4.1. IMPACT OF INTRA- AND INTER-SUBCARRIER NONLINEARITIES

Table 4.2: System parameters used in simulation.

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Specification</th>
</tr>
</thead>
<tbody>
<tr>
<td>Fiber</td>
<td>SMF</td>
</tr>
<tr>
<td>Distance</td>
<td>1500 km</td>
</tr>
<tr>
<td>Span length</td>
<td>75 km</td>
</tr>
<tr>
<td>DCF</td>
<td>none</td>
</tr>
<tr>
<td>Amplifier</td>
<td>EDFA</td>
</tr>
<tr>
<td>Gain</td>
<td>15 dB</td>
</tr>
<tr>
<td>Noise figure</td>
<td>4.5 dB</td>
</tr>
<tr>
<td>Dispersion coefficient</td>
<td>17 ps/nm/km</td>
</tr>
<tr>
<td>Dispersion slope</td>
<td>0 ps/nm²/km</td>
</tr>
<tr>
<td>Fiber effective area</td>
<td>80 µm²</td>
</tr>
<tr>
<td>Nonlinear index</td>
<td>$2.6 \times 10^{-20}$ m²/W</td>
</tr>
<tr>
<td>Field (Optilux results)</td>
<td>unique</td>
</tr>
</tbody>
</table>

For the intradyne coherent detection, the frequency offset between the transmitter and local oscillator lasers was set to 100 MHz and the linewidth of the local oscillator laser was 100 kHz. The performance of each subchannel was obtained by adjusting the frequency of the local oscillator laser.

The off-line signal processing after coherent detection included (i) matched filtering to maximize the signal-to-noise ratio, (ii) re-sampling, (iii) fixed frequency domain equalization for chromatic dispersion [45], (iv) digital square and filter clock recovery [44], (v) polarization recovery and residual distortion compensation using 13-tap adaptive equalizers in a butterfly configuration [45], (vi) carrier frequency
4.1. IMPACT OF INTRA- AND INTER-SUBCARRIER NONLINEARITIES

offset recovery using a spectral domain algorithm [58], (vii) carrier phase recovery using the blind phase search algorithm [50], and (viii) symbol decisions. The BER was obtained by direct bit error counting using rectilinear decision boundaries. Corresponding parameters for these DSP algorithms are listed in Table 4.3.

<table>
<thead>
<tr>
<th>Algorithms</th>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Dispersion compensation</td>
<td>Method</td>
<td>frequency domain</td>
</tr>
<tr>
<td>Adaptive equalizer</td>
<td>Method</td>
<td>CMA-RDA</td>
</tr>
<tr>
<td></td>
<td>Convergence</td>
<td>$1 \times 10^{-6}$</td>
</tr>
<tr>
<td></td>
<td>parameters</td>
<td>and $1.2^{-5}$</td>
</tr>
<tr>
<td></td>
<td>Number of</td>
<td>13</td>
</tr>
<tr>
<td></td>
<td>taps</td>
<td></td>
</tr>
<tr>
<td></td>
<td>Number of</td>
<td>25</td>
</tr>
<tr>
<td></td>
<td>iterations</td>
<td></td>
</tr>
<tr>
<td>Carrier phase recovery</td>
<td>Block length</td>
<td>optimized</td>
</tr>
<tr>
<td></td>
<td>Number of</td>
<td>32</td>
</tr>
<tr>
<td></td>
<td>test phases</td>
<td></td>
</tr>
</tbody>
</table>

4.1.2 Results

The simulation was first performed by VPItransmissionMaker. The dependence of the BER (average for all subcarriers) and the superchannel information rate on the per subcarrier launch power for transmission over 1500 km of SMF are shown in Figure 4.2 for 3, 5 and 9 subcarriers. All three configurations have a bandwidth occupancy of 162.5 GHz and transmission bit rate of 1.206 Tb/s.
4.1. IMPACT OF INTRA- AND INTER-SUBCARRIER NONLINEARITIES

The BER and information rate results have the well-known dependence on launch power, where the performance is limited by ASE noise at low launch powers and by fiber nonlinearities at high launch powers. For 3, 5 and 9 subcarriers the optimum launch powers are 0, -2 and -5 dBm, respectively. At optimum launch power, the 9-subcarrier configuration provides a total information rate that is 31 Gb/s (2.6%) and 46 Gb/s (3.8%) higher than that for the 5- and 3-subcarrier configurations, respectively.

The simulation parameters were also chosen to investigate the relative importance of the nonlinear effects by activating inter- and intra-subcarrier nonlinearities as the...
number of subcarriers and hence baud rate per subcarrier were changed.

To investigate the impact of the intra- and inter-subcarrier nonlinearities, the simulations were performed with only SPM (label Intra-subcarrier), only XPM (label Inter-subcarrier), and with both SPM and XPM (label Superchannel). Figures 4.3, 4.4 and 4.5 illustrate the dependence of the BER (average of all subcarriers) and superchannel information rate on the launch power after 1500 km of SMF when the nonlinearities are selectively chosen for 3, 5 and 9 subcarriers each modulated at 50.25 Gbaud, 30.15 Gbaud and 16.75 Gbaud, respectively.

Figure 4.3: BER and IR versus launch power for a 1.206 Tb/s superchannel after 1500 km transmission for 3 subcarriers modulated at 50.25 Gbaud.
4.1. IMPACT OF INTRA- AND INTER-SUBCARRIER NONLINEARITIES

Figure 4.4: BER and IR versus launch power for a 1.206 Tb/s superchannel after 1500 km transmission for 5 subcarriers modulated at 30.15 Gbaud.

Figure 4.5: BER and IR versus launch power for a 1.206 Tb/s superchannel after 1500 km transmission for 9 subcarriers modulated at 16.75 Gbaud.
4.1. IMPACT OF INTRA- AND INTER-SUBCARRIER NONLINEARITIES

The results demonstrate that at optimum launch power the intra-subcarrier effect is somewhat larger than inter-subcarrier effect for 3 and 5 subcarriers. When the number of subcarriers or launch power increases, the nonlinear interference among the subcarriers becomes more significant than that for the intra-subcarrier nonlinearities.

The dependence of the superchannel information rate at optimum launch power on fiber length for each configuration is shown in Figure 4.6. For an information rate of 1 Tb/s, by interpolation the 9-subcarrier configuration provides an increase in the achievable transmission distance of 102 km (5.9%) and 218 km (12.7%) compared to the 5- and 3-subcarrier configurations, respectively.

Figure 4.6: Dependence of 1.206 Tb/s superchannel information rate on fiber length for 3, 5 and 9 subcarriers.
4.1. IMPACT OF INTRA- AND INTER-SUBCARRIER NONLINEARITIES

The rest of the numerical investigations of this Chapter are based on the Optilux software, however the previous results will be used as reference values.

In Optilux, the type of propagation in a fiber is governed by a string of four characters. The first character is 'g' if GVD (i.e. $\beta_2$, $\beta_3$) is on or '-' in the absence of GVD. The second character is 'p' for propagation of a polarized field in presence of birefringence and PMD or '-' in the absence of such effects. The third is 's' if SPM is on or '-' in the absence of SPM. Likewise, the fourth character is 'x' or '-' in the presence/absence of XPM. The most complete case is 'gpsx' and corresponds to propagation in the presence of fiber with GVD+PMD+SPM+XPM.

Recently, it has been shown that the center subcarriers suffer from larger nonlinear phase perturbation, induced by inter-subcarrier nonlinearities, than the edge subcarriers in superchannels with more than 2 subcarriers [24]. The behaviour of the intra-subcarrier and inter-subcarrier nonlinearities on the center subcarrier was investigated for variation in the number of subcarriers and launch power per subcarrier. In unique field, simulations were performed for a multi-subcarrier superchannel in the presence of inter- and intra-subcarrier nonlinear effects and the signal after CPR was saved. Then, simulations were done for a single-subcarrier with intra-subcarrier effect. The difference between these two saved signals is an approximation of nonlinear phase perturbation on the center subcarrier due to inter-subcarrier nonlinearities (e.g., XPM). The effect of FWM was neglected. The random number generator seed was set to make the results repeatable. Figure 4.7 illustrates the dependence of the standard deviation of the nonlinear phase perturbation induced by inter-subcarrier nonlinearity ($\sigma_{XPM}$) on an increase above the corresponding optimum launch power for 3, 5 and 9 subcarriers.
4.1. IMPACT OF INTRA- AND INTER-SUBCARRIER NONLINEARITIES

Figure 4.7: Standard deviation of nonlinear phase perturbation induced by inter-subcarrier on center subcarrier versus increase above optimum launch power.

Although the $\sigma_{XPM}$ increases with an increase in the number of subcarriers and launch power, the superchannel BER performance of 9 subcarriers setup outperforms 5 and 3 subcarriers for each increased power above the optimum launch power. These results are consistent with several analytical, simulation and experimental papers recently reported on the existence of an optimum symbol rate that minimizes the overall nonlinear phase perturbation generated during signal propagation in a fiber [97–99,102].

Figure 4.8 illustrates the dependence of the superchannel BER after 1500 km of SMF on the standard deviation of the phase perturbation induced by
inter-subcarrier nonlinearity on the center subcarrier for 3, 5 and 9 subcarriers. For BER = 10^{-2}, by interpolation the 9-subcarrier configuration provides an increase in the nonlinear phase perturbation-tolerance, induced by inter-subcarrier XPM, of 0.07 Rad/\pi (58\%) and 0.047 Rad/\pi (39\%) compared to the 3- and 5-subcarrier configurations, respectively.

Figure 4.8: Superchannel BER versus standard deviation of phase perturbation on center subcarrier induced by inter-subcarrier.

For a subset of 1 sample per symbol of the received signal after carrier phase recovery, Figure 4.9 illustrates the trajectory of inter-subcarrier nonlinear phase noise of the center subcarrier. The standard deviations of nonlinear phase trajectory are 0.027\pi Rad, 0.037\pi Rad and 0.046\pi Rad for 3, 5, and 9 subcarriers, respectively.
Figure 4.9: Trajectory of nonlinear phase perturbation of center subcarrier induced by the inter-subcarrier XPM after 1500 km fiber.
4.1. IMPACT OF INTRA- AND INTER-SUBCARRIER NONLINEARITIES

The auto-correlation function (ACF) allows us to quantify the residual memory after propagation and DSP. If time correlations are found, they could be exploited by a subsequent algorithm to achieve an improvement in system performance. Recent studies have shown that the inter-subcarrier phase perturbation can be correlated over time under certain conditions [28, 32, 103–106]. These papers determined the ACF of the averaged phase perturbation based on the block-wise constant model, but it is an unrealistic approach for practical systems. These models are useful in order to track and model the inter-subcarrier XPM. Figures 4.10, 4.11 and 4.12 indicate the ACF of XPM-induced nonlinear phase noise on center subcarrier after 1500 km transmission for 3, 5 and 9 subcarriers setups, respectively. The AUTOCORR function of MATLAB was used for calculating the auto-correlation function (available in Econometrics Toolbox).

Figure 4.10: ACF of XPM-induced phase noise on center subcarrier in 3 subcarriers.
4.1. IMPACT OF INTRA- AND INTER-SUBCARRIER NONLINEARITIES

![Figure 4.11: ACF of XPM-induced phase noise of center subcarrier in 5 subcarriers.](image1)

![Figure 4.12: ACF of XPM-induced phase noise of center subcarrier in 9 subcarriers.](image2)
The partial ACF of the XPM-induced phase noise is shown as insets in Figures 4.10, 4.11 and 4.12. The fact that ACF is non-zero shows that there is indeed time correlation in XPM-induced nonlinear phase noise. The amount of memory increases with the number of co-propagating subcarriers.

While intra-subcarrier nonlinearity can be compensated by means of back-propagation, inter-subcarrier nonlinearity is treated as noise in the literature [32, 103, 104]. The modeling of inter-subcarrier nonlinear interference noise has therefore come to be one of the most important and extensively studied in the field of optical communications. One modeling approach that has attracted much attention in recent years is the so called Gaussian noise model [40] which assumes that inter-subcarrier nonlinearity is additive white Gaussian noise (AWGN). The motivation for this assumption comes mainly from the claim that in dispersion uncompensated coherent systems, the high accumulated dispersion allows considering the dispersed optical signal, as well the inter-subcarrier nonlinearity that accompanies it, as a white Gaussian process. By additionally assuming that inter-subcarrier nonlinearity is additive, one ends up with a simple and easy to use model that only requires the calculation of the inter-subcarrier nonlinearity variance [104].

4.2 Correlation of Modulated and Probe Signals

The use of narrow WDM channel spacing and high channel bit-rates lead to inter-subcarrier XPM nonlinearity presenting one of the main obstacles to achieving large transmission distances. It would be very useful if the XPM-induced penalties could be accurately estimated for a given system by measuring XPM nonlinearity of
a CW probe channel by modulated pump channels. This would lead to accurate experimental characterization of the system, isolating the effects of XPM from other impairments. Additionally, this allows us to calculate system penalties using straightforward analytical expressions.

In this research, simulation results quantify the degree of correlation for the phase perturbation induced by inter-subcarrier XPM between a center-modulated channel (denoted center-modulated) and a CW center channel (denoted center-probe) in a Nyquist-superchannel system. The aim is to determine the extent to which reliable statistical information about the inter-subcarrier nonlinear phase perturbation can be obtained from a probe signal either through simulation or experiment.

### 4.2.1 Simulation Setup

The assessment of the phase perturbation for both the center-modulated and center-probe subcarrier signals uses identical system setups (including noise seeds) so that the differences in the inter-subcarrier nonlinearity, i.e. XPM, between the two received signals are solely due to the type of center subcarrier (modulated or CW).

For a superchannel bit rate of 1.206 Tb/s (15% FEC coding overhead), configurations with 3, 5 and 9 subcarriers were considered each modulated at symbol rates of 50.25 Gbaud, 30.15 Gbaud and 16.75 Gbaud, respectively. The superchannel configurations are identical to those explained in the previous section except for the center-probe subcarrier signal the center subcarrier was not modulated. The signal processing included the same DSP algorithms applied in section 4.1.1. Corresponding parameters for the DSP algorithms were also identical to those shown in Table 4.3.
4.2.2 Results

To investigate the amount of phase perturbation induced by inter-subcarrier XPM on the center-modulated and center-probe subcarrier in superchannel signals, the simulations were performed for a multi-subcarrier superchannel signal with both inter- and intra-subcarrier effects and also for single-subcarrier with only intra-subcarrier. For each simulation, the received signal was saved after DSP in the coherent receiver.

For a subset of 2 samples per symbol of the received signal, Figure 4.13 illustrates the inter-subcarrier nonlinear phase fluctuation for the center-modulated and center-probe signals at high launch powers (3 dB above optimum launch power of each setup). The phase perturbations are offset by ± 0.05 rad/π for clarity. The results illustrate the correlation, albeit imperfect, between the center-modulated and center-probe signals.

For the case of 5 subcarriers, Figures 4.14 and 4.15 superimpose all the phase perturbations induced by the inter-subcarrier XPM as a function of launch power (X and Y polarizations). The results are for the center-modulated and center-probe signals, respectively. The center probe signal underestimates the phase perturbation to some extent. The standard deviation of the phase variations induced by inter-subcarrier XPM on the center-modulated/-probe signal is shown in Figure 4.16. Increasing the launch power per subcarrier increases the standard deviation of the phase perturbation induced by inter-subcarrier XPM. Figure 4.17 illustrates the correlation coefficients for the phase perturbations between the center-modulated and center-probe signals. The correlation exceeds 0.9 for launch powers up to 3 dB above the corresponding optimum launch powers.
4.2. CORRELATION OF MODULATED AND PROBE SIGNALS

Figure 4.13: Temporal phase fluctuation of center subcarrier induced by the inter-subcarrier XPM after 1500 km fiber at high launch powers for 3, 5, and 9 subcarriers.
Figure 4.14: Phase perturbation versus launch power for 5 subcarriers center-modulated.

Figure 4.15: Phase perturbation versus launch power for 5 subcarriers center-probe.
4.2. CORRELATION OF MODULATED AND PROBE SIGNALS

Figure 4.16: Standard deviation of phase perturbation induced by the inter-subcarrier XPM versus launch power.

Figure 4.17: Correlation coefficient for the phase perturbation induced by the inter-subcarrier XPM between the center-modulated and center-probe signals.
The aforementioned results provide an indication of the reliability of statistical information about the phase perturbation that can be obtained from a probe signal in a Nyquist-superchannel either through simulation or experiment.

4.3 Summary

The impact of the inter- and intra-subcarrier nonlinearities on the performance of a 1.206 Tb/s DP 16-QAM superchannel was evaluated in terms of the BER and information rate. At a fixed spectral efficiency of 7.4 b/s/Hz, an increase in the number of subcarriers or launch power causes the inter-subcarrier nonlinearities to increase in relative importance compared to the intra-subcarrier nonlinearity. For transmission over 1500 km, the 9-subcarrier configuration provided a total information rate that is 2.6% and 3.8% higher than the 5- and 3-subcarrier configurations, respectively. For an information rate of 1 Tb/s, the 9-subcarrier configuration yielded an increase in the achievable transmission distance of 5.9% and 12.7% compared to the 5- and 3-subcarrier configurations, respectively.

The results indicated that with an increase in launch power or the number of subcarriers, the standard deviation of the phase perturbation of center subcarrier becomes more significant than that for the intra-subcarrier nonlinearity. However, the 9-subcarrier configuration provided an increase in the maximum nonlinear phase perturbation-tolerance of 58% and 39% compared to the 5- and 3-subcarrier configurations, respectively.

For a 1.206 Tb/s Nyquist-superchannel using 3, 5, and 9 subcarriers, the degree of correlation for the inter-subcarrier nonlinear phase perturbation between center-modulated and center-probe subcarrier signals has been quantified. The
correlation between a center-modulated subcarrier signal and center-probe subcarrier signal exceeds 0.9 for per-subcarrier launch powers up to 3 dB above the corresponding optimum launch power.
Chapter 5

Conclusions and Future Works

5.1 Conclusions

In this research, the performance of a single-carrier 448 Gb/s DP 16-QAM signal and a three-carrier 1.206 Tb/s DP 16-QAM superchannel signal was assessed using MAP detection to mitigate transmitter and receiver based pattern-dependent distortion. Different decision rules for the MAP detector were considered based on multiple observations of the same symbol as the detection window advances through the received symbol sequence. For the single-carrier 448 Gb/s signal, a BER floor was not observed for the back-to-back system (at least within the constraints imposed by offline signal processing). A back-to-back OSNR sensitivity of 26.9 dB (BER = 10\(^{-3}\)) and transmission over 1200 km of SMF were achieved using 7-symbol minimum distance MAP detection. For the three-carrier 1.206 Tb/s superchannel signal, a back-to-back OSNR sensitivity of 26.4 dB (BER = 10\(^{-3}\)) and transmission over 1500 km of SMF with a spectral efficiency of 4.4 b/s/Hz were achieved. A launch power range of 2.7 dB was obtained and the achievable transmission distance
increased by about 7.6% with 7-symbol minimum distance MAP detection.

The performance of 1.206 Tb/s three-carrier DP 16-QAM superchannel was demonstrated by evaluating MI using LUT-MAP detector. The 7-symbol minimum distance LUT-MAP detector provided an increase in the MI of about 0.3 bits/symbol per polarization in the transmission systems.

Four techniques for mitigating the pattern-dependent distortion were compared in the context of a 1.206 Tb/s DP 16-QAM superchannel signal with three subcarriers. The LUT based NLPD with iterative calculation of the pre-distortion provided the best performance in back-to-back systems and transmission systems followed by the MAP detector, VNLE and sparse-VNLE. The NLPD does not require online computations. At the FEC threshold of \( 1.9 \times 10^{-2} \), the LUT based NLPD provided an increase in the transmission distance of 300 km compared to the sparse-VNLE, VNLE and MAP detector. For transmission over 1500 km of SMF, the VNLE and MAP detector exhibited additional OSNR margins of 0.16 dB and 0.47 dB relative to the sparse-VNLE. Compared to the VNLE, the sparse-VNLE exhibited a 55% reduction in number of kernel coefficients.

In the context of Nyquist-superchannel, for a 1.206 Tb/s DP 16-QAM superchannel, the impact of the number of subcarriers on the intra- and inter-subcarrier nonlinearities and on the maximum achievable transmission distance for an information rate of 1 Tb/s was determined by simulation. The superchannel bit rate and spectral efficiency of 7.4 b/s/Hz were kept constant for 3, 5 and 9 subcarriers which takes into consideration practical constraints for such systems. An increase in the number of subcarriers caused the inter-subcarrier nonlinearities to increase in relative importance compared to the intra-subcarrier
nonlinearity. For transmission over 1500 km, the 9-subcarrier configuration provided a total information rate that is 2.6% and 3.8% higher than the 5- and 3-subcarrier configurations, respectively. For an information rate of 1 Tb/s, the 9-subcarrier configuration yielded an increase in the achievable transmission distance of 5.9% and 12.7% compared to the 5- and 3-subcarrier configurations, respectively. Moreover, the 9-subcarrier configuration increased the maximum nonlinear phase perturbation-tolerance by 58% and 39% compared to the 5- and 3-subcarrier configurations, respectively.

The degree of correlation between center-modulated and center-probe subcarrier signals was quantified to determine the extent to which reliable statistical information about the inter-subcarrier nonlinear phase perturbation can be obtained from a probe signal either through simulation or experiment. The results illustrated the correlation, albeit imperfect, between the center-modulated and center-probe signals. The correlation exceeded 0.9 for per-subcarrier launch powers up to 3 dB above the corresponding optimum launch power. The results provide an indication of the reliability of statistical information about the phase perturbation that can be obtained from a probe signal in a Nyquist-superchannel either through simulation or experiment.

5.2 Future Work

5.2.1 Post-FEC Performance
As FEC technology continues to evolve it becomes increasingly difficult to consider transmission performance isolated from coding. Therefore, one fruitful area for future
research might be to investigate gains in achievable transmission reach as a function of correctable BER. Is it possible to gain more transmission distance by increasing FEC overhead? How does the implementation complexity compare between compensation algorithms and increased FEC overhead?

5.2.2 Probabilistic Constellation Shaping

Recently, probabilistic constellation shaping (PCS) has been proposed for WDM systems to increase the tolerance to nonlinear fiber effects \[107, 108\]. The PCS scheme cleverly uses constellation points with high amplitude less frequently than those with lower amplitude to transmit signals that, on average, are more resilient to noise and other impairments. In the context of high baud superchannels, it could be useful to investigate the pattern-dependent distortion tolerance for modulation based on PCS.

5.2.3 Experimental Validation

In the context of nonlinear phase noise in Terabit per second Nyquist-superchannels, the most straightforward next step would be the experimental validation of the results presented in Chapter 4. In simulation, there is flexibility to activate the inter- and intra-subcarrier nonlinearities selectively, whereas in experiments the nonlinear effects are inseparable.

In simulation, the effects of polarization mode dispersion were neglected for simplicity. It would be interesting to see how the XPolM evolves in conjunction of the nonlinear phase noise generated by inter-subcarrier XPM.
Bibliography


Appendix A

List of Equipments Used in Experiments

The main equipments used in experimental setup are listed as follows:

A.1 Transmitter

- Arbitrary Waveform Generator - M8195A - Keysight
- Bit Pattern Generator - 12103A
- Spectrum Analyzer - E4448A - Agilent
- Optical Modulator - 46213D

A.2 Optical

- Tunable Laser - N7714A - Agilent
A.3. **RECEIVER**

- Polarization Mux Emulator - Kylia
- Optical Filter - 4000S - Finisar Waveshaper
- Optical Amplifiers - JDS and PriTel
- Optical Attenuator - JDS

## A.3 Receiver

- Optical Spectrum Analyzer - Q8384 - Advantest
- Digital Signal Analyzer - DSA X93204A - Agilent