Fiber-Optic Systems with Coherent Detection  
and Four-Dimensional Modulation

MARTIN SJÖDIN

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Photonics Laboratory, Department of Microtechnology and Nanoscience,  
Chalmers University of Technology, SE-412 96 Göteborg, Sweden
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Abstract

To increase the data rates of fiber-optic communication systems, modulation formats with higher spectral efficiency and/or higher power efficiency than binary formats are of great interest. These formats require coherent detection and this thesis has been dedicated to investigating both novel modulation formats and two different types of coherent systems.

Apart from intradyne systems with a free-running local oscillator laser in the receiver, we have investigated self-homodyne coherent detection, in which the phase reference is a polarization-multiplexed pilot tone. The optical signal-to-noise ratio requirement of self-homodyne systems has been quantified and compared with intradyne systems. Using both experiments and numerical simulations, we also demonstrated the unique ability of self-homodyne systems to compensate for nonlinear distortion due to the Kerr effect and a scheme to increase the spectral efficiency in WDM transmission.

The thesis also explores power-efficient modulation formats in intradyne systems. Modulation formats with higher power efficiency and comparable spectral efficiency than polarization-multiplexed quadrature phase shift keying (PM-QPSK) and polarization-multiplexed 16-ary quadrature amplitude modulation (PM-16-QAM), can be obtained by optimizing constellations in the four-dimensional (4-D) signal space of the optical carrier. Polarization-switched QPSK (PS-QPSK) recently emerged as the most power-efficient modulation format in four dimensions. We were the first to generate PS-QPSK experimentally and we also demonstrated an algorithm for equalization and polarization demultiplexing. In addition, we investigated single channel and WDM transmission with both numerical simulations and a loop experiment, and proposed a scheme for differential coding.

Another interesting 4-D constellation is set-partitioning 128 PM-16-QAM (128-SP-QAM), exhibiting almost the same spectral efficiency as PM-16-QAM and higher power efficiency. We performed a numerical study of 128-SP-QAM and found that the transmission distance can be increased with more than 40% compared to PM-16-QAM, that the tolerance to laser phase noise is similar, and that differential data encoding can be used without joint phase estimation.

Finally, we demonstrate that a relative amplitude scaling between the symbols with even and the symbols with odd parity in Gray-coded polarization-multiplexed NPSK yields a new class of modulation formats, with higher asymptotic power efficiency than the original constellations. A 16-level format with 0.44 dB gain over PM-QPSK is obtained for an amplitude scaling equal to the golden ratio.

Keywords: coherent detection, self-homodyne, pilot tone, intradyne, 16-ary quadrature amplitude modulation, quadrature phase-shift keying (QPSK), polarization-switched QPSK, 128-SP-QAM, self-phase modulation, wavelength-division multiplexing, polarization multiplexing.
Thesis for the degree of Doctor of Philosophy

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Photonics Laboratory, Department of Microtechnology and Nanoscience,
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Phone: +46 (0) 31 772 1000

Front cover illustration: Three-dimensional projection of the constellation of
subset-optimized polarization-multiplexed quadrature phase-shift keying. The figure is
made by Erik Agrell from the Department of Signals and Systems.

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Finally, we demonstrate that a relative amplitude scaling between the symbols with even and the symbols with odd parity in Gray-coded polarization-multiplexed $N$PSK yields a new class of modulation formats, with higher asymptotic power efficiency than the original constellations. A 16-level format with 0.44 dB gain over PM-QPSK is obtained for an amplitude scaling equal to the golden ratio.

Keywords: coherent detection, self-homodyne, pilot tone, intradyne, 16-ary quadrature amplitude modulation, quadrature phase-shift keying (QPSK), polarization-switched QPSK, 128-SP-QAM, self-phase modulation, wavelength-division multiplexing, polarization multiplexing.
List of papers

This thesis is based on the following appended papers:


Publications by the author not included in this thesis:

Peer reviewed international conference papers:


Journal papers:


Patent applications:

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Martin Sjödin

Göteborg
April 2012
## Abbreviations used in the text

<table>
<thead>
<tr>
<th>Abbreviation</th>
<th>Description</th>
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<tbody>
<tr>
<td>ADC</td>
<td>Analog-to-digital converter</td>
</tr>
<tr>
<td>ASE</td>
<td>Amplified spontaneous emission</td>
</tr>
<tr>
<td>ASIC</td>
<td>Application-specific integrated circuit</td>
</tr>
<tr>
<td>BER</td>
<td>Bit error rate</td>
</tr>
<tr>
<td>BPF</td>
<td>Band-pass filter</td>
</tr>
<tr>
<td>BPSK</td>
<td>Binary phase-shift keying</td>
</tr>
<tr>
<td>CD</td>
<td>Chromatic dispersion</td>
</tr>
<tr>
<td>CMA</td>
<td>Constant modulus algorithm</td>
</tr>
<tr>
<td>DBPSK</td>
<td>Differential binary phase-shift keying</td>
</tr>
<tr>
<td>DCF</td>
<td>Dispersion compensating fiber</td>
</tr>
<tr>
<td>DD-LMS</td>
<td>Decision-directed least mean square</td>
</tr>
<tr>
<td>DEMUX</td>
<td>Demultiplexer</td>
</tr>
<tr>
<td>DFB</td>
<td>Distributed feedback</td>
</tr>
<tr>
<td>DSP</td>
<td>Digital signal processing</td>
</tr>
<tr>
<td>ECL</td>
<td>External cavity laser</td>
</tr>
<tr>
<td>EDFA</td>
<td>Erbium-doped fiber amplifier</td>
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<tr>
<td>FEC</td>
<td>Forward error correction</td>
</tr>
<tr>
<td>FFT</td>
<td>Fast Fourier transform</td>
</tr>
<tr>
<td>FMF</td>
<td>Few-mode fiber</td>
</tr>
<tr>
<td>FPGA</td>
<td>Field-programmable gate array</td>
</tr>
<tr>
<td>IF</td>
<td>Intermediate frequency</td>
</tr>
<tr>
<td>IFWM</td>
<td>Intrachannel four-wave mixing</td>
</tr>
<tr>
<td>IPDM</td>
<td>Interleaved polarization division multiplexing</td>
</tr>
<tr>
<td>IQM</td>
<td>IQ modulator</td>
</tr>
<tr>
<td>ISI</td>
<td>Intersymbol interference</td>
</tr>
<tr>
<td>LO</td>
<td>Local oscillator</td>
</tr>
<tr>
<td>MMF</td>
<td>Multi-mode fiber</td>
</tr>
<tr>
<td>MUX</td>
<td>Multiplexer</td>
</tr>
<tr>
<td>MZM</td>
<td>Mach-Zehnder modulator</td>
</tr>
<tr>
<td>OOK</td>
<td>On-off keying</td>
</tr>
<tr>
<td>OSNR</td>
<td>Optical signal-to-noise ratio</td>
</tr>
<tr>
<td>Abbreviation</td>
<td>Description</td>
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<tr>
<td>--------------</td>
<td>------------------------------</td>
</tr>
<tr>
<td>PBC</td>
<td>Polarization beam combiner</td>
</tr>
<tr>
<td>PBS</td>
<td>Polarization beam splitter</td>
</tr>
<tr>
<td>PDL</td>
<td>Polarization-dependent loss</td>
</tr>
<tr>
<td>PLL</td>
<td>Phase-locked loop</td>
</tr>
<tr>
<td>PM</td>
<td>Polarization-multiplexed</td>
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<tr>
<td>PM-QPSK</td>
<td>Polarization-multiplexed QPSK</td>
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<tr>
<td>PM-16-QAM</td>
<td>Polarization-multiplexed 16-QAM</td>
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<tr>
<td>PMD</td>
<td>Polarization-mode dispersion</td>
</tr>
<tr>
<td>PS-CMA</td>
<td>Polarization-switched CMA</td>
</tr>
<tr>
<td>PS-QPSK</td>
<td>Polarization-switched QPSK</td>
</tr>
<tr>
<td>QAM</td>
<td>Quadrature amplitude modulation</td>
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<tr>
<td>QPSK</td>
<td>Quadrature phase-shift keying</td>
</tr>
<tr>
<td>RF</td>
<td>Radio frequency</td>
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<tr>
<td>RS</td>
<td>Reed-Solomon</td>
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<tr>
<td>Rx</td>
<td>Receiver</td>
</tr>
<tr>
<td>SE</td>
<td>Spectral efficiency</td>
</tr>
<tr>
<td>SER</td>
<td>Symbol error rate</td>
</tr>
<tr>
<td>SNR</td>
<td>Signal-to-noise ratio</td>
</tr>
<tr>
<td>SOP</td>
<td>State of polarization</td>
</tr>
<tr>
<td>SPM</td>
<td>Self-phase modulation</td>
</tr>
<tr>
<td>SP-QAM</td>
<td>Set-partitioning QAM</td>
</tr>
<tr>
<td>SSMF</td>
<td>Standard single-mode fiber</td>
</tr>
<tr>
<td>Tx</td>
<td>Transmitter</td>
</tr>
<tr>
<td>ULAF</td>
<td>Ultra Large Area Fiber</td>
</tr>
<tr>
<td>WDM</td>
<td>Wavelength-division multiplexing</td>
</tr>
<tr>
<td>XPM</td>
<td>Cross-phase modulation</td>
</tr>
<tr>
<td>XPolM</td>
<td>Cross-polarization modulation</td>
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1 Introduction

The transmission of messages over significant distances for communication purposes is called telecommunications, and has had a huge impact on both the social life of people and on the world economy. The data rates of the first communication systems were low, as these were based on simple techniques like drumbeats, smoke signals, or signalling flags, but the development has taken us to a point where information can be transmitted with almost incomprehensible speed and movies in HD quality can be downloaded faster than we are able to watch them. Fig. 1.1, which is based on data from Ref. [1], shows how the number of internet users in the world and the corresponding percentage of the world’s population have changed since the end of 1995. The growth has been steady and is expected to continue until most people use the internet on a daily basis. Fiber-optic systems will play a key role, since they form the back-bone of telecommunication networks. Due to the low attenuation and the high carrier frequency of light (193 THz @ λ = 1.55 µm), optical fibers are excellent for transporting information. For comparison, the carrier frequencies of microwaves are about four orders of magnitude smaller, reducing the modulation bandwidth accordingly. On the other hand, microwaves enable efficient wireless communications, and fiber-optic systems and radio frequency (RF) systems thus complement each other and are not direct competitors. Additional advantages with the optical fiber are immunity to electromagnetic interference (which can be a problem in RF communication systems), low sensitivity to moisture, low weight, and high flexibility.

This chapter provides a short overview of the history of fiber-optic communications and a motivation for the work performed in this thesis. In addition, some important concepts are introduced to facilitate the reading of the other chapters.

1.1 The history of fiber-optic communications

Many technological advances are behind today’s ultra-fast communication systems. To avoid a too long historical overview, we exclude smoke signals and focus on the development during the last 50 years. As a starting point we choose the year of 1962, when the operation of electrically pumped semiconductor lasers was reported by four research groups [2–5]. The low-loss optical fiber was then proposed by Kao and Hockham in 1966 [6], and two crucial components for the realization of fiber-optic communications now existed. The road leading to commercial transmission systems was however still long and cumbersome, and more inventions were needed. Luckily, further progress was ensured by the supremely talented photonics researchers, and the first semiconductor laser working at room temperature was reported in 1970 [7]. The attenuation in state-of-the-art optical fibers was still about 20 dB/km [8] in the early seventies, but refinements of the manufacturing processes reduced it substantially. The modern fiber is, with attenuation of about 0.20 dB/km at λ = 1.55 µm, one of the most
transparent media on earth, and enables transmission over very long distances (∼100 km) before detection or amplification is required. With good lasers and low loss fiber available, the first commercial transmission systems were installed in the early eighties. These operated at 45 Mb/s [9], had 10 km reach, and used the on-off keying (OOK) modulation format, in which the data is coded by the presence and the absence of optical pulses.

Cost-effective long-haul transmission was the next challenge. In spite of the low attenuation, transmission over long distances calls for periodic amplification of the signals. In the early fiber-optic systems, *optoelectronic repeaters* were used to detect optical signals and re-transmit them at higher power. Such expensive devices were needed for every channel, which made it costly to utilize the fiber to its full potential. *Coherent detection* was investigated in the eighties to enable longer reach, by enhancing the receiver sensitivity\(^1\) with a high power local oscillator (LO) laser [10–12]. Coherent systems were however difficult to implement due to the requirement of a stable optical phase-locked loop (PLL) for synchronization. Optical PLLs are challenging to implement even today, and commercial coherent systems were never realized.

The invention of the *erbium-doped fiber amplifier* (EDFA) in the late eighties [13–15] was very important, since it enabled simultaneous amplification of C-band signals (\(\lambda = 1530–1570\) nm) with a single device, making optoelectronic repeaters and sensitivity enhancement with coherent detection obsolete. EDFAs in combination with *wavelength-division multiplexing* (WDM) made it possible to increase the data throughput by adding more channels (as long as the C-band was not fully occupied).

To characterize the performance of communication systems, the *bit-rate-distance product*, i.e. the aggregate data rate multiplied by the reach, is frequently used. Fig. 1.2 shows how it has been increased in optical communications by inventions like the EDFA and WDM. A similar Fig. can be found in Ref. [16], but here the results from some landmarks experiment

\(^{1}\)The receiver sensitivity is the minimum average received power required to operate at a specific bit error rate.
are included. The single-mode fiber (SMF) enabled much longer reach than the multi-mode fiber (MMF), as did the switch to an operating wavelength of $\lambda = 1550$ nm instead of 1310 nm. The trends indicated with dashed lines never led to commercial success, even though some soliton links have been installed. Looking at the year of 2009, we see the current world record for bit-rate-distance, set by Salsi et al. [17]. They were the first to achieve more than 100 Pb/s·km, an increase of about six orders of magnitude compared to 30 years before.

The motivation for pursuing higher data rates is simple: There is a strong demand pushing the developments further. The internet is used for both entertainment and professional purposes, millions of people are downloading music and movies, playing online games, and using live streaming. To enable even more people to use even more bandwidth-demanding services, the development of fiber-optic communications needs to continue. Currently we have reached the era of multilevel modulation formats, as indicated in Fig. 1.2. These formats outperform binary formats (e.g. OOK) in terms of spectral efficiency, i.e. the information rate per bandwidth unit. To achieve the world record, Salsi and co-workers used a multi-level format known as polarization-multiplexed quadrature phase-shift keying (PM-QPSK), which is discussed in section 5.2.1.

Multilevel formats have also been used to achieve records for data throughput (in excess of 100 Tb/s) in a single fiber, when the reach is not considered [18, 19]. The experiments achieving these results are often labelled hero experiments, and use the latest technologies in transmitters, receivers, fibers, and optical amplifiers, in addition to hundreds of lasers sources. Due to requiring large resources, they are normally performed by industry labs such as Alcatel Lucent, AT&T, and NTT. In the latest hero experiments [18, 19], and in most transmission experiments today, the performance was not measured in real-time. The
received signals are instead sampled with analog-to-digital converters (ADCs) and the data is processed offline with a computer. The latest hero experiment with real-time performance monitoring was presented at the Optical Fiber Communication Conference (OFC) in 2007 [20]. An aggregate data rate of 25.6 Tb/s was achieved by using PM-QPSK with differential direct-detection.

1.2 Coherent fiber-optic systems

Using multilevel formats has several important implications. The tolerance to transmission impairments is in general lower than for binary formats, and more complicated generation and detection schemes are required. The most interesting multilevel formats use the phase of the optical carrier to encode data (some use both the phase and the amplitude), and preserving the phase in the detection process calls for coherent detection, which has led to a renewed interest in coherent systems.

Apart from preserving the phase information, coherent detection provides the highest receiver sensitivity and enables mitigation of transmission impairments such as chromatic dispersion (CD) and polarization-mode dispersion (PMD) by using data post-processing. The most common coherent detection approach is intradyne detection [21–23], in which the phase reference is a free running LO laser, and phase synchronization and impairment mitigation are performed with digital signal processing (DSP). The work in this thesis about intradyne systems is mostly devoted to power-efficient modulation formats making good use of the four-dimensional signal space of the optical field. Such formats have been investigated both experimentally, analytically, and with numerical simulations. Algorithms for phase estimation and polarization demultiplexing have been suggested and implemented and we investigated differential coding schemes to remove the phase ambiguity after detection.

The thesis also deals with self-homodyne coherent systems [24], in which a pilot tone is co-propagated with the signal in the orthogonal polarization state. Since the pilot tone and the signal are correlated in phase, there is no need for high speed DSP or highly coherent laser sources. The benefits and drawbacks of self-homodyne coherent systems are discussed in chapter 3, and the performance has been quantified with both experiments and numerical simulations.

1.3 Basic concepts

Modulation formats

The term modulation format is frequently used in the thesis. It refers to a method for sending a sequence of data by manipulating an electromagnetic carrier wave, which here is the optical field. Modulation formats use a set of signal levels, commonly referred to as symbols. Each signal level represents a short sequence of bits of information and is in optical communications normally defined by the amplitude, the phase, and the polarization state of the wave. The number of bits carried by each symbol depends on the total number of symbols in the constellation. If there are M different symbols occurring with equal probability, the number of transmitted bits per symbol is \( \log_2(M) \) and is together with the power efficiency the most important figure-of-merit of the modulation format. The power efficiency is determined by how well separated the symbols are in the signal space for a given average optical power. The concept is explained in more detail in chapter 5.

\(^2\)a phase reference is required in all coherent systems.
Figure 1.3: Photocurrents and constellation diagram after coherent detection of a 28 Gbaud QPSK signal. (a) The in-phase current, $i_I = \Re(i)$, during 700 ps, i.e., 20 QPSK symbols. The red dots indicate sampling in the middle of each symbol slot. (b) The quadrature-phase current, $i_Q = \Im(i)$, in the same time frame. (c) The complex photocurrent, $i = i_I + j i_Q$, visualized in a constellation diagram with 1024 symbols (36.6 ns time frame) after sampling at one sample per symbol. The samples from (a) and (b) are shown with red dots.

**Constellation diagrams**

*Constellation diagrams* are used to visualize detected signals and the constellations of modulation formats. They provide information about both the amplitude and the phase, which is important since many modern lightwave systems use both properties for data encoding. To plot a constellation diagram, both the in-phase ($i_I$) and the quadrature-phase ($i_Q$) part of a signal are required. Figs. 1.3(a)-(b) show the photocurrents $i_I$ and $i_Q$ during 700 ps after coherent detection of a 28 Gbaud QPSK signal. The photocurrents are sampled at a rate of one sample per symbol and the sampling instants are indicated with red dots. The acquired samples can be used to visualize $i = i_I + j i_Q$ in the complex plane, which is done in Fig. 1.3(c) for 1024 QPSK symbols. The first 20 samples in the constellation diagram (the red dots) are those from Figs. 1.3(a)-(b).

**Bit error rate and symbol error rate**

Measuring the *bit error rate* (BER) is the most common approach to characterize the performance of a communication system. A *bit sequence* is sent from the transmitter and compared with the detected sequence in the receiver. Incorrectly identified bits are referred to as *bit errors*, and the BER is

$$\text{BER} = \frac{\# \text{ bit errors}}{\# \text{ transmitted bits}}. \quad (1.1)$$

Bit errors are caused by, e.g., noise, intersymbol interference, and crosstalk between channels. Due to the stochastic nature of the signal, it is also common to talk about the *bit error probability*. The bit error probability is the *expected* value of the BER, when measured during a sufficiently long time interval. In case of complete system failure, the BER $\approx 0.5$.

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$^3$In this thesis, $j$ is used instead of $i$ to represent the imaginary number, since $i$ denotes the photocurrent.
Alternatively, the symbol error rate (SER) can be used as a performance merit. A sequence of symbols is then transmitted and compared with the detected symbol sequence. The SER may, e.g., be used in case of non-trivial bit-to-symbol mapping.

1.4 Outline of the thesis

The organization of the thesis is as follows. Chapter 2 describes a typical fiber-optic communication system. It starts with the transmitter, continues with the optical fiber, and ends with the receiver. In addition, multiplexing techniques such as WDM and polarization multiplexing are described. The principles and properties of self-homodyne and intradyne coherent systems are then discussed in chapter 3 and chapter 4, respectively, and the modulation formats investigated in the thesis are presented in chapter 5. Suggestions for interesting topics for future research are given in chapter 6 and finally the appended papers are summarized in chapter 7.
2 Fiber-optic communication systems

Three essential units are present in any telecommunications system [25]:

1. A transmitter which modulates information onto an electromagnetic carrier wave, creating what hereafter will be referred to as a “signal”.

2. A transmission medium, also called the “physical channel”, which transports the signal. The signal may be distorted by the channel during propagation.

3. A receiver after the channel which after equalization of the signal extracts the information.

Fig. 2.1 shows a block diagram of a general communication system. In fiber optics the transmitter is most commonly a semiconductor laser which may be directly modulated (by changing the bias current), or operated in continuous wave mode. In the latter case an external modulator is used after the laser to modulate the carrier. This approach is more expensive, but preferable if the transmission distance exceeds a few kilometers, due to its better performance. In addition, external modulators are necessary to achieve phase modulation.

The signal propagates through the optical fiber to the receiver, guided by total internal reflection. Optical amplifiers along the path add noise, and the pulses are broadened by chromatic dispersion (CD) and perhaps also distorted by nonlinear effects. In addition, the polarization state of the signal will change randomly. When it reaches the receiver, the optical signal is converted to an electrical signal, a photocurrent. Different detection schemes can be used, depending on the modulation format and the acceptable cost. Direct detection is cheapest and has been used since the early days of fiber-optic communications. Barring some unforeseen technological advances, it will continue as the method of choice in short links (e.g., access networks) for the foreseeable future. Coherent detection, on the other hand, is becoming the dominating approach in long-haul transmission. The two detection schemes, in particular coherent detection, are described in section 2.3.

![Figure 2.1: A general communication system.](image-url)
Analog and digital signals

Signals are either analog or digital. An analog signal varies continuously and is sensitive to noise, i.e., random fluctuations due to, e.g., thermal motion of electrons (electrical signals), spontaneously emitted photons (optical signals), or the particle nature of electrons and photons. Analog signals generally need high signal-to-noise ratio (SNR) in the receiver.

On the other hand, in a digital signal the information is encoded as a set of discrete values, and unless the noise disturbance exceeds a certain threshold, the information can be exactly recovered. In addition, forward error correction (FEC) can be applied for full restoration of the data even at very low SNR. The drawbacks of digital communications (requiring more electronics and bandwidth) are of much less significance than the benefits. Anyone who has had the experience of living far away from the TV transmitter station and then switched from analog to digital broadcasts knows the benefits of digital communications. Fiber-optic communications mostly use digital signals.

The theory dealing with digital communications is known as information theory and was introduced by Claude Shannon in 1948 [26], in a paper that may be regarded as the starting point of the information era. Shannon proved that the capacity $C$ of an additive white Gaussian noise (AWGN) channel with a bandwidth of $B$ Hz and SNR of $S/N$ is

$$C = B \log_2 \left( 1 + \frac{S}{N} \right). \quad (2.1)$$

The capacity is the tightest upper bound on the amount of information that can be transmitted over a communications channel with arbitrarily small BER. Studying (2.1) we note that larger capacity can be achieved by increasing the SNR. It should however be remembered that (2.1) assumes an AWGN-limited channel, and due to the Kerr effect this is not always applicable to fiber-optic communications. The Kerr effect is discussed in section 2.2.4.

2.1 The transmitter

This section is dedicated to the semiconductor laser and the external modulator.

2.1.1 Semiconductor lasers

Light can be generated in different spectral ranges by using different types of lasers. In fiber-optic communications, semiconductor lasers are used since they are easy to integrate and have high coupling efficiency to the fiber, in addition to features such as low power consumption, high reliability, and compatibility with electronic circuits. Like other laser types, semiconductor lasers produce more or less coherent light through stimulated emission.

The gain region consists of a double heterostructure p-n-junction, in which the active region has smaller bandgap and higher refractive index than the p- and n-type materials. There are two major advantages with this structure. (i) The active region effectively confines electrons and holes due to its smaller bandgap, and (ii) the higher refractive index makes the gain region act like a dielectric waveguide (similar to the optical fiber), ensuring confinement of the photons as well. To obtain a single lasing mode and avoid modal dispersion, which greatly limits the transmission reach, it is common to use a grating close to the active region which gives a periodic variation of the refractive index. As opposed to using reflective facets, optical feedback is provided by the whole cavity and undesired modes are suppressed.

Fig. 2.2 shows the structure of a distributed feedback (DFB) laser utilizing such a grating. DFB lasers are sufficiently reliable and cheap to be used in nearly all commercial long-haul systems [27, p. 90].
Figure 2.2: A simple schematic of a DFB laser.

Laser phase noise

Lasers are often labelled as coherent light sources, implying that the output beam is deterministic, or monochromatic, in contrast to, e.g., light from the sun, which is a superposition of emissions from a countless number of atoms radiating independently of each other. Stimulated emission of particles known as photons are used in all lasers. Photons generated by stimulated emission are in-phase with each other, have the same amplitude, propagate in the same direction, and are co-polarized. If the phase, the amplitude, the frequency, and the direction of propagation are known for a coherent wave, they can be predicted at any other time and after any propagation distance. Such light sources would be very attractive, but unfortunately real laser beams are not perfectly coherent, and our knowledge of the phase is limited by phase noise. Assume we know the phase of a laser beam at a certain instant, and we want to predict the phase some time later. Due to phase noise, the inaccuracy of the prediction increases the longer we wait, and eventually we are only able to guess the phase. Laser phase noise is caused by, e.g., spontaneous emission of photons, the quantum nature of light, and instabilities in the laser cavity. The laser linewidth enables us to estimate for how long time the phase is stable. The power spectral density of the laser spectrum is in most cases well described by a Lorentzian function and the linewidth is defined as its 3-dB bandwidth.

Laser phase noise was not an important impairment in dispersion-managed on-off keying (OOK) systems, but before the invention of the dispersion compensating fiber (DCF) it limited the reach of repeater-less systems due to phase-to-amplitude conversion by CD [28, 29]. In coherent systems, the phase noise is an important problem and researchers have investigated its impact and different compensation schemes. Optical phase-locked loops [30], systems with inherent phase noise robustness [24,31], and compensation with digital signal processing (DSP) algorithms after detection [23] have been proposed. The DSP approach has mostly been studied with numerical simulations, where the phase noise is modelled as a Wiener process and the phase change during the time $T$ is

$$\psi(t + T) - \psi(t) \sim N\left(0, \sigma_p^2\right), \quad (2.2)$$

where $N$ is the normal distribution with zero mean and variance

$$\sigma_p^2 = 2\pi \Delta \nu T, \quad (2.3)$$

9
where $\Delta \nu$ is the combined linewidth of the local oscillator (LO) laser and the signal laser and $T$ is the sampling period. Fig. 2.3 shows examples of laser phases modelled by (2.2) during 1 $\mu$s (10000 symbols for a 10 Gbaud signal) as a function of time, for four different values of $\Delta \nu T$. The phase noise tolerance depends on the modulation format and the compensation algorithm. $\Delta \nu T = 10^{-3}$ is a large value and makes phase estimation difficult even for a relatively robust format like QPSK, while $\Delta \nu T = 10^{-6}$ has negligible impact on the performance even for 256-QAM for state-of-the-art phase estimation algorithms [32,33]. DFB lasers have traditionally had linewidths of several MHz, but recent advances will make them useful in systems with small phase noise tolerance too [34].

2.1.2 Modulation

Direct modulation (by changing the laser bias current) or an external modulator can both be used to modulate the optical carrier. The first approach is used for low-cost short range interconnects, while the more expensive external modulation is preferred in long-haul transmission since it gives higher signal quality. The principle of external modulation is to embed a waveguide in an electro-optical substrate, typically lithium niobate (LiNbO$_3$), and change its refractive index (and at the same time the carrier phase) by applying a voltage. The voltage dependence of the refractive index is approximately linear.

The Mach-Zehnder modulator

Phase modulators can provide intensity modulation when placed in interferometric structures. The principle of the Mach-Zehnder modulator (MZM) [35] is to split a signal into two branches with phase modulators and the interference after recombination then determines
the output levels. The phase modulators in dual-drive MZMs can be driven independently, in contrast to single-drive devices. Single-drive MZMs can only perform pulse carving or generate OOK, but dual-drive MZMs can, in principle, map bit sequences to anywhere in the complex plane [36]. The MZM is the most common modulator in high speed systems, due to its excellent performance and the possibility of modulating both the phase and the amplitude. Fig. 2.4(a) shows a simple schematic of a dual-drive MZM and Fig. 2.4(b) the dependence of the optical intensity and the optical field on the applied driving voltage. The voltage giving a $\pi$ phase shift in one interferometer arm is denoted by $V_\pi$. Changing the signal phase with $\pi$ requires $\pi$ phase-shifts in both arms, and subsequently the peak-to-peak voltage of the driving signal should be $2V_\pi$. Although a smaller amplitude can be used, the output signal is then more sensitive to noise and driving signal intersymbol interference (ISI). For optimal performance, the voltage swing should be $2V_\pi$.

Placing dual-drive MZMs in parallel facilitates the generation of advanced modulation formats. The modulator with two parallel dual-drive MZMs is known as the IQ-modulator (IQM). Fig. 2.5 illustrates how this device is used to generate QPSK by (i) applying independent binary data to each MZM to generate two binary phase-shift keying (BPSK) signals and (2) combining these with a relative phase-shift of $90^\circ$. Using additional MZMs

![Figure 2.4](image1.png)

**Figure 2.4:** (a) Schematic of a dual-drive MZM. (b) The transfer functions for the optical power and the amplitude of the optical field as a function of the applied voltage.

![Figure 2.5](image2.png)

**Figure 2.5:** Schematic of an IQ-modulator and the generation of a QPSK signal.
enable more advanced modulation formats to be generated with binary electrical signals. The quad-parallel modulator [37] is the optimal 16-QAM modulator and consists of two parallel IQMs (four MZMs). The IQM generate two QPSK signals which are combined with a 1:2 amplitude ratio to create 16-QAM. A 64-QAM modulator with six parallel MZMs has also been demonstrated [38].

2.2 The optical fiber

The optical fiber, the channel between the transmitter and the receiver, has many great properties: the attenuation in standard single-mode fibers (SSMF) can be as low as 0.15 dB/km [39] at 1.55 μm (but is often close to 0.20 dB/km), the intrinsic bandwidth is very large (≈ 35 THz), the flexibility is high, the weight is low, installation of optical fibers is easy, they are insensitive to moisture, and immune to electromagnetic interference.

There are several fiber types. Multi-mode fibers (MMFs) are used for transmission over short distances, such as within buildings or vehicles. MMFs are inexpensive and have good light coupling properties but suffer from intermodal dispersion, since the signal energy is carried by several different modes which are solutions to Maxwell’s equations and a set of boundary conditions. The modes propagate with different speed, which can cause significant pulse broadening even after a few hundred meters. There are also dispersion-shifted fibers, with very low dispersion around 1.55 μm, polarization-maintaining fibers, in which the state of polarization (SOP) of the signal is preserved during propagation, and allwave fibers, with larger bandwidth than SSMFs as a result of removal of the OH⁻ absorption peak around 1.4 μm. Many recent long-haul transmission experiments have used ultra large area fibers (ULAF), with lower attenuation, larger dispersion, and smaller nonlinear coefficient than SSMF.

Light propagation in optical fibers is modelled with the nonlinear Schrödinger equation (NLSE) [40, p. 40], which is considered to be very accurate. For fibers with rapidly and randomly varying birefringence (a good description of most sufficiently long transmission fibers) the NLSE can be generalized to the Manakov equation [41]

\[
j \frac{\partial A}{\partial z} = \frac{\beta_2}{2} \frac{\partial^2 A}{\partial t^2} + \frac{\beta_3}{6} \frac{\partial^3 A}{\partial t^3} - \frac{\alpha A}{2} - \gamma |A|^2 A,
\]

(2.4)

where \( A \) is the complex envelope of the two polarization components of the optical field, normalized such that \( |A|^2 \) represents the optical power, \( \beta_2 \) is the group velocity dispersion coefficient, \( \beta_3 \) is the third order dispersion, \( \alpha \) is the attenuation constant, \( z \) is the distance of propagation, \( t \) the time coordinate in a reference frame moving with the signal group velocity, and \( \gamma \) is the nonlinear coefficient. \( A \) is assumed to vary slowly compared to the carrier wave, a valid assumption in most cases, where the pulse duration is at least several ps.

2.2.1 Chromatic dispersion

The dispersion terms in (2.4) can be understood by studying the Taylor expansion of the dispersion relation of the fiber around the angular carrier frequency \( \omega_0 = 2\pi \nu_0 \)

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

(2.5)
where \( \beta \) is the wave number and
\[
\beta_1 = \left. \frac{d^2 \beta}{d \omega^2} \right|_{\omega=\omega_0}.
\] (2.6)

\( \beta_0 \) is the wave number of the carrier and absent in (2.4) since the equation does not account for such rapid phase variations. \( \beta_1 \) is related to the group velocity of the pulse envelope by \( v_g = 1/\beta_1 \) and is also absent in (2.4), due to a substitution of variables that makes the coordinate system move with the group velocity of the signal. On the other hand, the third term in (2.5), which contains the group velocity dispersion parameter \( \beta_2 \), is present in (2.4) and accounts for the frequency dependence of the group velocity. Physically this means that different spectral components of an optical pulse with finite spectral width \( \Delta \omega \) travel at different speeds through the fiber, which may cause either pulse compression (over short distances only) or pulse broadening. \( \beta_2 \) is related to the dispersion parameter of the fiber, \( D \), by
\[
D = -\frac{2\pi c}{\lambda^2} \beta_2.
\] (2.7)

CD-induced pulse broadening is a problem in fiber-optic communications since it introduces ISI. CD also causes walk-off in time between pulses with different carrier frequencies, which generally is beneficial, as it reduces the impact of interchannel nonlinear effects (section 2.2.4). Although a certain (symbol rate dependent) amount of residual dispersion can be tolerated in the receiver, there are several well-proven methods to compensate for CD. Currently, periodic dispersion compensation still dominates in long-haul transmission, but compensation with DSP after detection is expected to be preferred in future systems.

Fig. 2.6 shows the impact of CD on 28 Gbaud QPSK after 10 km and 100 km SSMF transmission (\( D = 17 \text{ ps/nm-km} \)). After 10 km, the uncompensated noise-less signal is still error-free, but with significantly reduced noise tolerance due to the large ISI. After 100 km, distinguishing the signal is impossible and the BER \( \approx 0.5 \) unless the CD is compensated.

### Periodic dispersion compensation

All-optical dispersion compensation with DCF, with opposite sign of \( D \) compared to SSMF, has been in commercial use since the mid-nineties. DCF should have large negative dispersion to enable the length to be short, and thereby keeping the loss and the impact of nonlinear effects as small as possible. \( D \approx -100 \text{ ps/nm-km} \) is a common value and the DCF modules are usually placed after the SSMF spans.

### Dispersion compensation with DSP

In case of coherent detection, CD can be compensated with DSP after transmission. This enables longer transmission reach than optical dispersion management, since power loss in DCFs is avoided and the impact from self-phase modulation and nonlinear phase noise reduced (see section 2.2.4). In addition, the tolerance to residual dispersion is higher if adaptive equalization is utilized. Figs. 2.7(a)-(b) show transmission links with CD compensation with DSP and DCF, respectively. As illustrated in Fig. 2.7(b) DCF-based compensation often requires an additional optical amplifier to maintain a high optical signal-to-noise ratio (OSNR), a further drawback compared to DSP-compensation.

### 2.2.2 Loss and amplification

Although the optical fiber is highly transparent, it has a small attenuation mainly due to Rayleigh scattering and infrared absorption [27, p. 56]. The fourth term in (2.4), which
contains the attenuation constant $\alpha$, accounts for fiber loss. $\alpha$ is typically 0.18–0.22 dB/km in a SSMF. The attenuation calls for periodic amplification of signals in long-haul transmission, which can be achieved by using, e.g., EDFAs or Raman amplification.

**Erbium doped fiber amplifiers**

EDFAs are indispensable for fiber-optic communications, since they enable amplification over a wide spectrum with large and polarization-independent gain, high pumping efficiency, modulation format transparency, and relatively low noise figure. An additional advantage with EDFAs is that they, due to the slow response time ($\sim$ ms), do not cause cross gain saturation, which in a communications context is a problem for parametric amplifiers [42] and semiconductor optical amplifiers [43].

The principle of the EDFA is to use a piece of optical fiber doped with erbium ions ($\text{Er}^{3+}$) as gain medium, and pump it with lasers at 980 nm and/or 1480 nm. This enables gain in the C-band (1530–1570 nm) and amplifiers with gain in the L-band (1570–1610 nm) and S-band (1480–1530 nm) have also been fabricated by using higher erbium concentration [44], or doping with thulium [45,46]. Incoming photons stimulate emission of new photons with the same phase, frequency, polarization, and propagation direction. In other words, the amplifier operates as a laser without feedback. A schematic of an EDFA is shown in Fig. 2.8. The optical isolators prevent amplification of reflections from, e.g., fiber connectors. The length of the doped fiber is normally $\sim$ 10 m.

![Figure 2.6: 28 Gbaud QPSK after 10 and 100 km SSMF transmission and after CD compensation.](image)

![Figure 2.7: Links with (a) CD compensation with DSP after coherent detection, and (b) periodic CD compensation with DCF modules.](image)
Amplified spontaneous emission noise

The main drawback of the EDFA (and with most optical amplifiers) is the generation of amplified spontaneous emission (ASE) noise, which limits the transmission distance. The ASE noise power per polarization after the amplifier is [27, p. 305]

\[
P_{\text{ASE}} = \int_{\text{C-band}} S_{\text{ASE}}(\nu) \, d\nu = \int_{\text{C-band}} n_{sp} \hbar \nu_0 |G(\nu) - 1| \, d\nu,
\]

(2.8)

where \( S_{\text{ASE}} \) is the power spectral density per polarization, \( n_{sp} \) the population inversion factor of the gain medium, \( G \) the gain, and the integration is over the C-band (1530–1570 nm). This noise power is generated each time the signal passes an amplifier with the same gain and \( n_{sp} \) factor. The ASE noise degrades the OSNR, which is defined as

\[
\text{OSNR}_{0.1\text{nm}} = \frac{P_{\text{sig}}}{2P_{\text{ASE},0.1\text{nm}}},
\]

(2.9)

where \( P_{\text{sig}} \) is the signal power. The factor of 2 in the denominator is due to equal amounts of ASE power in both polarization modes. The OSNR is measured with an optical spectrum analyzer and normalized to a 0.1 nm noise bandwidth. It is easily realized that a shorter amplifier spacing is beneficial to obtain a higher OSNR in the receiver. Fig. 2.9 shows how the OSNR depends on the transmission distance for amplifier spacings from 25 km and up to 150 km, assuming the signal power launched into each fiber span to be 1 mW and an EDFA noise figure of 5 dB (\( n_{sp} = 1.58 \)). If we want longer fiber spans while maintaining a high OSNR, we need to increase the launch power, use a fiber with lower attenuation, or amplifiers with smaller noise figures. Of these alternatives, the first is by far the easiest to accomplish, but unfortunately it increases the influence of the Kerr nonlinearity. The graphs in Fig. 2.9 do not provide any insight about the impact of nonlinear effects.

2.2.3 Polarization effects in the fiber

There are several polarization-related effects in fiber-optic transmission. They are briefly explained below.

Polarization drift

The signal’s SOP changes randomly during propagation due to, e.g., temperature variations and mechanical stress, which for some systems calls for polarization tracking in the receiver. In intradyne systems, the tracking is performed with DSP after detection, but in
Figure 2.9: The OSNR as a function of the transmission distance and the amplifier spacing. The EDFA noise figure is 5 dB, $\alpha = 0.2$ dB/km, and the signal power is 1 mW.

other coherent detection schemes, or in systems using polarization multiplexing and direct detection, optical polarization tracking is required. In an attempt to estimate the speed of SOP rotations, a field trial in 2005 investigated the speed of the polarization changes at the output of a transmission link (15 spans with 55 km of SSMF between the cities of Berlin and Darmstadt) [47]. Data of the SOP was collected over a period of 2.5 months and two events were registered in which the polarization angle exceeded a rotation of $\pi/10$ in less than 0.1 ms, corresponding to a polarization rotation of 3 krad/s. This number gives an idea of the maximum rotation speed one has to deal with in the receiver.

PMD

Polarization-mode dispersion (PMD) is due to the random birefringence of the fiber, which makes the refractive index polarization-dependent. Before practical implementation of coherent detection was feasible, PMD was considered a major limitation for high speed optical communications, since its influence is detrimental for short pulses and optical PMD compensation is difficult at high data rates. However, with digital coherent receivers, PMD can be perfectly compensated after detection [48–50], provided the equalization filter has a sufficient number of taps. In reality, the complexity of electronic equalizers is limited and as a consequence there are limits to the PMD tolerance too. Since PMD has a probability distribution with long tails [51] and the instantaneous differential group delay can vary significantly with time, system outage may occur in spite of the equalization.

PDL

Polarization-dependent loss (PDL) is a polarization related effect that cannot be well compensated with DSP [52, 53]. PDL causes the signal power and the OSNR to fluctuate and
may give the tributaries of a polarization-multiplexed signal unequal performance, which increases the risk for misconvergence of the polarization demultiplexing algorithms [54]. It has been shown with both experiments and numerical simulations that the worst case of PDL is when the lossy axis of a PDL element is aligned with one of the polarization tributaries of the signal. The case where the lossy axis is aligned equally to both tributaries causes depolarization, but gives the smallest OSNR penalty [53]. The explanation is that a good equalizer is efficient in separating the polarization tributaries, even with substantial loss of orthogonality. However, the depolarization should be more detrimental without electronic equalization.

2.2.4 The Kerr nonlinearity

The last term in (2.4) is due to the Kerr effect [40], which is the most important intrinsic limitation in fiber-optic communications. The presence of nonlinear distortion in the channel is a fundamental difference between optical communications and RF communications. The refractive index \( n \) of a fiber with nonlinear coefficient \( \gamma \) is weakly dependent on the light intensity \( I \)

\[
\Delta n \sim \gamma I,
\]

\[
(2.10)
\]

which constrains the signal power and causes a trade-off between nonlinear distortion and ASE noise accumulation. Although the optical power launched into a fiber typically is on the order of only 1–10 mW, the intensity is very high due to the small area of the fiber core (50–150 \( \mu \)m). The signal therefore affects the refractive index and the system has optimal performance at a specific operating power. Although a further power increase improves the OSNR, the nonlinear effects become more detrimental for the performance than the ASE noise. This trade-off is illustrated in Fig. 2.10, which shows numerical simulation results of the BER of a 112 Gb/s polarization-multiplexed 16-ary quadrature amplitude modulation (PM-16-QAM) signal as a function of the fiber span launch power and the transmission distance. Starting from the lowest launch power of \(-6.0 \) dBm, the BER first decreases as a function of launch power, but from around \(-3.5 \) dBm increasing the power degrades the performance.

Several phenomena originate from the Kerr effect, and they can be categorized into groups of intrachannel and interchannel nonlinear effects.

Intrachannel nonlinearities (SPM and IFWM)

Self-phase modulation (SPM) and intrachannel four-wave mixing (IFWM) are intrachannel nonlinear effects, and thus occur also for a single channel. Their importance depend on many parameters: the modulation format, the dispersion map, the signal power, and the data rate.

Starting with SPM, one of the consequences of (2.10) is that a signal propagating in a fiber with length \( L \) and attenuation \( \alpha \) induces a phase shift on itself [40, p. 80]

\[
\Delta \Phi_{\text{SPM}} = \gamma PL_{\text{eff}},
\]

\[
(2.11)
\]

where \( P \) is the optical power and \( L_{\text{eff}} = [1 - \exp(-\alpha L)] / \alpha \) is the effective fiber length, used to account for attenuation. The name SPM stems from the signal imposing the phase shift on itself. SPM gives frequency chirp, which under certain circumstances can be balanced by CD to create optical solitons [55,56], or be used for pulse compression. On the other hand, SPM may lead to signal distortion, which is particularly detrimental for modulation formats with multiple amplitude levels, since symbols with different amplitudes acquire
different nonlinear phase shifts. In addition, when interacting with ASE noise SPM creates nonlinear phase noise [57], which is a limitation especially in dispersion-managed long-haul systems using phase modulation and low symbol rates. Dispersion compensation with DSP after coherent detection reduces the influence of SPM, since the effect requires small accumulated dispersion to be efficient. Instead, IFWM gets increasingly important [58–60]. IFWM occurs when neighboring pulses overlap with each other due to dispersive broadening. When spectral components with slightly different frequencies interact, four-wave mixing [61] occurs and power is transferred between the pulses.

A numerical study of 40 Gb/s transmission, involving several different modulation formats [62], showed SPM to have stronger influence than IFWM in links with optical dispersion management and in case of multi-ring constellations. IFWM however dominated in all other cases, and increasing the data rates reduces the SPM impact, due to the greater dispersive broadening. This paved the way for a recently presented analytical theory for fiber propagation, taking both dispersion and nonlinear effects into account [63]. The theory assumes moderate signal power and large dispersion, which is valid for most practical cases.

Figs. 2.11(a)-(b) show one polarization tributary of received 112 Gb/s PM-16-QAM signals transmitted in a link with ten 80 km spans of SSMF. The fiber parameters are $D = 17$ ps/nm/km, $\alpha = 0.2$ dB/km, and $\gamma = 1.2$ /W-km and the launch power was 1 dBm ($-2$ dBm per polarization). The reason for the different appearances of the constellations is that different dispersion management has been used. In (a) the dispersion was compensated after coherent detection and the SPM distortion appears very small. On the other hand, in case (b) CD is compensated with DCF after each SSMF span, resulting in strong SPM impact. Since CD-induced pulse broadening depends quadratically on the symbol rate [27, p.
Figure 2.11: One polarization tributary of PM-16-QAM signals after transmission over 10 spans with 80 km of SSMF. The launch power is 1 dBm in all cases and the fiber parameters are $D = 17 \text{ ps/nm-km}$, $\alpha = 0.2 \text{ dB/km}$, and $\gamma = 1.2 \text{ /W-km}$. (a) 112 Gb/s (14 Gbaud) and dispersion compensation after coherent detection. IFWM has a stronger impact than SPM. (b) 112 Gb/s and dispersion compensation with DCF modules after each fiber span, which gives characteristic SPM distortion. (c) 448 Gb/s (56 Gbaud) and DSP compensation. There is no noticeable SPM distortion. (d) 448 Gb/s and DCF compensation. Due to the higher symbol rate, the SPM distortion is smaller here than in case (b).

48] the signal distortion is data rate-dependent. In Figs. 2.11(c)-(d), all parameters from (a)-(b) have been kept except for the data rate, which is 448 Gb/s. By comparing Fig. 2.11(b) and Fig. 2.11(d), it is obvious that the symbol rate increase has reduced the influence of SPM. In addition, the small distortion at 112 Gb/s is not visible in Fig. 2.11(c). The larger symbol spread in (c)-(d) is due to the inverse dependence of the SNR after detection on the data rate at a fixed launch power.
Interchannel nonlinear effects (XPM and XPolM)

Interchannel nonlinear effects such as cross-phase modulation (XPM) [40] and cross-polarization modulation (XPolM) [64] occur in systems with at least two wavelength channels. Generally, both intrachannel and interchannel nonlinearities affect the performance of WDM systems but the relative importance depends on the symbol rate, with intrachannel nonlinearities being more important at high rates and large channel spacing [65,66].

XPM is similar to SPM but involves at least two waves at different wavelengths, or two orthogonally polarized waves at the same wavelength. In the simplest case with two waves at different wavelengths, wave 2 induces a phase shift on wave 1 of [40, p. 230]

\[ \Delta \Phi_{\text{XPM}} = 2\gamma P_2 L_{\text{eff}}, \]  

(2.12)

where \( P_2 \) is the power at the second wavelength. The factor of 2 in (2.12) shows that XPM is twice as strong as SPM for a given power. The impact of XPM in WDM systems depends on the interaction time between pulses at different wavelengths and also on the overlapping symbol sequences, in particular for modulation formats with amplitude modulation. Traditionally, XPM has been mitigated by using fiber with nonzero dispersion to induce pulse walkoff [67]. The influence of XPM is also known to be strongly reduced when the channel spacing is increased. In other words, for a fixed spectral efficiency, XPM is mitigated by increasing the data rate [68,69].

XPolM, also known as nonlinear polarization scattering, refers to the modulation of the signal SOP by other channels. The effect coexists with XPM, in a sense that one effect will not occur without the other, with the exception of the interacting waves being either co-polarized or orthogonally polarized [64, pp. 656–658]. XPolM is particularly troublesome for polarization-multiplexed transmission, as it gives rise to depolarization, i.e. the SOP may change rapidly in a fashion dictated by the intensity of symbols from other WDM channels. The impact of XPolM on coherent systems has been investigated with numerical simulations by Xie [70–72] and by Bononi et al. [73]. XPolM was found to have a large impact in dispersion-managed coherent systems [71, 73], and to be more detrimental in case of polarization multiplexing than for single-polarization transmission. The reasons are (i) that the SOP of a polarization-multiplexed signal is data dependent, causing significant polarization-scattering in WDM transmission, and (ii) XPolM does not degrade polarization-insensitive receivers, but XPolM-induced crosstalk between the polarization components may cause large problems for polarization-multiplexed systems, since the process is too fast to compensate with digital equalization [70]. PMD was found to mitigate the impact from XPolM due to reducing the data dependence of the SOP [72]. Introducing a controlled amount of PMD in systems may therefore improve the performance.

2.3 The receiver

The receiver converts the optical signal to one or more (depending on the modulation format) electrical baseband signals. The receivers in fiber-optic communications use either direct detection or coherent detection.

2.3.1 Direct detection

In direct detection a single photodetector detects the optical signal and generates the photocurrent \( i(t) \)

\[ i(t) = R_d P_{\text{sig}}(t), \]  

(2.13)
where $R_d$ is the responsivity (measured in A/W) of the detector and $P_{\text{sig}}(t)$ the signal power. As shown in (2.13), the photocurrent depends linearly on the optical power and contains no phase information, implying that direct detection cannot be used for phase modulated signals. However, by placing a delay interferometer before the detector, we can detect signals using the relative phase for modulation, such as the family of $N$-level modulation formats known as DVPSK. DBPSK ($N = 2$) has about 3 dB higher receiver sensitivity than OOK, assuming ASE noise limited balanced detection. Unfortunately, the performance of DVPSK scale badly with $N$, partly due to a sensitivity penalty [74], and partly due to requiring precoders which are quite complicated for $N > 4$ [75]. Modulation formats using both the phase and the amplitude of the carrier normally require coherent detection for good performance, and there has been little focus on direct detection formats in recent years.

### 2.3.2 Coherent detection

In a coherent receiver the signal is mixed with a phase reference to preserve the phase information. Coherent detection has been used for many years in RF communications, but in fiber-optics it has been used commercially for only a short time. As mentioned in chapter 1, researchers investigated coherent systems in the eighties to boost the transmission distance and the receiver sensitivity [10–12]. The invention of the EDFA ended those activities, but 25 years later the need for more spectrally efficient modulation formats sparked a renewed interest. Fig. 2.12 shows an early coherent receiver using an optical phase-locked loop (PLL) for synchronization. This approach requires optical phase-locking of lasers with very narrow linewidths and is not interesting for modern coherent systems, since the availability of high speed electronics has made optical PLLs unnecessary.

**Phase diversity**

The coherent receiver shown in Fig. 2.12 lacks phase diversity and can therefore not detect signals with IQ modulation, i.e. signals using both the real and the imaginary part of the optical field to encode data. To achieve phase diversity and enable IQ modulation, an optical 90° hybrid, as shown in Fig. 2.13, is used. For reasons that are clarified below, the hybrid is normally followed by a pair of balanced detectors. In the following description of a phase diversity coherent receiver, intradyne detection with a LO laser is assumed, although it is also possible to use a pilot signal as a phase reference, as discussed in chapter 3.

The optical fields of the signal and the LO laser are

$$E_{\text{sig}} = A_{\text{sig}}(t) \exp(j\omega_{\text{sig}}t)$$

and

$$E_{\text{LO}} = A_{\text{LO}} \exp(j\omega_{\text{LO}}t),$$

where $A_{\text{sig}}$ and $A_{\text{LO}}$ are the complex amplitudes and $\omega_{\text{sig}}$ and $\omega_{\text{LO}}$ are the

![Figure 2.12: BPSK receiver with an optical PLL.](image-url)

- **Signal**
- **Coupler**
- **LO laser**
- **Optical PLL**

...
Figure 2.13: Coherent phase diversity receiver with an optical 90° hybrid followed by a pair of balanced detectors.

$\omega_{LO}$ the angular carrier frequencies. The amplitudes are related to the optical power by $P_{\text{sig}} = |A_{\text{sig}}|^2$ and $P_{\text{LO}} = |A_{\text{LO}}|^2$. Mixing $E_{\text{sig}}$ and $E_{\text{LO}}$ in the 90° hybrid generates

$$E_{I \pm} (t) = \frac{1}{2} (E_{\text{sig}} \pm E_{\text{LO}})$$

(2.14)

$$E_{Q \pm} (t) = \frac{1}{2} (E_{\text{sig}} \pm jE_{\text{LO}}) ,$$

(2.15)

which are incident on the detectors. Bearing in mind that $|z|^2 = z^*z$ for a complex number $z$, the photocurrents after single-ended detection are

$$i_{I+} (t) = R_d \left| \frac{E_{\text{sig}} + E_{\text{LO}}}{2} \right|^2 = \frac{R_d}{4} \left[ |E_{\text{sig}}|^2 + |E_{\text{LO}}|^2 + 2R(E_{\text{sig}}E_{\text{LO}}^*) \right]$$

$$= \frac{R_d}{4} \left\{ P_{\text{sig}} + P_{\text{LO}} + 2\sqrt{P_{\text{sig}}P_{\text{LO}}} \cos \left[ \omega_{IF} + \theta_{\text{sig}} (t) - \theta_{\text{LO}} (t) \right] \right\}$$

(2.16)

$$i_{I-} (t) = R_d \left| \frac{E_{\text{sig}} - E_{\text{LO}}}{2} \right|^2 = \frac{R_d}{4} \left[ |E_{\text{sig}}|^2 + |E_{\text{LO}}|^2 - 2R(E_{\text{sig}}E_{\text{LO}}^*) \right]$$

$$= \frac{R_d}{4} \left\{ P_{\text{sig}} + P_{\text{LO}} - 2\sqrt{P_{\text{sig}}P_{\text{LO}}} \cos \left[ \omega_{IF} + \theta_{\text{sig}} (t) - \theta_{\text{LO}} (t) \right] \right\}$$

(2.17)

$$i_{Q+} (t) = R_d \left| \frac{E_{\text{sig}} + jE_{\text{LO}}}{2} \right|^2 = \frac{R_d}{4} \left[ |E_{\text{sig}}|^2 + |E_{\text{LO}}|^2 + 2\Im(E_{\text{sig}}E_{\text{LO}}^*) \right]$$

$$= \frac{R_d}{4} \left\{ P_{\text{sig}} + P_{\text{LO}} + 2\sqrt{P_{\text{sig}}P_{\text{LO}}} \sin \left[ \omega_{IF} + \theta_{\text{sig}} (t) - \theta_{\text{LO}} (t) \right] \right\}$$

(2.18)

$$i_{Q-} (t) = R_d \left| \frac{E_{\text{sig}} - jE_{\text{LO}}}{2} \right|^2 = \frac{R_d}{4} \left[ |E_{\text{sig}}|^2 + |E_{\text{LO}}|^2 - 2\Im(E_{\text{sig}}E_{\text{LO}}^*) \right]$$

$$= \frac{R_d}{4} \left\{ P_{\text{sig}} + P_{\text{LO}} - 2\sqrt{P_{\text{sig}}P_{\text{LO}}} \sin \left[ \omega_{IF} + \theta_{\text{sig}} (t) - \theta_{\text{LO}} (t) \right] \right\} ,$$

(2.19)

where $R_d$ denotes the detector responsivity and $\omega_{IF} = \omega_{\text{sig}} - \omega_{\text{LO}}$. The balanced in-phase ($I$) and quadrature-phase ($Q$) photocurrents are then obtained by taking the difference between $i_{I+}/Q+$ ($t$) and $i_{I-}/Q-$ ($t$)

$$i_I (t) = i_{I+} (t) - i_{I-} (t) = R_d \sqrt{P_{\text{sig}}P_{\text{LO}}} \cos \left[ \omega_{IF} (t) + \theta_{\text{sig}} (t) - \theta_{\text{LO}} (t) \right]$$

(2.20)
\[ i_Q(t) = i_{Q+}(t) - i_{Q-}(t) = R_d \sqrt{P_{sig}P_{LO}} \sin \left[ \omega_{IF}(t) + \theta_{sig}(t) - \theta_{LO}(t) \right], \] (2.21)

which suppresses the DC components. This is beneficial, since the term containing the signal power otherwise may distort the constellation. In addition, balanced detection makes use of the entire available signal and LO power.

**Polarization diversity**

Coherent detection requires the signal and the LO to be co-polarized, which calls for polarization tracking of the SOP. Tracking in the optical domain can be employed before detection, but in intradyne systems polarization diversity receivers are preferred. In these, a polarization beam-splitter separates the polarization components, which are detected with a pair of phase diversity coherent receivers. The linear relation between the photocurrents and the optical signal permits SOP tracking after detection. Chapter 4 provides more information about intradyne systems and DSP algorithms used for, e.g., polarization demultiplexing.

**Laser frequency drift and phase noise**

The carrier frequency of a laser drifts randomly around its mean value and there is normally a small frequency offset between the transmitter laser and the LO laser. Consequently, the received constellation rotates and assumes the shape of a number (which depends on the modulation format) of concentric rings. In addition, as discussed in section 2.1.1, the finite laser linewidth causes random phase fluctuations. The phase drift needs compensation after detection, except for in some pilot tone-based detection schemes, such as those investigated in papers A–C.

### 2.4 Wavelength-division multiplexing

In wavelength-division multiplexing (WDM) systems, channels at different wavelengths are multiplexed together and co-propagate in the transmission link. The channel frequencies are standardized by the International Telecommunication Union to cover the S-, C-, and L-bands and the available spacings are 25 GHz, 50 GHz, and 100 GHz. Fig. 2.14 shows a schematic of a WDM system. To join and split the WDM channels, a multiplexer (MUX)
and a demultiplexer (DEMUX) are used, respectively. The same device can be used for both purposes, depending on the direction of propagation. A MUX/DEMUX can be diffraction-based or interference-based [27, pp. 238–239]. In the first type a diffraction grating disperses the incoming light into different wavelength components, while the second type uses optical filters and directional couplers.

Rejection of crosstalk in direct detection receivers is highly dependent on the DEMUX, but in coherent systems only the WDM channel with a carrier frequency near the LO’s is down-converted within the receiver bandwidth, and residual crosstalk is removed with digital filtering. Ref. [76] presents an experiment about real-time coherent detection of 46 Gb/s polarization-multiplexed QPSK (PM-QPSK), and less than 1 dB OSNR penalty was measured for up to ten incident channels without optical filtering. This enables highly flexible networks, since channel selection may be performed by tuning the LO frequency. Transponders can then be re-routed along any path without manual reconnection to demultiplexer ports in case of wavelength tuning [76].

The capacity of WDM systems is limited by the gain spectra of optical amplifiers. C-band EDFAs have about 5 THz (40 nm) bandwidth, enabling amplification of 100 WDM channels on a 50 GHz grid. On the other hand, the low-loss window in state-of-the-art fiber extends over 400 nm (from 1.3 µm to 1.7 µm) [27], which can accommodate 1086 WDM channels at the same spacing. Unfortunately, utilizing the whole fiber bandwidth is difficult. Gain over 100 nm can be achieved with Raman amplification [77] but Raman amplifiers require a lot of pump power and have polarization sensitive gain [27]. Currently they are mainly used to complement EDFAs in order to reduce link noise figures. Ultra wideband amplification can also be obtained with parametric amplifiers [78], but due to the fast response time of the Kerr effect (∼fs) these devices cause nonlinear WDM crosstalk when saturated [42], and they are thus unsuitable for WDM transmission.

### 2.5 Polarization multiplexing

Due to the circular symmetry of the optical fiber, the fundamental mode exhibits a twofold degeneracy. These two modes can be used simultaneously to double the data rate compared to single-polarization transmission, which is known as polarization multiplexing\(^1\), dual polarization, or polarization division multiplexing. To create polarization-multiplexed data, the signals from a pair of modulators are combined with a polarization beam-combiner (PBC), ensuring orthogonal SOPs. In experiments, polarization multiplexing is often emulated by splitting a single-polarization signal and recombining it with a relative path delay (for data decorrelation) with a PBC.

Tracking and compensation for the SOP variation is required in the receiver to separate the orthogonally polarized signals. This is often referred to as polarization demultiplexing. Intradyne systems use DSP after detection to achieve polarization demultiplexing, while optical polarization tracking is needed in direct detection systems (using, e.g., PM-D\(\text{N}\)PSK) or in case of self-homodyne coherent detection. Two examples of optical polarization tracking are found in [79,80].

\(^1\)Different notations for polarization multiplexing (“PM”, “PDM”, and “DP”) are used in the literature. In this thesis we use “PM”, and QPSK and 16-QAM with polarization multiplexing are referred to as “PM-QPSK” and “PM-16-QAM”.

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2.6 FEC coding

FEC coding reduces the BER by adding redundant bits in the transmitter, which are used to correct bit errors after detection. Historically, the target BER has been $10^{-9}$ or $10^{-12}$, but with FEC a much higher BER is tolerable. Reed-Solomon (RS)$(255,239)$ [81], the most common FEC code in optical communications, adds a 7% overhead to the signal (the payload is 239 bytes and the overhead 16 bytes), i.e. a 40 Gb/s signal is converted to 42.8 Gb/s. Up to eight bytes can then be fully corrected, meaning that the code can correct 8–64 bit errors. RS$(255,239)$ reduces a raw BER of $10^{-3}$ to $10^{-13}$ after decoding, provided the errors are random [82, p. 180]. FEC increases the transmission reach significantly, which will be exemplified with PM-QPSK. For this format the required OSNR is 5.8 dB smaller at a raw BER of $10^{-3}$ compared to $10^{-9}$ [83, p. 192]. Returning to Fig. 2.9 gives an idea of the impact on the transmission reach. Assuming a data rate of 112 Gb/s, a span length of 100 km, and differential coding, the required OSNR is 13.9 dB and 19.3 dB at BERs of $10^{-3}$ and $10^{-9}$, respectively\(^2\), which increases the reach from 2300 km to 8100 km.

FEC codes can be categorized into groups using “soft” or “hard” decisions. With hard decoding, binary decisions are made between ones and zeros, while a soft-decision decoder processes the signal to correct the errors. Soft decision FEC codes with about 20% overhead are expected in future systems with data rates of up to 1 Tb/s per channel. In addition, in some modern commercial FEC decoders [84] the decoding latency can be varied for a trade-off between error correction capability and latency. Latency is otherwise one of the drawbacks with FEC. For e.g. RS$(255,239)$, $255 \times 8 = 2040$ bits are required for each decoding iteration, meaning that the decoding of every bit takes, at least, 2040 bit time slots.

Some researchers already assume more advanced FEC codes to boost the transmission reach. In [85], a receiver sensitivity of 2.7 photons per bit and unrepeatered transmission over 370 km of ULAF was achieved with 6.23 Gb/s 4-pulse position modulation PM-QPSK with 19.25% coding overhead. Ref. [86] demonstrated Nyquist WDM transmission of 112 Gb/s PM-16-QAM, 7% overhead and an assumed “FEC threshold” of $3.8 \times 10^{-3}$ enabled a transmission reach of 1958 km with a net spectral efficiency of 7.12 bit/s/Hz. On the other hand, assuming 20.5% overhead and a FEC threshold of $1.5 \times 10^{-2}$, the system reached 3700 km at 6.32 bit/s/Hz.

\(^2\)The differential coding penalties are 0.55 dB and 0.15 dB at BER = $10^{-3}$ and BER = $10^{-9}$, respectively.
3 Self-homodyne coherent systems

This chapter is devoted to self-homodyne coherent systems. A brief background is given and the advantages and drawbacks are described in section 3.1. The remaining sections provide deeper insight into the properties of self-homodyne systems and some schemes that were proposed to improve their performance. The OSNR requirements and the impact of narrowband pilot tone filtering are discussed in section 3.2 together with the impact of PDL and SOP misalignment. The possibility to cancel SPM distortion is discussed in section 3.3 and a scheme to increase the spectral efficiency in WDM transmission is presented in section 3.4. The chapter concludes with some remarks on other pilot tone schemes.

3.1 Basic principle

Optical self-homodyne coherent detection was proposed in 2005 by Miyazaki et al. [24] from the National Institute of Information and Communications Technology (NICT) in Japan, to enable coherent detection without the need for high-speed DSP or optical PLLs. Self-homodyne systems use a polarization-multiplexed pilot tone as a phase reference in the receiver instead of a LO laser. In 2005 [24], 20 Gb/s transmission with 2 bits per symbol and real time detection was achieved by combining PSK with the inverse RZ format [87], and real-time detection of 20 Gb/s QPSK in a WDM system at 1.2 bit/s/Hz spectral efficiency was demonstrated [88]. In addition, the tolerance to DGD, residual dispersion and the optimal power ratio between the signal and the pilot tone was investigated [89]. The NICT researchers were inspired by Betti et al., who worked on phase noise tolerant coherent systems many years before [31]. However, Betti and coworkers proposed using both a polarization-multiplexed pilot tone and a LO laser for coherent detection, and electronic signal processing to recover the signal phase and SOP. Fig. 3.1 shows a typical transmitter in a self-homodyne system. The laser output is split into two branches by a polarization beam splitter (PBS). In the upper branch a signal is generated by an IQM (or any kind of modulator), but the lower branch is passive and the pilot tone passes unaffected. The signal and the pilot tone are recombined with a polarization beam combiner (PBC), ensuring orthogonal SOPs.

Advantages and motivation

The main advantage with self-homodyne coherent detection, and the main motivation to investigate it, is the phase correlation of the signal and the pilot tone, which gives great phase noise tolerance. Using lasers with up to 30 MHz linewidth, 160 km transmission of 20 Gb/s QPSK and 40 Gb/s 16-QAM was achieved in 2006 and 2008, respectively [90,91]. The symbol rate was 10 Gbaud in both cases, corresponding to $\Delta \nu T = 6 \times 10^{-3}$, at least two orders of magnitude higher than typical values in intradyne 16-QAM experiments [92–94].
Since phase recovery is not needed, coherent detection is possible without high-speed DSP, or with relaxed requirements on the signal processing. Real time BER measurements were also presented for 10 Gbaud 8PSK and 5 GBaud 64-QAM [95,96].

In the experiments cited above, integrated transmitter and receiver components were used to avoid phase drift (due to, e.g., temperature fluctuations) between the pilot tone and the signal when travelling in separate fibers. Note however that the very slow phase drift between the two polarization components still requires tracking in the receiver, which can be performed with, e.g., a phase shifter in the optical path of either the phase reference or the signal, or after detection if DSP is used. With higher-order QAM formats, ISI is often an important performance limitation and may need compensation to avoid BER floors. In [91,96], ISI compensation was performed in the transmitter [97] to enable real-time BER measurements. The arbitrary waveform generator used for driving signal generation limited the symbol rate to 10 Gbaud in these experiments.

An additional advantage with self-homodyne detection is the absence of cycle slips, which makes it possible to use non-differential coding even for higher-order QAM constellations and thereby reducing the required OSNR in the receiver.

**Drawbacks**

There are also a number of drawbacks with self-homodyne detection. Co-propagating the phase reference in the fiber link causes an OSNR penalty due to ASE noise, as shown in paper A. In addition, conventional polarization multiplexing cannot be used, since the pilot tone has the same carrier frequency as the modulated signal. To make self-homodyne systems commercially viable, schemes increasing the spectral efficiency are needed, as a factor of two is lost compared to intradyne systems otherwise.
3.2 OSNR requirement and pilot tone filtering

In paper A we proposed a scheme to increase the receiver sensitivity of self-homodyne detection by using narrow (sub symbol rate) band-pass filtering of the pilot tone after separating it from the signal, as illustrated in Fig. 3.2. The first band-pass filter (channel selection BPF) before the receiver selects one channel. The signal and the pilot tone of this channel are split and the pilot tone is filtered by an additional narrow-band filter (PT BPF). As shown in paper A, narrow filtering and optimization of the power ratio between the pilot tone and the signal reduce the required OSNR in self-homodyne detection, and makes the performance approach intradyne detection. On the other hand, there is a practical limitation of how narrowly a signal can be filtered and the intradyne limit is only obtainable in theory. Due to, e.g., thermal noise there is also a limitation on how low the pilot tone power may be in practise. Since the center frequencies of both the pilot tone and the narrow band-pass filter may drift with time, it is important to implement a control system in the receiver that maximizes the pilot tone power after the filter. Another scheme to improve the sensitivity, which is not mentioned in paper A, is to use the pilot tone for optical injection-locking [98] to generate a new phase reference at higher OSNR. Injection-locking was recently used to demonstrate pump recovery for phase-sensitively amplified links [99].

3.2.1 Tolerance to SOP misalignment and PDL

Since DSP-based polarization demultiplexing is not used in self-homodyne detection, it is of interest to explore the impact of polarization misalignment in self-homodyne receivers. Fig. 3.3(a) shows the results from numerical simulations of how the OSNR required to achieve a BER of $10^{-3}$ is affected by deviating from the optimal SOP orientation before the PBS. Gray-coded QPSK and 16-QAM at 10 Gbaud have been used. The symbol rate affects the required OSNR, but not the impact of the misalignment. QPSK is as expected more robust and a 0.5 dB OSNR penalty compared to the case with perfect alignment is obtained at an angle of 13 degrees, as compared to 5 degrees for 16-QAM.

Fig. 3.3(b) shows the impact of PDL on self-homodyne detection. A lossy element has been introduced before the receiver and the performance depends on the orientation angle of the signal and the pilot tone relative to this element. There are two extreme cases. (i) the orientation angle is 0° and the whole loss affects one polarization, e.g., the signal power is reduced but not the pilot tone power (or vice versa). In case (ii), the orientation angle is 45° and the signal and the pilot tone are attenuated equally, which leads to loss of orthogonality of the polarization vectors. This is known as depolarization, and can intuitively
Figure 3.3: (a) The tolerance to polarization misalignment before the PBS in the receiver. (b) PDL tolerance for different orientations of a PDL element before the receiver.

severely degrade the phase reference. The OSNR degradation is also plotted after sweeping the orientation angle with 5° steps and averaging the BER. For 16-QAM depolarization is clearly most detrimental, but for QPSK having all the loss in one tributary is worse, even though the difference is small compared to the other extreme. For worst case PDL, 0.5 dB OSNR penalty occurs for PDLs of 1.25 dB and 0.55 dB for QPSK and 16-QAM, respectively.

3.3 SPM tolerance and the Manakov model

An interesting feature of self-homodyne systems is the ability to cancel signal distortion caused by SPM. According to the Manakov model [41], the XPM phase shifts between two orthogonally polarized signals with the same center frequency is the same as the SPM phase shifts. This implies that the signal and the pilot tone in a self-homodyne system obtain the same nonlinear phase-shifts, and the SPM-shift of the signal is subsequently cancelled in the photodetection process. We found that cancellation indeed can be achieved and demonstrated this in 2009 [100]. The work was later extended in paper B. Unfortunately, the SPM cancellation is incompatible with narrow pilot tone filtering, since this removes the phase modulation on the pilot tone and reduces the efficiency of the cancellation of nonlinear distortion. As shown in paper B, the cancellation also depends on the CD, and it is beneficial to have small accumulated dispersion in the locations of the system where the optical power is high.

Fig. 3.4 shows numerical simulation results illustrating SPM cancellation. A 10 Gbaud 16-QAM signal was transmitted over a link with four spans of 80 km SSMF and periodic dispersion compensation with DCF. The power launched into each fiber span was −1 dBm for both the signal and the pilot tone. Figs. 3.4(a)-(b) show the signal and the pilot tone before the receiver. They are affected by both ASE noise and nonlinear distortion due to SPM and XPM from each other. Since the phase reference has a similar nonlinear phase-shift as the signal, the symbols obtain their correct positions after coherent detection, as shown in Fig. 3.4(c).

3.4 Interleaved polarization division multiplexing

The biggest drawback with self-homodyne systems is the incompatibility with conventional polarization multiplexing. Unless properly addressed, this issue reduces the spectral efficiency compared to intradyne systems with a factor of 2. Fig. 3.5(a) shows the spectra of a
WDM signal with polarization multiplexing. Such spectra cannot be used in self-homodyne systems due to the co-propagating pilot tones and a single polarization state is thus used to transmit data, as illustrated in Fig. 3.5(b). However, in 2010 we demonstrated the interleaved polarization division multiplexing (IPDM) scheme to increase the spectral efficiency of self-homodyne systems, by placing the signals as in Fig. 3.5(c) [101]. The channel spacing is the same in the polarization tributaries, but a relative wavelength shift is made which allows the pilot tones to fit between the data spectra. IPDM was then further investigated in paper C. Experimental and numerical simulation results show that in order to obtain high spectral efficiency in an IPDM system, pre-filtering of the data spectra is required. Overlap between pilot tones and neighboring data spectra has a negative impact on the performance since it is critical to have a high quality phase reference in a coherent system. In addition, narrow pilot tone filtering in the receiver is crucial to obtain good performance, and in paper C we investigated the impact of different filter shapes and bandwidths. The measured increase in spectral efficiency compared to self-homodyne WDM systems using a single polarization was 33%, at a 2 dB OSNR penalty compared to the single channel performance. As shown in paper C the penalty can be reduced with better filtering.

There are some additional complications compared to systems with polarization multiplexing and intradyne detection. The channels in an IPDM system need to be polarization-
aligned to minimize crosstalk. Polarization maintaining fibers should be used in the transmitter to deal with this. IPDM is also incompatible with conventional optical add-drop multiplexers and new solutions are required to use the scheme in optically routed networks.

## 3.5 Other pilot tone schemes

Pilot signals have been used in several experiments to enable transmission of higher-order QAM. The optical communications group at Tohoku University has presented experimental results for 128-QAM [102], 256-QAM [103], and 512-QAM [104], and used frequency-shifted pilot tones in these experiments to track the phase of the LO laser in the receiver under optical PLL operation [105]. The linewidth of the beat-signal between the pilot tone and the LO laser was measured to be less than 10 Hz [103]. Pilot tones may be essential for reliable phase estimation for higher-order QAM.

Another interesting scheme is to use a pilot tone in combination with a frequency comb from a single laser source. With the exception of the pilot, all subcarriers in the comb are modulated, and in the receiver a comb of phase references is generated from the pilot tone and used for homodyne detection.
4 Intradyne coherent systems

This chapter is devoted to intradyne coherent systems, which use DSP for impairment mitigation and polarization demultiplexing. To the best of our knowledge the term “intradyne” was first used by Davis et al., who in 1986 demonstrated a phase-diversity coherent receiver for amplitude-shift keying and presented experimental results for data rates of 320 and 680 Mb/s [106]. The system combined the advantages of homodyne (low electrical bandwidth) and heterodyne (no optical phase synchronization) detection by using a $3 \times 3$ coupler to combine the signal and the LO and generate three beat signals with $120^\circ$ phase difference. The data was reconstructed by squaring and adding the photocurrents after detection [107].

The first experiment on intradyne detection with digital demodulation was presented by Derr in 1991 [21]. 100 Mb/s QPSK was coherently detected and the in-phase and the quadrature-phase photocurrents digitized and processed with a lookup table to extract the phase difference between the QPSK signal and the LO. The phase difference was fed to a digital PLL which generated a feedback signal. Frequency-stabilized HeNe lasers with narrow linewidth were used in the experiment ($f_{IF} < \pm 6.25$ MHz) and the system reportedly worked for hours with BER $< 10^{-9}$. Derr’s motivation for pursuing multilevel modulation and coherent detection was not to increase the spectral efficiency, but to improve the performance of long-haul systems over dispersion limited channels and to reduce the bandwidth requirements of electrical components [22].

For reasons discussed in chapter 1, coherent systems did not attract much attention in the nineties, but with time a growing concern over the spectral efficiency of WDM systems, and how to mitigate transmission impairments, led to a renewed interest. Thirteen years after Derr’s QPSK experiment, Taylor demonstrated intradyne coherent detection with the first DSP compensation of CD [23]. After this pioneering work, tremendous efforts have been made to investigate intradyne coherent detection and systems operating at 100 Gb/s with real-time DSP are now commercially available. Section 4.1 introduces the principles of intradyne detection and briefly summarizes some important real-time implementations after Derr’s work. Section 4.2 discusses DSP and some important algorithms for equalization and phase estimation, as well as CD compensation.

4.1 Basic principle

In the intradyne receiver, the signal is mixed with the LO field in a $90^\circ$ optical hybrid, which provides I- and Q-signals that are detected by balanced photodetectors. The photocurrents are sampled by ADCs and DSP performs clock recovery, tracks the intermediate frequency (IF) between the signal and the LO laser, and compensates for transmission impairments such as CD, PMD, and nonlinear effects. As there is no need to phase lock the lasers,
Intradyne systems are more cost effective and easier to implement than the coherent systems from the eighties relying on optical PLLs.

Intradyne systems normally have phase and polarization diversity [107], implying that the signal can be demodulated regardless of its phase orientation or polarization state. The ambiguity in the launch SOP of a polarization-multiplexed signal can be circumvented by using pilot symbols, or the FEC, to find the correct alignment. In research laboratories, experiments on intradyne coherent systems almost exclusively use offline processing of the data, i.e. the photocurrents are digitized and downloaded to a PC, which performs the signal processing offline. There are some notable drawbacks with this, such as utilization of DSP algorithms that are too complex for real implementation, and difficulty to capture the impact of events occurring on long time scales, such as PMD fluctuations over time.

Real-time intradyne detection

Real time intradyne detection at high data rates requires high speed ASICs that are costly and challenging to develop. Alternatively, the DSP can be implemented in a field-programmable gate array (FPGA) circuit, which has lower performance but is useful for performance verification due to its lower cost and re-programmability, in contrast to an ASIC that cannot be modified after being manufactured. After the interest in coherent transmission was reborn, an important step forward for real-time intradyne detection was taken by Pfau et al. in 2006. [108]. They transmitted QPSK at 800 Mb/s over 63 km of SSMF and used an FPGA-based receiver for signal recovery. Parallel signal processing at a speed 16 times lower than the symbol rate was applied, but after phase estimation the data were recombined to the original bit stream, whose BER was measured in real-time. Laser phase noise was a major limitation in the experiment, largely due to the low symbol rate. $\Delta \nu T$ was 0.01 and differential data encoding (see section 5.1) was used to avoid cycle slip-induced error bursts. The same group managed to double the data rate later that year [109], but manual control of the SOP before the receiver was still required. Real-time detection of PM-QPSK at 2.4 Gb/s with DSP-based polarization tracking was instead presented in 2007 [110]. The BER under rapid polarization variation was similar as for a fixed polarization state before the receiver.

A milestone for fiber-optic communications was reached by Nortel (now Ciena), who demonstrated an ASIC-based intradyne coherent receiver for 40 Gb/s PM-QPSK [111]. Clock recovery, dispersion compensation, carrier recovery, polarization-tracking, and equalization were performed digitally, which required 12 trillion integer operations per second. The ASIC was manufactured using 90 nm CMOS technology and had 21 W of power dissipation. The new receiver was used by AT&T in an experiment investigating the impact of PMD on PM-QPSK in 80 channel WDM transmission [112]. The reported performance was promising, with 1.2 dB and 2.2 dB OSNR penalties with respect to back-to-back for instantaneous DGDs of up to 100 ps and 127 ps, respectively. The performance dependence on the launch polarization was in addition found to be minimal. This work was a large step forward for coherent communications, as most previous experiments had used offline data processing. At the present time, several companies offer commercial intradyne PM-QPSK transceivers.

Yoshida et al. from Tohoku university reached another milestone with their FPGA-based receiver for polarization-multiplexed 1 Gbaud 64-QAM (12 Gb/s), demonstrated in 2011 [113]. The receiver utilized the pilot tone assisted scheme from their previous work with offline data processing [105] and a back-to-back BER lower than $10^{-12}$ was obtained. After transmission over $2 \times 80$ km SSMF, a BER of less than $10^{-7}$ was measured.
4.2 Digital coherent receivers

Fig. 4.1 shows a schematic of a digital coherent receiver. The optical front-end consists of a polarization diversity $90^\circ$ optical hybrid followed by four balanced photodetectors. After digitization of the I- and Q-photocurrents from the two polarization tributaries (the sampling rate is 2 samples per symbol, to satisfy the Nyquist criterion [114]), post-processing of the data is performed. An excellent overview of the DSP blocks is provided in [115]. They may be listed as:

1. Compensation for front-end imperfections, such as unequal detector responsivities, or IQ angle error in the optical $90^\circ$ hybrids.
2. Low-pass filtering for noise suppression.
3. CD compensation.
4. Digital clock recovery to enable re-sampling of the signal at the optimal sampling instant.
5. Polarization demultiplexing, equalization, and sampling at one sample per symbol.

In the work described in this thesis, all the DSP blocks above have been developed and implemented, with the exception of clock recovery due to the use of synchronous sampling. Most of the efforts concern polarization demultiplexing and phase estimation and these blocks are, together with CD compensation, explained in more detail.

4.2.1 Compensation for chromatic dispersion

After transmission in links without optical dispersion management, CD compensation is one of the first DSP steps after coherent detection [115]. It may be performed in the time domain by using a finite impulse response filter, or in the frequency domain by multiplying the Fourier transform of the signal with the inverse of the transfer function of the fiber link

$$T_{\text{CD}} = \exp \left( \frac{jz\beta_2\omega^2}{2} \right), \quad (4.1)$$
where \( z \) is the total transmission distance and \( \omega \) the angular frequency of the carrier wave. CD compensation with DSP permits removing the DCF (thus increasing the reach) and the optical amplifiers placed before the DCF modules. An additional cost advantage for digital compensation is due to optical dispersion management requiring careful tuning of the DCF lengths to perform well at high data rates [116].

A potential problem with DSP-based compensation is the conversion of LO phase noise to amplitude fluctuations [28, 29]. The impact of this effect on PM-QPSK was investigated by Xie with numerical simulations [117]. For residual dispersion of 30000 ps/nm (1800 km SSMF if \( \lambda = 1550 \) nm), the maximum LO linewidths causing less than 0.5 dB OSNR penalty in the receiver were 3 MHz, 1.1 MHz, and 0.4 MHz for data rates of 42.8 Gb/s, 112 Gb/s, and 224 Gb/s. These results show that much narrower linewidths may be required in practise, than those obtained with numerical simulations of the back-to-back performance [32, 33].

Another drawback with DSP compensation is the power consumption. Roberts et al. reported that of the 10 W dissipated by their real-time coherent PM-QPSK receiver, 46% is due to the CD compensation [118]. This should however be balanced against the possibility of reducing the number of optical amplifiers compared to dispersion-managed links.

### 4.2.2 Polarization demultiplexing and equalization

Polarization demultiplexing and equalization is the second DSP task. Equalization compensates for linear transmission impairments, such as residual CD, PMD, and ISI induced by, e.g., receiver and transmitter components. Sampling at 1 sample per symbol is usually performed in this stage too. To equalize the signal a “cost function”, which depends on the modulation format, is used. Minimizing the cost function yields an equalized constellation and the updating rules for the taps of the equalization filter are found by differentiating the cost function with respect to the complex conjugated tap coefficients [119, Theorem 3].

In what follows, \( i_x \) and \( i_y \) are the photocurrents at two samples per symbol before equalization. The length of the vectors depends on the number of equalizer taps. The equalizer consists of four filters in a butterfly structure, and provides the output signals

\[
\begin{align*}
    i_x &= h_{xx}^T i_x + h_{xy}^T i_y, \\
    i_y &= h_{yx}^T i_x + h_{yy}^T i_y.
\end{align*}
\]

The equalization algorithms used in the work in this thesis are outlined below.

**The constant modulus algorithm**

The constant modulus algorithm (CMA) [120] is frequently used to perform polarization demultiplexing and equalization of modulation formats with a single amplitude level (constant modulus). As the CMA works independently of the carrier frequency and the phase, it is suitable for blind equalization, i.e. filter adaptation is performed without a training sequence, which is considered to be an attractive feature. The taps of the butterfly filter are updated to minimize the cost function

\[
J_{\text{CMA}} = E \left[ (|i_x|^2 - P_0)^2 + (|i_y|^2 - P_0)^2 \right],
\]

where \( P_0 \) is the average signal power in each polarization. It is worth noting that (4.4) is minimized even if the same current is output twice, which is known as the singularity problem of the CMA [115].
Figure 4.2: Measured photocurrents of PS-QPSK after 3320 km transmission and polarization demultiplexing and equalization with the PS-CMA proposed in paper D.

The updating rules are \[ h_{xx}^{(k+1)} = h_{xx}^{(k)} - \mu \left( |i_x|^2 - P \right) i_x^* i_x, \] \[ h_{xy}^{(k+1)} = h_{xy}^{(k)} - \mu \left( |i_x|^2 - P \right) i_x^* i_y, \] \[ h_{yx}^{(k+1)} = h_{yx}^{(k)} - \mu \left( |i_y|^2 - P \right) i_y^* i_x, \] \[ h_{yy}^{(k+1)} = h_{yy}^{(k)} - \mu \left( |i_y|^2 - P \right) i_y^* i_y, \] where $\mu$ is the step size (typically $10^{-4}$–$10^{-3}$) and $k$ the iteration number. In general, a large $\mu$ provides good polarization tracking capability, but may give bad steady-state performance. In the work in this thesis, $\mu$ values of $1 \times 10^{-4}$–$2 \times 10^{-4}$ were used. There are algorithms providing better polarization tracking/equalization than the CMA. Independent component analysis (ICA) can be used for both constant modulus formats and higher-order QAM. ICA has been shown to exhibit both faster convergence than the CMA for a given SNR penalty and a significantly lower probability of failure. In particular, ICA does not have a singularity problem. Still, the CMA is often preferred due to its good performance and low computational complexity compared to competing algorithms.

The polarization-switched constant modulus algorithm

The CMA does not work for polarization-switched QPSK (PS-QPSK, see section 5.3.1), even though this modulation format is a subset of PM-QPSK. The reason is that fulfilling the constant modulus criterion is not sufficient for proper polarization demultiplexing of PS-QPSK, and, e.g., a phase offset between the polarization tributaries is not tolerable. In paper D we proposed a new algorithm, the polarization-switched CMA (PS-CMA), which may be used instead of the CMA. Its cost function is

\[ J_{PS-CMA} = E \left[ \frac{\left( |i_x|^2 + |i_y|^2 - P \right)^2}{2} + |i_x|^2 |i_y|^2 \right], \]
where $P$ is equal to the total average power. (4.9) is minimized when a QPSK symbol in one tributary is accompanied by zero power in the orthogonal state. Similar to the cost function of the conventional CMA, it has a singularity problem, but for PS-CMA misconvergence is easy to detect, since it implies that one output signal has zero power. The filter taps in PS-CMA are updated according to

$$h^{(k+1)}_{xx} = h^{(k)}_{xx} - \mu \left[ |i_x|^2 + 2 |i_y|^2 - P \right] i_x i_x^*, \quad (4.10)$$

$$h^{(k+1)}_{xy} = h^{(k)}_{xy} - \mu \left[ |i_x|^2 + 2 |i_y|^2 - P \right] i_x i_y^*, \quad (4.11)$$

$$h^{(k+1)}_{yx} = h^{(k)}_{yx} - \mu \left[ |i_y|^2 + 2 |i_x|^2 - P \right] i_y i_x^*, \quad (4.12)$$

$$h^{(k+1)}_{yy} = h^{(k)}_{yy} - \mu \left[ |i_y|^2 + 2 |i_x|^2 - P \right] i_y i_y^*. \quad (4.13)$$

As we show in paper D, PS-CMA has similar tracking properties and SNR performance as the standard CMA has for PM-QPSK. In addition, a switch between the two algorithms may be performed by using a common set of updating rules for the filter taps and changing the value of one parameter. This is useful if a system is alternating between PM- and PS-QPSK, which may be the case if the conditions in a link calls for a more robust modulation format. Fig. 4.2 shows measured photocurrents (from the experiment reported on in paper F) for 20 Gbaud PS-QPSK after 3320 km transmission and equalization with the PS-CMA.

**Decision-directed equalization**

To equalize and polarization demultiplex $N$-ary quadrature amplitude modulation (QAM) formats with $N \geq 8$, the CMA is often used for pre-convergence in a first equalizer stage, and a decision-directed least-mean square (DD-LMS) equalizer [121, 124] is applied to fine tune the performance. The reason to use the CMA first is the sensitivity of the DD-LMS equalizer to the initial filter tap weights. The cost function of the DD-LMS equalizer is

$$J_{DD} = \mathbb{E} \left[ \frac{1}{2} |X_D - i_{x,y}|^2 \right], \quad (4.14)$$

where $X_D$ is the constellation point closest to $i_{x,y}$. The updating rules for the filter taps are

$$h^{(k+1)}_{xx} = h^{(k)}_{xx} - \mu [X_D - i_x] i_x^*, \quad (4.15)$$

$$h^{(k+1)}_{xy} = h^{(k)}_{xy} - \mu [X_D - i_x] i_y^*, \quad (4.16)$$

$$h^{(k+1)}_{yx} = h^{(k)}_{yx} - \mu [X_D - i_y] i_x^*, \quad (4.17)$$

$$h^{(k+1)}_{yy} = h^{(k)}_{yy} - \mu [X_D - i_y] i_y^*. \quad (4.18)$$

Filter adaptation with the DD-LMS equalizer is often combined with decision-directed phase tracking, with an algorithm such as the one in (4.21)–(4.23) in section 4.2.3.

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4.2.3 Phase noise and frequency offset estimation

Since the LO laser in an intradyne system is free running (i.e., not locked to the signal laser), IF tracking is required after detection to retrieve the data. The IF is defined by $f_{IF} = \nu_{\text{sig}} - \nu_{\text{LO}}$, where $\nu_{\text{sig}}$ and $\nu_{\text{LO}}$ are the carrier frequencies of the signal laser and the LO laser. Compensation is required for both the frequency offset and the phase noise.

Fig 4.3 shows the spectra of two lasers before and after detection. The offset is 20 MHz and both lasers have 2 MHz linewidth, giving a beating linewidth of 4 MHz. Frequency offset gives a photocurrent a ring-like shape, since a phase variation which increases linearly with time is superimposed on the data. The offset is easily removed by applying FFT on the signal, locating the strongest frequency component, and applying a linear phase shift with opposite sign. Laser phase noise (section 2.1.1) makes the carrier phase vary in a stochastic manner, and the received signal thus exhibits a random phase drift whose speed depends on the sum linewidth of the lasers [32]. The complexity of the phase noise tracking depends on the modulation format and, e.g., PM-16-QAM requires more complex algorithms than PM-QPSK. Figs. 4.4 (a)-(b) show simulated QPSK and 16-QAM constellations before and after tracking and compensation of frequency offset and laser phase noise. Due to its three amplitude levels, 16-QAM assumes the shape of three concentric rings in its uncompensated state.

The Viterbi-Viterbi estimator

Phase estimation for modulation formats with constant phase spacing between the symbols is performed with the well-known Viterbi-Viterbi estimator [125], in combination with block-based processing of samples. The samples are segmented into blocks of length $M$ and the phase is considered to be approximately constant within each block. The phase estimates are

$$\hat{\theta}_k = \frac{1}{4} \arg \left( \sum_{l=m_k}^{m_k+M-1} i_l^4 \right),$$

as the name implies, decision-directed equalization requires a constellation to be reasonably close to its true shape to begin with.
where $m_k$ is the index of the first symbol in block $k$ and $i$ is the complex photocurrent from the $x$- or the $y$-polarized signal. There is a trade-off between laser phase noise and OSNR, since a small $M$ allows tracking of rapidly varying phases, while a long block is desired if the OSNR is low, to reduce the impact from ASE noise on the phase estimates. The case $M = 1$ corresponds to a symbol-by-symbol estimator, which can track rapid phase changes but is very sensitive to ASE noise.

An interesting remark is that slightly better phase estimation than with the Viterbi-Viterbi algorithm can be achieved with

$$\hat{\theta}_k = \frac{1}{4} \text{arg} \left( \sum_{l=m_k}^{m_k+M-1} |i_l|^p e^{j4\angle i_l} \right),$$

(4.20)

with $p = 1$ or $p = 2$ for QPSK [126]. The difference in performance between (4.19) and (4.20) is small, but noticeable at low OSNR. Fig. 4.5 shows a comparison between the Viterbi-Viterbi algorithm (4.19) and the algorithm in (4.20) with $p = 1$ and $p = 2$. The combined laser linewidth is 2 MHz and the performance is compared at OSNRs giving BER = $10^{-2}$ and BER = $10^{-3}$ without phase noise. Better phase tracking than with (4.19) can clearly be achieved, especially at the lower OSNR. Still, even in that case the difference in required OSNR is only 0.05 dB, and at the higher OSNR value the performance improvement with (4.20) is negligible.

The Viterbi-Viterbi estimator is used for PS-QPSK as well. The principle is the same as for PM-QPSK, but a decision is first made on the launch SOP of the QPSK symbols by
Figure 4.5: Phase tracking results for 28 Gbaud PM-QPSK with differential coding, with the Viterbi-Viterbi algorithm (4.19) and with the algorithm in (4.20). The phase error variance has been simulated as a function of the block length for (a) an OSNR of 11.75 dB (BER = $10^{-2}$ without phase noise), and (b) an OSNR of 13.85 dB (BER = $10^{-3}$ without phase noise). $\Delta \nu = 2$ MHz and $2^{15}$ symbols were used in each simulation.

comparing the amplitudes of $i_x$ and $i_y$. The sample with the highest amplitude is selected for phase estimation, and block-based processing with (4.19) is applied.

**Phase estimation for high-order QAM formats**

For modulation formats with uneven phase spacing (e.g., 16-QAM) the Viterbi-Viterbi estimator is inaccurate (unless very long blocks are used) and other algorithms have been proposed that are more computationally demanding, but gives much better phase estimation. In this thesis the algorithm in [32] has been used for phase estimation for PM-16-QAM and 128-SP-QAM. The received signal $i_{x,y,k}$ is rotated with $B$ test angles for the carrier phase

$$\psi_b = \frac{b \pi}{2B},$$  \hspace{1cm} (4.21)

For each test angle a block of symbols is fed into a decision circuit and the squared distance to the closest point in the constellation, $|d_{k,b}|^2$, is calculated for each symbol in the block.

$$|d_{k,b}|^2 = |i_{x,y,k} \exp(j\psi_b) - X_{D,k}|^2,$$  \hspace{1cm} (4.22)

where $X_{D,k}$ is the closest constellation point. The sum

$$s_{k,b} = \sum_{n=-N}^{N} |d_{k-n,b}|^2,$$  \hspace{1cm} (4.23)
is then calculated, in which $N$ is the one-sided length of the block of samples. The choice of $N$ is a trade-off between noise suppression and tracking of fast phase variations. Compared to using the Viterbi-Viterbi estimator, this algorithm is significantly more accurate for higher-order QAM [33].
5 Multilevel modulation formats

OOK was for many years the only modulation format in optical communications, which is understandable since it exhibits advantages such as low implementation complexity and impairment robustness. The lone exception was DBPSK, offering 3 dB higher sensitivity and similar complexity. OOK and DBPSK are binary formats, encoding a single bit per symbol in the amplitude and the differential phase, respectively. More efficient bandwidth utilization can be achieved with modulation formats using the amplitude, the phase and the polarization of the optical field simultaneously, which calls for coherent detection. The modulation format attracting most attention so far in coherent communications is QPSK with polarization multiplexing (PM-QPSK), but since a few years rectangular polarization-multiplexed 16-QAM (PM-16-QAM) is also a hot research topic, and experimental results have been reported even for 128-QAM [127], 256-QAM [103], and 512-QAM [104]. These formats have been used in wireless links for a long time but implementing them in optical communications is challenging, due to the much higher data rates and larger carrier phase noise. A lot of research is required on both the transmitter and the receiver side, in addition to investigating the tolerance to transmission impairments. There is also a significant difference in the available processing power per information bit between optical systems and RF systems, and the latter can afford using much more complex DSP. Still, a lot of progress has been made in recent years, which has paved the way for, e.g., commercial PM-QPSK transceivers.

This chapter will discuss multilevel modulation in optical communications. The first section in this chapter introduces important concepts such as power efficiency, signal-to-noise ratio, and spectral efficiency. In the remaining sections, the modulation formats that have been important for this work are discussed. They are categorized as conventional (PM-QPSK and PM-16-QAM) and four-dimensional (PS-QPSK, 128-SP-QAM and subset-optimized PM-NPSK) formats, and described in section 5.2 and section 5.3, respectively.

5.1 Basic concepts

In the following discussion we assume a discrete-time channel with additive, circularly symmetric, white Gaussian noise as the dominant impairment\(^1\). The noise is ASE generated by optical amplifiers, and/or shot noise. It is also assumed that the symbols in a constellation occur with equal probabilities. The spectral efficiency of a modulation format\(^2\) with \(M\) levels

\(^1\)A good assumption for coherent communications without optical CD compensation [63], assuming nonzero dispersion and weak to moderate impact from nonlinear effects.

\(^2\)It is also common to talk about the spectral efficiency of WDM systems, defined as the data rate per channel divided by the channel spacing and given in bit/s/Hz.
is then $SE = \log_2 (M)$, with the unit *bits per symbol*.

## Power efficiency and SNR

The vectors describing the symbols of a modulation format are formed from the real and imaginary part of the optical field’s x- and y-polarized components as [128]

$$
\mathbf{c} = \{ \Re(E_x), \Im(E_x), \Re(E_y), \Im(E_y) \}.
$$

The constellation $C$ of a modulation format with $M$ symbols is defined as

$$
C = \{ \mathbf{c}_1, \ldots, \mathbf{c}_M \},
$$

and has an average energy per symbol of

$$
E_s = \frac{1}{M} \sum_{k=1}^{M} ||\mathbf{c}_k||^2.
$$

The Euclidean distance between $\mathbf{c}_k$ and $\mathbf{c}_j$ is denoted by $d_{kj} = ||\mathbf{c}_k - \mathbf{c}_j||$, and the smallest Euclidean distance between any pair of symbols in $C$ is $d_{\text{min}} = \min_{k \neq j} d_{kj}$. The asymptotic power efficiency of $C$ is [129]

$$
\gamma = \frac{d_{\text{min}}^2 \log_2 (M)}{4E_s} = \frac{d_{\text{min}}^2}{4E_b},
$$

where $E_b$ is the energy per bit. The factor of 4 in the denominator normalizes the power efficiency to 1 (0 dB) for BPSK, QPSK, and PM-QPSK. (5.4) compares modulation formats at the same data rate, but replacing $E_b$ with $E_s$ enables comparisons at the same symbol rate.

The SNR in digital communications is usually expressed as $E_b$ divided by $N_0$, the noise power spectral density. In case of ASE limited balanced coherent detection, it is related to the OSNR by

$$
\frac{E_b}{N_0} = \frac{2R_b \text{OSNR}}{2.1 \text{nm}},
$$

where $R_b$ is the bit rate. Hereafter, we assume that $E_b$ is fixed for each modulation format and the SNR is varied by changing $N_0$. Modulation formats with larger $M$ generally require higher SNR to achieve a specific BER, which causes a trade-off between spectral efficiency and power efficiency.

## BER and SER

Each symbol $\mathbf{c}_k$ in (5.2) is surrounded by a decision region, the *Voronoi region*, defined as all points in the signal space closer to $\mathbf{c}_k$ than to any other points in $C$. The error probability of $\mathbf{c}_k$ is the probability of judging it to be outside of the Voronoi region (due to, e.g., noise). There are analytic SER and BER expressions for some modulation formats, but others have too complex Voronoi regions to enable exact computations. However, a simple and useful SER approximation is the *union bound* [83, p. 184]

$$
\text{SER} \leq \frac{1}{M} \sum_{k=1}^{M} \sum_{j=1}^{M} \frac{1}{2} \text{erfc} \left( \frac{d_{kj}}{2\sqrt{N_0}} \right),
$$

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Figure 5.1: 16-QAM with (a) Gray coding. The most likely symbol errors cause a single bit error. (b) Natural coding. Some of the most likely symbol errors cause two bit errors.

which considers the pairwise probability of error from confusing the symbols $c_k$ and $c_j$ (easily calculated, since it depends only on $d_{kj}$). By modifying (5.6) to account for the number of bit errors for each symbol error, i.e. the Hamming distance $d_H$, we also obtain an upper limit for the BER

$$BER \leq \frac{1}{M} \sum_{k=1}^{M} \sum_{j=1, j \neq k}^{M} \frac{1}{2} \text{erfc} \left( \frac{d_{kj}}{2 \sqrt{N_0}} \right) \cdot \frac{d_H(c_k, c_j)}{\log_2(M)}. \quad (5.7)$$

Union bound expressions for the BER and the SER were derived in paper H to evaluate the performance of a modulation format whose Voronei regions did not permit exact computations.

Gray coding

The assignment of bit sequences to symbols is often performed using Gray coding [83, p. 100], for which a single bit error occurs when a symbol is mistaken for one of its nearest neighbors.\(^3\) Detecting a nearest neighbor is, by far, the most likely error event in the presence of circularly symmetric Gaussian noise, implying that $BER \approx SER/\log_2(M)$ with Gray coding. Fig. 5.1 illustrates the difference between Gray coding and natural coding\(^4\) for square 16-QAM. The most likely symbol errors are those that occur if a symbol is shifted to the other side of any one of the nearest detection thresholds (the dotted lines and the axis). By studying Fig. 5.1(a), it is easy to verify that such an operation gives one bit error in case of Gray coding. If we instead consider natural coding, shown in Fig. 5.1(b), some of the most likely symbol errors cause two bit errors.

Differential coding

Differential coding, proposed by Weber in 1978 [131], is used to overcome phase ambiguity in the carrier synchronization. For modulation formats with a fourfold phase ambiguity

\(^3\)symbols separated by $d_{\text{min}}$
\(^4\)the bit assignment with a particular transmitter setup in absence of coding
(which applies to all formats studied in this thesis), two bits per symbol can be encoded in the transitions between the quadrants in the complex plane, with the benefit that the data can be decoded regardless of the phase orientation of the signal. Differential coding is also important for coherent systems since it avoids error bursts due to cycle slips in the phase synchronization. Differential coding leads to an OSNR penalty, since one symbol error causes two differential errors, but the penalty is quite small and of little significance compared to the benefits.

Lattices

For spectrally efficient modulation (large $M$), the best constellations are regular structures, known as lattices, with spherical boundaries [132,134]. The optimal lattice in two dimensions is the hexagonal lattice, but implementing a modulation format based on this is difficult. Instead, the $Z_2$ lattice, also known as the square lattice, is frequently used. QPSK and 16-QAM belong to $Z_2$ when transmitted in a single polarization. In case of polarization multiplexing, their constellations consist of one and four hypercubes, respectively, and are part of the $Z_4$ lattice. Interestingly, it is possible to place an additional shifted $Z_4$ lattice inside $Z_4$ to create the $D_4$ lattice

$$D_4 = Z_4 \cup \left[ Z_4 + \left( \frac{1}{2}, \frac{1}{2}, \frac{1}{2}, \frac{1}{2} \right) \right],$$

which is the set of all four-dimensional (4-D) points whose integer coordinates have an even sum, and the densest lattice in four dimensions [132]. For a given $M$, the difference in $\gamma$ is 1.51 dB when comparing large constellations from $D_4$ and $Z_4$ [133].

Implementing constellations from lattices with spherical boundaries is difficult. To obtain a practical modulation format, a cubic cut can be made, in which $M$ points are cut out from the lattice with a cubic boundary. For a large $M$, the asymptotic power efficiency is reduced by 0.46 dB due to the cubic cut [134], which can be compared to 0.20 dB loss for circular cuts in two dimensions. With $Z_4$, a cubic cut gives the conventional QAM constellations (e.g., PM-QPSK and PM-16-QAM), while $M$-SP-QAM (e.g., 8-SP-QAM which is the same as PS-QPSK, and 128-SP-QAM) are obtained from $D_4$. SP-QAM is discussed in section 5.3.

5.2 Conventional modulation formats

Conventional modulation formats are optimized in a two-dimensional (2-D) signal space, but polarization multiplexing allow them to use the four degrees of freedom of the carrier, by transmitting independent constellations in the polarization components. Two such formats are addressed in this section, PM-QPSK and PM-16-QAM. Their constellations and transmitter realizations are described, and a brief historical overview (from a fiber optics perspective) is provided.

A few remarks should be made before we continue. Firstly, OOK and BPSK (which can be defined in a single dimension) may also be included among the conventional formats, but while being very important for optical communications (especially historically), they have not been used in this work and are not discussed further. Secondly, the decoding of PM-QPSK and PM-16-QAM is not described in detail, since it is based on a standard rectilinear grid of decision regions and hard decisions.

Please note that polarization multiplexing a 2-D constellation does not change its power efficiency. Therefore, e.g., QPSK has the same $\gamma$ as PM-QPSK.

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5.2.1 PM-QPSK

With the same receiver sensitivity as BPSK, spectral efficiency of four bits per symbol, and moderate implementation complexity PM-QPSK is becoming the mainstream format in long-haul transmission. Due to its maturity and performance, the format is frequently used as a benchmark for modulation schemes with similar spectral efficiency and/or power efficiency. In this thesis PM-QPSK was therefore used in paper D–F in comparisons with PS-QPSK, and in paper C to compare the IPDM scheme with a conventional WDM system.

The constellation and the generation of PM-QPSK

The PM-QPSK constellation is defined by the 16 levels

\[ C_{\text{PM-QPSK}} = \{ (\pm1, \pm1, \pm1, \pm1) \}, \] (5.9)

and has \( \gamma = 0 \) dB. Constellation diagrams of QPSK with non-differential and differential coding are shown in Figs. 5.2(a)-(b). Gray coding is applied in both cases and ensures that the most likely symbol errors cause a single bit-error. Although PM-QPSK is fairly robust against laser phase noise, differential coding is often preferred, as the coding penalty is moderate (0.55 dB at BER = 10\(^{-3}\)) and a single cycle slip otherwise may result in a long error burst.

PM-QPSK is generated with a pair of IQMs (one for each polarization) driven by binary electrical signals with peak-to-peak voltages of 2\( V_p \). Other transmitter implementations exist [135], but using IQMs is the best option since they take advantage of the nonlinear MZM transfer function, which reduces the impact from noise and driving signal distortion. In addition, Gray bit-to-symbol mapping is obtained without pre-coding. A coherent receiver with phase and polarization diversity (see Fig. 4.1) is required for detection. The CMA (section 4.2.2) is often used for polarization demultiplexing and equalization, and phase synchronization is performed with the Viterbi-Viterbi algorithm (section 4.2.3).

PM-QPSK in fiber optics

Although the first experiments on coherent transmission used BPSK, attention was soon shifted to PM-QPSK due to its higher spectral efficiency. The first generation and transmission of PM-QPSK was presented by Ly-Gagnon et al. in early 2006 [136]. A data rate
of 40 Gb/s was achieved and manual polarization control was used before the receiver. This landmark experiment was soon followed by others, and Pfau et al. demonstrated real-time intradyne detection of a 2.8 Gb/s signal (with DSP-based polarization demultiplexing) the year after [109]. Many great results have been achieved since. In 2009, Alcatel Lucent France demonstrated WDM transmission of $155 \times 100$ Gb/s over 7200 km of large effective area fiber with 90 km span length [17]. The 112 Pb-km/s bit rate distance product is still the world record. More recently the same group achieved WDM transmission over 12000 km with 200 Gb/s per channel [137]. To enable such long reach, the fiber span length was however only 50 km. A third important long-haul experiment was presented in 2010 by Cai et al., who applied narrow prefiltering in the transmitter and maximum likelihood sequence estimation, to demonstrate $96 \times 112$ Gb/s WDM transmission with spectral efficiencies of 3.0 bit/s/Hz and 4.0 bit/s/Hz over 10610 km and 4370 km, respectively [138].

In spite of all these great demonstrations, the most important PM-QPSK experiment is arguably the one demonstrated by Sun et al. four years ago [111], since it once and for all proved that coherent PM-QPSK transmission is practically feasible at high data rates.

### 5.2.2 PM-16-QAM

PM-16-QAM has attracted a lot of attention recently, since it is a promising next step after PM-QPSK to double the spectral efficiency, by transmitting eight instead of four bits per symbol. In this thesis the format was used in paper G as a benchmark for transmission reach and phase noise tolerance in a comparison with 128-SP-QAM.

#### The constellation and the generation of PM-16-QAM

Although there are other 16-QAM constellations, the abbreviation normally refers to square 16-QAM, which with polarization multiplexing is defined by the 256 levels

$$C_{PM-16-QAM} = \{ (\pm(1,3), \pm(1,3), \pm(1,3), \pm(1,3)) \}.$$ (5.10)
By using (5.4) we find that $C_{PM,16-QAM}$ has $\gamma = -3.98$ dB. Figs. 5.3 (a)-(b) show 16-QAM with non-differential coding and differential coding, respectively. Differential coding is important due to the format’s relatively low tolerance to phase noise. The coding penalty of 0.4 dB (at BER = $10^{-3}$) is smaller than for PM-QPSK, since only half the bits are differentially encoded.

The best PM-16-QAM transmitter has two parallel IQMs (quad-parallel MZM) per polarization, driven by binary electrical signals at $2V_p$. Such devices are difficult to fabricate and the first realizations could only generate constellations with noticeable distortion [37], but recent advances has enabled well-performing parallel modulators even for 64-QAM [38]. The advantages of the quad-parallel MZM are (i) it takes advantage of the nonlinearity of the MZM transfer function and (ii) smaller insertion loss than other schemes [139]. In spite of the progress in realizing the optimal solution, the most common transmitter implementation for PM-16-QAM is still to apply four-level electrical signals to a single IQM per polarization, as illustrated in Fig. 5.4. While this approach is well-proven, it has the drawbacks (compared to the quad-parallel MZM) of being more sensitive to noise and ISI, and higher insertion loss.

The receiver hardware for PM-16-QAM is similar as for PM-QPSK, but has more stringent requirements on the laser linewidths and the DSP. As mentioned in section 4.2.2, the CMA is often combined with a decision-directed equalizer to achieve satisfactory performance, and more complex phase tracking algorithms than the Viterbi-Viterbi estimator are preferred, such as the final algorithm in section 4.2.3.

PM-16-QAM in fiber optics

PM-16-QAM was first demonstrated in optical communications by Winzer et al. [140] in 2008. 10 channels at 112 Gb/s on a 25 GHz WDM grid were transmitted over 315 km. In an extension of this experiment the spectral efficiency was increased to 6.2 bit/s/Hz [92], and the paper included a more detailed description of the transmitter and the receiver. In 2010, the same group presented results from WDM transmission with 224 Gb/s per channel [93], and NTT set a new world record for high capacity transmission of 69.1 Tb/s, by using 432 channels with 171 Gb/s [141]. Generation of 448 Gb/s (56 Gbaud) PM-16-QAM and transmission over 1200 km of ULAF was then demonstrated [142]. The data rate was a record for single carrier systems not resorting to optical time-division multiplexing. The following year, Renaudier et al. reported on 1200 km WDM transmission of 40 channels at 448 Gb/s [143], which they accomplished by using fiber with ultra low loss and large effective area, and a free-space optics device for converting QPSK to 16-QAM.

![PM-16-QAM transmitter with one IQM per polarization tributary.](image)
In short, the experiments cited here have demonstrated the feasibility of WDM transmission of PM-16-QAM over distances up to \( \sim 1000 \) km. With new emerging technologies, more advanced FEC, and efficient schemes for compensation of nonlinear effects [94], longer reach will be possible. In 2012, Ciena also reported on a commercial dual-carrier PM-16-QAM 400 Gb/s system.


5.3 Four-dimensional modulation formats

While PM-QPSK and PM-16-QAM exhibit good performance they do not fully exploit the 4-D signal space of the carrier, due to being optimized in two dimensions and multiplexed onto orthogonal polarizations. More power-efficient formats can be found by considering the four dimensions simultaneously. 4-D modulation has a long history in wireless communications, but is a relatively new topic in fiber optics. Bülow was the first to present a practical constellation for fiber-optic communications that cannot be separated into a pair of polarization-multiplexed 2-D constellations [144]. He proposed polarization-QAM (POL-QAM), which is an extension of PM-QPSK from four to six polarization states and has a spectral efficiency of 4.5 bits per symbol. On the other hand, much of the work about modulation in RF communications, of which some was performed long before [144], is applicable to fiber optics as well. Already in 1973, Zetterberg and Brändström showed that taking the even or the odd parity subset from the 4-D hypercube creates a constellation with 1.76 dB higher asymptotic power efficiency than the hypercube itself [145]. In [145] the 4-D signal space was spanned by the phase and the amplitude of two different carriers, as opposed to two polarization components. The constellation is thus generated by launching QPSK on one carrier at a time, which from an information theory perspective is equivalent to using a selected polarization state for each symbol. The latter approach is in the optical community known as polarization-switched QPSK (PS-QPSK) [129,146], and happens to be the most power-efficient modulation format in signal spaces with up to four dimensions.

In contrast to PM-QPSK and PM-16-QAM, 4-D formats have correlation between the data in the polarization components, which makes it possible to increase the minimum distance between the symbols at a fixed average power. As a consequence, a slight increase in decoding complexity is required, since $i_x$ and $i_y$ should not be demodulated independently. This section presents decoding schemes for three 4-D formats. The conventional transmitters need modification as well to be used for 4-D formats, although the complexity is similar as for the conventional formats, since it is a matter of adding a few XOR gates or an additional amplitude modulator. Transmitters for 4-D modulation formats will also be demonstrated below.

5.3.1 PS-QPSK

It was previously believed that BPSK has the best ASE-limited sensitivity among practical modulation formats, but in 2009 an 8-level format, now known as PS-QPSK, was shown to have 1.76 dB higher power efficiency [129,146]. In addition, PS-QPSK can be generated with low complexity modulators. These features triggered a lot of interest and activities in the research community. PS-QPSK is the main topic in paper D–F.

The constellation and the generation of PS-QPSK

As the name implies, PS-QPSK can be described as QPSK transmitted in one selected polarization-state per symbol. Of the three bits per symbol, QPSK encodes the first two

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\[^{6}\text{The 4-D analogue of the cube, created by moving the cube one unit length into the fourth dimension.}\]

\[^{7}\text{From an information theory point-of-view, another implementation of PS-QPSK is as 2 pulse-position modulation QPSK (2PPM-QPSK), where two adjacent symbol slots in time are used instead of two polarizations [147].}\]
and the third determines the launch polarization. The constellation may be defined as

\[ \mathcal{C}_{\text{PS-QPSK,1}} = \sqrt{2} \{ (\pm 1, \pm 1, 0, 0), (0, 0, \pm 1, \pm 1) \} , \]  

which is QPSK transmitted in either the x- or the y-polarization. Fig. 5.5 shows the PS-QPSK realization in (5.11) and the bit-to-symbol mapping. Blue and red symbols indicate transmission in the x- (the third bit is a one) and the y-polarization (the third bit is a zero), respectively. Alternatively, PS-QPSK is obtained by separating PM-QPSK into the even and the odd parity subsets and choosing one of them. With the standard PM-QPSK transmitter (one IQM per polarization), the subset with even parity symbols is chosen by taking the data sequences \( d_1, d_2, d_3 \) and using a pair of XOR gates to generate the fourth driving signal

\[ d_{\text{par}} = \text{XOR}\{ \text{XOR} (d_1, d_2), d_3 \} . \]  

The PS-QPSK constellation is then

\[ \mathcal{C}_{\text{PS-QPSK,2}} = \pm \{ (1, 1, 1, 1), (1, 1, -1, -1), (1, -1, 1, -1), (1, -1, -1, 1) \} , \]  

which is equivalent to (5.11) after a 45° polarization rotation. The odd parity PM-QPSK subset is chosen by replacing one XOR gate with an AND gate.

Figs. 5.6(a)-(b) show two different PS-QPSK transmitters. The transmitter in Fig. 5.6(a) has one IQM per polarization and requires four driving signals, of which the fourth is generated with (5.12). The transmitter proposed in paper E is shown in Fig. 5.6(b) and requires three driving signals. Two drive an IQM, which is followed by a polarization-switching stage with a pair of amplitude modulators driven by data and data inverse of a third driving signal. The amplitude modulators may be replaced by a single polarization modulator [148]. At the post-deadline session of the Optical Fiber Communication Conference (OFC) 2012, NTT presented an integrated PS-QPSK modulator with three MZMs and a polarization-coupling circuit [149]. The modulator was free from the 3-dB IQ-coupling loss and had 23 GHz 3-dB bandwidth.

The receiver hardware is the same as for PM-QPSK, but there are differences in the DSP since the CMA does not work for PS-QPSK. In paper D we proposed the PS-CMA, which has similar tracking and SNR performance as the CMA has for PM-QPSK.
Decoding of PS-QPSK

The following methods are maximum-likelihood decoders for PS-QPSK.

1. The first scheme is based on the PS-QPSK realization in (5.11) and was used in paper E and paper F. Assume the received and equalized signals from the polarization tributaries x and y are $i_x = a(t) + jb(t)$ and $i_y = c(t) + jd(t)$. For each symbol a decision is made on the launch SOP by comparing $|a| + |b|$ and $|c| + |d|$, and the QPSK bits are obtained from the selected signal.

2. The PS-QPSK realization in (5.13) is assumed in the decoding and the decisions are made in the phase plane. This approach was used in e.g. [150].

3. This method is based on the PS-QPSK realization in (5.13). The decoding is similar as for PM-QPSK, but a parity check is made to take advantage of the correlation between the data in the tributaries. This approach is also used to decode 128-SP-QAM and is explained in more detail in section 5.3.2.

Differential coding for PS-QPSK

We demonstrated in paper F that PS-QPSK can be differentially encoded, with two bits encoded in the quadrant transitions and the third in the launch SOP. It is possible to

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8A maximum-likelihood decoder is the optimal decoder if the symbols are equiprobable [83, p. 163].
encode the third bit differentially too, but at the cost of a higher BER since a symbol error due to an incorrect decision on the launch SOP would give two bit errors instead of one. On the other hand, the scheme removes the polarization ambiguity in the receiver, which is an interesting feature.

In the decoding, a decision is first made on the launch polarization of the QPSK symbol, and the differential phase transition is found by comparing with the preceding symbol. To avoid error bursts due to cycle slips, joint phase estimation is required for $i_x$ and $i_y$. This applies to PS-QPSK with non-differential coding too.

**Experimental and numerical results**

Transmission of PS-QPSK was first investigated with numerical simulations in which it was compared with PM-QPSK at the same bit rate [151, 152]. Poggiolini et al. also suggested using the format as a fallback option to keep a degraded PM-QPSK-based link in operation [151]. The idea is to switch modulation format with maintained symbol rate, at the expense of a 25% bit rate reduction. It was reported in [151, 152] that PS-QPSK exhibits higher tolerance than PM-QPSK to nonlinear crosstalk between WDM channels. This was attributed to higher tolerance against XPolM for PS-QPSK [152], since it has two launch SOPs as compared to four for PM-QPSK. A conclusion supported by [71], where conventional PM-QPSK was compared with PM-QPSK with time interleaved return-to-zero, which exhibits data independent SOP. The time interleaving reportedly reduced XPolM significantly in dispersion-managed systems.

These interesting results were soon followed by experiments. By using a novel transmitter, PS-QPSK was generated experimentally for the first time in paper E, and the sensitivity advantage over PM-QPSK was confirmed. The format was also compared with PM-QPSK in single channel transmission and optical dispersion compensation was found to result in a smaller OSNR penalty in the receiver than DSP-based compensation, in agreement with numerical simulation results [62]. In other single channel experiments, Fischer et al. presented a comparison of PS- and PM-QPSK in an ULAF link over transoceanic distances [150] and demonstrated a significant reach improvement with PS-QPSK, and Lavery et al. demonstrated 20% improvement in transmission distance (from 4650 km to 5620 km at a BER of $3.0 \times 10^{-3}$) for PS-QPSK at 112 Gb/s, by using digital back-propagation after coherent detection [153].

Experimental WDM results comparing the two formats have also been presented. In [154], 30% longer transmission reach was achieved for PS-QPSK for a BER of $3.8 \times 10^{-3}$ in the receiver, in a system with seven 40 Gb/s 50 GHz-spaced channels (0.8 bit/s/Hz spectral efficiency). In [148], PS-QPSK at 40.5 Gb/s tolerated 1.6 dB higher launch power per channel to achieve a BER of $10^{-3}$ in the receiver in 10 × 100 km 50 GHz-spaced WDM transmission. Renaudier et al. reported on a comparison between the formats at 28 Gbaud, in which PS-QPSK achieved 3 dB higher Q factor than PM-QPSK after transmission in a 4800 km link with hybrid Raman-Erbium optical repeaters [155], and Nölle et al. demonstrated 10% increase in transmission reach at a BER of $3.8 \times 10^{-3}$ in the receiver, in a WDM system with eight 50 GHz-spaced channels at 112 Gb/s [156]. At OFC 2012, Nölle and co-workers also reported on WDM transmission of 112 Gb/s PS-QPSK with 2.0 bit/s/Hz and 2.5 bit/s/Hz spectral efficiency, and transmission reach of 11000 km and 9000 km of ULAF, respectively, at the assumed FEC limit ($3.8 \times 10^{-3}$) [157].

In paper F we presented the results from an experimental and numerical comparison of 20 Gbaud PS-QPSK and PM-QPSK at the same bit rate and the same symbol rate. We investigated long-haul transmission with three different SSMF span lengths (83, 111 and 136 km), both for a single channel system and a WDM system with nine 50 GHz-
spaced channels. In the WDM experiment we found more than 41% and 21% increase in transmission reach with PS-QPSK for all span lengths at a BER of $10^{-3}$ in the receiver, when comparing the formats at the same symbol rate and the same bit rate, respectively. The simulations indicated a slightly larger performance difference in favor of PS-QPSK, but good overall agreement with the experiment, especially for the longer span lengths. In both the experiment and the numerical simulations, the relative difference in performance of the modulation formats was similar for the three span lengths. These results show the feasibility of long-haul transmission of PS-QPSK even when using standard fiber and very long spans.

In the simulations we also made the first comparison of PS- and PM-QPSK with differential coding. Compared to PM-QPSK at the same data rate, PS-QPSK then exhibits an additional reduction in required OSNR of 0.3 dB at a BER of $10^{-3}$, yielding a further improvement in relative transmission reach.

5.3.2 128-SP-QAM

Coelho and Hanik [158] introduced two modulation formats to optical communications by applying the Ungerboeck set-partitioning scheme [159] to PM-16-QAM. The formats were called set-partitioning 32 PM-16-QAM (32-SP-QAM) and set-partitioning 128 PM-16-QAM (128-SP-QAM), with the names stemming from the number of PM-16-QAM symbols that remain after the set partitioning operations. 128-SP-QAM is an interesting competitor to PM-16-QAM since it has comparable spectral efficiency (7 bits per symbol) and 2.43 dB higher $\gamma$ [160]. In paper G we presented the first detailed comparison of 128-SP-QAM and PM-16-QAM in long-haul transmission scenarios, by using numerical simulations of the nonlinear Schrödinger equation. We found a significant improvement in transmission reach with 128-SP-QAM (more than 40% for all cases studied) and also showed that the format can be differentially encoded and that, even though it is a 4-D constellation, joint phase estimation is not required for $i_x$ and $i_y$.

The constellation and the generation of 128-SP-QAM

128-SP-QAM is obtained by taking the even parity subset or the odd parity subset from PM-16-QAM. Hereafter we assume the subset with even parity (the PM-16-QAM symbols with an even number of ones) is used. 128-SP-QAM belongs to the $D_4$ lattice and has 1.55 dB lower asymptotic power efficiency than PM-QPSK [160]. The symbols represent codewords with eight bits, seven information bits $\{d_1, \ldots, d_7\}$ and a single ”parity bit”, $d_{par}$. The parity bit is obtained by performing XOR operations on the information bits

$$d_{par} = \text{XOR}\{d_1, \ldots, d_7\}. \quad (5.14)$$

An example of a 128-SP-QAM symbol is shown in Fig. 5.7. The data sequence to be transmitted is 1101011 and to fulfill the even parity criterion, the eight bit becomes a one. We thus end up with the sequence 11010111 and consequently launch the 16-QAM symbols with the coordinates $(3, -3)$ and $(-3, 3)$ in the x- and the y-polarization, respectively, as shown in Fig. 5.7(b). A 128-SP-QAM transmitter is shown in Fig. 5.8. The difference compared to PM-16-QAM is that one of the eight information bits in each code word is replaced with $d_{par}$.

Decoding of 128-SP-QAM

To decode 128-SP-QAM with non-differential coding, algorithm 2 from [161] is used. With words, it can be described as follows
Figure 5.7: Example of a 128-SP-QAM symbol. (a) Gray-coded 16-QAM. (b) We want to transmit the bit sequence 1101011. Due to the odd number of ones the parity bit is a one as well, and the 128-SP-QAM symbol is created by sending the 16-QAM symbols that represent 1101 (in x-pol.) and 0111 (in y-pol.).

1. Find two 16-QAM symbols independently by using the signals in the two polarizations. Decode the Gray-coded 16-QAM symbols to obtain eight bits.

2. Manipulate the detected 4-D signal by moving it over the closest decision threshold. (This will invert the most uncertain bit.) Decode the 4-D signal as above.

3. Check the parity of the two bit sequences (which will be different). Keep only the one with even parity. Discard the parity bit.

As shown in paper G, the algorithm can also be used to decode 128-SP-QAM with differential coding, if the following modification is made: The symbol corresponding to the even parity bit sequence should be saved and used in the decoding of the following symbol, to avoid an error in the next differential transition. It was also shown in paper G that the DSP algorithms for PM-16-QAM can be used for 128-SP-QAM, with similar performance. The required receiver hardware is also the same for the two formats.

Figure 5.8: Transmitter for 128-SP-QAM.

\[ d_{\text{Par}} = \text{xor}\{d_1, \ldots, d_7\} \]

\[ d_{\text{Par}} = \text{xor}\{d_1, \ldots, d_7\} \]
5.3.3 Subset-optimized PM-NPSK

PS-QPSK has shown the potential of optimizing constellations in four dimensions for optimal uncoded performance. The format is however less spectrally efficient than PM-QPSK, and one may ask whether there are low complexity modulation formats with higher power efficiency than PM-QPSK, and similar or higher spectral efficiency. The 16-point analogy of PS-QPSK, obtained by combining polarization-switching with 8PSK, is less power-efficient, with $\gamma = -2.3$ dB [145]. Since PS-QPSK is obtained from Gray-coded PM-QPSK by choosing the even or the odd parity subset, it can be suspected that combining the two subsets with a relative amplitude scaling is worth trying out. In paper H we reported on a new modulation format, subset-optimized PM-QPSK (SO-PM-QPSK), which is obtained by such amplitude ratio optimization. The constellation of SO-PM-QPSK is defined as

$$C_{SO\text{-}PM\text{-}QPSK} = \left[ \pm \{(1, 1, 1, 1), (1, 1, -1, -1), (1, -1, 1, -1), (1, -1, -1, 1)\}\right]$$

and has a squared minimum distance of

$$d_{\text{min}}^2 = \min \{8, 8A_r^2, (1 + A_r)^2 + 3(1 - A_r)^2\},$$

where $A_r$ is the amplitude ratio between the two PM-QPSK subsets. Inserting (5.16) in (5.4) together with $E_b = 2(A_r^2 + 1)/4$, we find $\gamma = 0.44$ dB for $A_r = \varphi$ and $1/\varphi$, where $\varphi = (\sqrt{5} + 1)/2 \approx 1.618$ is the golden ratio. The 4-D 16-level constellations with $\gamma = 1.11$ dB and $\gamma = 0.9$ dB are difficult to generate and decode [162, 163]. SO-PM-QPSK can be generated by placing an amplitude modulator before a PM-QPSK transmitter, and is therefore the first published practical 4-D 16-level constellation more power-efficient than PM-QPSK.

Similar optimizations for other Gray-coded PM-NPSK formats (with $N$ a power of 2 and $N \geq 4$) yields SO-PM-NPSK. To obtain a general expression for the squared minimum distance of these formats, we consider two neighboring symbols in NPSK. One of these may, for simplicity, be located on the real axis with coordinates $(b, 0)$ in the x-y-plane, where $b = \sqrt{2}$. It has two nearest neighbors, and we consider the one in the counter clockwise direction, defined by $(c = b\cos \phi, d = b\sin \phi)$, where $\phi = 2\pi/N$ is the angle between neighboring NPSK symbols. By considering the two cases (i) the minimum distance within each subset, and (ii) the minimum distance between symbols belonging to different subsets, we obtain a squared minimum distance of

$$d_{\text{min}}^2 = \min \{X, A_r^2X, (b - bA_r)^2 + (b - cA_r)^2 + (dA_r)^2\},$$

where $X = 2(b^2 + d^2)$. For $N = 8$ (PM-8PSK), $A_r = \sqrt{2}$ or $1/\sqrt{2}$ improves $\gamma$ with 1.25 dB. There is however another 64-level format with similar complexity, star-8-QAM, exhibiting an additional 0.35 dB asymptotic gain over PM-8PSK [83, p. 197].

Decoding

The decoding of SO-PM-NPSK shows similarities to the decoding of partitioning constellations [161]. The following maximum-likelihood decoding algorithm for SO-PM-QPSK was used in paper H.

1. The received symbol $r$ is first decoded as for PM-QPSK. We denote the resulting vector as $c_1$. 

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2. The most uncertain bit in $c_1$ is inverted to create the vector $c_2$, which will have different parity than $c_1$.

3. In the final step, we find $\min\{\|r - c_1\|, \|r - c_2\|\}$, which gives the decoded codeword.

**Performance at low SNR**

Fig. 5.9 shows the BER and the SER for SO-PM-QPSK and SO-PM-8PSK for $A_r = 1$ (corresponding to PM-QPSK and PM-8PSK) and for the amplitude ratios that maximize $\gamma$ ($\varphi$ and $\sqrt{2}$, respectively). The SER and the BER of SO-PM-8PSK are calculated with (5.6) and (5.7), and the expressions from paper H are used for SO-PM-QPSK. Union bound approximations of the BER and the SER for PM-8-QAM are included for comparison, as is the exact BER for PM-QPSK (same as for QPSK [83, p. 192]) which as expected shows very good agreement with the union bounded BER for $\text{BER} \leq 10^{-3}$.

SO-PM-QPSK with $A_r = \varphi$ performs better than conventional PM-QPSK for $\text{BER} \leq 2.5 \times 10^{-5}$. Using a smaller $A_r$ improves the performance of SO-PM-QPSK at higher BER, but below $E_b/N_0 \approx 5 \text{ dB}$, the gain from the amplitude scaling is negligible and the optimal constellation converges to PM-QPSK. The BER curves of SO-PM-8PSK and PM-8-QAM follow each other closely within the investigated SNR range, which also applies to their SER curves and the SER for PM-8PSK. On the other hand, the BER performance of PM-8PSK catches up with the other formats at low SNR, due to the fewer neighboring symbols at short distance in PM-8PSK. The gain in asymptotic power efficiency of 4-D constellations comes at the cost of more symbols being close to each other, which has a negative impact especially at low SNR.
5.3.4 Four-dimensional modulation formats and FEC coding

While 4-D formats have very good sensitivity in the signal space naturally provided by the carrier wave, FEC coding is used in many systems. Krongold et al. [164] argues that from a coding and information theory point of view, the way to achieve a certain power efficiency is to apply an appropriate FEC code and that uncoded modulation is not a sufficient indicator of the performance. The signal space dimensionality becomes much larger with coding, implying that the performance needs to be evaluated with FEC included. While this is a valid point, there are still many reasons to investigate uncoded modulation.

1. Most codes are block codes and to decode them correctly it has to be known where the block starts and ends. This is normally handled by using a specific bit pattern, which is uncoded by definition. If this bit pattern is not received correctly, the FEC decoder will fail.

2. In the case of data-aided transmission the updating of adaptive equalizers is performed with detected, uncoded symbols.

3. Uncoded modulation is of great importance if low latency is required, which may be the case in, e.g., control systems, cloud computing, video conferencing/telephony, and the transfer of stock market data. The demand for ultra low latency networking is growing rapidly, and in particular financial applications can be sensitive to microsecond delays [165]. Some vendors have options for transponders without FEC, capable of operating at much lower latencies [166]. The latency due to FEC is in Ref. [167] estimated to be in the range 15–150 µs, and disabling FEC is pointed out as a strategy to achieve high speed data transfer.

In addition, while [164] indeed compares PM- and PS-QPSK with equal rates and bandwidths, the FEC codes for the two formats have different complexity, and so, the comparison is not fair in that aspect. Nevertheless, investigating the performance with FEC included is an important next step in evaluating 4-D constellations.
5.4 Modulation format comparison

This section provides an overview of the performance of several interesting modulation formats. The power efficiencies and the spectral efficiencies are presented for different modulation format families. For PS-QPSK, PM-QPSK, 128-SP-QAM, and PM-16-QAM, we also show numerical results for the BER as a function of $E_b/N_0$ and the tolerance to laser phase noise.

Asymptotic power efficiency and required SNR

For PSK- and QAM-formats with $N$ symbols per polarization, the asymptotic power efficiencies are [162]

$$\gamma_{N-PSK} = \sin^2(\pi/N) \log_2(N), \quad (5.18)$$

and

$$\gamma_{N-QAM} = \frac{3 \log_2(N)}{2(N-1)}, \quad (5.19)$$

and are unaffected by polarization multiplexing. For set partitioning formats, the power efficiency relations depend on the spectral efficiency (in bits per symbol) [160]

$$\gamma_{SP-QAM} = \frac{3SE}{2(\text{SE} + 1)/2 - 1}, \quad \text{if } \frac{SE + 1}{2} \text{ is even,} \quad (5.20)$$

$$\gamma_{SP-QAM} = \frac{3SE}{2(\text{SE} + 1)/2 - 1/2}, \quad \text{if } \frac{SE + 1}{2} \text{ is odd.} \quad (5.21)$$

By using pulse-position modulation (PPM), spectral efficiency can be traded for increased power efficiency. The spectral efficiency and the power efficiency of PM- and PS-QPSK with $K$-fold PPM are shown in Table 5.1, which is inspired by [147, Table 1]. PPM may be useful for applications where spectral efficiency is not a top priority, such as space communications and unrepeated fiber transmission [168]. As seen in Table 5.1, there are two PPM-implementations for PM-QPSK. (i) The PPM time slots are chosen independently in the polarization components, or (ii) the same time slot is used for each PM-QPSK symbol. The former approach is more spectrally-efficient, since it doubles the number of PPM-encoded bits.

Fig. 5.10 shows $\gamma$ as a function of the spectral efficiency for some modulation format families using two polarization components. Apart from $M$-SP-QAM, PM-NPSK, SO-PM-NPSK, and PM-N-QAM, two- (OOK), four-, and eight-level intensity modulation are included, as are PM- and PS-QPSK with $K$-PPM. It is apparent that multilevel intensity modulation yields low power efficiency. On the other hand, a very high $\gamma$ can be achieved by combining PPM with PS- or PM-QPSK, but the spectral efficiency approaches zero for large

<table>
<thead>
<tr>
<th>Format</th>
<th>Dimensions</th>
<th>levels</th>
<th>SE [bits/symbol]</th>
<th>$\gamma$ [dB]</th>
</tr>
</thead>
<tbody>
<tr>
<td>$K$-PPM-PM-QPSK$_1$</td>
<td>$(2K)^2$</td>
<td>$(4K)^2$</td>
<td>$2\log_2(4K)/(K)$</td>
<td>$\log_2(4K)/2$</td>
</tr>
<tr>
<td>$K$-PPM-PM-QPSK$_2$</td>
<td>$4K$</td>
<td>$16K$</td>
<td>$\log_2(16K)/(K)$</td>
<td>$\log_2(16K)/4$</td>
</tr>
<tr>
<td>$K$-PPM-PS-QPSK</td>
<td>$4K$</td>
<td>$8K$</td>
<td>$\log_2(8K)/(K)$</td>
<td>$\log_2(8K)/2$</td>
</tr>
</tbody>
</table>

Table 5.1: Spectral efficiency and $\gamma$ for PM- and PS-QPSK with $K$-PPM.
SO-PM-NPSK has a power efficiency advantage over PM-NPSK, starting at 0.44 dB for 16 levels and increasing with $N$. The gain of SO-PM-16PSK with respect to PM-16PSK is 2.0 dB, but this is mainly of theoretical interest since the power efficiency still is quite low and, e.g., 2.2 dB worse than for PM-16-QAM. Comparing the curves for SP-QAM and PM-N-QAM at a given spectral efficiency, we note a difference of approximately 1.5 dB for large constellations. This is due to the different lattice densities of $\mathbb{Z}_4$ and $\mathbb{D}_4$, as pointed out in section 5.1. We also conclude that a 16-level constellation in $D_4$ with higher spectral efficiency than PM-QPSK and SO-PM-QPSK is obtainable. By removing 16 points from 32-SP-QAM and shifting the center of the obtained 16-level constellation to minimize its energy (by subtracting 0.1875 (1, 1, 1, 1) from each symbol), we find a power efficiency gain of 0.76 dB over PM-QPSK. The constellation is however rather complex, with four amplitude levels. Excluding the shift of the center point yields $\gamma = 0.58$ dB, and the number of different amplitudes is reduced to three. In this case, the 16-level constellation can be generated with four-level electrical signals.

The spectral efficiency and the asymptotic power efficiency of some interesting modulation formats are shown in Table 5.2. 32-SP-QAM has the same sensitivity as PM-BPSK and PM-QPSK and higher spectral efficiency. On the other hand, the format is more complicated to generate, since the most convenient approach is to use four-level driving signals. 2048-SP-QAM consists of the even (or odd) parity symbols of PM-64-QAM and can thus be implemented with comparable complexity. The format is, with more than 2.7 dB gain in power efficiency and 11 bits per symbol, an interesting alternative for systems targeting high spectral efficiency. Among the formats listed in Table 5.2, three are more power-efficient than PM-QPSK and these are all 4-D. Two of them, SO-PM-QPSK and POL-QAM, have equal or higher spectral efficiency than PM-QPSK.
<table>
<thead>
<tr>
<th>Format</th>
<th>No. of symbols $M$</th>
<th>$SE$ [bits/symbol]</th>
<th>$\gamma$ [dB]</th>
</tr>
</thead>
<tbody>
<tr>
<td>PM-BPSK</td>
<td>4</td>
<td>2</td>
<td>0</td>
</tr>
<tr>
<td>PS-QPSK</td>
<td>8</td>
<td>3</td>
<td>1.76</td>
</tr>
<tr>
<td>PM-QPSK</td>
<td>16</td>
<td>4</td>
<td>0</td>
</tr>
<tr>
<td>SO-PM-QPSK</td>
<td>16</td>
<td>4</td>
<td>0.44</td>
</tr>
<tr>
<td>POL-QAM</td>
<td>24</td>
<td>4.5</td>
<td>0.51</td>
</tr>
<tr>
<td>32-SP-QAM</td>
<td>32</td>
<td>5</td>
<td>0</td>
</tr>
<tr>
<td>SO-PM-8PSK</td>
<td>64</td>
<td>6</td>
<td>-2.33</td>
</tr>
<tr>
<td>PM-8-QAM</td>
<td>64</td>
<td>6</td>
<td>-1.98</td>
</tr>
<tr>
<td>128-SP-QAM</td>
<td>128</td>
<td>7</td>
<td>-1.55</td>
</tr>
<tr>
<td>PM-16-QAM</td>
<td>256</td>
<td>8</td>
<td>-3.98</td>
</tr>
<tr>
<td>512-SP-QAM</td>
<td>512</td>
<td>9</td>
<td>-3.68</td>
</tr>
<tr>
<td>2048-SP-QAM</td>
<td>2048</td>
<td>11</td>
<td>-5.82</td>
</tr>
<tr>
<td>PM-64-QAM</td>
<td>4096</td>
<td>12</td>
<td>-8.45</td>
</tr>
</tbody>
</table>

Table 5.2: Spectral efficiency and asymptotic power efficiency of some modulation formats relevant for optical communications.

Fig. 5.11 shows Monte Carlo simulation results for the BER as a function of $E_b/N_0$ for PS-QPSK, PM-QPSK, 128-SP-QAM and PM-16-QAM with both differential and non-differential coding. AWGN noise is the only impairment. The difference in required SNR between the conventional formats and the 4-D constellations decreases for higher BER, since PM-QPSK and PM-16-QAM are Gray-coded, while every symbol in PS-QPSK has six neighbors at the minimum distance and eight symbols in 128-SP-QAM have 24 neighbors.

Figure 5.11: BER vs. $E_b/N_0$ for PS-QPSK, PM-QPSK, 128-SP-QAM and PM-16-QAM with differential and non-differential coding.
Figure 5.12: Required OSNR for a BER of $10^{-3}$ for 112 Gb/s PS-QPSK, PM-QPSK, 128-SP-QAM and PM-16-QAM as a function of $\Delta \nu T$.

at the minimum distance. Since future systems may use FEC tolerating a higher raw BER before decoding, e.g., $10^{-2}$, it is interesting to compare modulation formats also at BERs in this region.

**Tolerance to laser phase noise**

Fig. 5.12 shows Monte Carlo simulations of the OSNR required to achieve a BER of $10^{-3}$ for PS-QPSK, PM-QPSK, 128-SP-QAM, and PM-16-QAM at 112 Gb/s, as a function of $\Delta \nu T$. Differential coding is used to avoid error bursts due to cycle slips and the performance is shown for the optimal block length for each combination of OSNR and $\Delta \nu T$. Shorter block lengths induce larger static penalties due to ASE noise, but may be preferable in case of rapid phase fluctuation.

For PS-QPSK, block lengths in the range 8–64 give less than 0.5 dB OSNR penalty for $\Delta \nu T$ of $2.0 \times 10^{-4}$ to $7.5 \times 10^{-4}$. A symbol-by-symbol estimator gives a static OSNR penalty of approximately 1 dB, but the best tracking performance. However, due to the high cycle slip probability, it gives catastrophic performance in case of non-differential coding. For PM-QPSK, block lengths in the same range (8–64) give less than 0.5 dB OSNR penalty for $\Delta \nu T$ in the range $8.5 \times 10^{-5}$ to $2.4 \times 10^{-4}$. In these simulations, phase estimation was performed separately for the two PM-QPSK tributaries. Slightly better tracking is achieved with joint estimation, for which less than 0.5 dB penalty is given (for the same block lengths) for $\Delta \nu T$ in the range $1.1 \times 10^{-4}$ to $4.9 \times 10^{-4}$.

128-SP-QAM and PM-16-QAM are, due to their denser constellations, more sensitive to phase noise. It was however shown in paper G that block lengths of 16 and 32 give good phase noise tracking over a wide range of phase noise variances for both formats. A 0.5 dB OSNR penalty threshold compared to the low phase noise regime was found at $\Delta \nu T = 10^{-4}$. 

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6 Future directions

Possible future research directions for increasing the data throughput and the performance of fiber-optic communication systems are discussed in this chapter.

Pilot tones

Pilot tones may be of interest for future systems especially for phase synchronization purposes, as has been demonstrated by Nakazawa’s group in several experiments \([103–105, 127, 169]\). Systems with pilot tones introduce more complexity in the optical domain in order to reduce the DSP complexity after detection. Presently, optical components tend to be expensive for large product volumes, while development of ASICs is expensive for small volumes but cost-effective if there are many customers. However, optical components may be cheaper in the future and the advances of silicon photonics could generate new possibilities.

Mode-multiplexing

Recently an interest has emerged in using more than a single mode in long-haul transmission. In contrast to short range transmission over conventional MMF, only a few modes are used and the novel technique is called \textit{mode-multiplexing}. The optical fiber in the upcoming systems is known as \textit{few-mode fiber} (FMF), since the number of supported modes exceeds one but is typically only two or three. Few-mode transmission is currently a hot research topic and a lot of progress is made. In \([170]\), \(3 \times 112\) Gb/s PM-QPSK transmission over 33 km of FMF was achieved, and \([170]\) reported on two-mode transmission of \(2 \times 100\) Gb/s PM-QPSK signals over 40 km of FMF. The key components in these experiments were a few-mode MUX/DEMUX, FMF with low mode-crosstalk, and a polarization- and mode-diversity coherent receiver.

The challenges of FMF transmission concern both the DSP- and the hardware complexity. To compensate for mode crosstalk, a multiple-input-multiple-output equalizer is required and the different modes need simultaneous detection and processing \([170,171]\), which requires a lot of computational efforts and one coherent receiver per mode. In particular, it is difficult to compensate for inter-mode DGD. However, with \(3 \times 80\)-Gb/s PM-QPSK a record transmission distance for FMF transmission of 1200 km was demonstrated recently, by concatenating multiple fiber segments with DGD of opposite sign \([172]\). In addition, progress is made on the FMF and fibers with lower DGD and lower mode coupling are being developed \([173]\).
**Digital backpropagation**

Digital backpropagation (DBP) has been proposed for joint compensation of linear and nonlinear impairments after coherent detection [174–176]. The principle of DBP is to solve the inverse of the NLSE through the fiber to estimate the transmitted signal. While DBP has enabled longer transmission reach by making it possible to increase the power launched into the fiber spans, it exhibits drawbacks such as large computational complexity and reduced efficiency in the presence of PMD. In addition, to compensate for interchannel nonlinear effects, DBP has to be applied jointly on the WDM channels which interacted strongly in the link. This will most likely never happen, as it seems too complex for real-time implementation at high data rates. On the other hand, with increasing data rates per channel, the balance between intra- and interchannel nonlinear effects is shifted, so that the former has stronger impact [65]. In this case, DBP applied on each channel separately may be able to compensate for most distortion.

**Superchannels with frequency combs**

To maximize the spectral efficiency in WDM transmission, high frequency stability of the channels is required to avoid linear crosstalk. Using a frequency comb and techniques such as duobinary shaping [177], or Nyqvist WDM [178, 179] could be a viable approach to achieve high spectral efficiency. Such a scheme was recently demonstrated in [180], where a $11 \times 112$ Gb/s PM-QPSK superchannel was generated and transmitted over a fiber link. The drawback with superchannels is that the component cost does not scale well with the data rate, since the subcarriers require separate transmitters and receivers.

**Modulation formats and coding**

It is of both practical and fundamental interest to find and investigate novel modulation formats with better performance than the traditional. While many of the modulation formats in use today have been known for a long time, recent work has shown it is still possible to find new constellations with better power efficiency. PS-QPSK and SO-PM-QPSK are two examples for coherent systems using a four-dimensional signal space. For systems with intensity modulation, Karout et al. have suggested new formats with improved performance compared to traditional ones [181, 182]. Many of the modulation formats in Fig. 5.10 have not been generated experimentally yet, and some of them could be worth investigating. As a final remark, it would be interesting to also consider the performance of newly proposed formats when FEC is applied.
7 Summary of papers

The thesis includes eight appended papers, which will be outlined below.

**Paper A**

A theoretical and experimental study of the OSNR requirement for self-homodyne coherent systems is performed. It is shown that self-homodyne systems have a 6 dB OSNR penalty compared to intradyne systems for a unity power ratio between the signal and the pilot tone. This is due to including the pilot tone in the OSNR measurement, and that it is affected by amplified spontaneous emission noise. However, narrow band-pass filtering of the pilot tone and optimization of its power yield the same performance limit as for intradyne detection.

**My contributions:** I came up with the idea of narrowband pilot tone filtering, helped develop the theory, set up and performed the experiment, and contributed in writing the letter.

**Paper B**

In this letter we investigate the possibility to cancel nonlinear phase distortion in self-homodyne transmission systems. Numerical simulations are used to quantify the impact of dispersion and noise and the concept is also validated in an experiment for 16-QAM signals at both 5 Gbaud and 10 Gbaud.

**My contributions:** I came up with the idea of cancellation of nonlinear phase distortion, set up and performed the experiment, and contributed in writing the letter.

**Paper C**

Polarization multiplexing cannot be used in self-homodyne transmission due to the pilot tones, which reduces the potential spectral efficiency with a factor of two compared to WDM systems using intradyne detection. In this paper we investigate a scheme to improve the spectral efficiency of self-homodyne systems. We evaluate it with an experiment and with numerical simulations, and measure a 33% spectral efficiency increase compared to transmitting data in a single polarization. The simulation results show that this number can be increased with better optical filtering.

**My contributions:** I came up with the idea of interleaved polarization division multiplexing together with Erik Agrell, I set up and performed the experiment, wrote the code for and performed the simulations, and wrote the paper.
Paper D

In this paper we present an algorithm for polarization demultiplexing and equalization of PS-QPSK, which were developed since the CMA cannot be used for this format. The algorithm is evaluated with numerical simulations, showing that the SNR penalty and the convergence time are similar to when the CMA is used for PM-QPSK.

My contributions: I participated in finding the cost function and derived the updating rules for the filter taps.

Paper E

This paper presents the first generation of PS-QPSK. We proposed a transmitter with an IQ-modulator cascaded with two parallel amplitude modulators performing as a polarization switch. PS-QPSK at 30 Gb/s was found to require 0.7 dB less OSNR than PM-QPSK at the same data rate. In addition, DCF compensation of dispersion resulted in a smaller OSNR penalty in the receiver than DSP compensation after coherent detection.

My contributions: I invented the PS-QPSK transmitter, set up and performed the experiment, wrote most of the DSP code, wrote the code for and performed the simulations, and wrote the paper.

Paper F

PS- and PM-QPSK are compared in an experiment and with numerical simulations for both single channel transmission and in a WDM system with nine 50 GHz-spaced channels. We use spans with 83, 111, and 136 km of standard single-mode fiber, and in all scenarios we measure an increase in reach with PS-QPSK over PM-QPSK of more than 41% at the same symbol rate and 21% at the same bit rate. The simulations show good agreement with the experimental findings, but the transmission reach is longer due to the absence of non-ideal effects and higher back-to-back sensitivity. We also propose differential coding for PS-QPSK and compares with PM-QPSK with differential coding. The power efficiency advantage of PS-QPSK increases with approximately 0.3 dB at a bit error rate of $10^{-3}$.

My contributions: I set up and performed the experiment (got assistance in setting up the loop), wrote most of the DSP code, wrote the code for and performed the numerical simulations, and wrote the paper.

Paper G

In this paper we use numerical simulations to compare 128-SP-QAM with PM-16-QAM for single channel and WDM transmission, at data rates of 112 Gb/s and 224 Gb/s. Differential coding is implemented for 128-SP-QAM, which avoids error bursts due to cycle slips in the phase estimation. We report on more than 40% increase in reach with 128-SP-QAM in all investigated transmission scenarios and find the formats to have similar tolerance to laser phase noise.

My contributions: I found the scheme for differential coding of 128-SP-QAM, wrote the code for and performed the numerical simulations (but got the code for the DD-LMS equalizer from Jianqiang Li), and wrote the paper.
Paper H

In this letter we present subset-optimized PM-QPSK, which is obtained by using different amplitudes for PM-QPSK symbols with even and odd parity. An amplitude ratio equal to the golden ratio yields 0.44 dB asymptotic SNR improvement over PM-QPSK. We also show that such optimization results in performance gain also for PM-NPSK constellations of higher-order.

My contributions: I invented the new modulation format. I implemented and performed the Monte Carlo simulations, collected all the data, and wrote the letter.
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Papers A–H
Paper A

OSNR Requirements for Self-Homodyne Coherent Systems

M. Sjödin, P. Johansson, Z. Tong, M. Karlsson and P. A. Andrekson,

Paper B

Cancellation of Nonlinear Phase Distortion in Self-Homodyne Coherent Systems

P. Johannisson, M. Sjödin, M. Karlsson, E. Tipsuwannakul, and P. A. Andrekson,

Paper C

Filter Optimization for Self-Homodyne Coherent WDM Systems Using Interleaved Polarization Division Multiplexing

M. Sjödin, E. Agrell, P. Johannisson, G-W. Lu, M. Karlsson and P. A. Andrekson,

Paper D

Modified constant modulus algorithm for polarization-switched QPSK

P. Johannisson, M. Sjödin, M. Karlsson, H. Wymeersch, E. Agrell, and P. A. Andrekson,

Paper E

Comparison of polarization-switched QPSK and polarization-multiplexed QPSK at 30 Gbit/s

M. Sjödin, P. Johannisson, H. Wymeersch, P. A. Andrekson, and M. Karlsson,

Paper F

Transmission of PM-QPSK and PS-QPSK with different fiber span lengths

M. Sjödin, B. Puttnam, P. Johannisson, S. Shinada, N. Wada, P. A. Andrekson, and M. Karlsson,

Paper G

Comparison of 128-SP-QAM with PM-16-QAM

M. Sjödin, P. Johannisson, J. Li, P. A. Andrekson, E. Agrell, and M. Karlsson,

Paper H

Subset-Optimized Polarization-Multiplexed QPSK for Fiber-Optic Communication Systems

M. Sjödin, E. Agrell, and M. Karlsson,

Submitted to IEEE Communications Letters, April 2012