

Photonic and Signal Processing Technologies for High Capacity Short Reach Optical Interconnects

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List of Abbreviations

ADC	Analogue-to-Digital Converter
AIR	Achievable Information Rate
APD	Avalanche Photodiode
ASE	Amplified Spontaneous Emission
AWG	Arbitrary Waveform Generator, Arrayed Waveguide Grating
AWGN	Additive White Gaussian Noise
BER	Bit Error Ratio
BPD	Balanced Photodiode
B2B	Back-to-Back
CD	Chromatic Dispersion
CH	Carrier Heating
CMA	Constant Modulus Algorithm
CW	Continuous Wave, Continuous Wavelength
DAC	Digital-to-Analogue Converter
DBP	Digital Backpropagation
DBR	Distributed Bragg Reflector
DD	Decision-Directed
DFB	Distributed Feedback
DLM	Dual Laser Module
DML	Directly Modulated Laser
DP	Dual Polarisation
DSP	Digital Signal Processing
DWDM	Dense Wavelength Division Multiplexing
EAM	Electro-Absorption Modulator
ECL	External Cavity Laser
EDFA	Erbium-Doped Fibre Amplifier
EGN	Enhanced Gaussian-Noise
EML	Externally Modulated Laser
FEC	Forward Error Correction
FFE	Feed-Forward Equaliser
FFT	Fast-Fourier Transform
FlexPAM	Flexible Pulse Amplitude Modulation
FP	Fabry-Perot
FSR	Free Spectral Range
FWHM	Full-Width Half-Maximum
FWM	Four-Wave Mixing
GMI	Generalised Mutual Information
GS	Geometrical Shaping
HD	Hard Decision

HPGS	Hybrid Probabilistic and Geometrical Shaping
IMDD	Intensity Modulation with Direct Detection
IQ-Tx	IQ Transmitter
ISI	Inter-Symbol Interference
ITL	Integrated Tunable Laser
LEAF	Large Effective Area Fibre
LMS	Least-Mean Squares
LO	Local Oscillator
MB	Maxwell-Boltzmann
MG	Gain Section of Master Laser
MI	Mutual Information
MIMO	Multiple-Input Multiple-Output
ML	Master Laser
MR	Reflector Section of Master Laser
MRR	Micro-Ring Resonator
MZM	Mach-Zehnder Modulator
NCG	Net Coding Gain
NGMI	Normalised Generalised Mutual Information
NLE	Nonlinear Equaliser
NLIN	Nonlinear Interference Noise
NLSE	Nonlinear Schrödinger Equation
OBPF	Optical Band Pass Filter
OLT	Optical Line Terminal
ONU	Optical Network Unit
OOK	On-Off Keying
PAM	Pulse Amplitude Modulation
PBS	Polarisation Beam Splitter
PC	Polarisation Controller
PC-AWGN	Partially Coherent Additive White Gaussian Noise
PD	Photodiode
PIC	Photonic Integrated Circuit
PIN	<i>p-i-n</i> photodiode
PMD	Polarisation Mode Dispersion
PNA	Parametric Noise Amplification
PON	Passive Optical Network
PPG	Pulse Pattern Generator
PR	Polarisation Rotator
PS	Probabilistic Shaping
P2MP	Point-to-Multipoint
P2P	Point-to-Point
QAM	Quadrature Amplitude Modulation
RC	Raised Cosine
RDA	Radius-Directed Equaliser
RIN	Relative Intensity Noise
RLS	Recursive Least Squares
ROP	Received Optical Power
RRC	Root-Raised Cosine
RTO	Real-Time Oscilloscope

SBS	Stimulated Brillouin Scattering
SD	Soft Decision
SDM	Space Division Multiplexing
SG	Gain Section of Slave Laser
SHB	Spectral Hole Burning
SL	Slave Laser
SMF	Single-Mode Fibre
SMSR	Side Mode Suppression Ratio
SNR	Signal-to-Noise Ratio
SOA	Semiconductor Optical Amplifier
SPM	Self-Phase Modulation
SR	Reflector Section of Slave Laser
SRS	Stimulated Raman Scattering
SSFM	Split-Step Fourier Method
TDM	Time Division Multiplexing
TEC	Thermo-Electric Cooler
TOF	Tunable Optical Filter
TWDM	Time/Wavelength Division Multiplexing
VCSEL	Vertical-Cavity Surface-Emitting Laser
VNLE	Volterra Nonlinear Equaliser
VOA	Variable Optical Attenuator
WDM	Wavelength Division Multiplexing
XGM	Cross-Gain Modulation
XPM	Cross-Phase Modulation

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Abstract

The development of integrated photonic technologies has enabled the miniaturisation and mass production of optical components for fibre-based optical transmission systems. This resulted in an exponential growth in bandwidth availability over the last few decades due to drastic cost reductions and increases in energy efficiency. However, as the world transitions to a carbon-neutral economy, the requirements imposed on the optical network get more and more challenging to meet. The use of intensity modulation with direct detection (IMDD) has historically been the preferred solution for short-reach applications, but coherent technologies have proven in recent years that they can outperform IMDD even after a few tens of km.

Some of the main technologies competing in the short-reach market, ordered in increasing complexity and performance, are directly modulated lasers (DML), externally modulated lasers, and coherent transmission. DMLs provide high power and energy efficiency, but the frequency chirp induced by the modulation makes it challenging to achieve high data rates and transmission distances. External modulation allows for improved performance at the expense of additional insertion losses, while coherent transmission provides great improvements in reach and spectral efficiency at the expense of increased complexity, component cost and energy consumption.

This thesis explores the use of photonic integrated circuits and low-complexity digital signal processing to enable cost-effective and energy-efficient increases in transmission capacity/reach in short-reach optical transmission systems. We study the use of the three technologies mentioned above for different use cases relevant for current and future optical links ranging from intra-data centre point-to-point links to point-to-multipoint systems for optical access networks.

Introduction

The increasing popularity of online services like streaming platforms, cloud storage, cloud computing, or video call services is driving an unprecedented rise in the bandwidth requirements of optical networks, from the short reach intra and inter-data centre links and optical access networks to the long haul transoceanic transmission links. This growth in data usage is expected to continue in the coming years, as more people around the globe get internet access, a bigger number of companies and businesses adopt a remote work model for their employees, and bandwidth-intensive applications such as autonomous vehicles and telemedicine are deployed. Additionally, as the number of internet users grows and the world transitions to a carbon-free energy system, optical transmission systems are required to steadily reduce their power consumption cost-effectively.

In order to increase the capacity of an optical transmission system, the three solutions currently employed are increasing the baud rate, increasing the spectral efficiency by moving to more advanced modulation formats, and employing more spectral channels in a wavelength division multiplexing (WDM) system. Transmission rates and channel performance can also be improved in the digital domain by introducing forward error correction (FEC) and digital signal processing (DSP). FEC consists of adding an overhead of redundant data to the transmitted signal to enable the full recovery of the original message from a noisy/distorted received signal. DSP techniques help mitigate linear and nonlinear impairments, such as bandwidth limitations or chromatic dispersion, via filtering, predistortion, or equalisation among others.

This thesis will focus on the use of integrated photonics components with diverse DSP techniques, tailored for the different applications, to enhance the performance of current and future high-capacity optical communications systems. The studies are performed based on experimental demonstrations and full system simulations accounting for the main impairments coming from different types of transmitters, receivers and transmission channels. The key contributions of the thesis are:

- Successful demonstration of a 30 Gbit/s transmission over 75 km of single-mode fibre using a fully integrated multisection directly modulated laser (DML). The transmission was enabled by the use of optical injection and reduced-complexity nonlinear equalisation.
- Detailed experimental and simulation-based study for flexible pulse amplitude modulation (FlexPAM) formats to allow for fine optimisation of the spectral efficiency as a function of signal-to-noise ratio in intensity modulation with direct detection (IMDD) systems.
- Demonstration of the potential capabilities of probabilistic shaping for mitigating the nonlinear impairments introduced by semiconductor optical amplifiers (SOA) in IMDD systems.
- Experimental demonstration of high bit rate coherent transmission using an integrated high-speed tuneable laser in the C-band. Switching times in the nanosecond ranges were obtained for 90 GBaud operation and high-order modulation formats.
- Development and simulation-based demonstration of an adaptive nonlinear equaliser for SOA-amplified short-reach transmission systems based on coherent technology.

The structure of this PhD thesis is as follows:

In Chapter 1, an introduction to diverse topics regarding optical communications systems is presented. Firstly, a review of the capacity limits in optical transmission

systems is given, with emphasis on current and future ways to increase transmission capacity. Secondly, optical sources and photodetectors are introduced. Then, an overview of the optical transmission systems' general structure is given, and the different requirements of the optical systems depending on the application are explained. The next two subsections explain in detail the IMDD and coherent formats and technologies. In Chapter 2, the basic coding and DSP techniques used for optical transmission systems are reviewed. First, an introduction to FEC is given. Second, DSP for IMDD systems is explained. Then, DSP for coherent is given. Finally, constellation shaping techniques are explained as a way of increasing the system's achievable information rates. Chapter 3 shows the characterisation of an integrated multi-section DML and its transmission performance at 30 Gbit/s. A fibre transmission simulation using a DML with and without optical injection was performed to characterise the influence of the external injection on the performance. Chapter 4 introduces the FlexPAM formats and their look-up table-based symbol-to-bit mapping for short-reach IMDD systems. A widely tunable, ultra-low noise integrated dual laser module was used for experimentally demonstrating the FlexPAM formats. A full-system simulation was executed to study the influence of different system upgrades on the performance. This chapter also includes an experimental study of the advantages of probabilistic shaping in IMDD systems affected by SOA nonlinearity. Finally, Chapter 5 shows the characterisation of a fully integrated tunable laser providing switching times in the order of nanoseconds and demonstrates the use of coherent technology for high-data-rate burst mode operation for short-reach applications. Additionally, an adaptive equaliser capable of mitigating nonlinear distortions induced by an SOA in a coherent signal is introduced and demonstrated on a simulation basis.

Chapter 1

Introduction to Optical Transmission Systems

Optical transmission systems use light to transmit information over a wide range of distances, from a few mm for optical networks on chip up to several thousands of km for submarine links. Due to its characteristics of reasonably low loss and nonlinearity, optical fibre is the main transmission medium currently used. Given the classical model, i.e. Maxwell's equations, a light wave of a given central wavelength propagating through a medium (e.g. an optical fibre) can be fully described by the amplitude, the phase and the polarisation state of its electric field. Based on this description of an optical signal, there are three degrees of freedom in which the information can be encoded, but they are commonly grouped as amplitude and phase (or complex amplitude) for two linear, independent, orthogonal polarisations (X and Y).

This chapter provides an overview of fibre-based optical transmission systems. In section 1.1, the fibre channel is introduced and the capacity limits on a communications system are explained. Section 1.2 describes some common optical sources and techniques for their characterisation. Section 1.3 gives the basics of end-to-end optical transmission systems, and subsections 1.3.1 and 1.3.2 explain in more detail the modulation formats and use cases for intensity modulation with direct detection

(IMDD) and coherent systems respectively.

1.1 Capacity Limits in Fibre Optics Systems

The main limitations of a transmission system are usually noise and bandwidth, as determined by the well-known Shannon limit, i.e. the maximum transmission capacity (C) of a system with a given bandwidth (BW) and signal-to-noise ratio (SNR) is given by:

$$C = BW \times \log_2(1 + \text{SNR}) \quad (1.1)$$

The Shannon limit assumes the only impairment on the system is additive white Gaussian noise (AWGN), which is a good approximation to some real-life noise sources such as thermal noise, coming from the random thermal motion of electrons in the medium, or amplified spontaneous emission (ASE), produced in a medium with population inversion due to the amplification via stimulated emission of random photons generated by spontaneous emission. Shot noise is another common source of noise that affects both electrical currents and optical signals, and its origin comes from the discrete nature of photons and electrons.

However, a fibre-based system cannot be modelled as an AWGN channel, and a more realistic description is required. Under some assumptions, the propagation over single-mode fibre (SMF) of a single polarisation optical signal, given the complex amplitude of the envelope of the electric field E , can be accurately modelled by the Nonlinear Schrödinger Equation (NLSE) [1]:

$$\frac{\partial E}{\partial z} = j \frac{\beta_2}{2} \frac{\partial^2 E}{\partial t^2} + \frac{\beta_3}{6} \frac{\partial^3 E}{\partial t^3} - \frac{\alpha}{2} E - j\gamma |E|^2 E \quad (1.2)$$

Where β_2 and β_3 are the group velocity dispersion parameter and the third order dispersion parameter respectively, α is the attenuation constant and γ is the non-linear coefficient. All these parameters are characteristic of the fibre and are not dependent on the input signal. In general, the NLSE cannot be solved analytically

for any arbitrary waveform and transmission distance. However, there are solutions for the special cases of linear regime (the fourth term on the right side in (1.2) is neglected) and of no dispersion regime (β_2 and β_3 terms are neglected). Based on these analytical solutions to the special cases, an iterative solution can be obtained for the full NLSE by applying the Split-Step Fourier Method (SSFM) [2]. When both polarisations are present, the NLSE is replaced by the averaged Manakov equations [3]

$$\frac{\partial E_{X/Y}}{\partial z} = j \frac{\beta_2}{2} \frac{\partial^2 E_{X/Y}}{\partial t^2} + \frac{\beta_3}{6} \frac{\partial^3 E_{X/Y}}{\partial t^3} - \frac{\alpha}{2} E_{X/Y} - \frac{8}{9} j \gamma (|E_X|^2 + |E_Y|^2) E_{X/Y} \quad (1.3)$$

Where E_X and E_Y are the complex amplitude of the envelope electric field for the X and Y polarisations respectively, and an $8/9$ factor multiplying the nonlinear term is introduced to account for the averaging effects of random birefringence throughout the fibre [4].

Given equations (1.2) and (1.3), some transmission impairments can be highlighted:

- Attenuation (α). Despite the low loss offered by optical fibres, some power gets lost during transmission, resulting in a degradation of the SNR of the signal. The main processes causing this attenuation are Rayleigh scattering and infrared absorption [5], and additional losses are obtained in practice due to material impurities and bending losses. Modern fibres have typical attenuation values below 0.2 dB/km at a wavelength of 1550 nm.
- Chromatic dispersion (β_2, β_3): Due to the wavelength-dependence of the group refractive index of non-ideal media, optical signals get distorted when propagated through fibre. Fibre chromatic dispersion is obtained as the addition of the material dispersion and the waveguide dispersion [6]. In general, dispersion results in the signal getting spread in the time domain, which can severely limit the transmission distance. Modern fibres have typical dispersion coefficients of around 16 ps/nm/km at a wavelength of 1550 nm, but more specialised fibres

can have their dispersion tailored to optimise system transmission.

- Nonlinearity (γ): It is produced by Kerr effect in the core of the fibre (i.e. the refractive index of the silica depends on the optical power of the light), and in general, it manifests itself as a combination of self-phase modulation (SPM), cross-phase modulation (XPM) and four-wave mixing (FWM) in an optical transmission system. SPM produces a phase rotation that depends on the optical power of the channel under study (it depends on the transmitted sequence over the channel). XPM produces a phase rotation in the channel under study that depends on the optical power of other channels (transmitted at different wavelengths in a WDM system) in the fibre. FWM produces mixing products at frequencies determined by the channel spacing in a WDM system, some of which fall on top of the channel under study and effectively interact with it as noise. If FWM is strong, the signal power also gets noticeably depleted due to the nonlinear interaction. SPM (intra-channel nonlinearity and is always present in a transmission system), and XPM and FWM (inter-channel nonlinearities only present in multichannel transmission) all originate from the same nonlinear phenomenon and are mathematically described by the equations mentioned above [7]. Modern fibres have typical nonlinear coefficients of around $1.3 \text{ W}^{-1} \text{ km}^{-1}$ at a wavelength of 1550 nm.

While Kerr-induced nonlinearity is commonly studied as additive noise due to the complexity associated with its compensation, all SPM, XPM and FWM are fully deterministic and can, in principle, be ideally corrected using adequate signal processing. However, nonlinear interactions between the signal and the noise fields result in additional noise terms that cannot be corrected. The main nonlinear interaction between signal and noise is called parametric noise amplification (PNA) [8].

Other important phenomena to consider in an optical fibre transmission system not described by equations (1.2) and (1.3) are polarisation mode dispersion (PMD), stimulated Raman scattering (SRS) and stimulated Brillouin scattering (SBS) [9]:

-
- PMD: It is produced by random birefringence variations in non-ideal optical fibres, and causes pulse broadening due to differences in propagation velocity for the different polarisation states. Modern fibres have typical PMD parameters below $0.1 \text{ ps}/\sqrt{\text{km}}$, which makes this effect relevant only in long haul, high baud rate optical transmission systems.
 - SRS: It originates from the interaction between propagating light and the vibrations of silica molecules in the fibre, and depletes shorter wavelength signals while amplifying the longer wavelengths. Its effects on a multichannel signal can be very complex and significantly degrade the performance of all the transmitted channels. SRS sets a limit to the product of total transmitted power times the total occupied optical bandwidth.
 - SBS: It originates from the interaction between propagating light and the vibrations of the fibre, and results in the transmitted signal getting depleted and a back-propagating scattered signal being generated. It is especially strong when low linewidth lasers are employed, and limits the maximum transmitted power per channel. Unlike SRS, it does not depend on the total bandwidth of the system.

Aside from the distortions introduced by the fibre transmission, many common nonlinear impairments in a transmission system come from the components used in the system, such as directly modulated lasers (DML) or semiconductor optical amplifiers (SOA), among others. These devices and their effect in an optical system are explained in more detail in section 1.3.

1.2 Optical Sources

The vast majority of optical transmission systems are based on laser sources, which consist of a resonant cavity with an active medium in which electrons can be pumped into an excited state. Under the right conditions, stimulated emission is produced

and highly coherent light is obtained [10]. This lasing process is always accompanied by the generation of ASE.

Some of the main types of lasers used for communication systems are:

- Fabry-Perot (FP) lasers: The simplest laser type, they are composed of an active region surrounded by two mirrors. Due to the non-wavelength-selective reflection of the mirrors, FP lasers produce a multimode optical output with a free spectral range (FSR) that depends on the refractive index and length of the cavity.
- Distributed Feedback (DFB) lasers: They are composed of an active region built with a grating structure that enables single-mode lasing. The emitting frequency from a DFB laser is thermally tunable.
- Distributed Bragg Reflector (DBR) lasers: They are built with an active region between one or two Bragg gratings that reflect light selectively. They are single-mode and have wide wavelength tunability.
- Vertical-Cavity Surface-Emitting Lasers (VCSEL): They are formed with a very small active region and high reflectivity DBRs. Their main characteristic is the generation of light perpendicular to the surface of the chip.
- External Cavity Lasers (ECL): They consist of an active region attached to an external, usually large cavity. They usually show wide tunability and low levels of phase noise.

In most practical cases, the main source of performance degradation in an optical communication system coming from the laser source is ASE, which affects both the amplitude and the phase of the generated signal.

Amplitude noise is usually defined in terms of the relative intensity noise (RIN), which quantifies the noise power for a given laser average output power. The RIN measurement, explained in [11], depends on the generated photocurrent, as thermal and shot noise cannot be separated from the laser noise. The measurement can

be limited by shot noise and thermal noise at low photocurrents, resulting in an observed RIN higher than the actual value.

Laser phase noise is usually divided into two separate processes [12]. The first one, caused by ASE, is a random walk-like process where each instantaneous phase increment comes from a random Gaussian distribution. It has a flat power spectral density and results in a Lorentzian spectrum. It is commonly described by its 3-dB full-width half-maximum (FWHM) spectral width, known as intrinsic linewidth. The second process, called 1/f noise, has its origin in electronic fluctuations in the driving current on relatively large time scales. Its power spectral density is inversely proportional to the frequency and produces a Gaussian spectrum. The presence of both Lorentzian and 1/f noise generally results in a Voigt spectrum. The linewidth can be measured using a delayed self-heterodyne setup [13], where the laser output is divided in two paths, decorrelated and recombined before detecting with a photodiode. This measurement depends on the record length due to the presence of slow 1/f drifts, and can sometimes give an inaccurate estimate for the phase noise. A more detailed noise characterisation is obtained from the FM noise measurement, where the phase noise can be measured as a function of frequency, allowing the direct measurement of the 1/f and Lorentzian components.

1.3 Optical Transmission Systems

The basic structure of a fibre optical transmission system can be seen in figure 1.1. The system starts with the digital information at the transmitter. The information is converted into the electronic (analogue) domain using a digital-to-analogue converter (DAC), and into the optical domain by modulating the properties of a carrier. The light is then transmitted through the channel, converted into the electronic domain using photodetectors at the receiver, and into the digital domain using an analogue-to-digital converter (ADC).

The transmitter side can consist of a single laser source for single-channel transmission, or multiple lasers for WDM transmission. One or multiple electrical bit

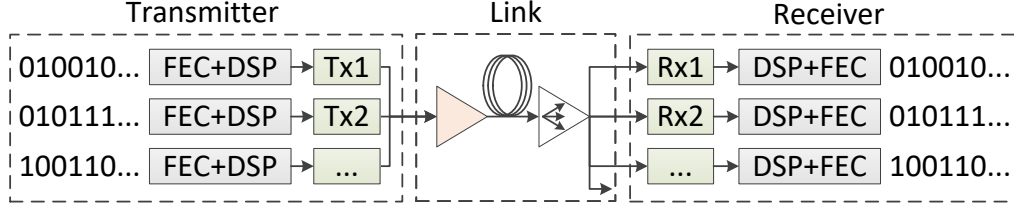


Figure 1.1: General structure of an optical transmission system. FEC: Forward error correction, DSP: Digital signal processing.

streams are fed into a forward error correction (FEC) algorithm and some digital signal processing (DSP) steps are applied (topics developed in Chapter 2). The digital sequences obtained after FEC and DSP are fed into a DAC, whose output can be amplified using electrical amplifiers. The electrical signals can be used to drive a DML, or to drive an external modulator, where a variety of modulation formats can be applied to the light of a laser operating in continuous wave (CW).

In terms of distance, the channel characteristics can range from a few metres of fibre within a data centre, to a few km, to several thousand km of fibre for submarine links. Regarding the number of users, it can connect the transmitter to a single receiver in a point-to-point (P2P) configuration, or connect to multiple users in a point-to-multipoint (P2MP) scenario. Arrayed waveguide gratings and passive splitters can be used for splitting the signal wavelength dependently and independently respectively. An optical link can be unamplified (typical for short-reach applications) or have in-line amplifiers to boost the signal power (e.g. long-haul systems). Two common types of optical amplifiers are the erbium-doped fibre amplifier (EDFA) and the SOA. While EDFAs offer high gain and linear behaviour, they are also bulky and limited to C-band operation. SOAs, on the other hand, can be integrated and can be designed to cover a much wider bandwidth [14], but typically have higher noise figures, lower saturation power, and fast dynamics (with recovery times in the order of 100 ps) that can introduce nonlinear distortions in the signal. Basic modelling of the SOA dynamics can be done with a lumped model, where the output field of the SOA $E_{out}(t)$ is obtained, given the input field $E_{in}(t)$, as shown in equation (1.4) [15]:

$$E_{out} = E_{in} \cdot \exp \left(\frac{h_N(t)}{2} (1 - j\alpha_H) + h_{CH}(t) (1 - j\alpha_{CH}) + h_{SHB}(t) - \frac{\alpha_L L}{2} \right) \quad (1.4)$$

Where $h_N(t)$, $h_{CH}(t)$ and $h_{SHB}(t)$ are the contributions of the carrier density number, carrier heating (CH), and spectral hole burning (SHB) to the integrated gain over the length of the SOA. α_H and α_{CH} are the gain-phase and CH-phase coupling parameters, and $\alpha_L L$ is the total waveguide scattering loss. The temporal evolution of $h_N(t)$, $h_{CH}(t)$ and $h_{SHB}(t)$ depend on $h_N(t)$ and the total input power, and their explicit equations can be found in [15]. The lumped model of an SOA can also be written for two polarisations.

The receiver architecture is usually dependent on the type of transmitter and modulation format selected. Two common types of receiver approaches are direct detection, where the optical power is detected using a photodiode (PD), and coherent detection, where the optical field is transformed to the electrical domain using a full coherent receiver. The electrical signals can be amplified with an electrical amplifier before converting them to the digital domain using an ADC. DSP can be applied to the digital signals before sending them through the FEC algorithm to recover the original binary sequences.

In the following subsections, two types of optical transmission systems will be explained: IMDD and coherent systems.

1.3.1 IMDD Transmission systems

Intensity modulation of light is the simplest way to encode information for an optical transmission system. The operating principle consists of transmitting the data in the optical intensity $|E|^2$ (i.e. the instantaneous power), which allows for the use of simple and inexpensive equipment for signal generation. Moreover, the detection of an intensity-modulated signal can be performed using a direct detection receiver consisting of a single photodiode, reducing cost and complexity also at the receiver

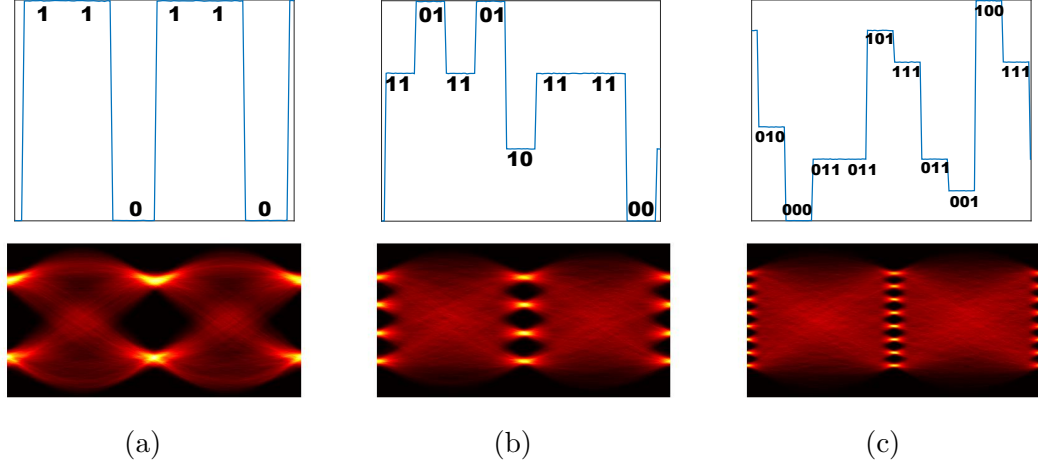


Figure 1.2: Example of time traces (top) and eye diagrams (bottom) for OOK (a), PAM4 (b), and PAM8 (c). The time traces show the symbol to bit allocation of the stream.

side. However, those cost and complexity reductions come at the expense of a significant limitation in spectral efficiency. As the photodetection process erases the phase and polarisation information from the incoming signal, IMDD systems only take advantage of one out of the four degrees of freedom that light possesses. Additionally, chromatic dispersion can impose significant penalties on the performance over fibre transmission, as it cannot be effectively compensated in the digital domain without knowledge of the received optical field, and power fading is produced due to the direct detection process [16].

The most basic IMDD modulation format is called on-off keying (OOK), and consists of encoding binary information by assigning the 1's and 0's to the on and off of a 2-level signal. OOK is robust against noise and nonlinear distortions, but requires the use of high bandwidth components in order to achieve high transmission rates due to its low spectral efficiency. To overcome this problem, M-level pulse amplitude modulation (PAM-M) formats can trade off noise resiliency for spectral efficiency by encoding more than one bit per symbol. The most common PAM formats currently used are PAM4 and PAM8, which encode 2 bit/symbol and 3 bit/symbol respectively. Figure 1.2 shows examples of OOK (a), PAM4 (b), and PAM8 (c) signals and eye diagrams.

In order to encode the information in an IMDD system, the main types of trans-

mitters are based on either direct or external modulation, each one with its own advantages and disadvantages:

- Directly modulated lasers: The operating principle of a DML is based on driving the current with the data signal into the active medium of the laser, producing a change in the optical output capable of encoding the information in the intensity of the light. The main advantages of DML-based transmitters are their high output power and their low cost and complexity. However, the optical signal obtained from a DML is, in general, highly nonlinear due to the complex laser dynamics, which especially affects the higher-order PAM formats. It also suffers from adiabatic and dynamic frequency chirp which can significantly degrade the signal quality after transmission over fibre [17; 18]. While frequency chirp is in general detrimental to fibre transmission, it has been proposed to use the interaction between chirp and chromatic dispersion to increase the achievable transmission distance [19]. To improve the quality of IMDD signals generated with a DML, the bias current into the device can be increased and the modulating current reduced. This results in a more linear response but reduces the extinction ratio and overall energy efficiency of the generation. Alternatively, optical injection locking (injecting a small amount of CW light from a master laser into the slave DML) can help reduce the frequency chirp and improve the overall response of the device [20]. A set of typical rate equations for a DML with optical injection can be written for the carrier density in the cavity N , the photon density S , and the output phase ϕ , as in equations (1.5), (1.6), and (1.7) respectively [21; 22]. The optical gain g is obtained as in equation (1.8), and the definition of the different parameters involved in the equations can be found in table 1.1.

$$\frac{dN}{dt} = \frac{I}{qV} - \frac{N}{\tau_c} - v_g g S \quad (1.5)$$

$$\frac{dS}{dt} = \Gamma v_g g S - \frac{S}{\tau_p} - \Gamma \frac{R_{sp}}{V} + \frac{2}{T} \cos(\phi) \sqrt{S \cdot S_i} \quad (1.6)$$

Table 1.1: Parameters involved in the DML rate equations

Symbol	Definition
N	Carrier density
I	Injected current
q	Elementary charge
V	Volume of the cavity
τ_c	Carrier lifetime
v_g	Group velocity
g	Optical gain
S	Photon density
Γ	Optical confinement factor
τ_p	Photon lifetime
R_{sp}	Spontaneous emission rate
T	Round trip time in the cavity
ϕ	Output phase
S_i	Injected photon density
ω_s	Optical frequency of the slave laser
ω_m	Optical frequency of the master laser
β	Linewidth enhancement factor
a	Differential gain
N_{th}	Threshold carrier density of the slave laser without optical injection
N_0	Transparency carrier density
ε	Nonlinear coefficient

$$\frac{d\phi}{dt} = \omega_s - \omega_m + \beta\Gamma v_g a \frac{N - N_{th}}{2} - \frac{\sin(\phi)\sqrt{S \cdot S_i}}{T} \quad (1.7)$$

$$g = \frac{a(N - N_0)}{1 + \varepsilon S} \quad (1.8)$$

- Externally modulated lasers (EML): EMLs solve the downsides of direct modulation by operating the laser in CW and using an external modulator to encode the information onto the optical carrier. They usually offer higher modulation bandwidth and better performance than DMLs, but they come at a higher cost and complexity, and a reduced energy efficiency. Two of the main types of external modulators are the electro-absorption modulator (EAM) and the Mach-Zehnder Modulator (MZM). EAM operation is based on changing the absorption coefficient of a semiconductor with an electric field. Some of the advantages of EAMs are that they require low driving voltages, can be fabricated with high bandwidths, and can be easily integrated on-chip, while

their main drawbacks are that they can achieve limited extinction ratios, have relatively high insertion loss, and introduce transient chirp in the signal [23]. MZMs work based on interference, as opposed to absorption, and their operating principle consists of splitting the light into two paths and modulating the phase in one (or both) of the arms before recombining. This process allows the tailoring of the amount of constructive/destructive interference of the signals, enabling chirp-free modulation with high extinction ratios, but at the cost of high insertion losses and a larger footprint than that of an EAM.

At the receiver side of an IMDD system, the main approaches for converting the optical signal into the electrical domain consist of using either a *p-i-n* photodiode (PIN) or an avalanche photodiode (APD):

- PIN photodiodes: They are composed of an undoped region *i* between a *p*-doped and an *n*-doped region. The *p-n* junction produces an electric field that accelerates the electrons produced by the incoming light via the photoelectric effect, while the *i* layer provides a longer region for photon absorption [10]. They produce photocurrents proportional to the incoming optical power $|E|^2$, erasing all information contained in the phase or the polarisation of the signal. Typical responsivity values (ratio between the photocurrent and the optical power) for PIN photodiodes are around 1 A/W.
- APD: They are designed so the electron-hole pairs generated by the incoming photons can generate new electron-hole pairs, producing a cascading effect that amplifies the obtained photocurrent [10]. Depending on the semiconductor material used for fabricating the APD, responsivity values as high as 100 A/W can be obtained. However, the avalanche effect also introduces noise and limits the bandwidth of the device.

Due to their low cost, low power consumption and simplicity, as well as their limited reach over fibre, IMDD systems are mostly used in short-reach applications. The main use case is intra data centre links of less than 2 km (Figure 1.3(a)), where

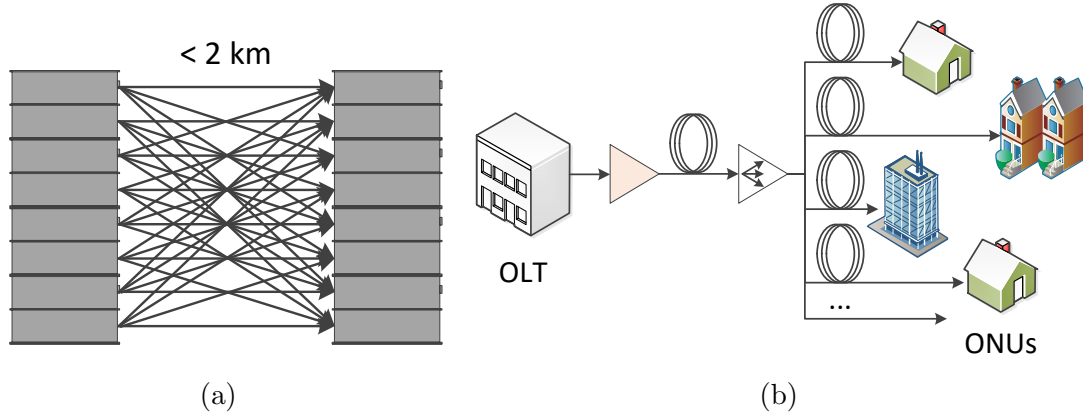


Figure 1.3: (a) Intra data centre network composed of a large number of short-reach P2P links. (b) PON consisting of a P2MP connection between the optical line terminal (OLT) and several optical network units (ONU) via a passive splitter.

thousands of servers are interconnected and huge numbers of transponders need to be deployed, making cost and power consumption the main factors to consider when designing the system. Optical access networks, connecting the optical fibres from the central distribution office to the end users (e.g. homes, buildings), are currently based on passive optical networks (PON) employing IMDD formats in order to minimise the cost and power consumption for the users [24]. To maximise the number of optical network units (ONU) connected to an optical line terminal (OLT), PONs rely on time division multiplexing (TDM) and passive splitting (Figure 1.3(b)). This P2MP configuration imposes big constraints on the required power budget (>30 dB), as the splitting losses increase with the number of users. Additionally, PONs need to operate both from the OLT to the ONUs (downstream) and from the ONUs to the OLT (upstream), and the different path losses for each ONU impose dynamic range requirements of over 10 dB. Those constraints are challenging to meet with IMDD systems as the data rate requirements increase, and coherent technology is expected to be introduced in the PON in the near-to-medium future [25; 26].

1.3.2 Coherent Transmission Systems

Coherent transmission systems use amplitude, phase, and polarisation of the light to encode information. One of the most common modulation formats in coherent sys-

tems is the M-ary quadrature amplitude modulation (M-QAM), which distributes the constellation symbols on a square grid with independent in-phase (I) and quadrature (Q) components. For dual-polarisation transmission, a QAM signal is encoded into each of two orthogonal polarisations X and Y . Typical M-QAM formats use values of M of the form 2^{2n} , for example, 4QAM i.e. quadrature phase shift keying (QPSK), 16QAM, 64QAM, or 256QAM. The constellations corresponding to each of those M-QAM formats are shown in figure 1.4.

At the transmitter side, the modulation is performed with a coherent transmitter (Figure 1.5(a)). The light from a linearly polarised laser source is split into two paths using a 50/50 coupler. Each path is then split into two again, and each of the four beams is sent into an independent MZM for amplitude modulation. The MZMs modulate the light with the four tributaries of the signal corresponding to the I and Q components of the X and Y polarisations, i.e. XI, XQ, YI, YQ. The quadrature components are rotated by $\pi/2$ radians before recombining them with their corresponding in-phase component. The Y -branch is sent through a polarisation rotator (PR), and both polarisations are then combined using a polarisation beam splitter (PBS). The combination of MZMs, couplers and phase/polarisation rotators is usually called an IQ modulator.

At the receiver side, a coherent receiver (Figure 1.5(b)) is used to recover all the information encoded in the optical field. This is done with the help of a strong local oscillator (LO) operating in CW at relatively high power and an emission wavelength close to that of the incoming signal (intradyne detection). The received signal is sent through a PBS, and the Y polarisation is rotated and aligned with

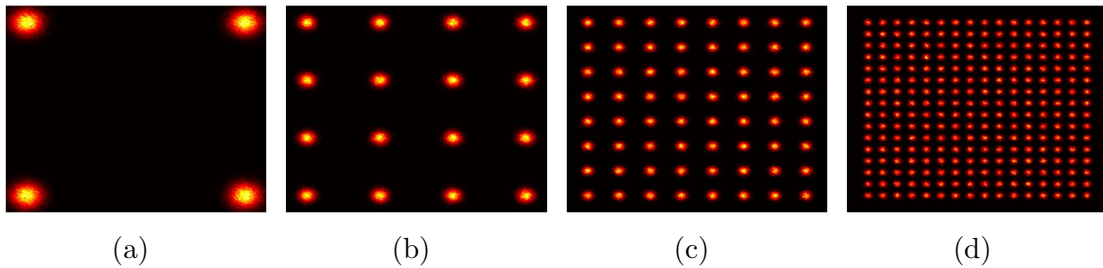


Figure 1.4: Constellations for (a) QPSK, (b) 16QAM, (c) 64QAM, and (d) 256QAM.

the X polarisation. It can be noted that the X and Y polarisations of the incoming signal will in general not match those of the transmitted signal due to random birefringence in the channel. Each branch is then mixed independently with the LO via a 90° hybrid, whose four outputs can be written as a function of the two inputs S and LO as shown in equation (1.9).

$$Out_k = \frac{S + j^{k-1} \cdot LO}{2} \quad \text{for } k = 1, 2, 3, 4 \quad (1.9)$$

So when detecting together Out_1 and Out_3 , and Out_2 and Out_4 using two balanced photodiodes (BPD), the electrical currents I obtained encode the information of the optical field as per equation (1.10).

$$\begin{bmatrix} I_{1,3} \\ I_{2,4} \end{bmatrix} = C \cdot |LO| \cdot |S| \cdot \begin{bmatrix} \cos(\varphi_S - \varphi_{LO}) \\ \sin(\varphi_S - \varphi_{LO}) \end{bmatrix} \quad (1.10)$$

Where C is a constant dependent on the responsivity of the photodiodes, φ_S is the instantaneous phase of the incoming signal, and $\varphi_{LO} = \omega_{LO} \cdot t$. The use of a strong LO operating in CW means the term $|LO|$ is a large constant that provides gain to the detection.

Coherent transmission systems offer several advantages with respect to IMDD.

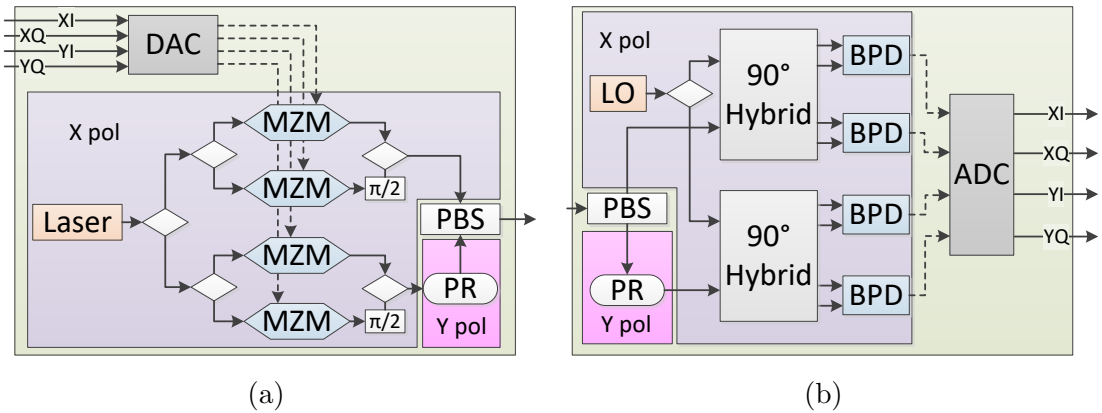


Figure 1.5: (a) Dual polarisation coherent transmitter. (b) Dual polarisation coherent receiver. The purple and pink regions show the areas where the light is X and Y polarised respectively. PBS: Polarisation beam splitter, PR: Polarisation rotator, BPD: Balanced photodiode.

First, it enables the use of amplitude, phase, and polarisation state of the light, allowing for much higher spectral efficiencies than with IMDD. Second, coherent receivers have much higher receiver sensitivities than direct detection-based receivers (in part thanks to the coherent gain provided by the strong LO), increasing the power budget of the system. Third, the detection of different channels on a WDM system can be performed by tuning the emitting frequency of the LO, without the need for tunable optical filters. Lastly, the information on the full field (amplitude, phase, and polarisation) enables the use of powerful DSP techniques capable of compensating large amounts of chromatic dispersion, PMD, and fibre-induced nonlinear distortions. However, the laser phase noise requirements in coherent systems are usually more strict than those of IMDD systems (as the latter do not encode information in the phase of the signal). Additionally, the need for an LO and the more complex transmitter and receiver architectures make the deployment of coherent systems more expensive than that of IMDD systems. Given the trade-offs between coherent and IMDD, coherent technology is the preferred solution for optical systems with high accumulated chromatic dispersion, expensive to deploy optical fibres, or where high spectral efficiency is required. One example of an optical transmission system where coherent technology is preferred is long-haul transmission (Figure 1.6), where thousands of km of fibre are deployed in multiple spans with in-line amplifiers.

Over the last few years, economies of scale, improvements in photonic integration, and the development of more energy-efficient DSP techniques have been driving down the transmission distance for which it is economically feasible to use coherent instead of IMDD technology. Metro data centre interconnects in the 10s of km

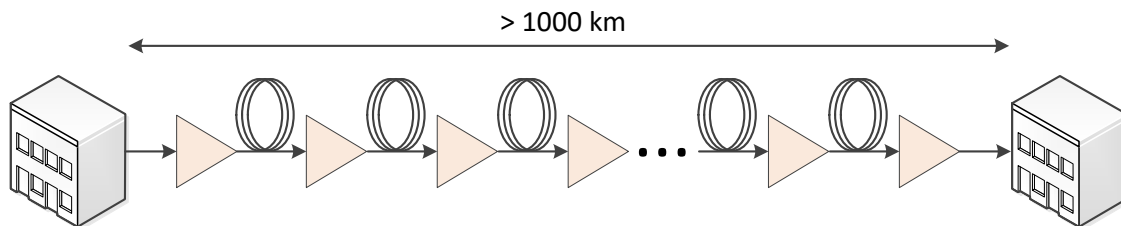


Figure 1.6: P2P long-haul link composed of multiple fibre spans and in-line amplifiers.

already use coherent technology, while campus data centre interconnects with distances up to 10 km are currently close to the threshold where IMDD stops being the preferred solution [27].

1.4 Conclusion

The use of optical wavelengths to transmit information has enabled the explosion in data rates over the last few decades thanks to the much larger bandwidth available compared to radio/microwave transmission. However, the saturation of the transmission capacity over an SMF can already be seen in today's optical networks [28]. Advanced modulation formats allow for higher spectral efficiencies, but also make the signal more sensitive to noise, transceiver distortions, and other channel impairments. In that situation, the currently preferred solution for capacity increases consists of adopting more advanced FEC algorithms and DSP techniques capable of boosting the performance of the already deployed systems. Fibre nonlinearity compensation also plays a major role in increasing the channel capacity, as it enables improving the SNR by using higher launch powers. Some alternative solutions have been proposed over the years to keep up with the increase in demand and avoid the capacity saturation, like space division multiplexing (SDM) based on multimode and multicore fibres [29], multiband transmission beyond the current C+L bands [30], or hollow-core fibres [31]. Each of those approaches has its advantages, but their higher cost and complexity have prevented their widespread use.

Short-reach networks, where cost and power consumption are the main drivers, are currently not fully optimised and close to their maximum capacity, and inexpensive and energy-efficient components are usually the preferred solution. DMLs provide simpler and more efficient solutions than EMLs for IMDD systems, but they introduce distortions due to the nonlinear dynamics of the device. Similarly, optical amplification using an SOA can be cheaper and more efficient than using an EDFA, but at the cost of introducing nonlinear distortions. DSP can extend the usability of those devices by compensating for their impairments in the digital domain. In

the next chapter, we will review some of the main techniques employed in the digital domain to improve performance and increase data rates in optical transmission systems, from FEC coding to DSP and constellation shaping.

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Chapter 2

Coding and DSP for Optical Transmission Systems

High baud rates and advanced modulation formats are required to increase transmission rates and spectral efficiency in modern optical systems. However, higher baud rates also have more noise falling in-band with the signal, reducing the signal-to-noise ratio (SNR). Additionally, advanced modulation formats with multilevel constellations have higher SNR requirements due to the reduced distance between symbols. Given those factors, as the system throughput is pushed closer to its total capacity, achieving reliable transmission becomes more and more challenging. Different levels of reliability are required depending on the application and are typically defined in terms of the bit error ratio (BER), i.e. the probability of any received bit being decoded wrong. Some common reliability thresholds require BERs below 10^{-9} , 10^{-12} , or 10^{-15} , and to achieve them at high data rates, forward error correction (FEC) and digital signal processing (DSP) techniques are usually required. On top of using FEC and DSP, constellation shaping is routinely used to maximise channel performance, increasing transmission distances and data rates. Section 2.1 will present the different types of FEC. Section 2.2 will explain the basics of DSP for intensity modulation with direct detection (IMDD) systems. Then, section 2.3 will introduce the standard DSP used in coherent systems. Finally, section 2.4 will

cover the motivation and basics of constellation shaping for coherent systems, with some notes on shaping for IMDD systems.

2.1 Forward Error Correction

FEC introduces redundancies in the information at the transmitter so that the receiver can detect and correct some errors. Depending on the type of FEC and the data overhead, a wide range of received signal quality can be corrected to be error-free. Two broad families of FEC techniques are the hard-decision FEC (HD-FEC), in which the information is decoded at the receiver and fed into the FEC algorithm, and the soft-decision FEC (SD-FEC), where the likelihoods of each reading (as opposed to the symbol decision) are fed into the algorithm [1].

HD-FEC codes offer relatively low power consumption, and their net coding gain (NCG) and latency vary depending on the implementation. Iterative decoders, such as product codes [2] and staircase codes [3], can trade off latency and NCG by using different decoding windows and iterations. Under the assumption of randomly distributed bit errors, the performance of a system after HD-FEC decoding can be predicted using the BER before FEC (Pre-FEC BER), leading to the definition of an "FEC limit" as the maximum Pre-FEC BER that leads to an error-free signal after a given FEC.

While iterative HD-FEC algorithms can work on relatively high Pre-FEC BERs, SD-FEC algorithms are proven to have significantly higher NCG and are the preferred solution for many current high-speed systems [4]. Unlike with HD approaches, the Pre-FEC BER has a poor prediction capacity in SD-FEC systems, and information theory-based metrics to estimate the achievable information rates (AIR), such as mutual information (MI) or generalised MI (GMI), are required [5]. When probabilistically shaped constellations are used, the normalised GMI (NGMI) has been proven to provide the most reliable metric for prediction [6].

Despite the high NCG provided by SD-FEC algorithms, they also have drawbacks preventing them from fully replacing the HD-FEC: The use of soft information

increases significantly the power consumption and the complexity of the decoder, and some SD-FEC algorithms show an error floor above the desired levels. For those reasons, it is common practice to implement an inner SD-FEC code to partially recover the signal using soft information, and an outer HD-FEC code to remove the error floor while reducing the overall power consumption. Other approaches can be found in the literature to try to reduce the energy requirements of the FEC while keeping a high NCG. Multilevel coding can be used to reduce the throughput of the SD-FEC [7; 8], and soft-aided HD-FEC can bridge the gap between HD and SD-FEC by trading off complexity and performance [9; 10].

2.2 DSP for IMDD

Some of the basic processing techniques performed in the digital domain, either at the transmitter or at the receiver side, are explained in this section:

- **Signal predistortion:** Consists of the modification of the pulse amplitude modulation (PAM) signal in the electrical domain at the transmitter to tailor the spectral profile or the level spacing after optical modulation. This technique can be used in systems based on directly modulated lasers (DML) [11], electro-absorption modulators (EAM) [12; 13], and Mach-Zehnder modulators (MZM) [14]. The aim of signal predistortion can be to obtain equidistant optical levels, mitigate nonlinearities, flatten the optical spectrum, or tailor the level spacing to optimise performance. The use of unevenly distributed optical power levels will be explained in subsection 2.4.1.
- **Digital filtering:** Consists of removing high-frequency components of a signal. It can be applied at the transmitter, at the receiver, or both, and is useful for reducing the spectral width, minimising inter-symbol interference, and removing out-of-band noise. Two commonly used filters are the raised cosine (RC) and the root-raised cosine (RRC). They are defined in the frequency domain given their bandwidth (BW) and the roll-off factor (β) as shown in equation

(2.1).

$$H(f) = \begin{cases} 1 & \text{if } f \leq (1 - \beta) \cdot BW \\ 0 & \text{if } f > (1 + \beta) \cdot BW \\ \left(\frac{1}{2} + \frac{1}{2} \cdot \cos \left[\frac{\pi \cdot (|f| - (1 - \beta) \cdot BW)}{2 \cdot BW \cdot \beta} \right] \right)^k, & \text{otherwise} \end{cases} \quad (2.1)$$

The factor k equals 1 for an RC filter and 0.5 for an RRC. The normalised frequency response of the RC and the RRC filters for different values of β can be seen in figure 2.1(a) and (b) respectively. It can be noted that the frequency response of the different filters converges at a value of 0.5 for the RC case (i.e. the 3-dB bandwidth), but at a value of 0.707 for the RRC. RRC filters are typically used both at the transmitter and the receiver end so the total combined effect of the filters results in a full RC filter.

- Feed-forward equalisation: Commonly used at the receiver side, feed-forward equalisers (FFE) are implemented by performing a linear combination of some neighbouring samples to obtain an equalised output [15]. It is a simple and efficient way of overcoming bandwidth limitations and other linear impairments in the received signal. The output y at the m -th sample given the input col-

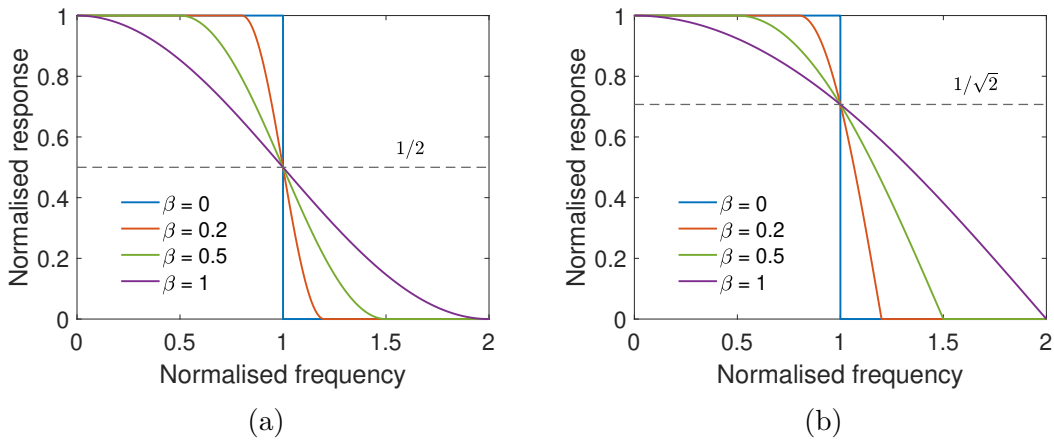


Figure 2.1: Normalised frequency response for different values of β for (a) an RC filter and (b) an RRC filter.

umn vector $\mathbf{E}$, the weights column vector $\mathbf{w}_1$, and the memory length L_1 , can be expressed as in equation (2.2).

$$y(m) = \mathbf{E}_m^T \cdot \mathbf{w}_1^* = \begin{bmatrix} E(m - L_1 + 1) \\ E(m - L_1 + 2) \\ \dots \\ E(m - 1) \\ E(m) \end{bmatrix}^T \cdot \begin{bmatrix} w_1(1) \\ w_1(2) \\ \dots \\ w_1(L_1 - 1) \\ w_1(L_1) \end{bmatrix}^* \quad (2.2)$$

The superscripts ^{T} and * represent the transpose and conjugate operations respectively. While all terms in (2.2) are real in IMDD equalisation, the complex notation generalises its validity also for coherent signals. The weights column vector $\mathbf{w}_1$ can be adaptively updated using an adaptation algorithm like the least-mean squares (LMS) or the recursive least squares (RLS). Training sequences can be used for weight pre-convergence, and pilot symbols can be added periodically into the signal to continuously adapt the weights over time. The adaptation can also be run fully blind by adapting the weights in a decision-directed (DD) manner, i.e. using the hard decision of the decoder as the ideal symbol. Equation (2.3) shows the weight adaptation equation using the LMS algorithm given the step size parameter μ and the m -th symbol decision $D(m)$.

$$\mathbf{w}_1(m + 1) = \mathbf{w}_1(m) + \mu \cdot (D(m) - y(m))^* \cdot \mathbf{E}_m \quad (2.3)$$

- Volterra nonlinear equalisation: If the system is affected by nonlinear impairments, such as fibre nonlinearity, complex dynamics from a DML or a semiconductor optical amplifier (SOA), or response saturation on the photodiode or electrical amplifier due to high received optical power, an FFE may not be enough to recover the received signal. The effect of nonlinearities in the system performance grows rapidly as the modulation format increases from on-off

keying (OOK) to PAM4/8/16, and nonlinear equalisers can help mitigate the distortions at the expense of higher computational complexity. Among nonlinear equalisers, the Volterra nonlinear equaliser (VNLE) is one of the most common approaches. Given the main nonlinear impairments in an optical system, and to keep the computational complexity bounded, VNLEs of up to 2^{nd} or 3^{rd} order are usually considered. Some approaches have been proposed to minimise the complexity of the 2^{nd} order VNLE without sacrificing performance, such as pruning or limiting the number of nonlinear beating terms [16]. Complexity reduction of the 3^{rd} order VNLE has also been proposed in the literature [17].

For a 2^{nd} order VNLE, the output y at the m -th sample can be expressed as shown in equation (2.4), where $\mathbf{W}_2$ is the second order weights triangular matrix and L_2 is the nonlinear memory length. The matrix $\mathbf{W}_2$ is triangular in order to avoid repeating terms due to the symmetry of the expression. Given this matrix representation of the 2^{nd} order VNLE, the complexity reduction technique presented in [16] can be easily introduced by limiting the number of diagonals different to zero in $\mathbf{W}_2$, e.g. if only the main diagonal has non-zero elements, it corresponds to a 2^{nd} order polynomial nonlinear equaliser, and each additional diagonal adds one beating neighbour. The weight values of $\mathbf{W}_2$ can be updated using the same algorithms as for $\mathbf{w}_1$.

$$\begin{aligned}
y(m) = \mathbf{E}_{1,m}^T \cdot \mathbf{w}_1 + \mathbf{E}_{2,m}^T \cdot \mathbf{W}_2 \cdot \mathbf{E}_{2,m} = & \begin{bmatrix} E(m - L_1 + 1) \\ E(m - L_1 + 2) \\ \dots \\ E(m - 1) \\ E(m) \end{bmatrix}^T \cdot \begin{bmatrix} w_1(1) \\ w_1(2) \\ \dots \\ w_1(L_1 - 1) \\ w_1(L_1) \end{bmatrix} + \\
& \begin{bmatrix} E(m - L_2 + 1) \\ E(m - L_2 + 2) \\ \dots \\ E(m - 1) \\ E(m) \end{bmatrix}^T \cdot \begin{bmatrix} w_2(1, 1) & w_2(1, 2) & \dots & w_2(1, L_2) \\ 0 & w_2(2, 2) & \dots & \dots \\ \dots & 0 & \dots & w_2(L_2 - 1, L_2) \\ 0 & \dots & 0 & w_2(L_2, L_2) \end{bmatrix} \cdot \begin{bmatrix} E(m - L_2 + 1) \\ E(m - L_2 + 2) \\ \dots \\ E(m - 1) \\ E(m) \end{bmatrix} \quad (2.4)
\end{aligned}$$

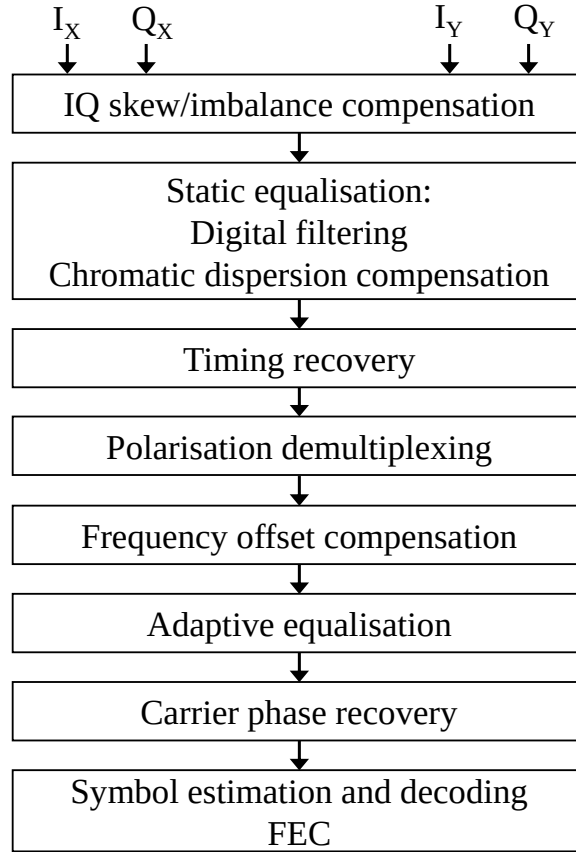


Figure 2.2: Flowchart of a typical coherent receiver DSP. The inputs of the first block correspond to the four outputs of a coherent receiver as in figure 1.5(b).

2.3 DSP for Coherent

As mentioned before, the access to the electric field provided by the coherent receiver enables the use of much more powerful DSP techniques than with IMDD systems. Most of the DSP steps in a fibre-optic coherent system are performed at the receiver, while the transmitter DSP usually consists of bit-to-symbol mapping (after FEC encoding), preemphasis/predistortion and digital filtering. Chromatic dispersion compensation can be implemented at the transmitter, at the receiver side, or can be split between them. The standard receiver DSP flowchart is shown in figure 2.2, where a dual-polarisation coherent signal is assumed to be detected for the DSP to process it [18]. The different steps are explained in more detail in subsection 2.3.1. Then, the digital backpropagation (DBP) algorithm for fibre nonlinearity compensation is explained in subsection 2.3.2.

2.3.1 Standard Coherent DSP Steps

- IQ imbalance compensation: Compensates mismatches between the I and Q components of the received signal. A common technique is the Gram-Schmidt orthogonalisation procedure [19].
- Digital filtering: Similar to those of IMDD systems, digital filters in coherent systems can remove out-of-band noise and reduce the spectral width of the signal. Dense wavelength division multiplexing (DWDM) systems, with tightly packed channels, use very small values of β to allow for minimal channel spacings without overlapping.
- Chromatic dispersion compensation: Most of the inter-symbol interference affecting a signal comes from chromatic dispersion during propagation and can be effectively corrected by assuming a linear regime in equation (1.2). By neglecting the nonlinear term, the propagation can be analytically reversed in the frequency domain, and the transfer function of the inverse fibre can be expressed as shown in equation (2.5) [20].

$$H_{CD}(\omega) = \exp\left(\frac{j\beta_2\omega^2 L}{2}\right) \quad (2.5)$$

Where β_2 is the group velocity dispersion parameter and L is the total fibre length. Third-order dispersion was also neglected in the expression above. In terms of implementation, the size of the Fast-Fourier Transform (FFT) can be optimised to minimise computational complexity for a given system [20]. It can be noted that on a long-haul transmission system, chromatic dispersion compensation can be the main source of energy consumption of the DSP flow [21], while short-reach applications do not need a chromatic dispersion compensation block [22].

- Timing recovery: Corrects the error in timing phase and frequency between the transmitter and receiver clocks. A common algorithm is the 4th power

method, which can be implemented both in time [23] and in frequency domain [24].

- **Polarisation demultiplexing:** Random polarisation rotations in the fibre scramble the X and Y components of the signal. The polarisations can be separated effectively by looking at the signal in the Stokes space [25]. This technique can be applied to any coherent signal generated with a coherent transmitter in the form of figure 1.5(a), where the power is evenly split into two orthogonal polarisations, and is based on the observation that the direction of minimum average power in the Poincaré sphere is aligned with the original X and Y polarisations. A simple implementation consists of calculating the Stokes parameters for multiple samples and fitting a 3-D plane to them. The polarisation rotation required for the signal is then obtained from the angle of the plane.
- **Frequency offset compensation:** Estimates the frequency difference between the received signal and the local oscillator (LO) and compensates it [26]. Most coherent systems have a free-running LO, i.e. not frequency locked to the transmitter laser, resulting in frequency offset values that can be in the order of a few GHz.
- **Adaptive equalisation:** Residual inter-symbol interference and time-varying perturbations are corrected in this step. Adaptive equalisers can also be used for polarisation demultiplexing in dual-polarisation signals. The basic equaliser for a single polarisation signal can be implemented as in equation (2.2), and it can be extended into the dual-polarisation equivalent as shown in equation (2.6).

$$\begin{bmatrix} y_X(m) \\ y_Y(m) \end{bmatrix}^T = \begin{bmatrix} \mathbf{E}_{X,m} \\ \mathbf{E}_{Y,m} \end{bmatrix}^T \cdot \begin{bmatrix} \mathbf{w}_{1,X,X} & \mathbf{w}_{1,X,Y} \\ \mathbf{w}_{1,Y,X} & \mathbf{w}_{1,Y,Y} \end{bmatrix}^* \quad (2.6)$$

The equalisers presented in equations (2.2) and (2.6) are known as strictly linear, as they treat the real and imaginary parts of $\mathbf{E}$ jointly. Improved per-

formance can be obtained in most systems by using a widely linear equaliser based on a real-valued 4×4 multiple-input multiple-output (MIMO) configuration, as opposed to the complex-valued 2×2 strictly linear MIMO equaliser used for a dual-polarisation [27]. The output of the widely linear equaliser can be seen in equation (2.7), where the subscripts r and i indicate the real and the imaginary parts respectively.

$$\begin{bmatrix} y_{X_r}(m) \\ y_{X_i}(m) \\ y_{Y_r}(m) \\ y_{Y_i}(m) \end{bmatrix}^T = \begin{bmatrix} \mathbf{E}_{X_r,m} \\ \mathbf{E}_{X_i,m} \\ \mathbf{E}_{Y_r,m} \\ \mathbf{E}_{Y_i,m} \end{bmatrix}^T \cdot \begin{bmatrix} \mathbf{w}_{1,X_r,X_r} & \mathbf{w}_{1,X_r,X_i} & \mathbf{w}_{1,X_r,Y_r} & \mathbf{w}_{1,X_r,Y_i} \\ \mathbf{w}_{1,X_i,X_r} & \mathbf{w}_{1,X_i,X_i} & \mathbf{w}_{1,X_i,Y_r} & \mathbf{w}_{1,X_i,Y_i} \\ \mathbf{w}_{1,Y_r,X_r} & \mathbf{w}_{1,Y_r,X_i} & \mathbf{w}_{1,Y_r,Y_r} & \mathbf{w}_{1,Y_r,Y_i} \\ \mathbf{w}_{1,Y_i,X_r} & \mathbf{w}_{1,Y_i,X_i} & \mathbf{w}_{1,Y_i,Y_r} & \mathbf{w}_{1,Y_i,Y_i} \end{bmatrix} \quad (2.7)$$

An adaptive equaliser can update its weights based on the amplitude of the signal, like the constant modulus algorithm (CMA) or the radius-directed equaliser (RDA), or can use the full signal (i.e. amplitude and phase), like a DD equaliser [18]. Amplitude-directed approaches have the advantage of being insensitive to phase and frequency noise, so they can be implemented as a standalone DSP block either before or after frequency offset compensation and carrier phase recovery. Meanwhile, DD equalisers require the phase recovery and the symbol decision steps to be integrated into the feedback loop. The update equation using the LMS algorithm for a DD equaliser is the same as in equation (2.3), while for an RDA, the update equation can be written as equation (2.8), where $R(m)$ is the closest radius to $y(m)$.

$$\mathbf{w}_1(m+1) = \mathbf{w}_1(m) + \mu \cdot (|R(m)|^2 - |y(m)|^2) \cdot y^*(m) \cdot \mathbf{E}_m \quad (2.8)$$

- Carrier phase recovery: It tracks and corrects the phase noise from the laser at the transmitter and the LO. Various algorithms have been proposed with different levels of computational complexity, latency, and modulation format flexibility [28; 29]. Alternatively, one common approach is to place pilot symbols

along the signal and use linear interpolation to calculate the phase correction required for each symbol [30].

2.3.2 Digital Backpropagation

On top of the standard DSP blocks, additional processing can be implemented to mitigate specific impairments. One of the major impairments in many current optical transmission systems, and hence one of the points where more potential gain is available, is fibre nonlinearity. Kerr nonlinearity distorts the signal at high launch powers, and its mitigation enables the increase of the signal power, allowing higher SNR values. Various DSP-based techniques for nonlinearity mitigation have been proposed, like using the inverse Volterra series transfer function of the fibre [31; 32] or perturbation theory-based approaches [33; 34]. One technique that has been extensively studied, and commonly used as a benchmark, is known as digital backpropagation (DBP) [35]. DBP replaces the chromatic dispersion compensation step with a numerical solution of equation (1.2) in reverse, compensating simultaneously for chromatic dispersion and fibre nonlinearity. The Split-Step Fourier Method (SSFM) is a well-studied implementation to solve the Nonlinear Schrödinger Equation (NLSE) (or the Manakov equations in dual polarisation), where the analytical solution for the dispersion (in frequency domain) and for the nonlinearity (in time domain) are implemented separately for small fibre lengths. The FFTs and inverse FFTs required for the SSFM dominate its computational complexity, and adaptive step sizes are required to optimise the accuracy of the solution [36].

In a WDM system, where each channel is affected by intra-channel self-phase modulation (SPM) and inter-channel cross-phase modulation (XPM) and four-wave mixing (FWM), the DBP can be implemented on a channel-by-channel basis (i.e. correcting only SPM), backpropagating a system of coupled differential equations as shown in equation (2.9) (correcting SPM and XPM), or combining all channels into a single (in general highly oversampled) field and backpropagating it (potentially

correcting all Kerr-induced signal-signal nonlinearity) [37].

$$\frac{\partial E_i}{\partial z} = -j\frac{\beta_2}{2}\frac{\partial^2 E_i}{\partial t^2} - \frac{\beta_3}{6}\frac{\partial^3 E_i}{\partial t^3} + \frac{\alpha}{2}E_i + j\gamma\left(|E_i|^2 + 2\sum_{k \neq i}|E_k|^2\right)E_i \quad (2.9)$$

In reality, different channels might use independent digital signal processors without sharing information, limiting the DBP to the modest gains offered by intra-channel nonlinearity mitigation [38]. Even when all channels are combined into a single field and backpropagated together, other factors can limit the achievable gain from DBP, namely the bandwidth of the transmitter/receiver, the carrier frequency uncertainty and the signal-noise nonlinear interaction.

The receiver bandwidth limits the nonlinearity mitigation when DBP is performed at the receiver because the channels generate out-of-band FWM tones as they propagate through the fibre, and those tones contain information required for the full recovery of the original signal. If the bandwidth of the receiver is not wide enough to capture all out-of-band information, it cannot reverse the propagation to the original state. If DBP is performed at the transmitter, where all the spectral content of the signal is known, the digitally generated out-of-band FWM tones need to be transmitted alongside the signal, which requires dedicated lasers and IQ modulators.

Independently of the system and simulation bandwidth, the mitigation of inter-channel nonlinearity can be severely affected by the frequency drift of the lasers at the transmitter and the LOs [39]. When separate lasers are used, the independent frequency drifts result in a random, slowly varying channel spacing that affects the walk-off between channels and distorts the nonlinear interactions. When DBP is performed assuming nominal channel spacing, if the digital channel walk-offs do not match the physical ones, the nonlinearity mitigation gets degraded. Optical frequency combs can be used to solve this issue, as their lines offer high frequency and phase correlation. The use of an optical comb as LOs is enough when DBP is performed at the receiver, even when independent lasers are used at the transmitter,

as the real channel spacing is encoded in the captured signals referenced to the frequency-locked LOs. Analogously, when DBP is performed at the transmitter, an optical comb is only required at the transmitter, and the frequency uncertainty of the LOs does not affect the recovery.

The signal-noise nonlinear interactions are a fundamental limitation that becomes more noticeable as the transmission distance increases and the SNR degrades. Parametric noise amplification (PNA) is the first signal-noise nonlinear interaction to become significant, and DBP cannot compensate for it as it involves interactions with ASE generated along the full channel. However, PNA can be reduced by splitting the DBP in half between the transmitter and the receiver [40]. In this case, frequency-referenced sources like optical frequency combs and wide bandwidth transceivers capable of containing the full signal and out-of-band tones are required both at the transmitter and receiver ends.

2.4 Constellation Shaping

While FEC and DSP offer high capacity gains, the AIR of the system is limited to the modulation-constrained capacity, which for a squared quadrature amplitude modulation (QAM) constellation over a memoryless additive white Gaussian noise (AWGN) channel is ~ 1.53 dB lower than the Shannon capacity [41; 42]. Constellation shaping can bridge the gap between the absolute capacity and the modulation-constrained AIR by changing the position of the symbols in the complex plane, i.e. geometrical shaping (GS) [43], or the probability of occurrence of the symbols, i.e. probabilistic shaping (PS) [44]. Basic PS consists of generating two PAM-M sequences (corresponding to the I and Q components of the M^2 -QAM constellation) with an amplitude-dependent symbol probability. The Maxwell-Boltzmann (MB) probability distribution is optimal for AWGN channels, and it is obtained for a symbol x in the set $X = \pm 1, \pm 3, \dots, \pm(M-1)$ as in equation (2.10):

$$P(x) = \frac{\exp(-\lambda x^2)}{K} \quad (2.10)$$

where λ is the shaping parameter and K is a normalisation parameter. Increasing the shaping strength reduces the entropy rate (i.e. the average number of bits per symbol encoded in the signal, obtained as $H(X) = -2 \sum_X P(x) \log_2 P(x)$) of the transmitted signal, but increases the peak-to-average power ratio (PAPR), improving the performance for a given AWGN power. Figure 2.3(a) shows the probability distributions of the 1-D symbols of a PS 64QAM constellation with different entropy rates ranging from 5 to 6 bit/symbol. Figure 2.3(b)-(d) show a 64QAM constellation with an SNR of 20 dB and different entropy rates ranging from 6 bit/symbol to 5.5 bit/symbol. It can be seen how the symbol spreading due to noise gets reduced (with respect to the symbol separation) as the shaping increases. These factors result in an optimal value of λ , dependent on the SNR, that maximises the AIR.

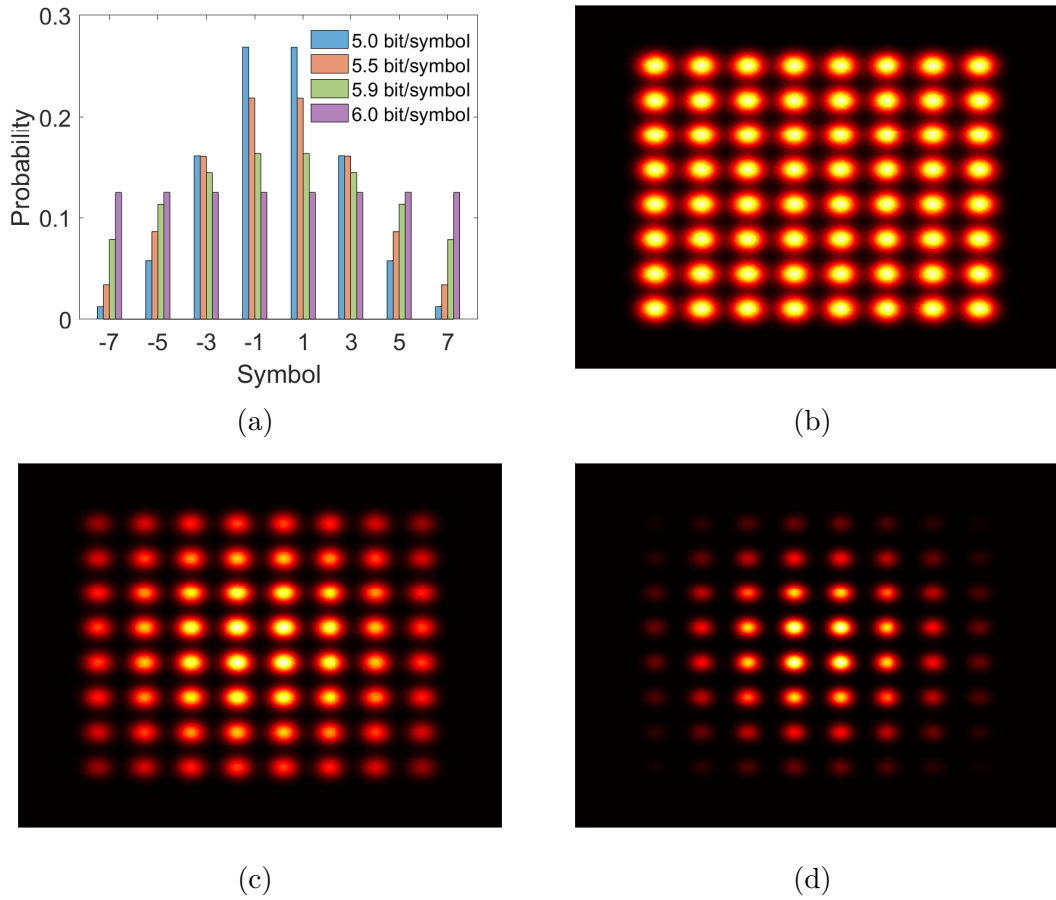


Figure 2.3: (a) 1-D symbol probabilities for a PS 64QAM constellation based on an MB distribution and different entropy rates. PS 64QAM constellations with an MB distribution for an SNR of 20 dB and an entropy rate of (b) 6 bit/symbol, i.e. unshaped 64QAM, (c) 5.9 bit/symbol, (d) 5.5 bit/symbol.

While GS does not reduce the signal entropy, potentially enabling the use of smaller constellations than with PS, the arbitrary symbol position does not allow the use of Gray mapping, which can introduce a penalty to the GMI with respect to the MI [42]. This is significant for most practical systems as the GMI is the relevant metric for systems using binary FEC. Additionally, the arbitrary symbol position imposes higher resolution requirements on the digital-to-analogue converter (DAC) at the transmitter.

As a first approximation, an amplified spontaneous emission (ASE)-dominated channel can be modeled as a memoryless AWGN channel, however, fibre-optic channels are in general nonlinear channels with memory for which the AWGN-optimised constellation shaping no longer provides the best results. The nonlinear interference noise (NLIN) model, also called the enhanced Gaussian-noise (EGN) model, [45; 46] quantifies the penalty from Kerr nonlinearity as an equivalent AWGN term, and predicts an increase in NLIN for higher shaping strengths. Using this model, a lower shaping strength is found to give optimal results [47]. Another relevant scenario where the predictions from the AWGN model break down is a phase noise-dominated channel [48]. In this case, the phase noise (usually described as a Wiener process) is corrected using a phase tracking algorithm, and the residual phase noise follows a Tikhonov distribution [49]. The model including an AWGN and a zero-mean residual phase noise term is called partially coherent AWGN (PC-AWGN), and the distinct channel transition probabilities [50; 51] result in different shaping to optimise performance. In general, obtaining the shaping that provides maximum AIR is an open problem and its calculation is performed by optimisation algorithms both for the nonlinear fibre channel [52; 53] and for the phase noise channel [54; 55].

As a next step beyond PS and GS, hybrid probabilistic and geometrical shaping (HPGS) combine both techniques to maximise the AIR [56; 57; 58] but requires in general complex optimisations to converge to the desired solution. 4-D constellation shaping techniques (as opposed to 2-D shaping of each polarisation separately) have also been proposed for further gains and increased nonlinearity tolerance [59]. Ma-

chine learning-based end-to-end system optimisations have shown good results over arbitrary channels [60; 61; 62], but they require computationally intensive training and do not provide further insights in the underlying physics due to the lack of a theoretical basis.

2.4.1 Shaping for IMDD

Unlike QAM-based systems, where the constellation is symmetric around the null point, PAM formats have a DC offset and the symmetry does not hold. In this scenario, two basic PS families can be defined: symmetric (or bipolar) PS and asymmetric (or unipolar) PS [63]. Analogous to the probabilities from equation (2.10) in coherent PS, asymmetric PS assigns the probabilities based on the symbol power, giving maximum probability to the lower symbol and minimum probability to the upper symbol. This approach simultaneously reduces the average signal power and the SNR requirements, offering no advantage in unamplified systems (as is the case in most IMDD systems). Symmetric PS, on the other hand, assigns minimum probability to the upper and lower symbols while keeping the average power constant. It does not affect the noise tolerance of the signal, but it has been shown to provide an advantage both in bandwidth-limited systems [63] and in the presence of nonlinear

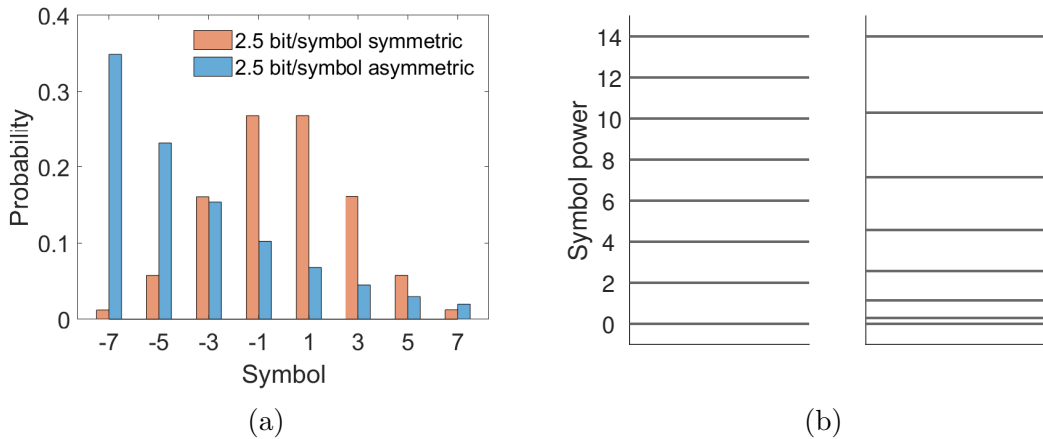


Figure 2.4: (a) Symbol probabilities for a PAM8 constellation with an entropy rate of 2.5 bit/symbol using a symmetric and an asymmetric distribution. (b) Symbol distribution for a GS PAM8 constellation with (left) equidistant power levels and (right) equidistant field amplitude levels. An extinction ratio of 30 dB was assumed for the equidistant amplitudes in the right plot.

distortions [64]. Figure 2.4(a) shows the symbol probabilities for a PAM8 signal with entropy rates of 2.5 bit/symbol using symmetric and asymmetric PS.

GS in PAM constellations offers a simple way of optimising performance in systems under both signal-independent and signal-dependent noise after photodetection. In channels dominated by signal-independent noise, such as receiver thermal noise, optimal performance is obtained from equidistant power levels (i.e. standard PAM). However, systems dominated by signal-dependent noise linearly proportional to the signal power, such as shot noise, operate optimally using constellations with equidistant amplitude levels (i.e. quadratically spaced power levels) [65]. Figure 2.4(b) shows an example of the symbol spacing on a PAM8 signal optimised for signal-independent noise (left) and for linearly proportional signal-dependent noise (right). Relative intensity noise (RIN)-dominated systems are characterised by signal-dependent noise quadratically proportional to the signal power, resulting in a different optimal symbol distribution. In a real system, the GS of the PAM constellation should be done considering all types of noise present to arrive at the ideal symbol distribution.

2.5 Conclusion

The use of advanced modulation formats and high baud rates to increase the channel throughput makes it more challenging to reliably recover the information at the receiver. Different techniques are applied in the digital domain, both at the transmitter and at the receiver end, to maximise the data rates ensuring error-free operation.

This chapter has discussed the use of FEC techniques, which introduce redundancies in the transmitted sequence to allow for the detection and correction of bit errors at the receiver. In recent years, the trend of the industry has been to use more powerful inner SD-FEC with higher overheads and NCG together with an outer HD-FEC. While binary FEC techniques, like those used for bit-interleaved coded modulation (BICM) [66], have been the dominant approach due to their flexibility

and ease of implementation, multilevel coding may grow in the coming years due to its potentially lower power consumption.

DSP is a core component in modern coherent systems, as well as in many PAM-based IMDD systems. Multiple algorithms have been developed and refined in the last decade to improve performance, reduce latency, and increase energy efficiency, and the evolution is expected to continue in the coming years. However, as the energy efficiency in the optical networks gains relevance, some of the signal processing load might shift from the digital to the analogue domain, where it can be performed with a fraction of the power. In that situation, a hybrid signal processing architecture can be expected to operate by combining photonic signal processing [67; 68], electronic signal processing [69], and DSP.

Constellation shaping has grown in importance over the last few years, and research is still being made to optimise it for real channels. As the AIR over these channels approaches their total capacity, the additional gains provided by further improvements become increasingly small, reaching a point of marginal returns. On the other hand, the maturing of its implementation can be expected to enable the use of constellation shaping for shorter-reach, lower-cost applications.

In the next chapter, we will look at an IMDD system based on a DML for low-cost short-reach optical access applications. The use of DSP techniques together with the on-chip optical injection provided by the novel device allowed us to achieve 30 Gbit/s PAM4 transmission over up to 75 km of fibre.

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Chapter 3

Optical Injection for DML-Based IMDD Systems

Intensity modulation with direct detection (IMDD) technology is still the preferred solution for short-reach, low-cost applications like intra-data centres and access networks, where the use of coherent technology is not economically feasible yet. Multiple demonstrations of IMDD transmission over 10s of km have been published in recent years using both direct modulation [1; 2; 3; 4; 5; 6], and external modulation [7; 8; 9; 10; 11; 12; 13]. Directly modulated lasers (DML) offer higher energy efficiency and simplicity than externally modulated laser (EML) solutions, but the frequency chirp produced by the direct modulation of the active medium usually degrades the performance beyond recoverable levels even with moderate levels of chromatic dispersion, limiting the usability to O-band operation or very short fibre lengths. As the passive optical network (PON) standards move towards faster data rates, longer distances, and C-band operation [14], the use of direct modulation becomes more and more challenging, and the required additional digital signal processing (DSP) to recover the signal can outweigh the cost/energy advantage of the DML. The main trend for increasing the data rates and transmission distances of IMDD systems while maintaining good performance consists of using external modulation. An alternative approach to moving to EMLs is using optical injection

locking. Injection locking can be obtained by feeding into a DML a low-power continuous wave (CW) beam close to the emission frequency. Under the right conditions, the light from the DML gets locked to the injected light, changing the properties of the output signal [15; 16].

Integrated DMLs with a master-slave configuration offer a scalable and cost-effective way of getting optical injection [17; 18]. They consist of two sections: a master laser (ML) operating in continuous wave and a slave laser (SL) which is directly modulated with the driving signal. In comparison with a standard DML, this configuration represents an increase in complexity. However, when comparing it to integrated EMLs with semiconductor optical amplifier (SOA) solutions [12], the proposed master-slave configuration offers a wider wavelength tunability range at considerably lower complexity and cost due to the simple regrowth-free fabrication process required [19; 18]. As the cost of photonic integrated circuits (PIC) decreases, the use of optical injection locking, in conjunction with an SOA, can become a more competitive solution for a low-chirp, high-output power transmitter for use in short-reach systems such as optical access networks. This can remove the need for discrete optical amplifiers and facilitate C-band time/wavelength division multiplexed (TWDM) systems operating over tens of km.

This chapter presents the characterisation and performance of a novel directly modulated multi-section laser with a master-slave configuration recently published in [20]. Section 3.1 explains the fabrication and characterisation of the device. Then, section 3.2 outlines the experimental setup for testing the device's performance and the simulation setup for assessing the results. Finally, section 3.3 shows the results of the experiments and simulations with injection-locking DMLs.

3.1 Integrated Master-Slave DML

The device was fabricated using AlGaInAs material with 5 strained quantum wells uniformly grown in the active region on an n-doped In-P substrate. The PIC is composed of four main sections including an ML composed of a gain section (MG)

between two reflector sections (MR_1 and MR_2), a variable optical attenuator (VOA), an SL composed of a gain section (SG) between two reflector sections (SR_1 and SR_2), and an SOA. A 2- μm ridge waveguide formed on zinc-doped p-type semiconductor material (AlGaInAs) was employed to couple light between adjacent sections and anti-reflective coating was placed on the front and back facets. The reflector sections of the ML and SL were created using high-order surface grating structures defined using electron-beam lithography with a slot depth of 1.35 μm and a slot width of 1.15 μm . The length of each of the sections of the PIC can be found in table 3.1. The VOA separates both laser sections and controls the power injected from the ML into the SL. The length of the slave gain section was kept short to reduce the photon lifetime and increase the laser bandwidth. The SOA, used to boost the output power, has an angled waveguide to reduce reflections. The schematic and a picture of the PIC can be seen in figure 3.1(a). A ruggedised packaging system (Figure 3.1(b)) that allows PIC characterisation and testing in the laboratory was designed. The solution is based on a typical metal butterfly package that accommodates a single 10G RF input for the gain section of the SL, several DC inputs to control the bias currents for the other sections and a thermo-electric cooler (TEC) for temperature control. In order to further reduce back-reflections, the package also includes an angled incident optical fibre.

The SL has a typical threshold current of 10 mA under free-running operation (Figure 3.1(c)), and of around 12 mA when the ML gain section is biased at 30 mA. The nonlinear response shown by the free-running laser is due to the saturation of the SOA section at the defined bias current of 30 mA for the SOA. Figure 3.1(d) shows the modulation response of the PIC when the SL is free-running and with injection from the ML, measured with the PIC on a probe station. An increase in the 3-dB bandwidth can be seen when using optical injection. The output frequency from the laser can be tuned over a range of hundreds of GHz by varying the different bias currents while keeping a side mode suppression ratio of around 50 dB (Figure 3.1(e)), making the device suitable for TWDM PON applications. The operating

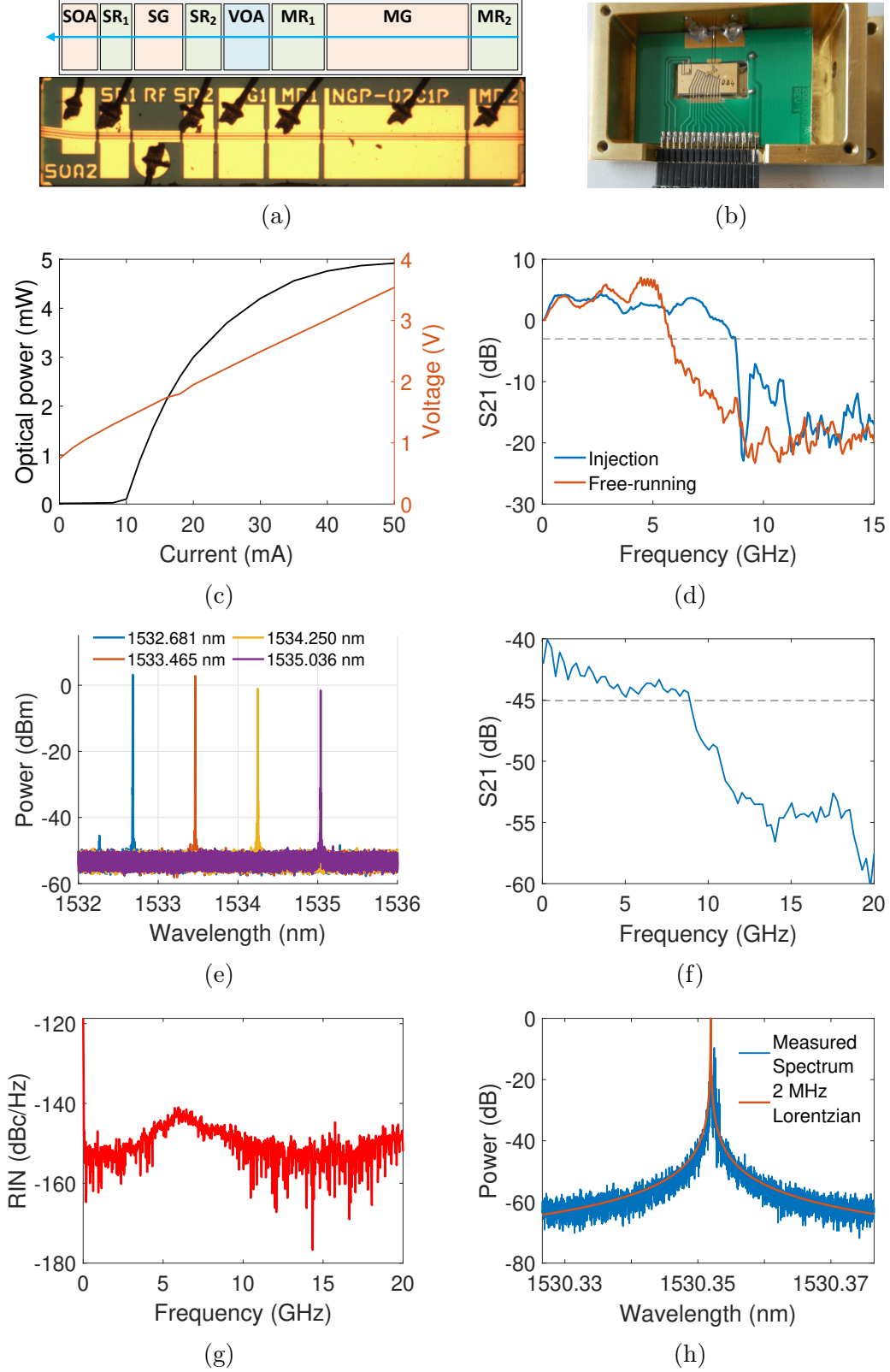


Figure 3.1: (a) Schematic and picture of the PIC used in this work. (b) Picture of the packaged device. (c) LIV plot of the free-running slave laser with the SOA biased at 30 mA. (d) Normalised modulation response of the device with and without optical injection measured using a probe station. (e) optical spectra of the PIC tuned to a wide range of wavelengths. (f) Modulation response of the packaged device at the operating point used for this work. (g) RIN measurement. (h) Linewidth measurement and Lorentzian fit.

Table 3.1: Length, bias current at the operating point and typical power consumption of each of the sections of the PIC.

	SOA	SR ₁	SG	SR ₂	VOA	MR ₁	MG	MR ₂
Length (μm)	233	135	200	135	200	221	600	221
Bias (mA)	35	41.2	37	41	1	59	31	70
Electrical Power (mW)	24	75.6	145.2	73.8	5	129.6	45.1	126

bias values used for this work can be found in table 3.1, together with typical values of power consumption per section (these values may vary depending on the operating point of the PIC). It can be noted that the power consumption of the PIC is in the order of hundreds of mW, so the total consumption from the device is dominated by the TEC.

With the SL injected by the ML, the packaged device shows a 3 dB bandwidth of around 8 GHz (Figure 3.1(f)). The relative intensity noise (RIN) from the laser was measured using the setup detailed in [21]. The obtained RIN values up to a frequency of 20 GHz are shown in figure 3.1(g). The laser relaxation oscillation at around 6 GHz can be seen, and values below -140 dB/Hz were obtained over the whole frequency range. The linewidth was measured by capturing the spectrum of the laser in continuous-wave operation with a high-resolution (20 MHz) optical spectrum analyser (Figure 3.1(h)). A 3-dB linewidth of around 2 MHz was estimated by fitting a Lorentzian profile to the measurement.

3.2 Transmission Experiment

The performance of the DML for transmission over distances relevant for PONs was tested for 30 Gbit/s 4-level pulse amplitude modulation (PAM) with two different experimental setups: Unamplified transmission over 25 km and 50 km of single-mode fibre (SMF), and amplified transmission over 75 km of SMF (Figure 3.2). An erbium-doped fibre amplifier (EDFA) was required at the transmitter to reach 75

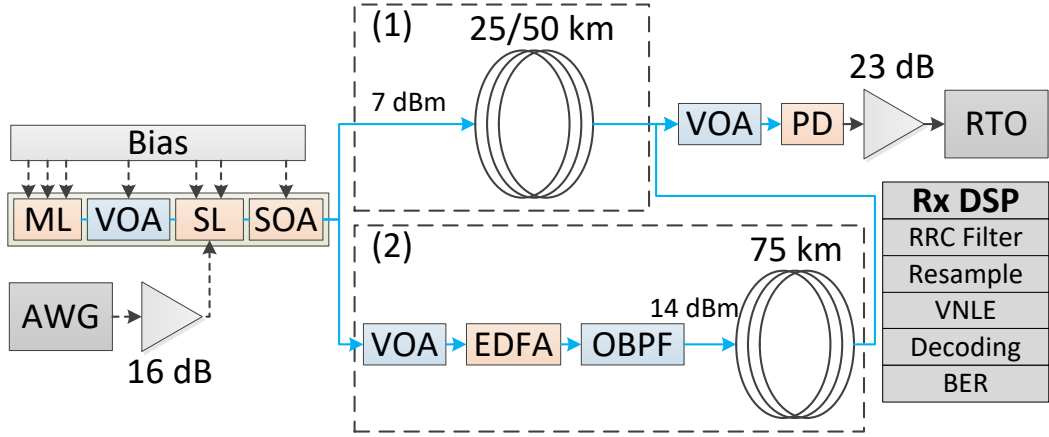


Figure 3.2: Experimental setup. Paths (1) and (2) show the unamplified and amplified transmission experiments respectively.

km of transmission due to optical fibre attenuation. A pseudorandom multilevel PAM4 sequence 2^{15} symbols long was used in all cases. The 15 GBaud multilevel signal was generated and filtered offline with a 10 GHz root-raised cosine (RRC) filter, and sent to an arbitrary waveform generator (AWG) operating at 90 GSa/s. The output of the AWG, was sent to a 16 dB gain linear RF amplifier before driving the SL of the DML (biased at 37 mA). The amplitude swing of the driving current was optimised for each experiment in order to obtain optimal performance at the receiver. A peak-to-peak current of 35 mA was set for the unamplified experiments, while it was reduced to 25 mA for the amplified transmission. The reduced drive current produces a more linear output at the expense of reducing the extinction ratio of the signal and hence degrading the signal-to-noise ratio (SNR). This trade-off was found to be beneficial when the amplified system was employed, as the use of the EDFA overcomes the degradation in SNR caused by increased link loss over 75 km. It should be noted that the use of a booster EDFA in this way results in additional fibre nonlinear effects as the optical launch power is increased. The resulting optical signal with an output power of 7 dBm from the DML was, for the unamplified experiment, transmitted over 25/50 km of SMF, while for the amplified one it was attenuated to -6 dBm before entering the EDFA which boosted the power to +22 dBm. An optical band pass filter (OBPF) was placed after the EDFA to reduce

out-of-band amplified spontaneous emission (ASE), resulting in a launch power of +14 dBm into 75 km of SMF. At the receiver, the transmitted optical signal entered a VOA which was used to sweep the received optical power (ROP) at the 20 GHz photodiode (PD). A 23 dB gain RF amplifier boosted the obtained photocurrent, which was captured with the real-time oscilloscope (RTO) at a sampling rate of 100 GSa/s. The captured signal was matched filtered offline with an RRC filter and resampled to 1 sample per symbol for digital equalisation.

In order to compare the relative impacts of chromatic dispersion, noise and nonlinearity on the performance of the 75 km transmission experiment, a 10 GBaud PAM8 (30 Gbit/s) signal of 2^{15} symbols long was generated and filtered offline (10 GHz RRC filter), and transmitted over 75 km of SMF. The increased number of levels of the PAM8 signal makes it more sensitive to nonlinear distortions coming from the complex laser dynamics, while the lower baud rate reduces the penalty coming from fibre dispersion. The same experimental setup and conditions as those associated with the 75 km PAM4 transmission described above were used; with the exception of the peak-to-peak value of the driving current driving the DML. This value was set to 30 mA due to the increased SNR requirements of the PAM8 signal.

To help study the influence of injection locking on the performance of the system, a simulation was run to replicate the experimental setup. The simulation setup is explained in the next subsection.

3.2.1 Simulation Setup

The simulation was set up in Matlab including some of the main impairments of the system: the bandwidth limitation of the different components, the ASE from the laser and amplifier, the electrical noise in the signal driving the laser, the laser dynamics, the linear and nonlinear distortions induced by fibre propagation, and the direct detection of the signal with a photodiode including shot noise, thermal noise and dark current. It should be noted that the purpose of the laser rate equations in this simulation is to provide insight into the influence of injection locking on the

Table 3.2: Parameters used for the simulation of the dynamics of the DML. The symbols used here correspond to the definitions from table 1.1.

Parameter	Value
Bias current	40 mA
S_i	$1.536 \times 10^{20} \text{ 1/m}^3$ (Injected power=80 μW)
V	$2 \times 10^{-17} \text{ m}^3$
τ_c	760 ps
v_g	$8.33 \times 10^7 \text{ m/s}$
Γ	0.25
τ_p	0.79 ps
R_{sp}	1.18 1/ps
T	4.802 ps
ω_s	195.94 THz
$\omega_m - \omega_s$	$-2\pi \times (4 \text{ GHz})$
β	3
a	$1.91 \times 10^{-20} \text{ m}^2$
N_{th}	$4.252 \times 10^{24} \text{ 1/m}^3$
N_0	$1.07 \times 10^{24} \text{ 1/m}^3$
ε	$1.48 \times 10^{-23} \text{ m}^3$

system performance and not to exactly replicate the dynamics and operation of the multisection integrated device used for the experiment.

The laser dynamics were modelled using the rate equations for a DML with optical injection (see subsection 1.3.1, equations (1.5) to (1.8)). The equations were solved by using a 4th order Runge-Kutta with the parameters shown in table 3.2. Figure 3.3(a) shows the steady-state output power and the carrier density of the noiseless free-running slave laser as a function of bias current. A threshold current of 18 mA can be observed. In figure 3.3(b), a colourmap showing the stability region of the injection locking is displayed. The yellow region shows the combinations of injected power into the slave laser and frequency detuning between master and slave that result in injection locking for a bias current of 40 mA and no modulation. A detuning of -4 GHz and a power of 80 μW were selected for the simulation to provide a stable operating point. The fibre propagation was modelled by solving the Non-linear Schrödinger Equation (NLSE) using the Split-Step Fourier Method (SSFM). The dispersion coefficient of the fibre was set to 16 ps/nm/km, the dispersion slope to 0.08 ps/nm²/km, the nonlinear coefficient to 1.3 1/W/km, and the attenuation

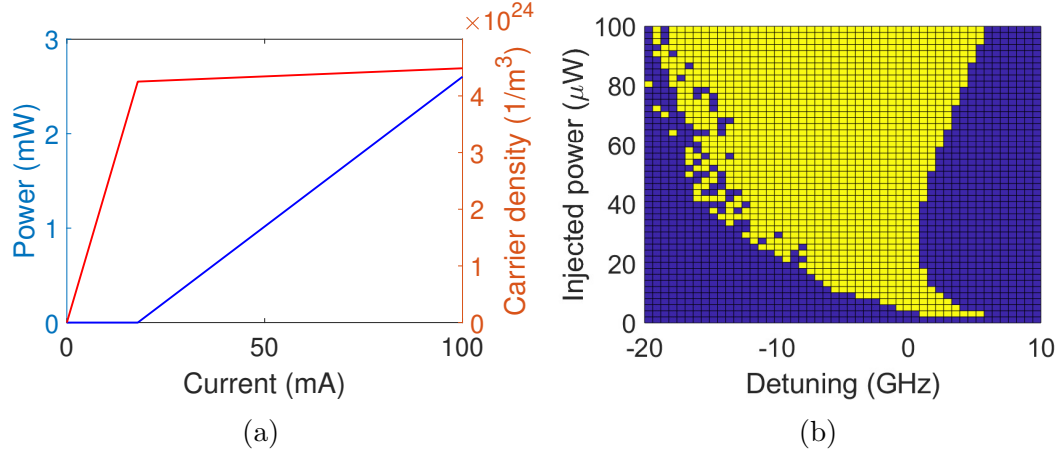


Figure 3.3: (a) Power and carrier density of the noiseless free-running DML as a function of bias current. (b) Colourmap showing the stable region (yellow) where the injected power and frequency detuning result in injection locking for a bias current of 40 mA and no modulating signal.

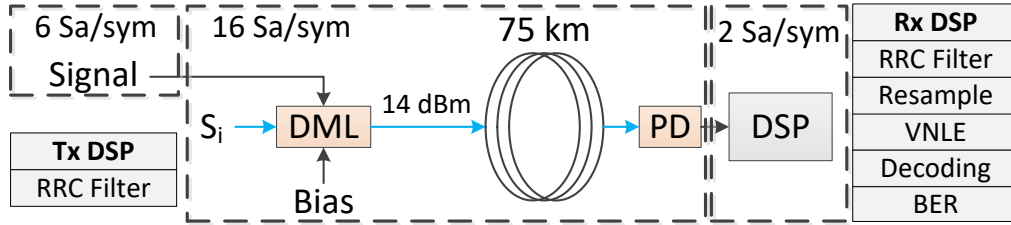


Figure 3.4: Schematic setup of the DML with optical injection simulation. The dashed lines show the samples per symbol used on each part of the simulation.

to 0.2 dB/km. An oversampling rate of 16 samples per symbol was used throughout the simulation to mimic the analogue behaviour of the physical signals. Figure 3.4 shows the schematic setup used for the simulation.

The simulation was used to compare the performance of 15 GBaud PAM4 and 10 GBaud PAM8 over 75 km of fibre with and without optical injection. Two pseudorandom multilevel sequences 2^{17} symbols long were used to generate the PAM4 and PAM8 signals, which were used to drive the DML with peak-to-peak currents of 12 mA and 16 mA respectively.

3.3 Results

The performance of the experimental system was studied, for transmission distances of 0 km, 25 km, 50 km and 75 km, in terms of the bit error ratio (BER) obtained by

counting errors over several captured sequences. In order to overcome the nonlinear distortions coming from the complex dynamics of the DML, a simplified second-order Volterra nonlinear equaliser (VNLE) was used. The number of linear taps L_1 was set to 21, the number after which the performance was found to stabilise. To achieve optimal performance at a minimal computational complexity, the number of nonlinear taps (L_2) required was obtained by analysing the performance of the PAM4 experiment at an ROP of -6 dBm. As shown in figure 3.5(a), the performance reaches a floor at 8 nonlinear taps, after which the performance gain becomes marginal. Given this optimal tap length, the computational complexity was further

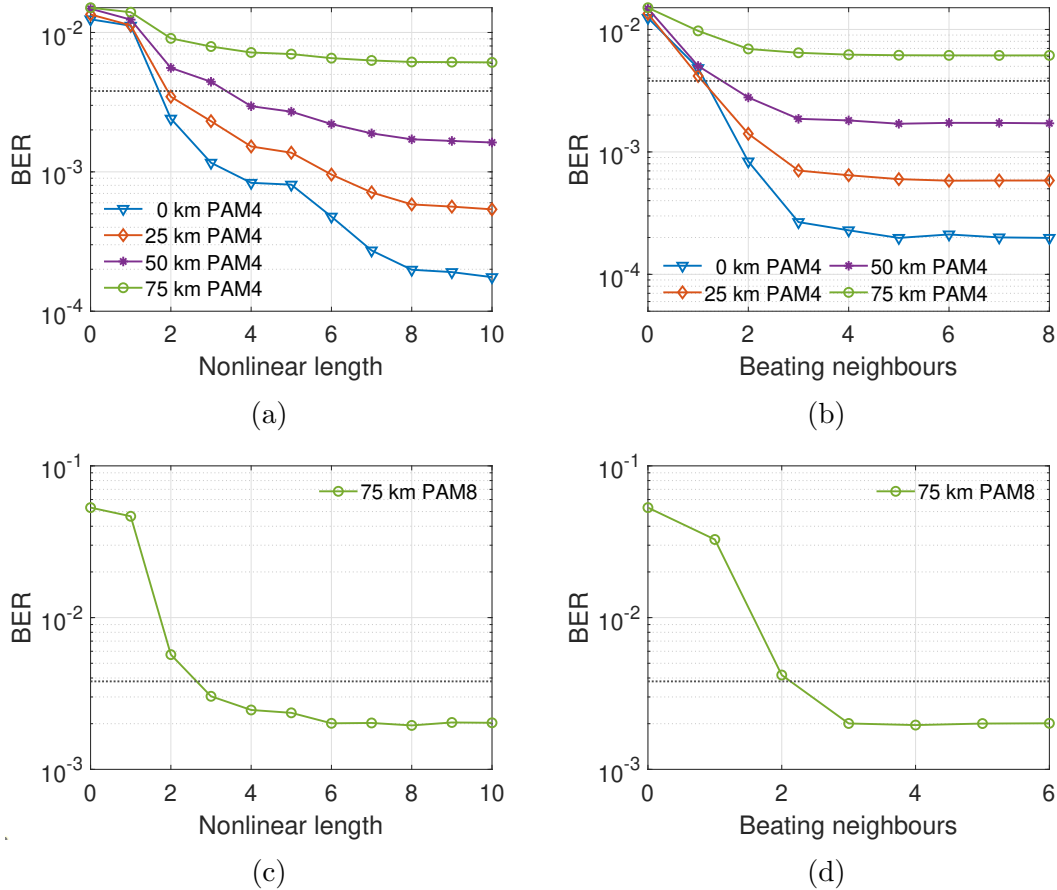


Figure 3.5: (a) BER vs. nonlinear length after different transmission distances from back-to-back (B2B) up to 75 km for PAM4 transmission at ROP -6 dBm. (b) BER vs beating neighbours after different transmission distances from B2B up to 75 km for PAM4 transmission with 8 nonlinear taps at ROP -6 dBm. (c) BER vs. nonlinear length after 75 km for PAM8 transmission at ROP -2 dBm. (d) BER vs beating neighbours after 75 km for PAM8 transmission with 8 nonlinear taps at ROP -2 dBm.

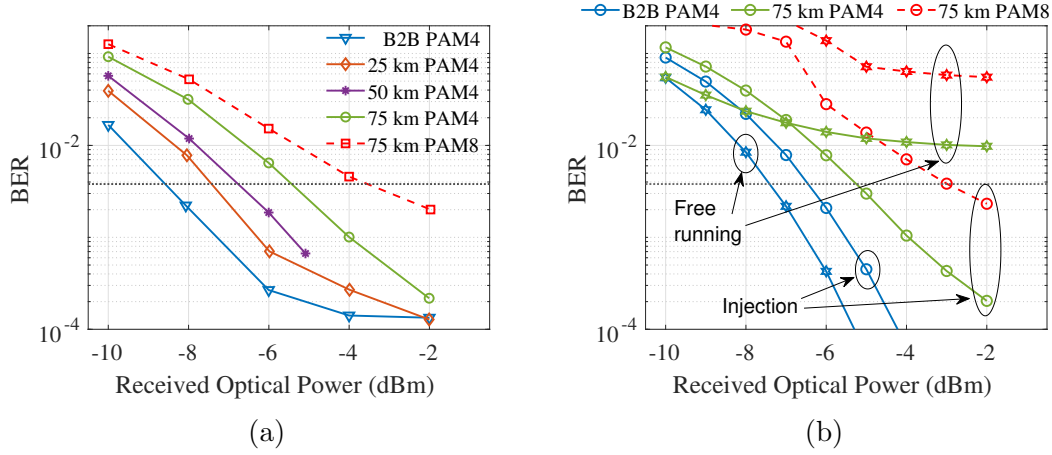


Figure 3.6: (a) BER vs. ROP after different transmission distances for PAM4 (solid lines) and PAM8 (dashed line) transmission experiments. (b) BER vs. ROP for transmission simulation with optical injection (circles) and free-running slave laser (stars) for PAM4 (solid lines) and PAM8 (dashed lines).

reduced by limiting the number of beating neighbours of the nonlinear multipliers (as explained in section 2.2). Figure 3.5(b) shows the BER obtained for each transmission scenario with a VNLE as the number of beating neighbours is increased from 0 (linear equalization) to its maximum value (the full 8 taps of the VNLE). It can be observed how the performance improvements after 3 beating terms are marginal in all cases, so a value of 3 was used (resulting in a complexity reduction of the nonlinear operations in terms of the number of real multiplications of over 55% compared to the full VNLE). A similar analysis of the required nonlinear length and beating neighbours was performed for the amplified 75 km PAM8 transmission at an ROP of -2 dBm, resulting in an optimal number of 6 nonlinear taps (Figure 3.5(c)) and 3 beating neighbours (Figure 3.5(d)) (corresponding to a complexity reduction of over 42%).

As shown in figure 3.6(a), BERs below the hard-decision forward error correction (HD-FEC) limit 3.8×10^{-3} are achieved at ROPs as low as -8.5 dBm for back-to-back (B2B) PAM4 transmission, while ROPs of around -7.5 dBm, -6.7 dBm and -5.5 dBm are required after PAM4 transmission distances of 25 km, 50 km and 75 km respectively. An error floor can be seen for the B2B and 25 km transmission at BERs around 10^{-4} , while the 50 km transmission results are shown up to a power

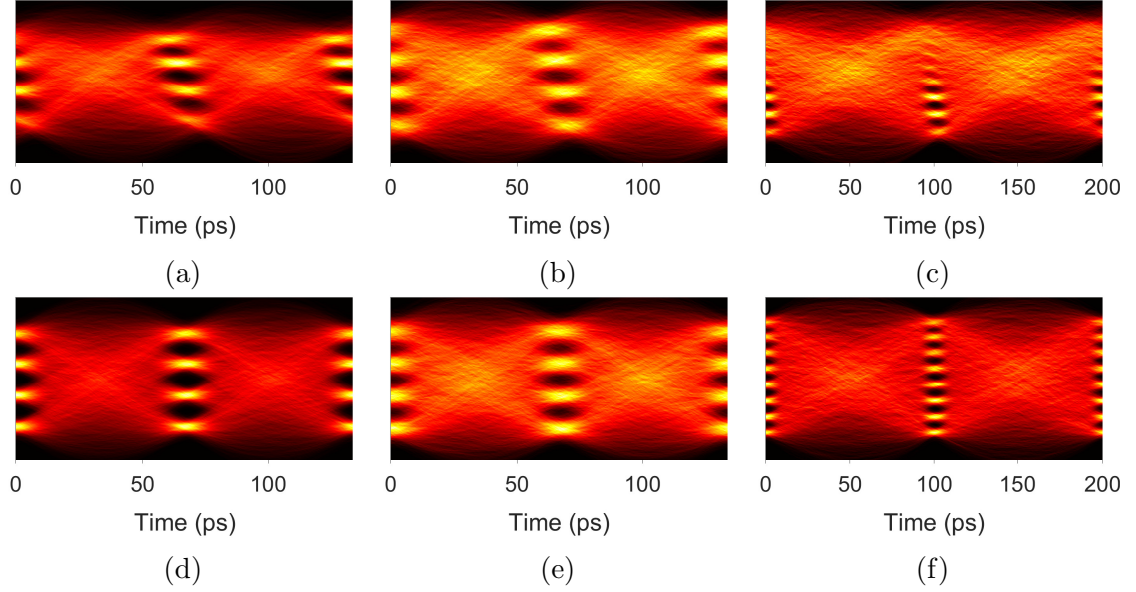


Figure 3.7: Experimental eye diagrams at the receiver for: (a) PAM4 B2B, feed-forward equaliser (FFE), (b) PAM4 75 km, FFE, (c) PAM8 75 km, FFE, (d) PAM4 B2B, VNLE, (e) PAM4 75 km, VNLE, (f) PAM8 75 km, VNLE. To account for the SNR requirements of each modulation format, PAM4 eye diagrams are shown at an ROP of -6 dBm, while PAM8 eye diagrams are shown at an ROP of -2 dBm.

of -5 dBm due to power limitations of the unamplified system. A power penalty of 2 dB at the HD-FEC limit can be seen for the 75 km PAM8 transmission compared to the 75 km PAM4 results, suggesting that noise and nonlinear distortions are a more critical impairment in the system than chromatic dispersion, even after 75 km of SMF. The trend of the 75 km PAM4 transmission results suggests that it may outperform the B2B experiment at high ROPs. This may be explained considering two different factors: First, the OBPF, introduced to remove ASE coming from the EDFA, also removes out-of-band ASE produced by the master and slave sections of the DML. Second, the lower driving current into the DML produces a more linear response with a lower extinction ratio, which degrades the performance at low ROPs, when the system is in the noise-limited region, but improves results at high powers where the nonlinear distortions become the main limiting impairment.

Figure 3.7 shows the received eye diagrams for some of the experiments performed. Figure 3.7(a,d) and figure 3.7(b,e) show the PAM4 eye diagrams at an ROP of -6 dBm for B2B and 75 km transmission respectively, while figure 3.7(c,f) show the PAM8 eye diagrams at a ROP of -2 dBm. Figure 3.7(a,d) show the eye

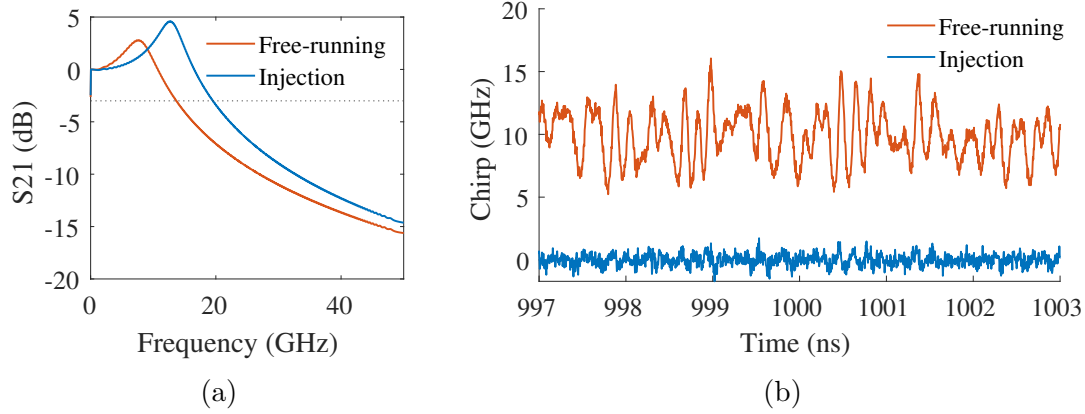


Figure 3.8: (a) Modulation response of the simulated laser using optical injection and free-running SL. (b) Simulated frequency chirp as a function of time for a 15 GBaud PAM4 signal using optical injection and free-running SL.

diagram on the PAM4 B2B case with a 21-tap feed-forward equaliser (FFE) and with the simplified VNLE respectively. A strong eye-skew coming from the nonlinear laser dynamic behaviour can be seen when the linear equaliser is used, which is properly compensated by the nonlinear equalisation implemented. Figure 3.7(b,e) show the eye diagrams after 75 km transmission with FFE and VNLE respectively. A clear eye-opening degradation can be observed due to the ASE noise added by the EDFA and additional fibre transmission, and eye-skew can be seen for the linear equaliser case. Figure 3.7(c,f) show the eye diagrams for PAM8 transmission using the FFE and the VNLE respectively. It can be seen how the nonlinear distortions from the system strongly degrade the upper levels when only linear equalisation is applied, while the VNLE is capable of recovering the full signal and achieving performance below the HD-FEC limit at an ROP of -2 dBm.

As previously mentioned, the simulation platform was developed to examine the impact of optical injection on the tested transmission scenarios. Figure 3.6(b) shows the BER vs. ROP results obtained by simulation for 15 GBaud PAM4 (B2B and 75 km transmission) and PAM8 (75 km transmission) after applying a simplified VNLE. Stars show the performance of a free-running SL and circles show the performance with injection locking. A performance improvement can be seen for the free-running case for the B2B transmission due to a higher extinction ratio, which is attributed to the increased modulation response without injection at low

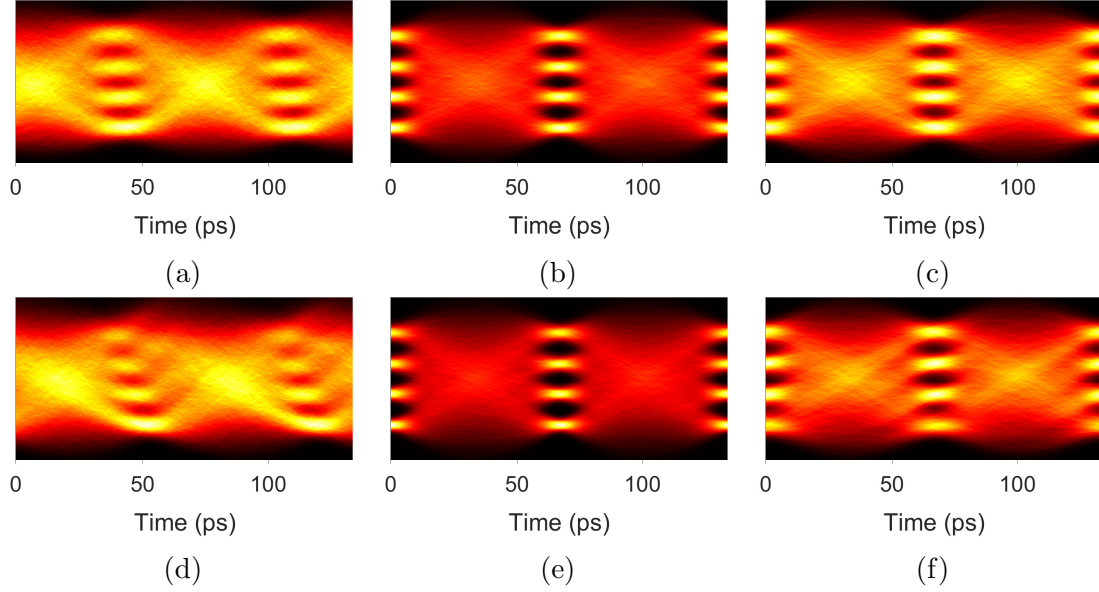


Figure 3.9: Simulated eye diagrams: (a) PAM4 with optical injection, transmitted analogue signal, (b) PAM4 B2B with optical injection, VNLE, (c) PAM4 75 km with optical injection, VNLE, (d) PAM4 free-running, transmitted analogue signal, (e) PAM4 B2B free-running, VNLE, (f) PAM4 75 km free-running, VNLE. The equalized eye diagrams are shown at an ROP of -6 dBm.

frequencies (we note that the external injection increases the relaxation oscillation frequency of the laser, which results in a greater overall modulation bandwidth but also produces a slight decrease in response around the free-running relaxation oscillation frequency in the 2 GHz to 10 GHz range, as illustrated in figure 3.8(a)). The lower error floor observed compared to the experiment can be attributed to different noise levels, which dominate the performance at high powers. After 75 km of fibre, the strong distortions induced by the interaction between frequency chirp and fibre dispersion result in significant performance degradation without external injection, while in the injected case good performance is achieved over the relatively long transmission distance both for the PAM4 and the PAM8 simulations as the external injection substantially reduces the level of frequency chirp. The difference in modulation-induced chirp that enables this is illustrated in Fig. 3.8(b), where the frequency chirp is shown for the optical injection case (blue) and the free-running case (red), showing a significant decrease in chirp when injection locking is used. Figure 3.9 shows eye diagrams generated using the system simulation described. Figure 3.9(a,d) present the analogue PAM4 eye diagrams at the transmitter for the

optical injection and free-running case respectively. A more complex dynamic behaviour and a noticeable eye-skew can be seen for the latter, and poor eye-opening due to inter-symbol interference (ISI) coming from limited device bandwidth is observed in both cases. Figure 3.9(b,e) show the B2B eye diagram at an ROP of -6 dBm after VNLE for the injected and free-running case respectively, where the increased extinction ratio of the free-running case results in a better eye diagram once the VNLE compensates the ISI and nonlinear distortions induced by the laser. Figure 3.9(c,f) show the equalised eye diagrams after 75 km transmission at an ROP of -6 dBm for the injected and free-running case respectively. A clear degradation in eye-opening can be seen in the free-running case due to complex chirp-induced distortions that the VNLE was not capable of compensating for.

3.4 Conclusion

We have shown the characterisation results of a multisection laser that can be used for directly modulated short-reach IMDD systems, with RIN levels of around -150 dB/Hz and an optical linewidth of 2 MHz. We have successfully demonstrated an unamplified 30 Gbit/s PAM4 transmission over up to 50 km of SMF in the C-band using a packaged, wavelength tunable, DML PIC employing optical injection. BER values below the HD-FEC threshold were obtained for ROPs as low as -6.5 dBm by using a low complexity VNLE operating at 1 sample per symbol. EDFA amplified transmission over 75 km of SMF was performed, obtaining below-FEC performance at -5.5 dBm and -3.5 dBm when using 15 GBaud PAM4 and 10 GBaud PAM8 respectively. The simplified VNLE employed in this work is shown to be capable of greatly improving the received signal quality, even after strong degradation from eye-skew effects and power-dependent nonlinearities. We have demonstrated via simulation that the aforementioned results were enabled by the external optical injection technology, which has proven to be suitable for high-speed transmitters capable of operating over longer distances. Improvements in PIC fabrication and packaging that increase the device bandwidth to around 15 GHz, should enable 50

Gbit/s transmission over tens of km of SMF. While the power budgets of up to 15 dB obtained with this device are low for optical access network applications, these results are still relevant in the context of optical fronthauling for radio access networks, as they can fulfil the data requirements imposed by the current trends without the added complexity of external modulators and optical amplifiers while allowing the potential implementation of low-cost WDM systems [22].

In the next chapter, we will introduce flexible PAM formats as a way of increasing the throughput granularity of IMDD systems. We will demonstrate the technique on a short-reach transmission experiment focused on intra-data centre links using a novel, fully integrated laser with ultra-low noise and wide tunability.

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Chapter 4

Flexible Pulse Amplitude Modulation Formats

Today's optical systems continue transitioning to higher data rates and increased levels of flexibility and reconfigurability. While advances in photonic integration have been driving significant reductions in cost, complexity and power consumption of coherent transmission systems, enabling their cost-effective use in ever-shorter links, it is still expected for intensity modulation with direct detection (IMDD) to remain the main technology for short-reach intra-data centre transmission in the medium future. However, as systems transition to higher baud rates to keep up with the rapid growth in demand [1], the bandwidth requirements become more difficult to meet with a directly modulated laser (DML), and the effect of chromatic dispersion cannot be neglected even after short fibre distances. In that context, the use of IMDD technology with external modulation provides a middle-ground between direct modulation and coherent transmission that enables low-cost and energy-efficient IMDD transmission at high baud rates and with advanced pulse amplitude modulation (PAM) formats.

Works on IMDD transmissions at rates beyond 100 Gbaud [2; 3] demonstrate that it is feasible to obtain single-lane data rates above 200 Gbit/s. However, the transition towards a wavelength division multiplexing (WDM) implementation, in

combination with some degree of optical switching (implemented using optical filters and multiplexers) [4; 5; 6], can result in a drastic variation in path loss present in individual server-to-server links in the data centre. This in turn leads to different achievable link data rates, which can be accommodated by operating at a constant baud rate with different modulation formats. With traditional PAM formats, where the number of levels is a power of two, the signal-to-noise ratio (SNR) requirements grow sharply as the number of levels increases from 4 to 8 and 16, limiting potential capacity increases. This constraint can be alleviated by using flexible PAM (FlexPAM) formats [7], in which a fractional number of bits is encoded in each symbol by using a non-power of two number of levels (e.g. PAM6). The use of FlexPAM enables finer granularity in the modulation format that gives the maximum throughput for a specific system SNR (which may change due to temperature variations that alter the laser output power and receiver noise, and different optical paths as optical switching is implemented in future data centre architectures) at the very modest computational cost of a lookup table.

Some examples of non-power of two PAM levels can be found in the literature: Ingham et al. [8] proposed the use of PAM3 and compared its performance with on-off keying (OOK) and PAM4 in a simulation study, while Zhou et al. [7; 9] recently studied the theoretical performance of the different PAM levels ranging from PAM4 to PAM8 in the context of intra-data centre links. Wei et al. [10; 11; 12] demonstrated an experimental comparison between PAM4, PAM6, and PAM8, showing transmission rates over 200 Gbit/s but relying on the use of complex digital signal processing (DSP) techniques such as a Volterra nonlinear equaliser (VNLE), which may not be suitable for low-power and low-complexity intra-data centre links.

This chapter presents the simulation and an experimental demonstration of a short-reach optical transmission system using FlexPAM formats ranging from OOK to PAM32 published in [13]. At the transmitter, a photonic integrated dual laser module based on InP semiconductor optical amplifiers (SOA) and low-loss Si_3N_4 feedback circuits is used as an optical source. 20 GBaud transmission over 1 km

of single-mode fibre (SMF) was demonstrated experimentally for 11 different PAM formats up to PAM16, achieving a bit error ratio (BER) below 3.8×10^{-3} for 10 formats, including PAM16. This work harnesses the properties of the laser described in [14] to enable experimental FlexPAM transmission demonstrations and performance analyses aimed at next-generation flexible data centre environments. In addition, the capabilities of the developed system are further explored through numerical simulations which, under the same conditions and baud rate as the experiment, closely match the experimentally measured results, and show that with an increase of 7 dB on the electrical signal SNR, the system is capable of supporting PAM32. The simulation also shows that with a 5 dB increase in SNR beyond that used in the experimental work and only a moderate increase in transmitter bandwidth, 53.5 GBaud transmission is feasible for modulation formats up to PAM16, reaching data rates as high as 214 Gbit/s on a single wavelength. Additionally, based on the work published in [15], probabilistic shaping (PS) is presented as an alternative to FlexPAM to enable fine transmission rate granularity. However, unlike FlexPAM, it does not change the number of levels of the signal, making it unsuitable for unamplified systems [16]. In systems where optical amplification can be beneficial due to high power budget requirements, such as passive optical networks (PON), the use of SOAs can provide more cost-effective and energy-efficient solutions than erbium-doped fibre amplifiers (EDFA) at the expense of nonlinear distortions of the signal. SOA nonlinearities get worse for big excursions of the signal power from its average, and minimising the probability of the outer symbols (via symmetric PS) can reduce their occurrence and boost the system's performance. Section 4.1 introduces the integrated dual laser module used for the experiment. In section 4.2, the symbol-to-bit mapping used for the non-standard PAM formats is detailed. Section 4.3 explains the experimental and simulation setup, and section 4.4 provides an overview and analysis of the results. Finally, section 4.5 presents an experimental demonstration of PS-PAM8 for optical transmission in systems with SOA-induced nonlinearity.

4.1 Integrated Dual Laser Module

The integrated tunable dual laser module (DLM) (Figure 4.1) used for the experiment contains two lasers, each one consisting of an InP gain section and Si_3N_4 phase and cavity sections. The thermo-optically tunable phase sections and the feedback circuits, consisting of two micro-ring resonators (MRR) in a Vernier configuration, create an MRR-based external cavity laser, allowing the module to cover a wavelength range of over 100 nm, covering the full C band, most of the S band, and part of the L band. Low-loss Si_3N_4 waveguides enable the introduction of long photon lifetimes into the laser resonator, which results in a significant increase of coherence properties, such as spectral coverage and output power with low intensity noise [17]. This device shows values of relative intensity noise (RIN) as low as -165 dB/Hz [18] over the whole spectrum (Figure 4.2), which has enabled us to demonstrate IMDD operation up to PAM16, and a high side mode suppression ratio (SMSR) of more than 50 dB. To achieve these conditions over different biases, the different sections of the laser need to be all tuned accordingly. This is unlike a conventional single-section laser where simply increasing the bias current will result in a decrease in the

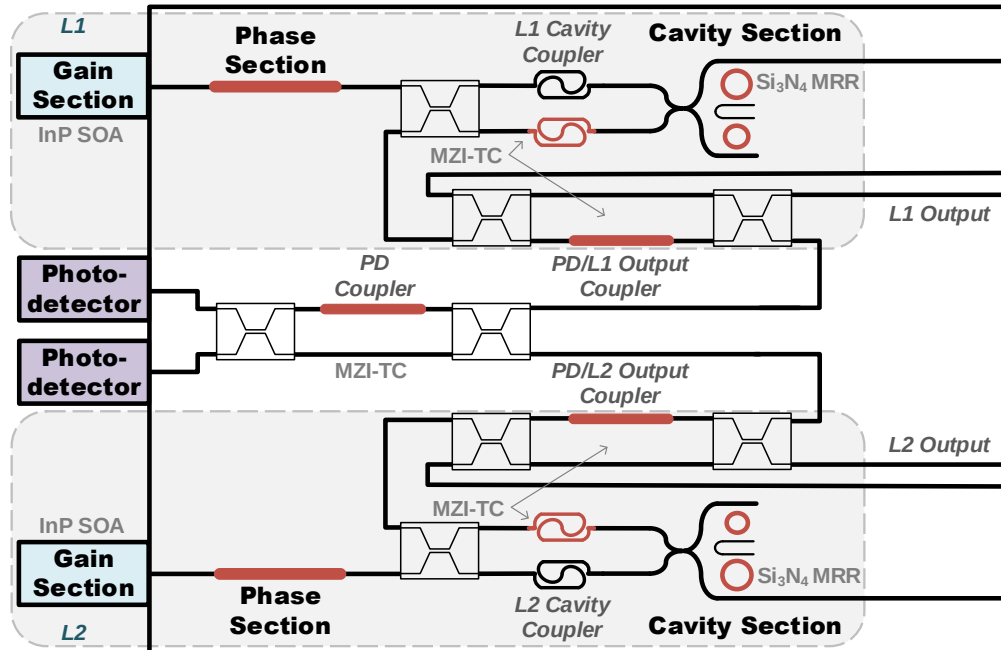


Figure 4.1: Schematic of the DLM.

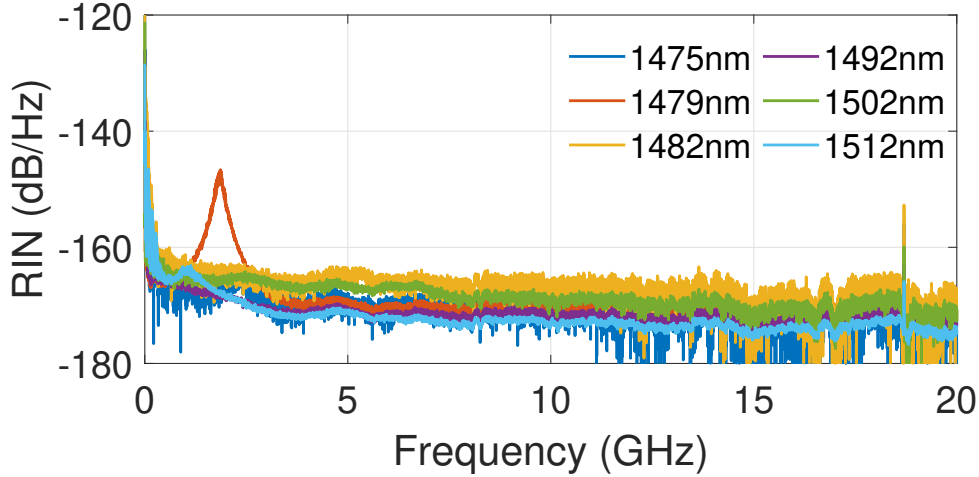


Figure 4.2: Measured RIN at different sample wavelengths for the tunable laser module with the gain section biased at 170 mA.

RIN. The advantages of extremely low RIN (well below the typical values of -140 dB/Hz for other available integrated widely tunable sources [19; 20]) over such a wide range, combined with a small footprint, make the dual tunable laser module a very compelling option for spectrally efficient WDM systems based on IMDD technology. A more detailed characterisation of the device can be found in [14]. For this work, we used a single laser output from the dual laser module, with the wavelength set to 1552.7 nm, an SMSR of more than 50 dB and an output power of 9.7 dBm.

4.2 Symbol-to-Bit Mapping

For the standard power of two PAM formats, we used Gray encoding to map each symbol to the corresponding bits, i.e. the bits were assigned so that the difference between any two adjacent symbols is always one bit. The non-standard PAM levels are encoded by grouping them in packets of symbols. Table 4.1 shows the maximum achievable bits per symbol ($\log_2 M$) of each PAM level, the number of symbols used for encoding (N_{Sym}), the bits per symbol (bit/Sym) obtained, and the total data rate in Gbit/s of the link when operated at 20 GBaud and at 53.5 GBaud (DR@20 and DR@53.5 respectively). We defined two different symbol-to-bit mapping techniques: The first one, which we will call asymmetric encoding, is realised by sorting all pos-

Table 4.1: Theoretical and real number of bits per symbol obtained for each PAM level.

M	$\log_2 M$	N_{Sym}	bit/Sym	DR@20 (Gbit/s)	DR@53.5 (Gbit/s)
2	1	1	1	20	53.5
3	1.59	2	1.5	30	80.25
4	2	1	2	40	107
5	2.32	4	2.25	45	120.375
6	2.59	2	2.5	50	133.75
7	2.81	4	2.75	55	147.125
8	3	1	3	60	160.5
10	3.32	4	3.25	65	173.875
12	3.59	2	3.5	70	187.25
14	3.81	4	3.75	75	200.625
16	4	1	4	80	214

sible N_{Sym} symbol combinations in terms of their average power, with bits assigned in ascending order. The bottom constellation in figure 4.3(a) shows the mapping of symbols to bits of a PAM3 signal using asymmetric encoding. The 2-symbol combination corresponding to two consecutive high power symbols is not used. PAM6 and PAM12 are generated the same way, with the symbol combinations arranged in a 2-D grid and discarding the symbols in the upper right corner (corresponding to the higher average power). PAM5, PAM7, PAM10 and PAM14 mapping follow the same principle on the corresponding 4-symbol combinations. This method ensures a lower density of high power levels, potentially improving performance in systems with high levels of RIN or other amplitude-dependent noise by reducing the occurrence probability of the higher power symbols most affected by noise. Alternatively, instead of reducing the density of the upper levels on the non-standard PAM formats, the mapping can be performed by removing the sharpest transitions from minimum to maximum (and vice versa), effectively reducing the density of upper and lower levels symmetrically. The constellation on top of figure 4.3(a) shows the mapping of symbols to bits of a PAM3 signal using symmetric encoding, where the 2-symbol

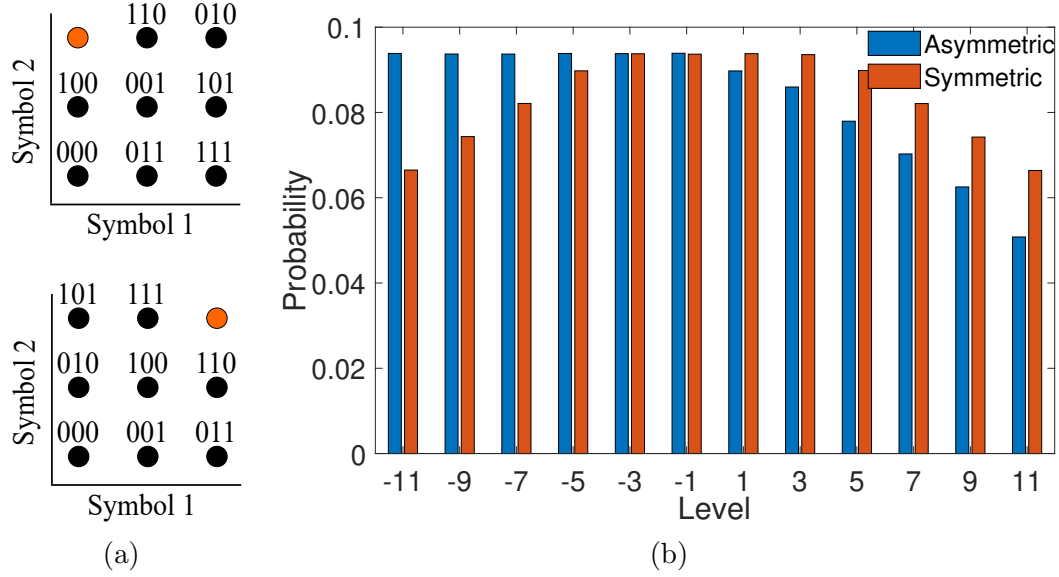


Figure 4.3: (a) Symmetric (top) and asymmetric (bottom) PAM3 symbol-to-bit mapping. (b) Probability per symbol of a PAM12 signal using symmetric (red) and asymmetric (blue) encoding.

combination corresponding to a low power symbol followed by a high power symbol is not used. Figure 4.3(b) shows an example of the probability distribution of each level in a PAM12 sequence, when reducing the density of higher levels (asymmetric) and when removing the sharp transitions (symmetric), and table 4.2 shows the symbol-to-bit mapping of the symmetric and asymmetric PAM6.

The use of non-standard PAM formats and the proposed mapping techniques have two main drawbacks. First, the use of non-Gray coding increases the average number of bit errors per symbol wrongly decoded. Second, as the modulation format increases, the complexity of the bit mapping increases rapidly (especially with a high number of encoding symbols). This makes some of the modulation formats less appealing for their use in intra-data centre links due to an increased latency at the decoding step (e.g., PAM14 requires the mapping of 15 bits into $14^4=38416$ symbol combinations).

4.3 Experimental Setup

The experiment was performed using the setup shown in figure 4.4(a). A pseudo-random multilevel sequence of 2^{15} symbols was generated offline for each PAM level

Table 4.2: Symbol-to-bit mapping for the symmetric and asymmetric PAM6.

Sym₁	Sym₂	Symm	Asymm		Sym₁	Sym₂	Symm	Asymm
1	1	00000	00000		4	1	11000	00110
1	2	00111	00010		4	2	10001	01011
1	3	10010	00101		4	3	01010	10001
1	4	11011	01001		4	4	00011	10111
1	5	-	01110		4	5	01101	11100
1	6	-	10100		4	6	10111	-
2	1	00110	00001		5	1	11110	01010
2	2	00001	00100		5	2	11001	10000
2	3	01001	01000		5	3	10011	10110
2	4	10100	01101		5	4	01100	11011
2	5	11100	10011		5	5	00100	11111
2	6	-	11000		5	6	01111	-
3	1	10000	00011		6	1	-	01111
3	2	01000	00111		6	2	11111	10101
3	3	00010	01100		6	3	11010	11010
3	4	01011	10010		6	4	10101	11110
3	5	10110	11000		6	5	01110	-
3	6	11101	11101		6	6	00101	-

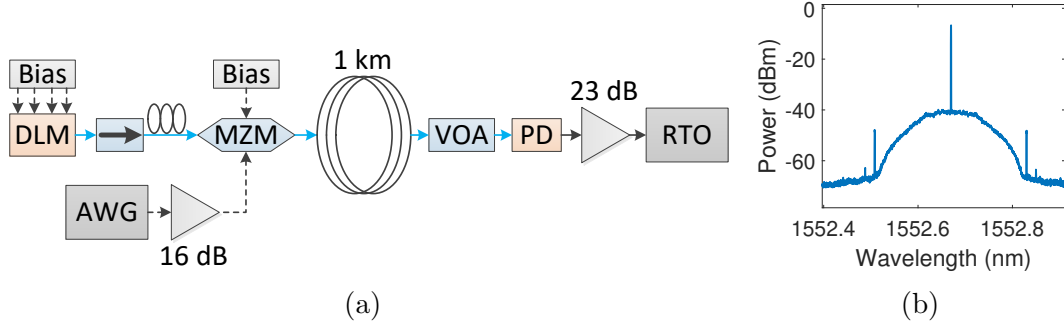


Figure 4.4: (a) Experimental setup. (b) Optical spectrum of a 20 GBaud PAM8 signal before the PD.

using MATLAB, and pre-distorted to compensate for the nonlinear transfer function of the LiNbO₃ Mach-Zehnder modulator (MZM) from Sumitomo (40 Gbit/s MZM). Even though a discrete MZM was used for this experiment, integrated technologies currently available [21] can enable the use of fully integrated transmitter solutions in short-reach links. Asymmetric symbol-to-bit mapping was used for the non-standard PAM signals. The 20 Gbaud signals were generated using an arbitrary waveform generator (AWG) (3-dB bandwidth of 33 GHz) operating at 90 GSa/s. A 16 dB gain electrical amplifier (3-dB bandwidth of 25 GHz) was placed before the MZM to match its V_π of ~ 2.3 V. An isolator and a polarisation controller (PC) were placed between the laser and the MZM. After transmission over 1 km of SMF, a VOA was used to sweep the received optical power (ROP) falling on the photodiode (PD) (3-dB bandwidth of 30 GHz). A 23 dB gain electrical amplifier (3-dB bandwidth of 55 GHz) was used before the signal was captured for offline processing with a real-time oscilloscope (RTO) (3-dB bandwidth of 110 GHz) operating at a sampling rate of 256 GSa/s. Figure 4.4(b) shows the optical spectrum of the signal captured before the PD.

The DSP performed on the received signals consisted of matched filtering with a 20 GHz root-raised cosine RRC filter, resampling to 2 samples per symbol, synchronisation, adaptive equalisation, and decoding. The normalisation of the non-power of two PAM formats was performed including an offset in the average value to account for the non-uniform probability distribution of the symbols. A decision-directed

least-mean squares algorithm was used for updating the weights of the 13-tap feed-forward equaliser (FFE). The BER was calculated by counting errors over several captured sequences.

4.3.1 Simulation Setup

A simulation was developed in Matlab for modelling the full experimental system. To compare its performance with the measurements from the experiment, the 3-dB simulation bandwidth value of all electrical and optoelectronic components was set to the corresponding experimental value. The laser's RIN was set to -160 dB/Hz and modelled as additive white Gaussian noise, while the linewidth was set to 100 kHz and modelled as a Wiener process. The sampling frequencies of AWG and RTO were set to 90 GSa/s and 256 GSa/s respectively. The SNR of the electrical signal driving the MZM was set to 27 dB (integrating the noise over the whole bandwidth of the AWG), the value at which the simulation results closely match those observed from the experiment. This SNR accounts for both the noise of the AWG and the electrical amplifier, as well as any degradation produced by nonlinear distortions, quantisation noise and other interactions. An oversampling rate of 16 samples per symbol was used to mimic the analogue electrical and optical signals in the simulation.

The fibre propagation was modelled by solving the Nonlinear Schrödinger Equation (NLSE) using the Split-Step Fourier Method (SSFM) setting a launch power of 0 dBm. The dispersion coefficient of the fibre was set to 16 ps/nm/km, the dispersion slope to 0.08 ps/nm²/km, the nonlinear coefficient to 1.3 1/W/km, and the attenuation to 0.2 dB/km.

The simulation was run for the same conditions as the experiment, i.e. 20 GBaud transmission using asymmetric mapping for the non-standard PAM formats. The simulation was also carried out with a baud rate of 53.5 GBaud and an electrical SNR of 32 dB (5 dB higher than what was used for the base simulation, which may be achieved by using a low drive voltage modulator [21] driven directly by the

AWG). All electrical bandwidths in the system were set to 30 GHz, and a sampling rate of 107 GSa/s was set for both AWG and RTO. The RRC filter bandwidth at the DSP side was also set to 30 GHz. Additionally, a comparison between the symmetric and asymmetric mapping techniques was performed by running all FlexPAM formats with the two different mapping schemes using the same conditions as the experiment.

4.4 Results

The experimental performance of each PAM format in the system was compared in terms of BER vs. ROP. The results after 1 km of SMF (Figure 4.5(a)) achieve performance below the forward error correction (FEC) limit of 3.8×10^{-3} for all modulation formats except PAM14. Figure 4.6 show the eye diagrams corresponding to PAM6, PAM8, PAM12, and PAM16 signals. PAM16 reaches below FEC performance at ROP of -2 dBm, resulting in a receiver sensitivity penalty of 15 dB compared to OOK. Even though lower-order modulation formats are less affected by the SNR of the system, it can be seen that PAM7 performs similarly to PAM8, and PAM14 worse than PAM16. This is a result of the additional penalty introduced by the symbol-to-bit mapping, where standard PAM formats can take advantage of Gray encoding, while non-standard formats suffer from the fact that the encoding is performed in blocks of multiple symbols. In the latter case, a decision error produces a cascading effect in which bits corresponding to the whole block of symbols are affected. This becomes more significant as the number of symbols used for decoding increases, and leads to the previously mentioned behaviour of PAM7 and PAM14, where a 4-symbol format is adjacent to a standard PAM format (PAM8 and PAM16 respectively).

Figure 4.5(b) shows the results of the 20 GBaud simulation with an electrical SNR of 27 dB on the driving currents operating from OOK up to PAM16. The figure shows how the trend of the simulated results closely resembles that of the experimental measurements. However, a difference in performance between experiment and simulation of up to 2 dB at the FEC threshold, can be observed for lower-order

modulation formats at lower ROPs. This difference in performance - which increases as the number of levels decreases from PAM12 down to OOK - can be attributed to the power-dependent noise figure of the electrical amplifier at the receiver for lower ROPs, which was not accounted for in the simulation. Additionally, in order to analyse the system's performance if lower noise AWG and RF amplifiers were available, the value of the SNR was increased. The system was found to support PAM32, with performance below the FEC limit obtained for an enhanced electrical SNR of 34 dB.

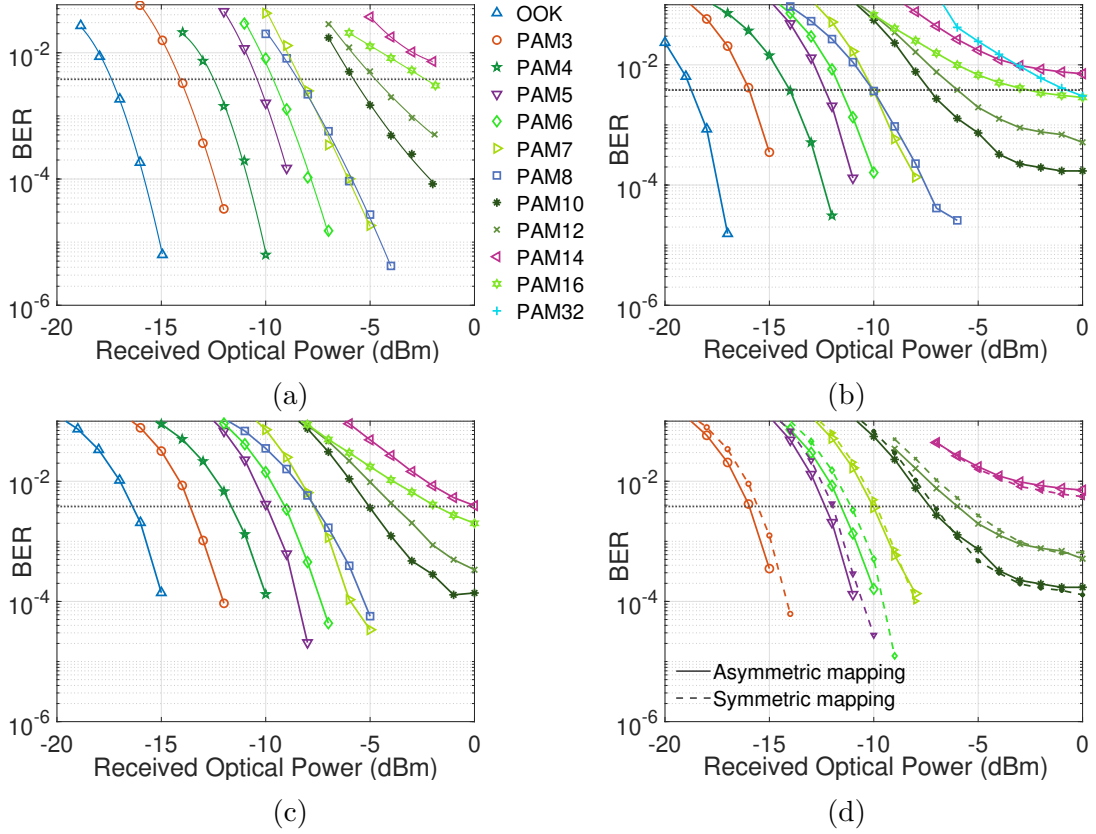


Figure 4.5: (a) Experimental results of BER vs. ROP of 20 GBaud transmission using various modulation formats ranging from OOK to PAM16 after 1 km of SMF. (b) Simulation results of BER vs. ROP of 20 GBaud transmission using various modulation formats ranging from OOK to PAM32 after 1 km of SMF. Electrical SNR of 27 dB from OOK to PAM-16, and 34 dB for PAM32. Asymmetric mapping was used for the non-standard formats. (c) Simulation results of BER vs. ROP of 53.5 GBaud transmission using various modulation formats ranging from OOK to PAM16 after 1 km of SMF. Electrical SNR of 32 dB. Asymmetric mapping was used for the non-standard formats. (d) Asymmetric and symmetric simulation results of BER vs. ROP of 20 GBaud transmission using non-standard PAM formats ranging from PAM3 to PAM14 after 1 km of SMF. Electrical SNR of 27 dB.

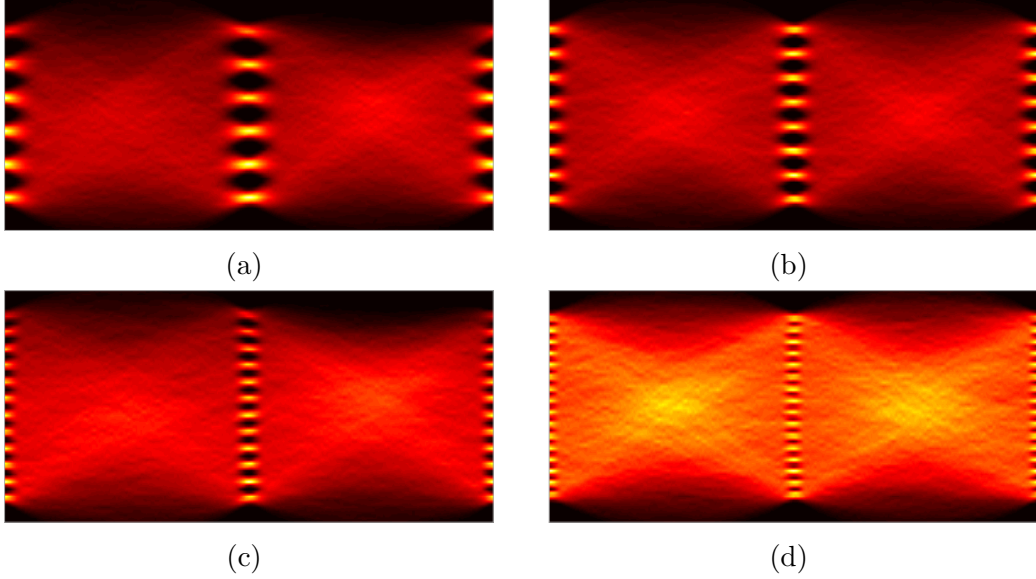


Figure 4.6: (a) PAM6 at ROP -7 dBm, (b) PAM8 at ROP -4 dBm, (c) PAM12 at ROP -2 dBm, (d) PAM16 at ROP -2 dBm. It can be noticed how the upper levels on (a) and (c) are less densely populated due to the asymmetric mapping used.

The results for the 53.5 GBaud simulation can be seen in figure 4.5(c), where performance below FEC was achieved for PAM16 at -2 dBm, providing a total transmission capacity of 214 Gbit/s. A receiver sensitivity penalty of up to 2 dB can be observed at the FEC limit for OOK with respect to the 20 GBaud simulation (OOK curves from figure 4.5(b) and (c)). This penalty is reduced as the modulation format increases, indicating how the choice of modulation format dictates whether system performance is ultimately limited by bandwidth or SNR. The ROP required to achieve below FEC limit performance in the experiment, the 20 GBaud simulation and the 53.5 GBaud simulation, can be seen in figure 4.7(a).

Comparing the performance of symmetric and asymmetric mapping of non-standard PAM formats at 20 GBaud (Figure 4.5(d)), it can be seen how the asymmetric encoding gives a marginal gain at low-order modulation formats, while no noticeable difference is observed at the higher orders. Since higher-order formats are more dependent on the SNR, and the asymmetric mapping is expected to perform better in the presence of increased signal-dependent noise, it can be concluded that the performance gain is small and may be negligible given the level of noise fluctuations in the system.

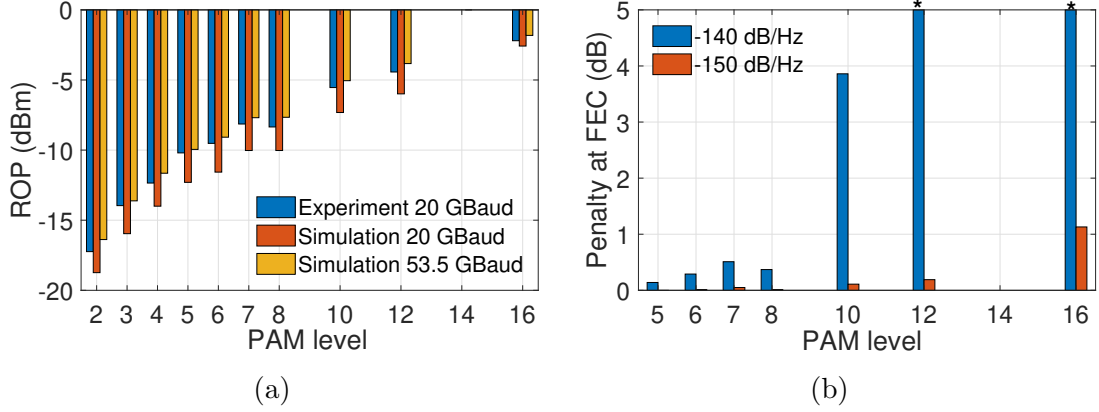


Figure 4.7: (a) ROP required to achieve below FEC performance for each PAM level according to the experimental results, the 20 GBaud simulation and the 53.5 GBaud simulation. Missing values at PAM14 correspond to the system not reaching the FEC limit. (b) Receiver sensitivity penalty at the FEC limit for different PAM levels for the 20 GBaud simulation. Missing values at PAM14 correspond to the system not reaching the FEC limit. (*) PAM-12 and PAM-16 do not reach the FEC limit at a RIN value of -140 dB/Hz.

The effects of RIN on the system's performance were studied by measuring the receiver sensitivity penalty at the FEC limit for the 20 GBaud simulation with a RIN of -140 dB/Hz and -150 dB/Hz with respect to the original results (-160 dB/Hz). As can be seen in figure 4.7(b), a power penalty of over 1 dB was obtained for PAM16 for a RIN value of -150 dB/Hz, but minimal differences can be seen for lower-order formats. At -140 dB/Hz significant degradations can be seen, and only modulation formats below PAM10 reach FEC limit performance.

4.5 PS for SOA-Amplified IMDD Systems

4.5.1 Experimental Setup

Four external cavity lasers (ECL), tuned to 192.65 THz, 192.85 THz, 193.05 THz and 193.25 THz, were operated in continuous wave and modulated with a single MZM with a 3 dB bandwidth of 20 GHz. The 37.5 GBaud PS-PAM8 signal was generated, filtered and predistorted offline using Matlab, and sent to an AWG operating at 84.375 GSa/s. A bipolar Maxwell-Boltzmann probability distribution was generated using a look-up table-based cost-minimising distribution matcher [22]. The output of

the AWG was amplified with a linear RF amplifier and sent into the MZM biased at quadrature. The output of the MZM was sent to an EDFA and a VOA to set the fibre launch power to 13 dBm. The amplified WDM signal was transmitted over 25 km of large effective area fibre (LEAF), after which two different scenarios were considered: (1) a tunable optical filter (TOF) with a suppression ratio of over 40 dB was used to select one of the middle channels, a VOA was used to set the optical power to -8 dBm and an SOA biased at 200 mA was used to amplify the signal to 10 dBm. (2) A VOA was used to set the optical power of the composite WDM signal to -9 dBm, an SOA amplified the signal to a power of 10 dBm, and a TOF was used to select one of the middle channels. In scenario (1), with a single channel being amplified, the main distortions introduced into the signal are power saturation, patterning effect, and self-phase modulation (SPM). However, when all four channels get amplified simultaneously, inter-channel nonlinearities such as cross-gain modulation (XGM), cross-phase modulation (XPM) and four-wave mixing (FWM) become a relevant performance limitation. Inter-channel nonlinearities are worse for the inner channels in a WDM system (as neighbouring channels on both sides surround them), so studying one of the middle channels corresponds to the worst-case scenario. The transmission over 25 km of LEAF results in a walk-off between adjacent channels of around 160 ps, corresponding to 6 symbol periods at 37.5 GBaud. This walk-off, while enough to decorrelate the symbol sequences between adjacent channels, is still in the order of the carrier recovery time, making the inter-channel nonlinearity not fully independent. The obtained channel from scenarios (1) and (2) was transmitted over 1 km of SMF and then a VOA was used to control the optical power falling into a 70 GHz photodiode (PD). The obtained photocurrent was amplified with a 23 dB gain linear amplifier and captured at 100 GSa/s with an RTO for offline processing. The schematic representation of the experimental setup is shown in figure 4.8.

The transmission experiments presented here show the performance of PS in SOA-amplified PONs. The use of LEAF, as opposed to the SMF commonly used in newly deployed short-reach links, was based on fibre availability at the moment

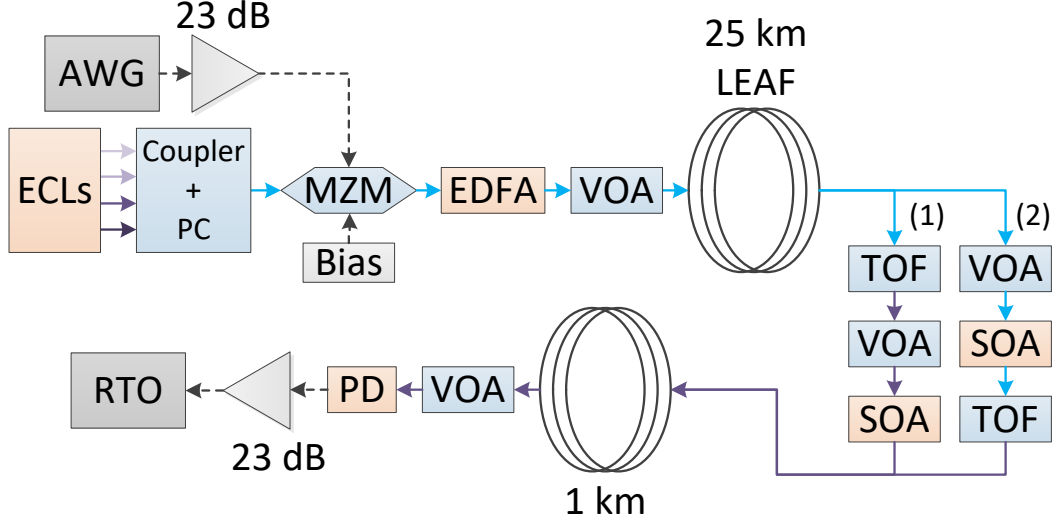


Figure 4.8: Experimental setup for the WDM PS-PAM8 demonstration.

of the experiment. The lower dispersion parameter of LEAF does not undermine the relevance of the experiment, as the amount of accumulated dispersion is still relevant considering optical network units (ONU) at a shorter distance to the optical line terminal (OLT), or in the context of O-band transmission. Additionally, the symmetric PS used here does not affect the impairments induced by the EDFA amplification and the fibre transmission, as the average power of the PAM signal remains unchanged, making the results easy to extrapolate to SOA-amplified intra-data centre links with shorter fibre distances.

4.5.2 Results

The entropy of the PS-PAM8 signal was set to 2.68 bit/symbol, giving a gross data rate of 100.5 Gbit/s per wavelength. Additionally, a 33.5 GBaud PAM-8 signal, generated at a sampling rate of 92.125 GSa/s, was used so that the performances of unshaped and PS-PAM8 under the same data rate could be compared. A simplified Volterra nonlinear equaliser operating at 1 sample per symbol was used to recover the signals. A nonlinear memory length of 15 taps, found to be optimal for scenario (1), was used for both scenarios. The number of beating neighbours was limited to 7. With this nonlinear memory length, the equaliser covers a time window of almost ± 200 ps, making it suitable for correcting intra-channel nonlinearities. However,

given the walk-off between channels, a small amount of inter-channel nonlinearities are also corrected by the equaliser in scenario (2). This is a result of the specific configuration of the experiment presented here, and would not happen on a real system with an independent data stream on each wavelength. However, it still allows us to make observations on the relative performance with and without PS. The BER results as a function of ROP for scenario (1) are shown in figure 4.9(a), where receiver sensitivity gains of more than 2 dB were obtained at the FEC limit by using PS. Figure 4.9(b) shows the eye diagrams at an ROP of -3 dBm for standard PAM8 (top) and PS-PAM8 (bottom). A total power budget of 30.25 dB was obtained accounting for the boost from the SOA. Figure 4.9(c) shows the BER results for scenario (2). In this scenario, with XPM, XGM and FWM also playing a role, higher values of BER are obtained for both standard PAM8 and PS-PAM8, with the standard PAM8 exhibiting an error floor at around 1.2×10^{-2} . A receiver sensitivity penalty of 2 dB with respect to scenario (1) was obtained for the PS-PAM8 transmission due to increased interchannel nonlinearity, reaching the FEC limit at -3 dBm ROP. Figure 4.9(d) shows the eye diagrams for scenario (2) at -3 dBm ROP for standard PAM8 (top) and PS-PAM8 (bottom), showing in both cases significantly less eye-opening than in scenario (1). The power budget obtained for the WDM amplification case was 28.75 dB.

Using a 37.5 GBaud signal, as opposed to 33 GBaud, increases the in-band noise and bandwidth requirements of the system, but also slightly reduces the symbol period, potentially lowering the nonlinear distortions. To test the influence of these factors, scenario (2) was tested using standard PAM8 at 37.5 GBaud and an ROP of -3 dBm, obtaining a similar BER value as for 33 GBaud. This confirms that the performance improvements come from the use of PS, and the mitigation of nonlinear distortions due to the reduced probability of the outer symbols of the PAM8 constellation (which deviate further from the average signal power).

While the normalised generalised mutual information characterises the system's performance in probabilistically shaped signals better than the BER, the results

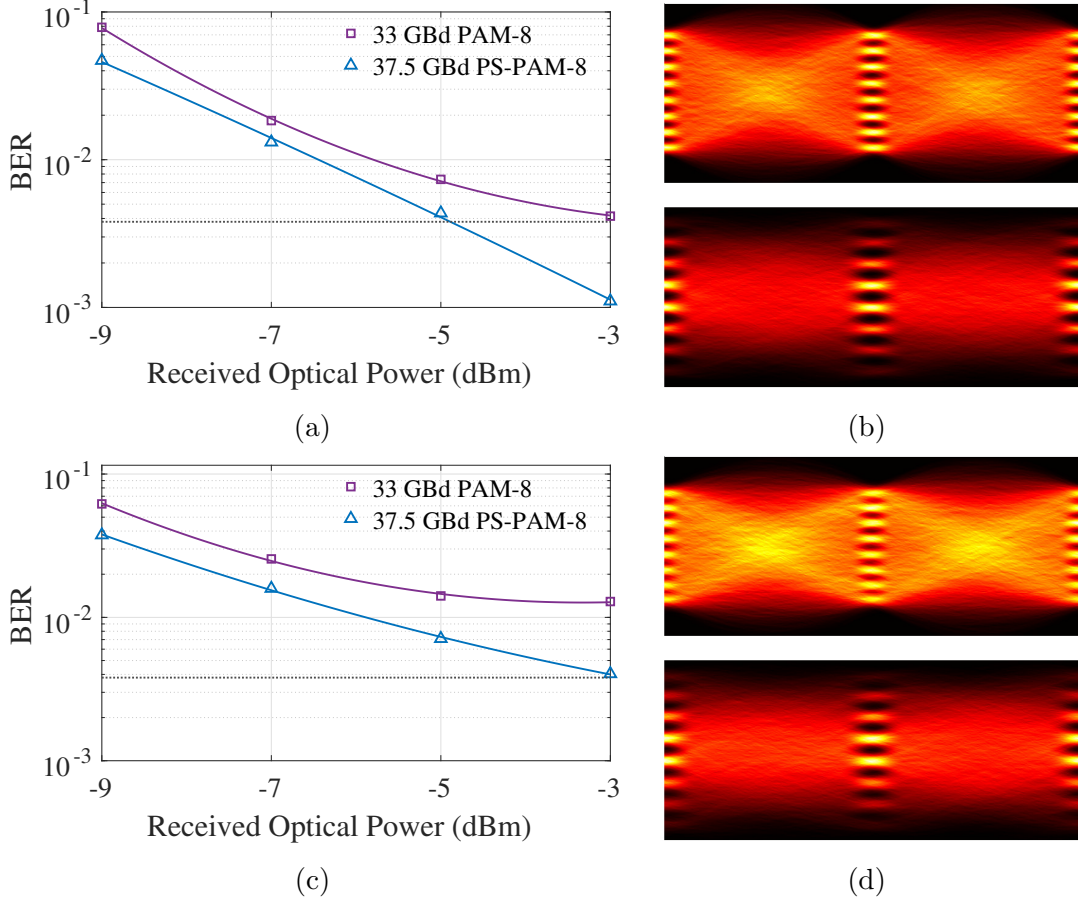


Figure 4.9: (a) BER vs ROP for scenario (1) PS-PAM8 (blue) and PAM8 (purple). (b) Eye diagrams for scenario (1) at an ROP of -3 dBm for PAM8 and PS-PAM-8 (top and bottom respectively). (c) BER vs ROP for scenario (2) PS-PAM8 (blue) and PAM8 (purple). (d) Eye diagrams for scenario (2) at an ROP of -3 dBm for PAM8 and PS-PAM-8 (top and bottom respectively).

shown in this section in terms of the raw BER are still relevant for comparing techniques. This is particularly true for systems with high code rates, where the pre-FEC BER has proven to be a relatively reliable predictor [23; 24].

4.6 Conclusion

This work has presented an experimental demonstration of a short-reach unamplified transmission system with a transmission rate granularity down to 0.25 bits per symbol ranging from OOK up to PAM16. The extremely low RIN of the integrated laser allowed for below FEC performance for modulation formats up to PAM16 with reduced DSP complexity compared to that of previous FlexPAM ex-

perimental demonstrations, making it especially suitable for intra-data centre links. The experiment was performed at a wavelength of 1552.7 nm, but given the results shown in [14] with the same dual laser module, it can be expected to work similarly over the full 100 nm range covered by the device. Additionally, we were able to show through simulation the feasibility of fulfilling the current standards for 100 Gbit/s, using a 53.5 GBaud PAM4 signal, with ROPs as low as -11 dBm, as well as a potential transmission capacity of up to 200 Gbit/s per wavelength, given a modest increase in transmitter bandwidth and a 5 dB SNR increase in our simulation, which could enable Tbit/s data centre links using integrated multi-wavelength transmitters. These results show that FlexPAM is a suitable option to cost-effectively maximise transmission rates of WDM-based intra-data centre links with optical switching and variable path losses.

In the PS experiment, we were able to show how symmetric PS can mitigate nonlinearities arising from the use of SOAs in systems using advanced PAM formats. Below the FEC limit performance was obtained for received powers as low as -5 dBm when amplifying a single channel, and -3 dBm when amplifying the full 4-channel WDM system. Those values, together with the power boost introduced by the SOA, result in a total power budget of 30.25 dB in scenario (1) and 28.75 dB for scenario (2). A simplified VNLE was used at the receiver, and the full equalization process was performed with the signal sampled at the baud rate, reducing the computational complexity and analogue to digital converter requirements. The high spectral efficiency and the low cost, footprint and power consumption provided by SOA amplification make the studied solution a potential candidate for optically amplified, short-reach networks based on IMDD technology.

In the next chapter, we will move from IMDD to coherent formats and look at using a fast tunable laser, for high-speed coherent transmission in continuous wave and burst-mode operation. We will show the capabilities of the novel device and demonstrate record performance for nanosecond switching devices.

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Chapter 5

Coherent Burst Mode

Transmission for Short Reach

Interconnects

The recent explosion in artificial intelligence and machine learning is driving a need for improved energy efficiency in data centre networks. Optical switching offers the potential for a less power-hungry alternative to electronic switching, and there has been renewed interest in the space since the deployment of optical circuit switching in a production hyperscale data centre network [1; 2]. The switching/reconfiguration speed in [1; 2] is of the order of seconds, however, the electromechanical-based technology can offer millisecond switching. Faster optical switching techniques such as wavelength switching can offer additional benefits such as reduced latency, and the ability to handle more bursty traffic. Changing the wavelength at an input port of a passive arrayed waveguide grating router switches the signal to a different output port, and fast-switching tunable lasers can be used to achieve nanosecond switching speeds. Such systems were widely studied in the 2000s [3; 4]. The tunable lasers available at the time offered very fast electro-optic switching, wide tunability, and despite poor linewidth were suitable for the intensity modulation with direct detection (IMDD) communication systems used at the time. As coherent communi-

cation grew in prominence, many of these widely tunable lasers were redesigned to use thermal tuning mechanisms, but in doing so, also sacrificed their fast switching capability.

Some demonstrations of fast optical switching in the C-band that can be found in the literature include nanosecond switching using on-off keying [5; 6], 10s-100s of ns switching using coherent systems [7; 8; 9; 10], and more recently, nanosecond switching using a coherent system [11]. These demonstrations were performed at relatively low data rates, mostly limited to low-order modulation formats. However, as the trend towards higher baud rates and spectral efficiency continues, tunable lasers offering a combination of low linewidth, nanosecond switching, and wide wavelength tunability are required.

As optical transmission systems move towards higher baud rates and utilize more advanced modulation formats, optical amplification in short-reach coherent systems becomes crucial in meeting power budget requirements, benefiting both data centre networks and passive optical networks (PON). Compared to erbium-doped fibre amplifiers (EDFA), semiconductor optical amplifiers (SOA) offer an integrated, more energy-efficient solution. Depending on the placement and operating regime of the SOA in the system, its influence can vary widely: Placed before modulation, an SOA adds noise to the beam, whereas if placed after modulation, it can add distortions to the signal when operated close to saturation. It can be useful to amplify the signal after modulation to overcome the losses from the modulator, and operating close to saturation helps maximise the launch power. In that situation, mitigation of the induced nonlinear distortions is important to recover the signal after detection. Some approaches include the use of a holding beam at the transparency wavelength to accelerate the recovery times of the medium [12; 13; 14], or the digital backpropagation of the signal using the dynamic equations of the SOA [15]. While both of those techniques can provide significant performance improvements, their added implementation complexity can make them unappealing: For the holding beam, a dedicated laser source is required at the right wavelength, and for the digital back-

propagation, an accurate characterisation of the internal parameters of the SOA is needed.

This chapter presents the characterisation of a widely tunable, low linewidth, fast-switching laser suitable for next-generation coherent optical access and data centre networks and the performance demonstration both in continuous wavelength (CW) operation and in burst mode recently published in [16; 17]. In section 5.1, we explain the device characterisation. Then, section 5.2 introduces the experimental setups used to quantify the performance of the laser on a high-baud rate, advanced modulation format scenario, and section 5.3 shows the results obtained from the experiments. Finally, section 5.4 introduces an adaptive Volterra-based nonlinear equaliser (NLE) for SOA-amplified signals compatible with the standard coherent digital signal processing (DSP) flow.

5.1 Device characterisation

The integrated tunable laser (ITL) structure consisted of an InGaAsP multi-quantum well gain section coupled with two micro-ring tuners and an electro-optic phase modulator, as shown in [18]. A wavelength coverage of 20 nm was observed with a side mode suppression ratio (SMSR) of more than 50 dB when a single ring was tuned, while a total coverage of 35 nm, from 1540 nm to 1575 nm, was obtained with both

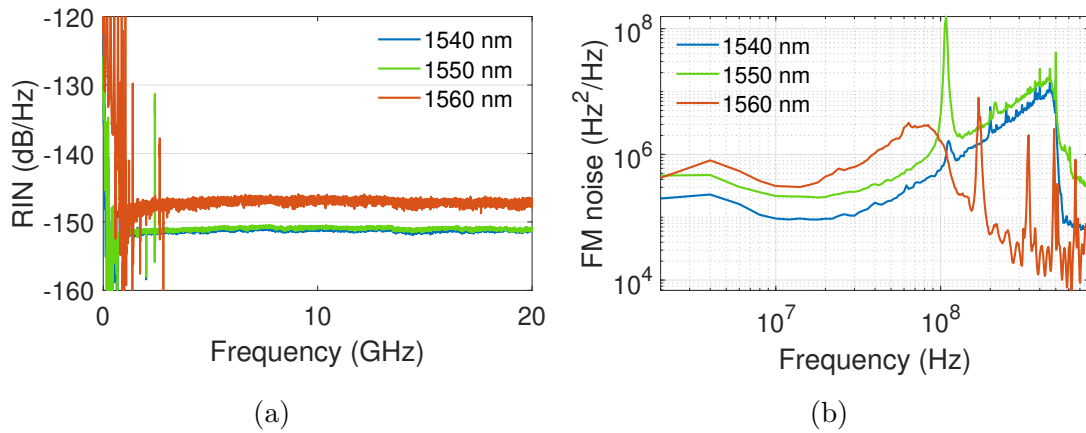


Figure 5.1: (a) RIN estimation for 1540 nm, 1550 nm and 1560 nm. (b) FM noise estimation for 1540 nm, 1550 nm and 1560 nm. The curves are scaled for the flat portions of the curves to correspond to the intrinsic Lorentzian linewidth.

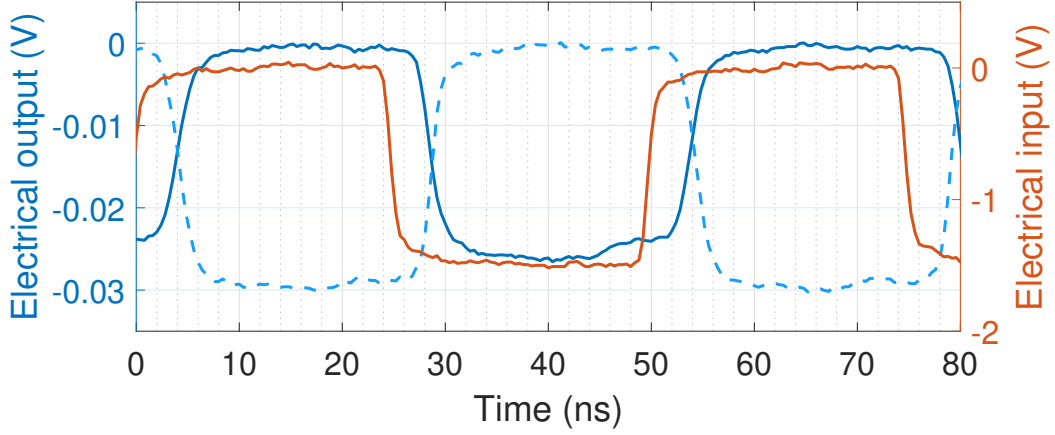


Figure 5.2: Switching demonstration using 25 ns electrical pulses to tune the laser between two target wavelengths spaced by 3.5 nm. The blue lines show the detected optical signal obtained by sending the output of the laser through a tunable optical filter centred at each of the switching wavelengths. Red solid lines: Electrical input. Blue solid lines: Electrical output at 1546.65 nm. Blue dashed lines: Electrical output at 1550.15 nm.

rings tuned simultaneously. The device was initially characterised for three different operating points accessible by tuning a single micro-ring tuner, corresponding to emission wavelengths of 1540 nm, 1550 nm, and 1560 nm. The relative intensity noise (RIN) was measured using the setup detailed in [19]. Values of -151 dB/Hz were obtained for 1540 nm and 1550 nm, and -147 dB/Hz for 1560 nm (Figure 5.1(a)). In all cases, the measurement was thermal and shot noise limited due to the output power from the device, meaning that the actual RIN value might be lower than the figures presented here. A frequency noise measurement was performed using a delayed self-heterodyne technique explained in [20]. The frequency noise results are presented in figure 5.1(b). The curves are scaled such that the intrinsic Lorentzian linewidth can be read directly from the flat portions of the curves. For this laser, linewidth values of 100 kHz, 200 kHz, and 300 kHz were obtained at 1540 nm, 1550 nm, and 1560 nm respectively. The curves were obtained using a digital low-pass filter to remove the instability caused by optical feedback. It must be noted that the device used for the characterisation did not have an integrated optical isolator, and recent versions of the device with an integrated isolator show linewidth values approaching the 100 kHz value over the whole operating range. Moreover, there

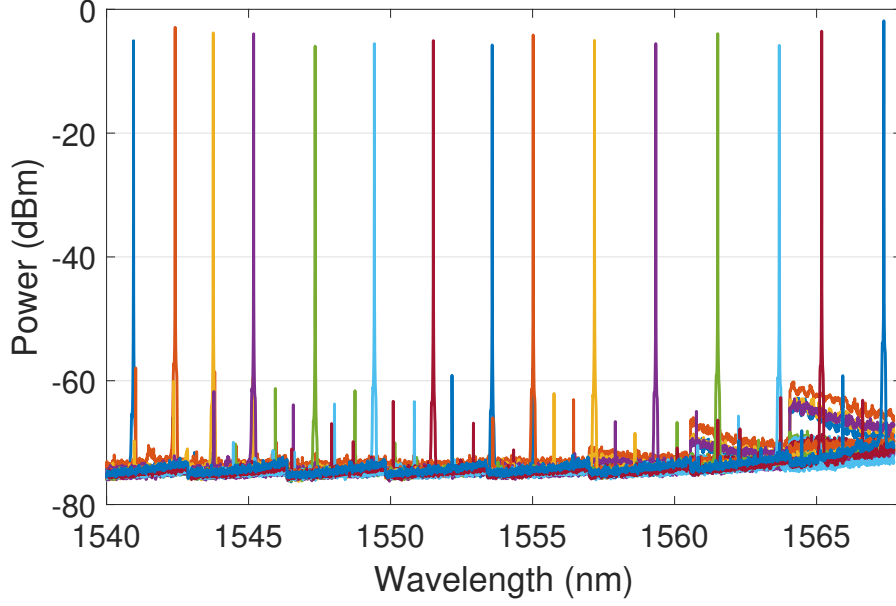


Figure 5.3: Overlapping spectra of several tuning points covering a 26 nm range in the C band.

is no evidence of deleterious $1/f$ frequency noise present above 1 MHz as was the case previously for fast-switching tunable lasers based on electronic current tuning [21]. Figure 5.2 shows a switching demonstration as the electrical signal to one of the micro-ring tuners is varied to switch the laser (without modulating data), from a wavelength of 1546.65 nm to 1550.15 nm. A switching time of around 2 ns was observed. Figure 5.3 shows an overlapping plot of optical spectra for different tuning points, all of them showing an SMSR of over 50 dB.

5.2 Experimental setup

This section explains the experiments performed with two iterations of the ITL. The first one, in subsection 5.2.1, shows the experimental setup used to assess the performance in CW operation of the ITL used for the characterisation in section 5.1 (without an integrated optical isolator). Subsection 5.2.2 shows the experimental setup used to study the burst mode operation performance of a second iteration of the ITL, including an optical isolator at the output end.

5.2.1 First iteration ITL

In order to demonstrate the performance of the device, a 60 GBaud dual polarisation (DP) 16-ary quadrature amplitude modulation (QAM) transmission experiment was performed, resulting in a gross transmission rate of 480 Gbit/s over 25 km of fibre. Four pseudorandom 4-level signals 2^{16} symbols long were generated offline, filtered with a root-raised cosine (RRC) filter with a roll-off factor of 0.2, predistorted to compensate for the bandwidth limitation and the relative delays between tributaries introduced by the RF circuit of the transmitter, and uploaded into an arbitrary waveform generator (AWG) Keysight M8194A operating at 120 GSa/s. The signals from the AWG were sent into a dual-polarisation IQ transmitter (IQ-Tx) with a 6 dB analogue bandwidth of 40 GHz, where they modulated the four tributaries (IX, QX, IY, QY) of the light from the laser. An external optical isolator was placed before the IQ-Tx to avoid backreflections. The wavelengths studied were 1544.1 nm, 1551 nm, 1556.6 nm and 1563 nm, accessible by tuning a single micro-ring tuner.

After the IQ-Tx, the signal was sent to an EDFA for power boosting. The total launch power over both polarisations was set to the powers of 3 dBm, 8 dBm and 12 dBm by tuning the gain of the EDFA, and the signal was transmitted over 25 km of large effective area fibre (LEAF) with a chromatic dispersion (CD) parameter of 4 ps/nm/km, attenuation coefficient of 0.19 dB/km and a nonlinear coefficient of around 0.7 1/W/km. At the receiver, an external cavity laser was used as a local

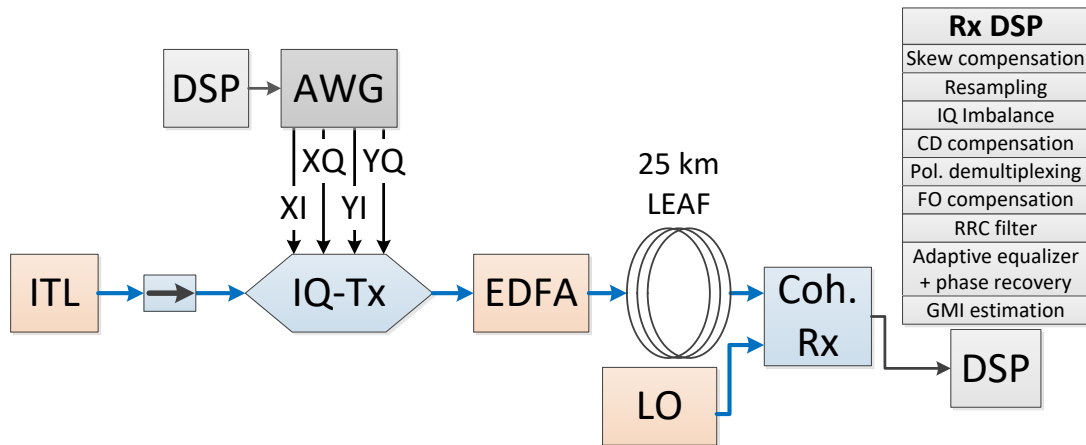


Figure 5.4: Experimental setup used for the first iteration of the ITL.

oscillator (LO) and a dual polarisation coherent receiver containing four balanced photodiodes (Finisar BPDV3120R) with a 3 dB bandwidth of 70 GHz converted the optical signal into electrical signals. Four real-time oscilloscopes (RTO) with a 3 dB bandwidth of 70 GHz operating at 200 GSa/s were used for capturing the electrical currents into digital signals for offline processing. A schematic representation of the experimental setup is shown in figure 5.4.

The DSP applied to the digital signals consisted of receiver skew compensation, resampling to two samples per symbol, signal normalisation, IQ imbalance compensation, CD compensation, frequency offset compensation, matched filtering, and adaptive decision-directed multiple input multiple output (MIMO) equaliser. The MIMO equaliser included widely linear feed-forward equalisation combined with a decision-directed carrier phase recovery for polarisation demultiplexing and compensation of residual distortions. The length of the equaliser was set to 31 taps, as beyond this, additional length increases did not result in noticeable performance gains. The tap weights were updated using a least means squares algorithm, and a short training sequence was used for initial convergence. The BER and the GMI of the equalised signals were obtained for each of the wavelengths by averaging over both polarisations and multiple captured sequences.

5.2.2 Second iteration ITL

The performance of the second iteration of the ITL, including an optical isolator on chip, was analysed by looking at the GMI in a back-to-back (B2B) configuration for burst mode operation of DP-16QAM at baud rates of 60 GBaud and 90 GBaud. The switching and settling times between four wavelengths were studied.

A pulse pattern generator (PPG) was used to generate two square-wave signals of amplitude 2 V and 1.25 V respectively, and period 12 μ s with a relative phase difference of $\pi/2$ rad. The two square-wave signals were used to drive the two micro-ring tuners in the ITL, resulting in the emission wavelength sequence 1545.06 nm, 1548.64 nm, 1550.04 nm, 1546.46 nm. The output of the ITL, with a power of around

-3 dBm, was then amplified using an SOA (Kamelian OPB-10-15-N-C-FA) before feeding it into the IQ-Tx for modulation. The IQ modulator (6-dB bandwidth of 40 GHz) was driven by four amplified electrical signals created, filtered (RRC filter with a roll-off factor of 0.01) and predistorted offline, and generated with an AWG operating at 120 GSa/s. Given the electrical bandwidth of the transmitter (limited by the IQ modulator), strong preemphasis was required at 90 GBaud for the higher frequencies, while the 60 GBaud signal performed better with a lower predistortion to avoid amplifying out-of-band noise. The optical spectrum of the 60 GBaud and 90 GBaud signals during switching can be seen in figure 5.6(a).

The DP-16QAM signal was amplified with an EDFA to a power of 7 dBm and sent to a 99/1 coupler. The 1% output of the coupler was sent into an optical spectrum analyser (OSA), and the 99% to the DP coherent receiver. After the balanced photodiodes, four RTOs operating at a sampling rate of 200 GSa/s were used for digitising the signals. An LO was tuned to each of the wavelengths separately for detection, and offline DSP was applied to the captured signals. Figure 5.5 shows the schematic of the experimental setup and the DSP sequence. The transmission performance of the wavelengths 1545.06 nm and 1546.46 nm was also evaluated in CW operation as a function of optical signal-to-noise ratio (OSNR) using an EDFA and a tunable optical filter as a source of amplified spontaneous emission (ASE), and

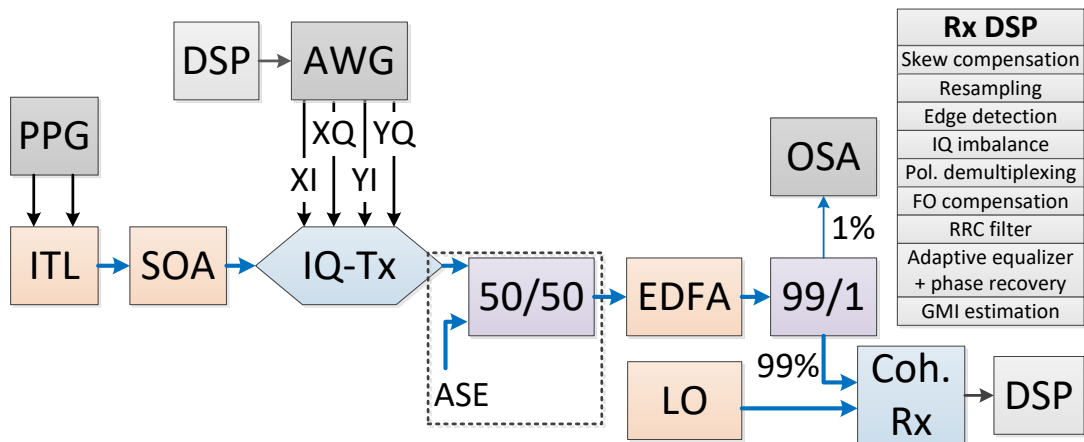


Figure 5.5: Experimental setup used for the second iteration of the ITL. The dotted box shows the noise loading setup used for the characterisation of the device as a function of OSNR.

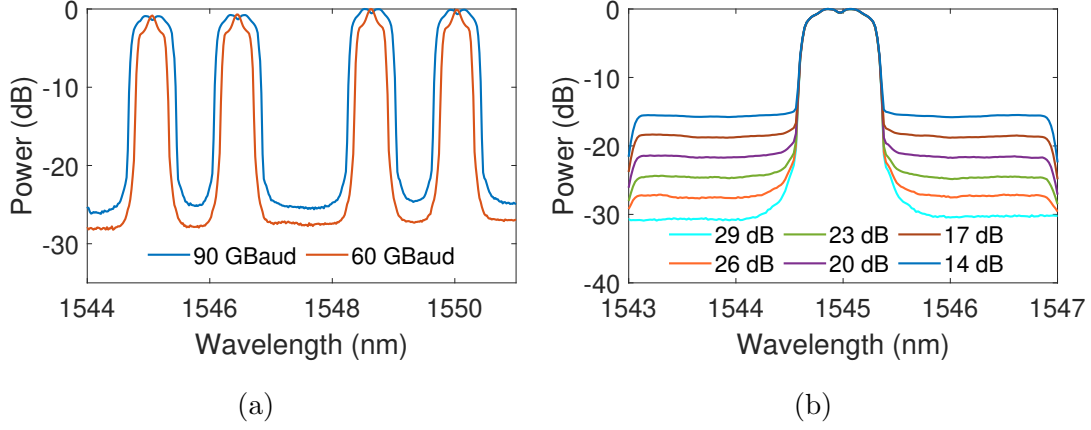


Figure 5.6: (a) Optical spectrum of the 60 GBaud (red) and 90 GBaud (blue) signals during the 4-wavelength switching. (b) Superimposed spectra of the 90 GBaud signal at 1545.06 nm for OSNR values ranging from 14 dB up to 29 dB.

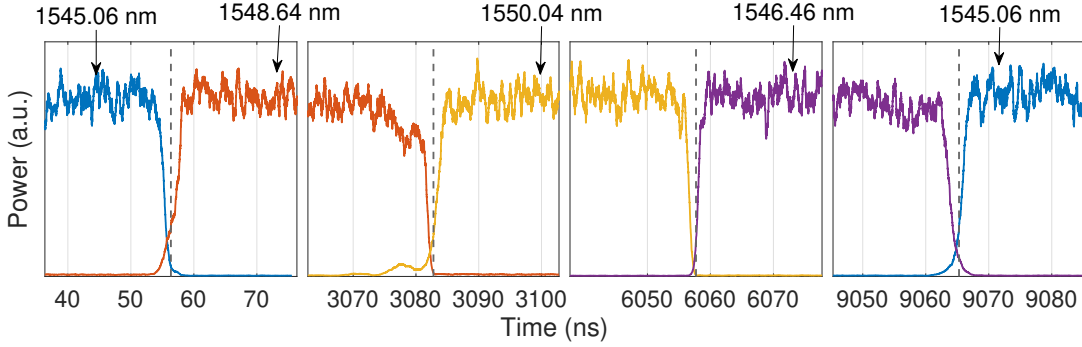


Figure 5.7: Superimposed time traces of the power captured by the coherent receiver with the LO tuned to the four different wavelengths. For display clarity, the traces were smoothed using an averaging window of 0.5 ns. The dashed lines in the plots show the switching instant detected by the edge detection function.

a 50/50 coupler to combine the ASE with the signal. Figure 5.6(b) shows the superimposed spectra of the 90 GBaud signal at 1545.06 nm for OSNR values ranging from 14 dB up to 29 dB. The GMI was calculated for each wavelength at different OSNR values averaging over a total time of 2^{16} ns.

For the burst mode experiment, the LO was tuned to each of the studied wavelengths separately. Figure 5.7 shows the superimposed time traces of the switching between the four wavelengths. The change power applied to the micro-ring tuners to switch wavelength is around 0.25 mW, giving a switching power (calculated as the ratio between the power swing in the ring tuners and the gross data rate) of 0.35 fJ/bit for the 90 GBaud (720 Gbit/s) signal, and of 0.52 fJ/bit for the 60 GBaud (480 Gbit/s). Sharp rise and fall times in the order of a few nanoseconds can be

observed, with transition times of less than 5 ns, consistent with the switching characterisation from figure 5.2. The variations in pulse duration can be attributed to a small deviation from $\pi/2$ rad of the phase difference between the square waves driving the micro-ring tuners of the ITL. The GMI was obtained for 2^{13} symbols using a blind equaliser with taps preconverged using a training sequence of 512 symbols (≈ 5.7 ns at 90 GBaud, ≈ 8.5 ns at 60 GBaud). This calculation was performed starting from different symbols after the burst detection (dashed black lines in figure 5.7) in steps of 200 symbols (≈ 2.2 ns at 90 GBaud, ≈ 3.3 ns at 60 GBaud). In order to obtain reliable statistics, 30 bursts were captured for each wavelength and baud rate.

5.3 Results

5.3.1 First iteration ITL

The BER results obtained in these experiments can be seen in figure 5.8(a). For reference, the figure shows as dashed lines the BER limit for error-free performance of a signal with a hard-decision (HD) forward error correction (FEC) code with 7% overhead, and a typical BER limit for a 17% overhead soft-decision (SD) FEC-coded signal [22]. The GMI, well known to be a better predictor for the performance of SD-FEC-coded signals, is shown in figure 5.9 together with the minimum achievable information rate (AIR) required for a 17% overhead SD-FEC code. BER values below the HD-FEC limit can be seen for 12 dBm launch power over the whole wavelength range. For a launch power of 8 dBm, the GMI values obtained are well above the SD threshold, and the pre-FEC BER is below the estimated SD-FEC limit. However, the 3 dBm results are limited by a low signal-to-noise ratio (SNR) and the obtained performance is below the SD-FEC requirements. While significant performance gains were obtained from increasing the launch power from 3 dBm to 8 dBm, only a modest improvement was observed from moving to 12 dBm. This trend can be explained by the effects of fibre nonlinearity limiting the performance at high

powers and indicates that additional reductions in BER can potentially be obtained with nonlinear equalization techniques such as digital backpropagation (DBP). It can be noted, however, that the AIR of the received signals is mostly limited by the entropy of the DP-16QAM signal, and higher-order modulation formats would be required to take advantage of the additional gains provided by the DBP. The obtained results were found to be consistent over the whole wavelength range, with just minor performance variations among wavelengths mainly due to differences in output power (affecting the required gain from the EDFA) and, to some extent, due to different linewidth values from the laser.

Figure 5.8(b)-(e) show various constellations under different wavelengths and launch power conditions. Clear constellations can be seen for launch powers of 8 dBm and 12 dBm independently of the operating wavelength, while the constellation corresponding to a launch power of 3 dBm shows a wide spreading of the

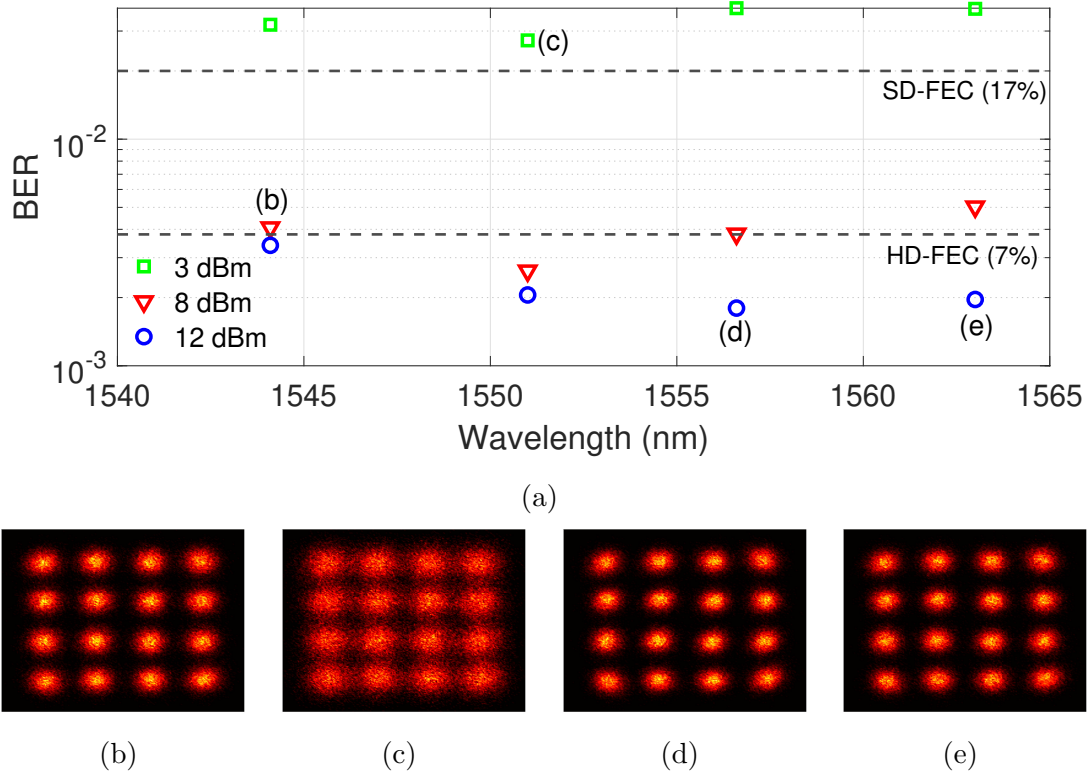


Figure 5.8: (a) BER obtained by averaging over both polarisations for different launch powers and carrier wavelengths. Dashed lines show typical values for the BER limit of a 7% overhead HD-FEC and a 17% overhead SD-FEC. (b)-(e) show the constellation diagrams of the equalised output at 1544.1 nm (8 dBm), 1551 nm (3 dBm), 1556.6 nm (12 dBm) and 1563 nm (12 dBm) respectively.

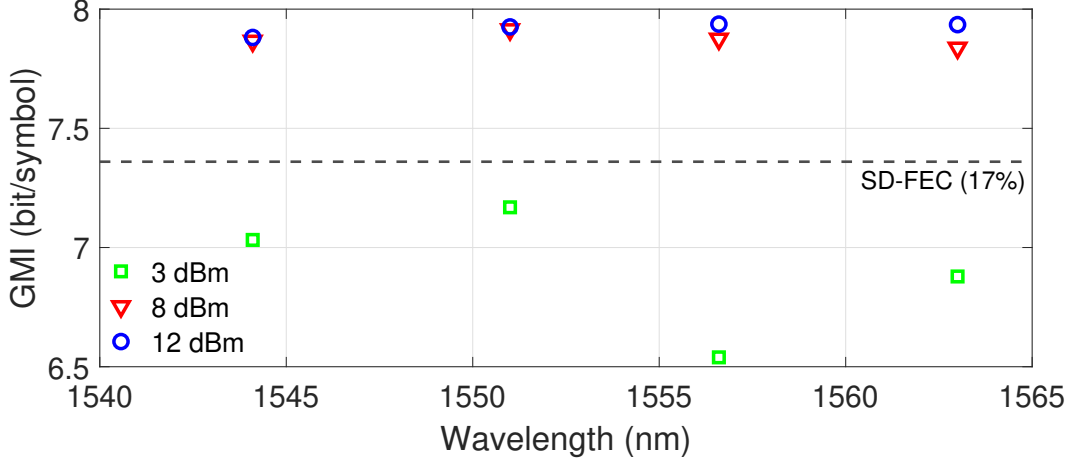


Figure 5.9: GMI obtained over both polarisations for different launch powers and carrier wavelengths. Dashed lines show typical values for the AIR required for a 17% overhead SD-FEC.

constellation points due to the degradation in SNR.

5.3.2 Second iteration ITL

Figure 5.10 shows the GMI as a function of OSNR for the 60 GBaud and 90 GBaud signals with the laser operating in CW at the wavelengths 1545.06 nm and 1546.46 nm. For the 60 GBaud signals, both wavelengths show similar performance with the GMI reaching its maximum of 8 bit/symbol for OSNR values above 20 dB. On the other hand, for the 90 GBaud signals, there is a difference in performance between wavelengths, with 1545.06 nm having slightly higher AIRs than 1546.46 nm over the whole OSNR range. The GMI values at 90 GBaud for high OSNR are around 7.9 bit/symbol.

The obtained GMI as a function of waiting from the burst detection time before starting the processing can be found in Figure 5.11. The ‘×’ symbols show the individual values of GMI obtained for each of the 30 bursts, while the black solid lines show the trend of the average value. At 60 GBaud, the GMI shows a sharp increase after 10 ns for 1545.06 nm (Figure 5.11(a)), with the values approaching 8 bit/symbol after 20 ns. For the other wavelengths, the settling time of the laser required for this 60 GBaud operation is smaller than the duration of the training sequence and the results in figure 5.11(c, e, g) show GMI values close to the steady state value from the

switching instant. A small penalty of less than 0.05 bit/symbol in steady-state AIR can be seen for 1548.64 nm with respect to the other wavelengths. At 90 GBaud, the GMI shows a sharp increase over the first few ns for the four wavelengths studied. They reach values above 7.8 bit/symbol after 10 ns and stable performance after less than 40 ns. At a wavelength of 1548.64 nm (Figure 5.11(d)), steady-state GMI values above 7.85 bit/symbol are obtained, while for wavelengths of 1545.06 nm, 1550.04 nm and 1546.46 nm (Figures 5.11(b,f,h) respectively) steady-state values above 7.9 bit/symbol can be observed. The total switching time, from the power drop on one wavelength to the end of the training sequence on the other, is in all cases less than 20 ns.

Similar trends can be seen for both baud rates, with 1545.06 nm giving the worse transient performance, and 1548.64 nm giving the worse steady-state performance, both wavelengths obtained from the same voltage level on the second ring. From the CW characterisation of the performance at 1545.06 nm and 1546.46 nm, tuning only the voltage on the second ring, better performance can be seen for the former wavelength, while the latter shows a higher GMI and faster transients in the switching experiment. That suggests the voltage levels of the square wave driving the second ring can be further optimised, potentially achieving faster, more stable switching for all wavelengths. It is also notable that the switching times do not increase for wavelength jumps of 3.6 nm compared to 1.4 nm.

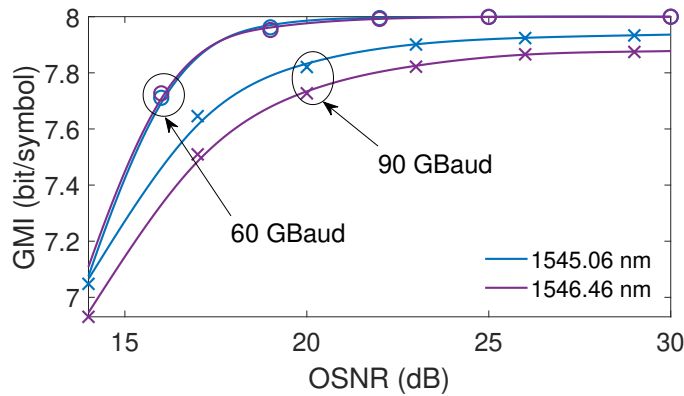


Figure 5.10: GMI as a function of OSNR in CW operation for 60 GBaud (○) and 90 GBaud (×) signals with the ITL tuned to 1545.06 nm (blue) and 1546.46 nm (purple).

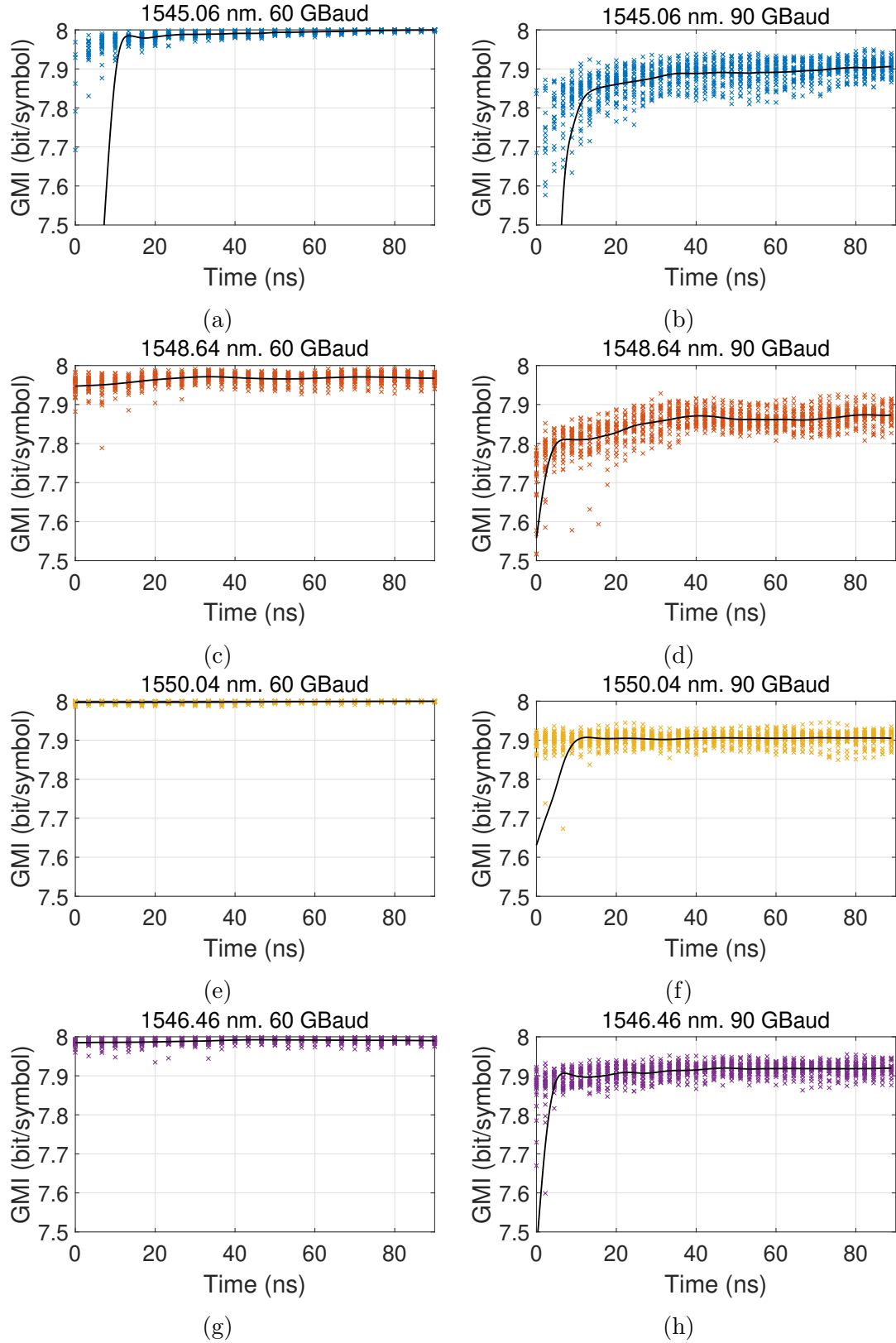


Figure 5.11: GMI as a function of waiting time from the burst instant. The ‘x’ symbols show the GMI for each burst, and the black line shows the trend of the average value for (a) 1545.06 nm at 60 GBaud, (b) 1545.06 nm at 90 GBaud, (c) 1548.64 nm at 60 GBaud, (d) 1548.64 nm at 90 GBaud, (e) 1550.04 nm at 60 GBaud, (f) 1550.04 nm at 90 GBaud, (g) 1546.46 nm at 60 GBaud, (h) 1546.46 nm at 90 GBaud.

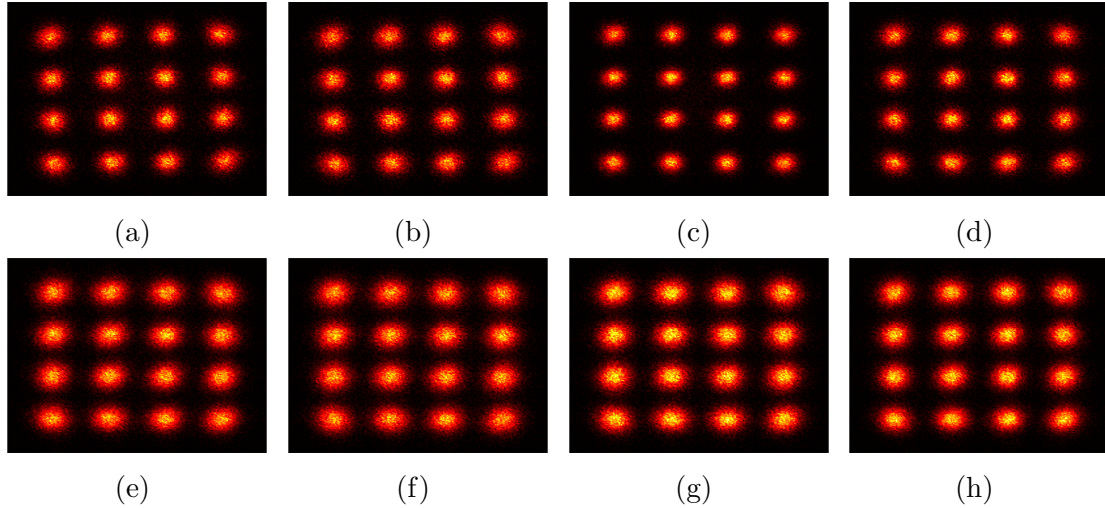


Figure 5.12: Constellation plots for a waiting time of 10 ns operating at: (a) 1545.06 nm, 60 GBaud, (b) 1548.64 nm, 60 GBaud, (c) 1550.04 nm, 60 GBaud, (d) 1546.46 nm, 60 GBaud, (e) 1545.06 nm, 90 GBaud, (f) 1548.64 nm, 90 GBaud, (g) 1550.04 nm, 90 GBaud, (h) 1546.46 nm, 90 GBaud.

Figure 5.12 shows the constellation plots for each wavelength for a waiting time of 10 ns from the switching instant for 60 GBaud (a-d) and 90 GBaud (e-h), showing that no significant excess phase noise is present even after such a short waiting period after the switching instant.

5.4 DSP for SOAs in Coherent Systems

The SOA used for the experiment in figure 5.5 was used to boost the power before modulation, hence not introducing any nonlinear distortions. However, for certain short-reach applications, it can be useful to place an SOA after the modulator, where the light already contains the information encoded in its amplitude and phase. In this section, we present an NLE for SOA-amplified systems compatible with the standard coherent DSP and study its performance with a single polarisation 16QAM simulation.

5.4.1 Properties of SOA nonlinearity

Neglecting carrier heating, spectral-hole burning, and scattering losses in the cavity, the lumped model of a single-polarisation SOA can be written as [23]:

$$E_{out}(t) = E_{in}(t) \cdot \exp\left(\frac{h_N(t)}{2} (1 - j\alpha_H)\right) \quad (5.1)$$

$$\frac{dh_N(t)}{dt} = \frac{h_0 - h_N(t)}{\tau_c} - \frac{\exp(h_N(t)) - 1}{P_{sat}\tau_c} |E_{in}(t)|^2 \quad (5.2)$$

Where $h_N(t)$ is the contribution of the carrier density number to the integrated gain over the length of the SOA, α_H is the gain-phase coupling parameter, $h_0(t)$ is the small signal gain, τ_c is the carrier lifetime, and P_{sat} is the saturation power.

By looking only at the dependencies of the different variables, the constants in equation (5.1) can be ignored, obtaining:

$$E_{out}(t) \propto E_{in}(t) \cdot \exp(h_N(t)) \quad (5.3)$$

Additionally, assuming that the relative changes in the carrier density are small compared to the input power fluctuations, the terms depending on $h_N(t)$ in equation (5.2) can be treated as constants and ignored too. This assumption holds for unsaturated SOAs, as the input signal can have high extinction ratios while generating a small carrier density depletion. In that scenario, we obtain:

$$\frac{dh_N(t)}{dt} \propto |E_{in}(t)|^2 \quad (5.4)$$

So taking the time derivative of equation (5.3) we obtain:

$$\begin{aligned} \frac{dE_{out}(t)}{dt} &\propto \frac{dE_{in}(t)}{dt} \cdot \exp(h(t)) + C \cdot E_{in}(t) \cdot \exp(h(t)) \cdot |E_{in}(t)|^2 \\ &\propto \frac{dE_{in}(t)}{dt} + C \cdot E_{in}(t) \cdot |E_{in}(t)|^2 \end{aligned} \quad (5.5)$$

Where C is a proportionality constant. On the right-hand side of equation (5.5), we can see the terms $dE_{in}(t)/dt$, which contains the information modulating the input signal, and $E_{in}(t) \cdot |E_{in}(t)|^2$, introducing a power-dependent distortion. This term has the typical form of self-phase modulation, however, since the constant C does

not have to be purely imaginary, it can potentially encode any power-dependent distortion affecting both amplitude and phase. The term $\exp(h(t))$ is neglected as for the assumption mentioned above.

5.4.2 Nonlinear Equaliser

Given the relation shown in equation (5.5), the second-order Volterra NLE introduced in equation (2.4) in the context of IMDD systems can be modified to obtain the nonlinear taps of the equaliser as a field-power beating, as opposed to a field-field beating:

$$y(m) = \mathbf{E}_{1,m}^T \cdot \mathbf{w}_1^* + \mathbf{E}_{2,m}^T \cdot \mathbf{W}_2^* \cdot |\mathbf{E}_{2,m}|^2 \quad (5.6)$$

Where, unlike in equation (2.4), $\mathbf{W}_2$ is not a triangular matrix as the symmetry of the expression was removed by including the optical power. $\mathbf{W}_2$ is defined, given the nonlinear memory length L_2 , as:

$$\mathbf{W}_2 = \begin{bmatrix} w_2(1,1) & w_2(1,2) & \dots & w_2(1,L_2) \\ w_2(2,1) & w_2(2,2) & \dots & \dots \\ \dots & \dots & \dots & w_2(L_2-1,L_2) \\ w_2(L_2,1) & \dots & w_2(L_2,L_2-1) & w_2(L_2,L_2) \end{bmatrix} \quad (5.7)$$

And the complexity reduction from [24] can also be applied to it by limiting the number of non-zero diagonals.

Equation 5.6 can be extended to dual polarisation signals as:

$$\begin{aligned} y^{X/Y}(m) = & \left(\mathbf{E}_{1,m}^{X/Y} \right)^T \cdot \left(\mathbf{w}_1^{XX/YY} \right)^* + \left(\mathbf{E}_{1,m}^{Y/X} \right)^T \cdot \left(\mathbf{w}_1^{XY/YX} \right)^* \\ & + \left(\mathbf{E}_{2,m}^{X/Y} \right)^T \cdot \left(\mathbf{W}_2^{XX/YY} \right)^* \cdot \left| \mathbf{E}_{2,m}^{X/Y} \right|^2 + \left(\mathbf{E}_{2,m}^{X/Y} \right)^T \cdot \left(\mathbf{W}_2^{XXY/YYX} \right)^* \cdot \left| \mathbf{E}_{2,m}^{Y/X} \right|^2 \\ & + \left(\mathbf{E}_{2,m}^{Y/X} \right)^T \cdot \left(\mathbf{W}_2^{XYX/YXY} \right)^* \cdot \left| \mathbf{E}_{2,m}^{X/Y} \right|^2 + \left(\mathbf{E}_{2,m}^{Y/X} \right)^T \cdot \left(\mathbf{W}_2^{XYY/YXX} \right)^* \cdot \left| \mathbf{E}_{2,m}^{Y/X} \right|^2 \end{aligned} \quad (5.8)$$

Table 5.1: Parameters used in the SOA lumped model

Symbol	Definition	Value
h_0	Small signal gain	6 (a.u.)
$\alpha_L L$	Scattering losses	3 (a.u.)
α_H	Gain-phase coupling parameter	3 (a.u.)
P_{sat}	Saturation power	20 mW
τ_c	Carrier lifetime	100 ps
ε_{SHB}	Inverse of SHB saturation power	1 W ⁻¹
ε_{CH}	Inverse of CH saturation power	2 W ⁻¹
α_{CH}	CH-phase coupling parameter	-1 (a.u.)
n_{sp}	Spontaneous emission inversion factor	2 (a.u.)

5.4.3 Simulation

In order to assess the effectiveness of the NLE presented here, a simulation was executed for the amplification with an SOA of a single polarisation 16QAM signal. Symbol rates of 10 GBaud, 25 GBaud and 50 GBaud were studied. The SOA lumped model from equation (1.4) was solved using a 4th order Runge Kutta for input powers ranging from -10 dBm to 5 dBm. The adiabatic approximation for the evolution of the gain from carrier heating (CH) and spectral hole burning (SHB) was used [23]. The SOA parameters used for the simulation can be seen in table 5.1, and the rest of the components in the system were ideal. A root-raised cosine (RRC) filter with a bandwidth equal to half the baud rate and a roll-off factor of 0.1 was applied at the transmitter and receiver. An oversampling rate of 32 samples per symbol was used in the simulation to mimic the analogue behaviour of the physical signals. The schematic setup used for the simulation can be seen in figure 5.13.

The performance of the system was studied for a feed-forward equaliser (FFE) with a memory length of 21 taps, and 5 NLE with a linear memory length of 21 taps, a nonlinear memory length of 9 taps, and 0, 1, 2, 3, and 4 beating neighbours (NLE₀, NLE₁, NLE₂, NLE₃ and NLE₄ respectively), i.e. containing 1, 3, 5, 7, and 9 non-zero diagonals in $\mathbf{W}_2$. The GMI was estimated for each input power into the

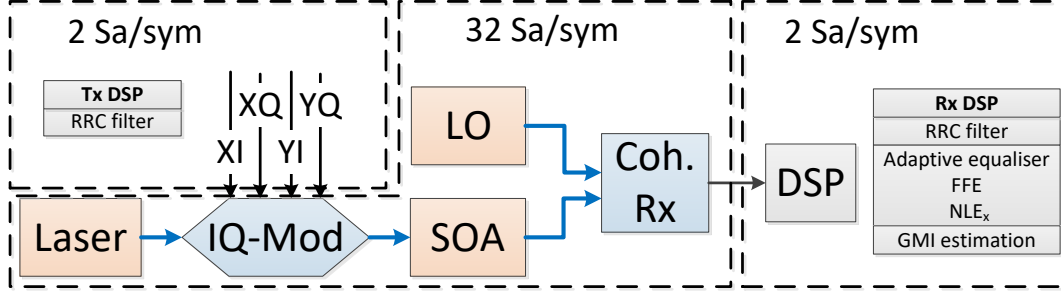


Figure 5.13: Schematic setup of the SOA amplification simulation. The dashed lines show the samples per symbol used on each part of the simulation.

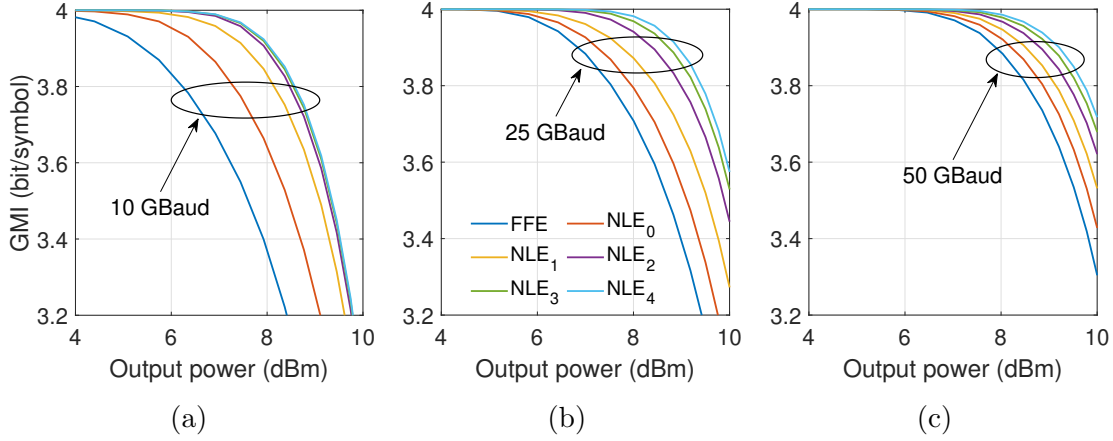


Figure 5.14: GMI as a function of SOA output power for different equaliser types at (a) 10 GBaud, (b) 25 GBaud, and (c) 50 GBaud.

SOA and equaliser type using the partially coherent additive white Gaussian noise (PC-AWGN) model.

The GMI as a function of the output power of the SOA for the different baud rates and equalisers is shown in figure 5.14. Figure 5.14(a) shows the results for the 10 GBaud simulation. In this regime, with a symbol duration similar to the carrier lifetime, nonlinear distortions are severe and the signal starts degrading at relatively low powers. Gains of over 1 dB can be seen from using the NLE_0 compared to the FFE, and further gains of close to 1 dB are obtained from NLE_1 . However, additional beating neighbours beyond this point show diminishing returns and the total gain from the NLE saturates at around 2.3 dB. Figure 5.14(b) and (c) show the results for the 25 GBaud and the 50 GBaud simulations respectively. In this situation, with symbol durations shorter than the carrier recovery times, increases in the baud rate result in a more linear signal amplification and hence, less penalty

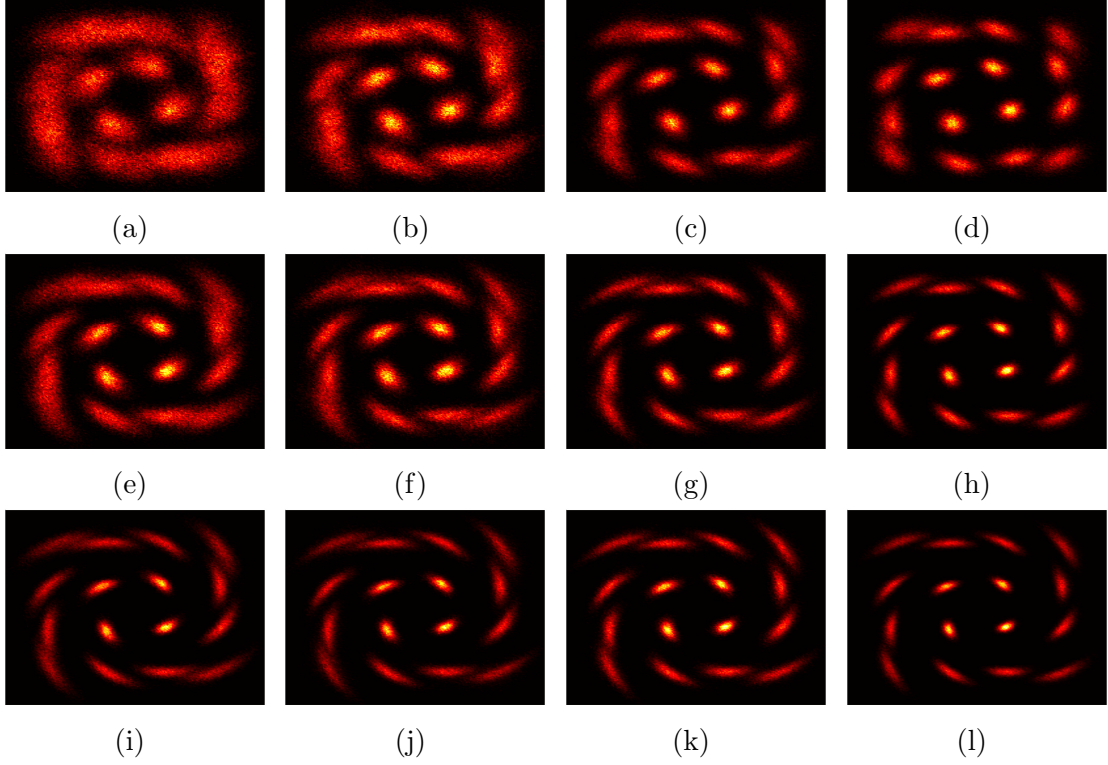


Figure 5.15: Constellation plots for an input power into the SOA of 0 dBm (output power of around 8.4 dBm) for (a-d) 10 GBaud, (e-h) 25 GBaud, (i-l) 50 GBaud using: (a, e, i) FFE, (b, f, j) NLE_0 , (c, g, k) NLE_1 , and (d, h, l) NLE_4 .

even at higher output powers. Incremental improvements can be seen as the number of beating neighbours is increased from 0 to 4, and while the NLE at these baud rates does not show the same gain saturation as for 10 GBaud, the gain from additional beating terms becomes smaller. The presence of a sharp saturation only for the 10 GBaud signal can be explained by the ratio between the symbol duration and the carrier lifetime: each additional beating neighbour adds half a symbol duration to the equaliser time window (i.e. the window grows faster for lower baud rates), and once that time window is significantly larger than the recovery times of the medium, the nonlinear interaction between additional terms is negligible.

Figure 5.15 shows the constellation plots at an input power into the SOA of 0 dBm, corresponding to an output power of around 8.4 dBm, for: (a-d) 10 GBaud, (e-h) 25 GBaud, (i-l) 50 GBaud using: (a, e, i) FFE, (b, f, j) NLE_0 , (c, g, k) NLE_1 , and (d, h, l) NLE_4 . Sharper constellation points can be seen as the number of beating neighbours increases, but a significant amount of residual phase noise is

still observed in all cases. Distortions in the centroids of the constellation points are present, with more significant rotations for the lower rates. The NLE seems to be unable to correct those distortions, which may fundamentally originate from the processes neglected in subsection 5.4.1. In order to account for the displacement of the constellation points, the GMI was estimated using the centroids of each symbol cloud, as opposed to the ideal symbol position.

In the cases presented here, the signals entering the SOA are free of inter-symbol interference, i.e. they are not dispersed. If the signals entering the SOA have accumulated dispersion, the nonlinear interactions spread between many symbols. In that case, long nonlinear lengths with large numbers of beating neighbours are required for the NLE, making it computationally unfeasible. In some cases, like an SOA as a pre-amplifier at the receiver, this can be solved by precompensating at the transmitter for the chromatic dispersion on the link before the SOA.

5.5 Conclusion

We have presented the basic device characterisation and system performance of a fast-switching wavelength-tunable laser. This laser could be suitable for wavelength-switching applications in short-reach systems such as data centres or PON. RIN and frequency noise characterization over a range of wavelengths covering part of the C-band present RIN values in the order of -150 dB/Hz and linewidths in the order of 100 kHz. We presented the performance of the first iteration of the device in a high-speed communication system operating at 480 Gbit/s 16QAM over a distance of 25 km of LEAF, demonstrating the feasibility of using this laser for short-reach applications. We then presented the burst mode operation of the second iteration on the device including an optical isolator on-chip. Successful nanosecond class wavelength switching of a 720 Gbit/s (90 GBaud DP-16QAM) coherent channel was demonstrated for four different wavelengths in the C-band. GMI values of around 7.9 bit/symbol, switching powers of less than 0.4 fJ/bit, and switching times in the order of 10 ns, limited by the bandwidth of the electrical signal driving the

micro-ring tuners, were observed for all wavelengths.

The device can readily be integrated with additional components such as Mach Zehnder modulators and booster SOAs on the same monolithic InP platform to create a highly integrated fast optical switching transmitter. The results presented open the possibility to optically switch bursty traffic within the data centre, whether that traffic is IMDD or coherent. The device can support the deployment of high-speed fast reconfigurable optical networks based solely on optical switching. The introduction of optical switching into photonic networks can avoid electronic detection and retransmission of payload data at each node within the network resulting in lower transient times (latency) and power consumption while maintaining a small footprint.

The Volterra-based NLE introduced in this chapter demonstrated the potential of high gains with a fully adaptive equaliser compatible with the standard coherent DSP. This NLE enables the use of SOAs as boosters closer to saturation, potentially increasing the launch power (and achievable power budget) with a moderate increase in computational complexity. The total complexity of the DSP block can be finely tuned by the equaliser length and the number of beating neighbours, making it suitable for a wide range of systems with different complexity requirements.

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Chapter 6

Conclusion and Future Work

6.1 Conclusion

The ever-growing data rate requirements and the need for energy-efficient solutions in optical networks are driving the evolution of photonic integrated technologies. For short-reach applications, characterised by the massive deployment of components, cost is also a main factor. In this situation, the growth in system capacity needs to be met with scalable, cost-effective and energy-efficient solutions.

Intensity modulation with direct detection (IMDD) has been the best solution for the last decades due to its simplicity and low power consumption compared to coherent technology. As photonic integrated technologies mature, and coherent hardware and digital signal processing (DSP) get more efficient and standardised, the use cases for which IMDD is still the most economically feasible solution are getting reduced. However, a significant fraction of the market is still dominated by IMDD, and current trends suggest it will remain for the coming years (particularly for intra-data centre networks).

In this thesis, we reviewed the basics of optical fibre transmission systems, as well as the basic DSP and coding techniques used in IMDD and coherent systems. Then, we studied different techniques of increasing complexity involving photonic integrated circuits for short-reach optical transmission systems, starting with direct modulation-based IMDD, to externally modulated IMDD to coherent technology.

In Chapter 3, the use of optical injection for improving the performance of directly modulated lasers (DML) was studied on an experimental and simulation basis using a 15 GBaud 4-level pulse amplitude modulation (PAM4). Our results show significant reductions in bit error ratio (BER) over fibre due to the enhancement in bandwidth and chirp reduction from the optical injection. The device used for the experiment is a fully integrated, widely tunable multisection laser that can be fabricated cost-effectively compared to an integrated laser with an electroabsorption modulator (EAM).

In Chapter 4, we explained the use of flexible PAM (FlexPAM) formats for increased granularity in spectral efficiency and signal-to-noise ratio (SNR) requirements. We studied through simulation and experimentation the change in required received optical power for formats ranging from on-off keying (OOK) up to PAM16, showing the ability of this lookup table-based technique to maximise transmission rates for a given SNR by providing an entropy granularity as low as 0.25 bit/symbol. We then show an experimental demonstration of a PAM8 system using probabilistic shaping (PS) to reduce the penalty from the nonlinear distortions induced by a semiconductor optical amplifier (SOA) operating close to saturation. We show that symmetric PS improves performance in SOA-amplified systems with respect to unshaped PAM. This improvement, obtained from the reduced probability of the outer symbols that produce the largest power excursions, was observed both for constant baud rate and constant data rate.

In Chapter 5, we presented the characterisation of a fast-switching tunable laser covering a wavelength range of over 25 nm in the C band. Switching times in the order of nanoseconds were obtained for various wavelengths, and settling times in the 10s of nanoseconds were observed for 60 GBaud and 90 GBaud dual-polarisation 16-ary quadrature amplitude modulation (DP-16QAM). Switching powers below 0.4 fJ/bit were obtained, making the device particularly suitable for low-power reconfigurable networks requiring high data rates and low latency. Then, we introduced a Volterra-based nonlinear equaliser (NLE) tailored for SOA-amplified coherent sys-

tems and showed on a simulation basis its effectiveness in increasing the dynamic range of the transmission. Our results for different baud rates show gains in output power of over 1 dB at 50 GBaud (lower nonlinearity) and up to 2 dB at 10 GBaud (higher nonlinearity). The effectiveness of the NLE, together with its full compatibility with the standard coherent DSP and the ease of tailoring its computational complexity, make this technique a promising solution for maximising the usage of SOAs in systems requiring cost-effective, efficient and integrated optical amplification.

6.2 Future Work

This thesis has shown some use cases and techniques that can be useful in various situations regarding short-reach optical transmission systems. However, many different areas related to this topic still require further investigation:

- High-speed IMDD transmission using a multisection DML:

In order to compete with alternative approaches, DML-based solutions must provide high data rates to satisfy the demand. Improvements in fabrication and packaging on the device used in Chapter 3 can provide higher bandwidths, enabling transmission rates of 50/100 Gbit/s and offering a competitive solution for energy-efficient, low-cost and low-complexity fibre optical systems. The experimental study and simulation-based modelling of a next-iteration DML with improved capabilities would be of interest in the context of this thesis.

- FlexPAM for optical access networks:

In Chapter 4, FlexPAM formats were introduced and demonstrated in the context of an intra-data centre network with reconfigurable links inducing variable path losses. However, another use case where the high granularity provided by FlexPAM is the passive optical network (PON). Users within a

PON experience distinct path losses due to differences in fibre lengths and splitting ratios connecting them to the central office, and FlexPAM formats can help maximise the data rate depending on their specific path loss. In a time division multiplexing (TDM) scenario with flexible time slot allocation, this approach could allow for a higher average speed for all users in the network.

- Nanosecond scale burst mode demonstration using a fully integrated transmitter:

By integrating an external modulator with the tunable laser from Chapter 5 (and potentially an SOA for boosting the output power), a fully integrated transmitter can be obtained. The device could be used for IMDD (with an integrated Mach Zehnder modulator) or coherent transmission (with an IQ modulator) for high-speed reconfigurable optical transmission systems based on PAM or QAM formats respectively.

- Experimental demonstration of the NLE for SOA-amplified systems

Given the promising simulation results from the NLE in Chapter 5 for correcting nonlinear distortions from an SOA close to saturation, it would be relevant to perform an experimental demonstration. Different types of SOAs (e.g. bulk, quantum dot, ...) and different modulation formats (e.g. 16QAM, 64QAM, ...) can be studied as a function of optical power and bias current.

Appendix A

List of Publications Arising From This Work

A.1 Journal Papers

1. Devika Dass, **Marcos Troncoso Costas**, Liam P. Barry, Sean O'Duill, Chris G. H. Roeloffzen, Douwe Geuzebroek, Guillermo Carpintero, Robinson Cruzoe Guzman, and Colm Browning, "28 GBd PAM-8 transmission over a 100 nm range using an InP-Si₃N₄ based integrated dual tunable laser module," Opt. Express 29, 16563-16571 (2021)
2. **M. Troncoso-Costas**, D. Dass, C. Browning, F. J. Diaz-Otero, C. G. H. Roeloffzen and L. P. Barry, "Intra-Data Centre Flexible PAM Transmission System Using an Integrated InP-Si₃N₄ Dual Laser Module," in IEEE Photonics Journal, vol. 14, no. 1, pp. 1-6, Feb. 2022, Art no. 7205106, doi: 10.1109/JPHOT.2021.3128767.
3. A. G. Reza, **M. Troncoso-Costas**, C. Browning, F. J. Diaz-Otero and L. P. Barry, "Single-Lane 54-Gbit/s PAM-4/8 Signal Transmissions Using 10G-Class Directly Modulated Lasers Enabled by Low-Complexity Nonlinear Digital Equalization," in IEEE Photonics Journal, vol. 14, no. 3, pp. 1-9, June 2022, Art no. 7226909, doi: 10.1109/JPHOT.2022.3168683.

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4. **Marcos Troncoso Costas**, Ahmed Galib Reza, Hima Putta, Ankit Sharma, Gaurav Jain, Mark Smyth, Frank Smyth, John MacDougall, Colm Browning, Francisco Diaz Otero, and Liam Barry, "30 Gbit/s PAM4 transmission using an 8-GHz directly modulated multi-section laser," *Opt. Express* 30, 20194-20202 (2022)
 5. Ahmed Galib Reza, **Marcos Troncoso-Costas**, Colm Browning, Sean O'Duill, and Liam P. Barry, "Mitigation of SOA-Induced Nonlinearities With Recurrent Neural Networks in 75 Gbit/s/ λ PAM-4 IM/DD WDM-PON Transmission Systems," *J. Lightwave Technol.* 41, 3967-3975 (2023)
 6. **M. Troncoso-Costas**, G. Jain, Y. Li, M. Patel, L. N. Venkatasubramani, S. O'Duill, F. Smyth, A. Ellis, F. Diaz-Otero, C. Browning, and L. Barry, "Experimental demonstration of 480 Gbit/s coherent transmission using a nanosecond switching tuneable laser," *Optics Communications*, vol. 554, p. 130164, Mar. 2024. [Online]. Available: <https://www.sciencedirect.com/science/article/pii/S0030401823009124>

A.2 Conference Papers

1. **M. T. Costas**, C. Browning, F. D. Otero, A. Ellis and L. Barry, "Dynamic Constellation Adaptive Equalizer for Coherent Optical Transmission Systems," 2021 IEEE Photonics Conference (IPC), Vancouver, BC, Canada, 2021, pp. 1-2, doi: 10.1109/IPC48725.2021.9592974.
2. A. G. Reza, **M. T. Costas**, C. Browning and L. Barry, "High-speed PAM-4 Signal Transmissions with Directly Modulated Lasers for the Next-Generation Passive Optical Networks," 2021 International Conference on Information and Communication Technology Convergence (ICTC), Jeju Island, Korea, Republic of, 2021, pp. 747-749, doi: 10.1109/ICTC52510.2021.9621187.
3. A. G. Reza, **M. T. Costas**, L. Barry and C. Browning, "54-Gbit/s PAM-

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- 8 Transmission in Next-Generation Passive Optical Networks using Directly Modulated Lasers with Machine Learning Techniques,” 2021 13th International Congress on Ultra Modern Telecommunications and Control Systems and Workshops (ICUMT), Brno, Czech Republic, 2021, pp. 278-281, doi: 10.1109/ICUMT54235.2021.9631606.
4. A. G. Reza, **M. T. Costas**, C. Browning, F. D. Otero, and L. Barry, ”67.5 Gbit/s PAM-8 Signal Transmissions Over 25-km SMF With a 1550-nm 10G-Class DML Using Machine Learning,” in Conference on Lasers and Electro-Optics, Technical Digest Series (Optica Publishing Group, 2022), paper SM3J.4.
 5. **M. Troncoso Costas**, A. Galib Reza, C. Browning, F. Diaz Otero and L. Barry, ”High-speed PAM4 Transmission Using a 12.5G Class DML Enabled by Low Complexity DSP,” 2022 Conference on Lasers and Electro-Optics (CLEO), San Jose, CA, USA, 2022, pp. 1-2.
 6. **M. T. Costas**, A. G. Reza, C. Browning, F. D. Otero, and L. Barry, ”Beyond 100 Gbit/s per Wavelength PAM-8 Transmission over 25 km Using Low Complexity Equalization,” in Conference on Lasers and Electro-Optics, Technical Digest Series (Optica Publishing Group, 2022), paper JTh3B.21.
 7. **M. T. Costas**, A. G. Reza, C. Browning and L. Barry, ” 4×75 Gbit/s WDM PAM-8 Transmission Over 25 km for Future Optical Access Communications,” 2022 Photonics North (PN), Niagara Falls, ON, Canada, 2022, pp. 1-1, doi: 10.1109/PN56061.2022.9908330.
 8. A. G. Reza, **M. T. Costas**, L. Barry and C. Browning, ” 4×75 -Gbit/s PAM-4 Transmission for WDM-PON Systems Using Low-Complexity Digital Post-Equalizations,” 2022 27th OptoElectronics and Communications Conference (OECC) and 2022 International Conference on Photonics in Switching and Computing (PSC), Toyama, Japan, 2022, pp. 1-3, doi: 10.23919/OECC/PSC53152.2022.9850121.
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A.3 Patents

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