“Digital coherent transceiver for optical communications. From Design to Implementation”

-Master Thesis-

Esdras Anzuola Valencia

February 2012
“Digital coherent transceiver for optical communications. From Design to Implementation “

Author: Esdras Anzuola Valencia
Director: Aniceto Belmonte Molina
Tutor: Aniceto Belmonte Molina

Departament of Communication and Signal Theory
Technical University of Catalonia (UPC)

February 2012
Keywords

Coherent Detection, Optical Communications, Phase Shift Keying, Shot Noise, Phase estimation, Frequency estimation.

Abstract

Recent coherent optical communication systems address modulation and detection techniques for high spectral efficiency and robustness against transmission impairments. As a consequence of this technical reality and its potential achievements, the proposed objective of this project is to develop and demonstrate the coherent optical infrastructure and signal processing to produce robust high-capacity links over the 1.5-micron spectral band.

In this project we analyze the theoretical models of optical coherent communication systems as well as the front-end architectures used to implement them. Key concepts as balanced photo detection and quantum limit are explained and studied. Complex modulation schemes maximize spectral efficiency and power efficiency by encoding information in two degrees of freedom. Homodyne and heterodyne downconversion are shown to be linear processes that can fully recover the received signal field. When optical downconverted signals are sampled, compensation of transmission impairments can be performed using digital signal processing (DSP). Clock recovery, frequency offset compensation and phase offset compensation algorithms are studied and their performance is shown.

Based on the theory analyzed, an optical coherent transceiver is designed using commercially available devices. A system and device characterization is performed. Implementation effects as bandwidth limitations, laser source deviation and different noise sources are studied. Modulation and demodulation impairments introduced by real devices are analyzed in order to evaluate their penalization over the signal quality at the receiver. The experimental set-up is then implemented and parameterized. The system performance is validated and the receiver robustness is tested under the presence of additive white gaussian noise (AWGN).
Chapter 1. Introduction

1.1. Introduction
1.2. Motivation and thesis objectives
1.3. Thesis organization

Chapter 2. Basic concepts of optical coherent systems

2.1. Optical hybrids and balanced photo detection
2.2. Generic block system for optical coherent systems
2.3. Homodyne detection
2.4. Heterodyne detection
2.5. Modulation formats

Chapter 3. Design of a digital coherent system

3.1. Digital system overview
3.2. Digital transmitter
3.3. Digital receiver
  3.3.1. Carrier frequency estimator
  3.3.2. Clock recovery
  3.3.3. Phase offset compensation
3.4. System performance evaluation block

Chapter 4. Experimental set-up

4.1. System architecture
  4.1.1. Optical transmitter
  4.1.2. Optical receiver
4.2. Device Characterization
4.3. Implementation limits and impairment effects
  4.3.1. Bandwidth limitations
  4.3.2. Laser line width effects
  4.3.3. Central frequency laser deviation and path difference fading
  4.3.4. Modulation imbalance factors
  4.3.5. Effects of asymmetric hybrids
4.4. System parameters

Chapter 5. System validation

Chapter 6. Conclusions and future work

Bibliography

Appendix
GLOSSARY

Analog-to-Digital Converter, ADC:

An analog-to-digital converter is a device that converts continuous signals to discrete digital numbers.

Balanced detection:

A balanced detector, or balanced receiver, is a device that measures the difference in the intensity of two laser beams. It is a part of a receiver front-end architecture that provides better power efficiency for the receiver.

Band-Pass Filter, BPF:

A band-pass filter is a device that passes frequencies within a certain range and rejects (attenuates) frequencies outside that range.

Birefringence:

The double refraction of light in a transparent, molecularly ordered material, which is manifested by the existence of orientation-dependent differences in refractive index.

Bit Error Rate, BER:

Number of erroneous bit per transmitted bits.

Bit Rate, Rb:

The number of bits transmitted per unit of time.

Coherence Time:

The time over which a propagating wave (especially a laser) may be considered coherent. In other words, it is the time interval within which its phase is, on average, predictable.

Digital Signal Processing, DSP:

The group of numerical techniques that treat a digital signal. These techniques are very useful because of their stability due to error detection and correction and their reduced vulnerability to noise. In the design of the digital coherent receiver DSP is used for compensation of chromatic dispersion, polarization mode dispersion and tracking the phase of the received signal.
**Direct Detection Receiver, DD Receiver:**

A receiver that works converting optical power directly to electrical domain. These kinds of detectors are not able to recover the information on the phase of the optical signal.

**Intersymbol Interference, ISI:**

The overlapping between symbols, in which one symbol interferes with subsequent symbols due to a distortion in the received signal that causes the spread of the symbols. In optical communication the distortion that causes this effect is dispersion.

**Local Oscillator, LO:**

Laser located in the receiver. Its lightwave is mixed with the incoming signal in the coherent optical receiver. The mixing lets to obtain the phase of the received signal.

**Mach-Zender modulator, MZ modulator:**

External optical modulator able to modulate phase or and amplitude of an optical lightwave.

**Optical Coherent Detection Receiver:**

A receiver that mixes the incoming optical signal with a local optical source lightwave before converting the signal into electrical domain.

**Optical Coupler:**

A passive device for branching or coupling an optical signal. Generally, a coupler is centralized by joining the two fibers together so that the light can pass from the sender unit to the two receivers, or else it can be made by juxtaposing the two "receiver" fibers which will then be aligned and positioned so as to be facing the "sender" fiber. In our design the optical coupler is used to mix the incoming lightwave and the local oscillator lightwave.

**Phase shifter:**

Passive device that introduces a delay into an optic path. The phase delay can be fixed or simply adjusted with a controllable voltage.
**Phase Diversity Receiver:**

A coherent receiver front-end architecture that is able to extract phase information from the incoming signal by mixing it with a local oscillator signal and it is independent of the phase difference between them.

**Quantum Limit:**

A physical low bound for the BER that an optical can achieve due to the quantum nature of light. In optics, the quantum limit is imposed by the shot noise that occurs when a sufficient small number of photons generate an occurrence of independent and significant random events described by a Poisson distribution.

**Responsivity of photodiode, R:**

A parameter of the photodiode that represents the ability of the device to generate an electron hole when light hits its surface.

**Spectral Efficiency:**

A concept that refers to the information bit rate that can be transmitted over a given bandwidth.

**Shot Noise:**

A fundamental noise mechanism responsible for current fluctuations in all optical receivers even when the incident optical power Pin is constant. It is a manifestation of the fact that an electric current consists of a stream of electrons that are generated at random times.

**Thermal Noise:**

A variable current generated by the photodiode due to the fact that, at a finite temperature, electrons move randomly in any conductor. Random thermal motion of electrons in a resistor manifests as a fluctuating current even in the absence of an applied voltage. Thermal noise sets a fundamental lower limit to what can be measured.
Chapter 1

1.1. Introduction

In this thesis an optical coherent communication system is designed and implemented. These types of systems have some advantages over others based on direct detection, which have been used historically in optical communications. One of the properties of a coherent receiver is that it provides the possibility of using digital signal processing (DSP) once the signal has been detected. Through these methods we are able to compensate the impairments of optical coherent communications as clock recovery, carrier frequency and phase estimation, modulation imbalance impairments or additive Gaussian noise.

In optical communications there are two major kinds of detectors: direct detection and coherent detection. The direct detection is so named because the incoming signal is detected directly with the photodiode which is the element in charge of converting the optical power into a current (Figure 1). These detectors can only obtain the amplitude of the signal, losing its phase. With the direct detection only the amplitude of the signal can be obtained, losing its phase.
Figure 1. Direct detection receiver scheme. The incoming optical signal \( E_s(t) \) is band pass filtered \( E_{fs}(t) \) and its intensity is directly detected with a photodiode, which converts the optical power into a current \( I(t) \).

On the other hand, for coherent detection the incoming light wave is mixed with other light beam coming from the local oscillator (LO) before being detected by the photodiode (Figure 2). The signal detected by the coherent detector preserves both the amplitude and the phase of the signal.

Figure 2. Coherent receiver generic scheme. The coherent detector (CohD RX) uses a local oscillator signal \( E_{lo} \) to convert the optical signal into a currents. These current or set of currents are sent to the Digital Receiver, which is responsible for the demodulation process. The CohD together with the Digital Receiver is called Digital Coherent Receiver.

For coherent detection there are two basic schemes depending on how the downconversion from optical frequencies to baseband frequencies is performed. These schemes are called heterodyne detection and homodyne detection.

In homodyne detection the local oscillator is tuned so that the output of the optical mixer is at baseband frequencies (Figure 3).

In heterodyne detection, a signal of interest at some frequency is non-linearly mixed with a reference local oscillator source (LO) that is set at a close-by frequency. The outcome is centered at the difference frequency, which carries the information in amplitude, phase or frequency of the original higher frequency signal, but oscillating at intermediate carrier frequency (Figure 4) which can be handled easily.
Figure 3. Homodyne signal spectrum. The local oscillator is tuned so that it matches the optical frequency obtaining a baseband current at the output of the coherent receiver.

The translation to baseband frequencies can be performed then using electrical techniques or numerical methods after analog to digital conversion.

Figure 4. Heterodyne signal frequency spectrum. A signal of interest at some frequency is non-linearly mixed with a reference local oscillator source (LO) that is set at a close-by frequency. The outcome of the coherent receiver is the centered at the intermediate frequency, which carries the information in amplitude, phase or frequency of the original higher frequency signal.

Coherent systems have several advantages over direct detection systems, but for many years it has not been developed or used due basically to two reasons: the complexity of the optical and digital systems used in coherent communications and the reasonable bandwidth provided by the direct detection, enough for the requirements of many applications.

But coherent optical detection systems have important potential advantages over direct detection methods: a greater wavelength selectivity, increased sensitivity in the reception stage [1], and so on. In theory this achieves greater distances in optical links [2] plus a higher spectral efficiency [3].

During the mid 80’s and 90’s of last century there was a great activity on the research and development of optical coherent communications systems, which decreased gradually due mainly to the appearance of optical amplifiers and to the great technological limitations imposed by phase noise of optical sources [4]. Recently,
however, has revived the interest in such systems [5], on a quest to increase the capacity and in view of new technological developments in the area of optical sources, balanced photoreceptors, digital processing of high-speed signals [6] and applying innovative techniques of coding and synchronization. The current trend in coherent optical communications is primarily oriented to the digital processing and compensation of phase perturbations in optical systems with phase modulation [7][8].

1.2. Motivation and Thesis Objectives

It has been shown that there are many advantages that coherent reception can bring to optical communications [2]. The present project has as its main objective the design, implementation and test of an optical coherent transceiver.

The following objectives are targeted to be accomplished in this project:

- Describe and study the basic concepts involved in coherent communications.
- Analyze and model the optical coherent communication system as well as the front-end architectures used to implement the homodyne and heterodyne detection methods.
- Study the theoretical limits and performance of an optical coherent communication system using different modulation formats.
- Implementation and performance study of the different compensation techniques needed in a real optical coherent receiver such as:
  - Clock recovery
  - Frequency offset estimation
  - Phase offset estimation
- Design of an experimental set-up in order to implement an optical coherent communication system using commercially available devices.
- Characterize the physical devices involved on the system, as well as measure and quantify the impairments introduced by these non-ideal devices.
- Study the implementation limits introduced by the devices used as:
  - Bandwidth limitations
  - Laser source deviation.
  - Modulation imbalance factors
  - Asymmetric demodulation factors.
- Implementation of an optical coherent communication system.
Test our communication system and study its performance in the presence of additive white gaussian noise (AWGN).

1.3. Thesis Organization

In Chapter 2 the basics of optical coherent communications systems will be described, explaining the main concepts of homodyne and heterodyne detection, and exposing the theory and the main differences between the two coherent methods. Also some concepts about modulation formats, shot noise and quantum limit will be introduced.

In Chapter 3, the coherent system algorithms will be studied. We will indicate the compensating modules that must be present on a real coherent system as well as their theoretical fundamentals. The performance of each module will be tested in different scenarios in order to characterize their response in real systems.

In Chapter 4 we will describe the experimental set-up, showing the system overview, device characterization, hardware limitations and impairments and system parameters.

In Chapter 5 we present our analysis results. It is dedicated to study each module involved in the system. We will test their individual properties and we will characterize the whole system performance in different scenarios.

In Chapter 6 the conclusions obtained are presented and the ideas for future work on coherent systems are extracted from the project.
Chapter 2 – Basic Concepts of Optical Coherent Systems

In this section the basics of coherent detection are explained. Concepts about homodyne and heterodyne systems as well as their mathematical models are covered. The different system architectures will be shown and we will describe the main advantages of coherent detection over direct detection methods, such as sensitivity and system performance.

We will also discuss the modulation formats used for this project, as well as a brief description of the signal nature and spectrum, in order to better understand effects explained on further chapters.
2.1. Optical Hybrids and Balanced Photo detection

There are two important devices that must be explained before describing how a coherent system is designed. First, an optical hybrid is generally described as a device that is able to mix light beams. Second, balanced photo detectors are devices which optimize the translation of optical fields into currents.

2.1.1. Optical Hybrids

In coherent systems we need an optical hybrid in order to mix an optical signal with a local oscillator source. The simplest hybrid is a 3dB coupler, also called 50/50 beam splitter or 180° hybrid. This device can be modeled as shown in Figure 5.

\[
E_{o1}(t) = E_{i1}(t) + jE_{i2}(t)
\]

\[
E_{o2}(t) = jE_{i1}(t) + E_{i2}(t)
\]

*Figure 5. Optical 180° hybrid. Each input field is split into two output ports. A phase shift of 180 degrees is introduced in one of the branches.*

Here, each one of the two input fields is split into the two output ports and a phase shift of 180 degrees is introduced in one of the branches. Mathematically this device can be modeled as:

\[
\begin{bmatrix}
E_{o1} \\
E_{o2}
\end{bmatrix} = \frac{1}{\sqrt{2}} \begin{bmatrix}
1 & j \\
j & 1
\end{bmatrix} \begin{bmatrix}
E_{i1} \\
E_{i2}
\end{bmatrix}
\]

where \(E_{i1}, E_{i2} \ [V/m]\) are the input fields and \(E_{o1}, E_{o2} \ [V/m]\) are the output fields. Its diagram is shown in figure 5.

A 90° optical hybrid is a six-port device that is used for coherent signal demodulation (*Figure 6*). It would mix the incoming signal with the four quadrature states associated with the reference signal in the complex-field space.
2.1.2. Balanced photo detection

Balanced photo detection has been used due to its high sensitivity in compare to simple detection. The reason is that it is able to measure or detect low power signals where the dominant noise is additive and it is present in both branches [10].

In an optical coherent system, the use of balanced photo detectors allows reducing or eliminating the noise from the electrical signal, as well as it enables to maximize the use of the optical power generated by the local oscillator. The balanced configuration consists on using photo detectors with identical quantum efficiency in each one of the output coupler ports. The resulting signals can be subtracted in order to eliminate the noise present in both branches (Figure 7).

Furthermore, the influence of the fluctuations of the input signal power is not really significant because it is negligible in compare to the local oscillator power. But, in the other hand, the intensity noise from the local oscillator can strongly diminish the quality of the output intensity [11]. Using a balanced detection, a \( \pi \) factor is introduced.
in one of the branches, so that the common signal is subtracted and the differential signal is amplified.

Also, in a simple photo detector a significant part of the power is lost. In the balanced photo detector almost all the power is exploited, increasing the receptor sensitivity in compare to a simple detector [12].

In practice, the performance of a balanced photo detector is not perfect due to differences on the detectors responsitivities, electrical path length or other non-idealities. So the performance is characterized in terms of the CMRR (Common Mode Rejection Ratio), which is defined as the capacity of attenuate the common terms and amplify the differential ones [13].

\[
CMRR = 20 \log \left( \frac{A_d}{A_s} \right) \quad [dB]
\] (2.2)

Here \(A_d\) is the differential gain and \(A_s\) is the common gain.

2.2. Generic Block System for Coherent Communications

The most basic idea on a coherent system is that, in the reception stage, the modulated optical signal is mixed with a local oscillator. Figure 2.1 shows the basic block diagram of a coherent system.

In transmission, the light coming from the laser source is shaped by an electrical signal by the modulator. This electrical signal is generated based on the data to be transmitted and on the modulation format used. The transmitted optical signal \(E_S(t)\) \([V/m]\) can be represented as:

\[
E_S(t) = \sqrt{P_S} e^{i\phi_S} e^{iws_t}
\] (2.3)
where $P_S [W], w_S [rad/s]$ and $\phi_S [rad]$ are the power, frequency and phase of the optical signal, respectively.

The signal is sent through the channel and in the receiver stage is mixed with the light generated by the local oscillator (LO), which can be represented as:

$$E_{LO}(t) = \sqrt{P_{LO}} \cdot e^{j\phi_{LO}} \cdot e^{jw_{LO}t}$$  \hspace{1cm} (2.4)

Here, $P_{LO} [W], w_{LO} [rad/s]$ and $\phi_{LO} [rad]$ are the power, frequency and phase of the local oscillator signal [9].

Both signals are mixed in an optical hybrid and its outputs are sent to the digital receiver. As we have mentioned in the introduction, depending on the relationship between the optical frequencies of the signal and the local oscillator we will have two different demodulation schemes. In heterodyne detection $w_{LO} \neq w_S [rad/s]$ and the downconversion is performed in two separated steps. The coherent detector is a $180^\circ$ hybrid which outputs are converted into a current $I_{Het}(t) [A]$ by one single balanced photo detector. The output intensity is centered at a frequency $w_{IF} = w_S - w_{LO}$ [rad/s]. This current is digitally processed and the baseband conversion is performed by a digital downconversion. The resulting basebands currents are $I_{Het,i}(t) [A]$ and $I_{Het,Q}(t) [A]$, which correspond to the in-phase and quadrature branches of a complex demodulator (Figure 9).

![Figure 9. Coherent Heterodyne demodulation scheme. The incoming optical signal $E_s(t)$ is mixed with the local oscillator light beam $E_{lo}(t)$ in a $180^\circ$ Hybrid. The outputs of the coupler are connected to a balanced photo detector, which generates a current $I_{het}(t)$. An analog to digital converter transforms $I_{het}(t)$ that is translated to baseband frequencies by a digital oscillator.](image)

In homodyne detection both frequencies are exactly matched ($w_{LO} = w_S [rad/s]$) and the downconversion is achieved in one single step. The coherent detector is a $90^\circ$ hybrid which output fields are converted into two currents, $I_{Hom,i}(t)$ and $I_{Hom,Q}(t)$, generated by two balanced photo detectors. The output baseband signals are sent to the digital receiver (Figure 10).
Figure 10. Coherent Homodyne demodulation scheme. The incoming optical signal $E_s(t)$ is mixed with the local oscillator light beam $E_{lo}(t)$ in a 90° Hybrid, which is composed by four 180° hybrids and a 90° phase shifter. The outputs of the hybrid are connected to two balanced photo detectors, which generate two baseband currents $I_{\text{hom},i}(t)$ and $I_{\text{hom},q}(t)$. An analog to digital converter transform both currents into a digital signal, which can be processed by the digital demodulator.

Both schemes are able to demodulate correctly the transmitted signal but they accomplish the task in two different ways, which means that after this point we will have to study both systems separated.

### 2.3. Homodyne detection

In this coherent detection technique, the local oscillator frequency $w_{LO}$ is chosen to match exactly the information signal frequency.

The incoming optical signal $E_s(t)$ [V/m] is mixed in the receiver with the local oscillator signal $E_{LO}(t)$ [V/m] using a 90 degrees hybrid. This hybrid is formed by four hybrids of 180° and a 90 degrees phase shifter.

Figure 11. Coherent homodyne optical downconversion scheme. The incoming optical signal $E_s(t)$ and local oscillator signal $E_{lo}(t)$ enter the 90° hybrid through ports 1 and 2. The output fields of the hybrid, corresponding to ports 5, 6, 9 and 10, are translated into the baseband currents $I_{\text{hom},i}(t)$ and $I_{\text{hom},q}(t)$ by two balanced photo detectors.
In Figure 11 the homodyne downconversion scheme is shown. For each port there is an electrical field associated. The ports 1 and 2 are the device inputs, which correspond to the incoming modulated field and the local oscillator field:

\[ E_1(t) = E_S(t) = \sqrt{P_S} e^{j\phi_S} e^{j\omega_S t} \]  
\[ E_2(t) = E_{LO}(t) = \sqrt{P_{LO}} e^{j\phi_{LO}} e^{j\omega_{LO} t} \]  

Considering the ideal coupler equation,

\[ \begin{bmatrix} E_{d1} \\ E_{d2} \end{bmatrix} = \frac{1}{\sqrt{2}} \begin{bmatrix} 1 & j \\ j & 1 \end{bmatrix} \begin{bmatrix} E_{i1} \\ E_{i2} \end{bmatrix} \]  

the electrical intermediate electrical fields at ports 3 and 4, \( E_3(t) \) [V/m] and \( E_4(t) \) [V/m], can be expressed as:

\[ E_3(t) = \frac{1}{\sqrt{2}} E_s(t) \]  
\[ E_4(t) = \frac{1}{\sqrt{2}} E_{LO}(t) \]  

So the resulting output fields of the coupler, \( E_5(t) \) [V/m] and \( E_6(t) \) [V/m], can be expressed as:

\[ E_5(t) = \frac{1}{\sqrt{2}} (E_3 + jE_4) = \frac{1}{2} (E_s(t) + jE_{LO}(t)) \]  
\[ E_6(t) = \frac{1}{\sqrt{2}} (jE_3 + E_4) = \frac{1}{2} (jE_s(t) + E_{LO}(t)) \]  

From equation 2.3, the photo detector intensities \( I_5(t) \) [A] and \( I_6(t) \) [A] can be obtained as:

\[ I_5(t) = R |E_5|^2 = \frac{R}{4} (|E_s|^2 + |E_{LO}|^2 + 2Im\{E_s E_{LO}^*\}) + I_{n,5}(t) \]  
\[ I_6(t) = R |E_6|^2 = \frac{R}{4} (|E_s|^2 + |E_{LO}|^2 - 2Im\{E_s E_{LO}^*\}) + I_{n,6}(t) \]  

Here, \( I_{n,1}(t) \) [A] is the LO shot noise associated to each photodiode, which have been modeled as additive zero-mean Gaussian noise [15]. Assuming \( P_{LO} \gg P_S \) [W], \( I_{sh,3}(t) \)
[A] has a two sided psd of \( S_{n,t}(f) = q R_i \frac{E_{LO}}{2} \left| \frac{A^2}{N_0} \right| \), where \( q \) is the elementary charge of an electron and \( R_i \) [A/W] is the photo detector responsivity. This fact means that we are working on the shot noise limited scenario, where we can neglect other noise sources as thermal noise and dark current photo detector noise [14].

Using the same procedure, we can obtain the electrical fields at the intermediate ports 7 and 8 of the device.

\[
E_7(t) = \frac{j}{\sqrt{2}} j E_1(t) = -\frac{1}{\sqrt{2}} E_5(t) \tag{2.14}
\]
\[
E_8(t) = \frac{j}{\sqrt{2}} E_2(t) = \frac{j}{\sqrt{2}} E_{LO}(t) \tag{2.15}
\]

Applying equation 2.9, the output electrical fields at ports 9 and 10 can be expressed as:

\[
E_9(t) = \frac{1}{\sqrt{2}} (E_7 + j E_8) = \frac{1}{\sqrt{2}} \left[ -\frac{1}{\sqrt{2}} E_5(t) - \frac{1}{\sqrt{2}} E_{LO}(t) \right] \tag{2.16}
\]
\[
E_{10}(t) = \frac{1}{\sqrt{2}} (j E_7 + E_8) = \frac{1}{\sqrt{2}} \left[ -\frac{j}{\sqrt{2}} E_5(t) + \frac{j}{\sqrt{2}} E_{LO}(t) \right] \tag{2.17}
\]

From these equations we can obtain the intensities \( i(t) = |E|^2 \) generated by the photo detectors placed at the output of ports 9 and 10:

\[
I_9(t) = R |E_9|^2 = \frac{R}{4} (|E_s|^2 + |E_{LO}|^2 + E_s E_{LO}^* + E_{LO} E_s^*) + I_{n,9}(t)
\]
\[
= \frac{R}{4} (|E_s|^2 + |E_{LO}|^2 + 2\text{Re}\{E_s E_{LO}^*\}) + I_{n,9}(t) \tag{2.18}
\]

\[
I_{10}(t) = R |E_{10}|^2 = \frac{R}{4} (|E_s|^2 + |E_{LO}|^2 - E_s E_{LO}^* - E_{LO} E_s^*) + I_{n,10}(t)
\]
\[
= \frac{R}{4} (|E_s|^2 + |E_{LO}|^2 - 2\text{Re}\{E_s E_{LO}^*\}) + I_{n,10}(t) \tag{2.19}
\]

Using balanced detection at the output of both couplers, \( I_{hom,j}(t) \) [A] and \( I_{hom,q}(t) \) [A] can be expressed as:

\[
I_{hom,j}(t) = R \sqrt{P_s} \sqrt{P_{LO}} * \cos(\phi_s - \phi_{LO}) + I_{sh,j}(t) \tag{2.20}
\]
\[
I_{hom,q}(t) = R \sqrt{P_s} \sqrt{P_{LO}} * \sin(\phi_s - \phi_{LO}) + I_{sh,q}(t) \tag{2.21}
\]
where $I_{sh,i}(t)$ [A] and $I_{sh,q}(t)$ [A], considering equal responsivities and independent noise for each photodiode, are white zero-mean additive noises with two-sided PSD of $[15]$: 

$$S_{sh}(f) = qR \frac{P_{LO} A^2}{2} \frac{1}{Hz}$$  \hspace{1cm} (2.22)$$

Since it can be shown that thermal noise is always negligible compared to shot noise [9], and that we will not consider ASE noise in our system (no optical amplification will be used), we have neglected this terms in equations (2.20) and (2.21)

Finally, the two resulting intensities are the in-phase and quadrature components of the transmitted signal, which allows the use of complex modulation formats and the associated bandwidth gain. As we can see, the intensities on both branches are directly proportional to the local oscillator power, which allows improving the SNR at the receiver and reach the shot noise limit [9].

In the other hand, the control of the local oscillator phase $\phi_{LO}$ [rad] is needed in order to extract all the information from the incoming signal. This is because, as we see in equations (2.20)(2.21), the presence of this term will affect directly the output photo detector currents. This means that an additional module is needed in order to compensate this effect.

Also the local oscillator optical frequency has to match exactly the signal optical frequency, condition that introduces strict hardware design requirements. Some of these problems are avoided using heterodyne detection, which will be discussed in the next section.

2.4. Heterodyne detection

In heterodyne detection the local oscillator frequency $w_{LO}$ [rad/s] is not matched with the signal central frequency $w_s$ [rad/s] so the detected signal is present around an intermediate frequency $w_{f_i}$ [rad/s] (Figure 12).
Figure 12. Signal spectrum heterodyne optical downconversion. The incoming optical signal \( E_s(w) \) centered at \( w_s \) is translated from optical frequencies (THz) to an intermediate frequency by mixing it with a local oscillator source centered at \( w_{LO} \). After photo detection, \( I_{het}(w) \) is obtained.

The equation analysis is based on the demodulation scheme that uses an optical coupler, which mixes the incoming signal and the local oscillator generated field (Figure 13).

\[
E_3(t) = \frac{1}{\sqrt{2}} (E_1 + jE_2) = \frac{1}{\sqrt{2}} (E_s + jE_{LO})
\]

(2.23)

\[
I_3(t) = R \cdot |E_3(t)|^2 = \frac{R}{2} (|E_s|^2 + |E_{LO}|^2 + 2 Re\{E_2E_{LO}^*\}) + I_{sh,3}(t)
\]

(2.24)

Here, \( I_{sh,3}(t) \) [A] is the LO shot noise associated to the first photodiode and it has a two sided PSD of \( S_{sh,3}(f) = qR P_{LO} \left[ \frac{A^2}{Hz} \right] \) assuming a shot noise limited scenario. We get \( I_4(t) \) [A] using the same analysis:

\[
E_4(t) = \frac{1}{\sqrt{2}} (jE_1 + E_2) = \frac{1}{\sqrt{2}} (jE_s * e^{j\omega_st} + E_{LO} * e^{j\omega_{LO}t})
\]

(2.25)
\[ I_4(t) = R \cdot |E_4(t)|^2 = R \left( E_4(t) \ast E_4^*(t) \right) \]

\[
= \frac{R}{2} \left( |E_s|^2 + |E_{LO}|^2 + jE_sE_{LO}^* - j(E_sE_{LO}^*)^* \right) 
= \frac{R}{2} \left( |E_s|^2 + |E_{LO}|^2 - 2\text{Im}\{E_sE_{LO}^*\} \right) + I_{sh,4}(t) \tag{2.26}
\]

Applying the balanced photo detection:

\[ I_{het}(t) = I_3(t) - I_4(t) = 2R \ast \text{Im}\{E_sE_{LO}^*\} + I_n(t) \tag{2.27} \]

where \( I_n(t) = I_{sh,3}(t) - I_{sh,4}(t) \) is approximated as a zero-mean Gaussian noise.

Assuming that we are working on the shot noise limit, the noise PSD can be expressed as [16]:

\[ S_{In}(f) = qR(P_{LO} + P_S) \left[ \frac{A^2}{Hz} \right] \equiv qRP_{LO}\left[ \frac{A^2}{Hz} \right] \tag{2.28} \]

The PSD’s of the two noise processes can be added because they are generated by different photodiodes, which are independent. Substituting in (2.29), we get:

\[ I_{het}(t) = 2R\sqrt{P_S} \sqrt{P_{LO}} \ast \text{Im} n[(\omega_{RF} t + (\phi_S - \phi_{LO})] + I_n(t) \tag{2.29} \]

It is feasible to send information in amplitude, phase and frequency using this signal. Also, we can still use the local oscillator power to amplify the received signal improving the SNR. After optical down-conversion to an intermediate frequency, we need an electrical demodulation in order to extract the I and Q components. For that purpose, the widely known scheme Figure 14 is used:

**Figure 14.** Digital demodulation scheme for coherent heterodyne. The input current \( I_{het}(t) \) is multiplied by a sine signal \( I_{RF}(t) \) generated by a digital RF oscillator. After low pass filtering, the baseband in-phase (I) current \( I_{het, I}(t) \) is generated. To generate the quadrature (Q) current \( I_{het, Q}(t) \), the incoming current \( I_{het}(t) \) is multiplied by \( I_{RF}(t) \) shifted \( \pi/2 \) [rad] and low pass filtering is applied.

Here, the different intensities can be expressed as:

\[ I_{RF}(t) = I_{RF} \ast \sin (\omega_{RF} t + \phi_{RF}) \tag{2.30} \]

\[ I_{Het, I}(t) = I_{Het}(t) \ast I_{RF}(t) \tag{2.31} \]
Assuming a perfect frequency match between \( w_{RF} \) [rad/s] and \( w_{IF} \) [rad/s] we can express the resulting baseband signals and after low pass filtering we obtain the intensities for each branch:

\[
I_{HET,I}(t) = I_{RF} \ast \sin(w_{RF}t + \varphi_{RF} - \frac{\pi}{2})
\]

(2.32)

\[
I_{HET,Q}(t) = I_{RF} \ast \cos(\varphi_{S} - \varphi_{LO} - \varphi_{RF}) + I_{sh,I}(t)
\]

(2.33)

\[
I_{HET,Q}(t) = I_{RF} \ast \sin(\varphi_{S} - \varphi_{LO} - \varphi_{RF}) + I_{sh,Q}(t)
\]

(2.34)

where \( I_{sh,I}(t) \) [A] and \( I_{sh,Q}(t) \) [A] have a resulting PSD of \( S_{sh}(f) = qR \frac{P_{LO}}{2} \frac{A^{2}}{Hz} \). This means that the resulting baseband signals for the heterodyne case have exactly the same expression that the homodyne cases, as well as all the noises have the same PSD's. Hence, the heterodyne and the two-quadrature homodyne down converters have the same performance [17].

A difference between heterodyne and homodyne down conversion only occurs when the transmitted signal occupies one quadrature (e.g. 2-PSK) and the system is LO shot noise limited. This fact enables the use of a single-quadrature homodyne down converter that uses exactly the same scheme as the heterodyne case, which implies that the signal term is doubled (four times the power), while the shot noise power is only increased by two [18], obtaining a sensitivity improvement of 3 dB compared to the heterodyne and two-quadrature case.

The main advantage of using a heterodyne optical system is that we are using only one balanced photo detector and the simple 180° optical hybrid is used. However, the photocurrent in the heterodyne case has a larger bandwidth than the homodyne case due to the fact that the information signal is modulated in an intermediate frequency. Typically, this frequency \( w_{IF} \) is chosen to be close to the signal bandwidth (BW), which implies a total required bandwidth of twice the bandwidth for the heterodyne case. This implies an extra bandwidth requirement, doubling the required bandwidth of the homodyne case.

### 2.5. Modulation formats

As we saw in the previous chapter, coherent receivers maintain the phase of the received signal, so information can be sent in the phase, the frequency or in the amplitude of the signal. Many studies have been performed and the discussion of
the modulation formats has been widely covered. Our objective in this project is to
develop a practical system that offers us the best advantages of a coherent system,
so we need to define the key parameter in order to choose our modulation format.

Typical parameters are the bit-error-rate (BER), sensitivity and spectral
efficiency. Having a look to previous work [19], [20], [21] we obtain the following
results shown in Figure 15.

![Figure 15. Sensitivity (photons/bit) VS. Modulation formats.](image)

As we see, PSK and QAM modulation formats exhibit advantages in all the
parameters considered.

Phase-shift keying (PSK) is a digital modulation scheme that conveys data by
changing or modulating the phase of a reference signal. In the case of PSK format the
information is coded on the phase of the optical signal while the amplitude and
frequency are kept constant.

For binary PSK, the phase takes two values: 0 and π depending on the bit
transmitted. The sensitivity of such modulation is even better than the quantum
limit. An interesting aspect of PSK modulation is that the optical intensity remains constant and therefore, the amplitude decision thresholds remain constant.

![Figure 16. M-PSK IQ constellations. A) M=2 b) M=4](image)

The use of the PSK format requires that the phase of the optical carrier remains stable so that the phase information can be extracted at the receiver without ambiguity; this requirement puts a stringent condition on the tolerable line widths of the lasers involved. The laser line width can be seen as an additive phase noise, represented by $\varphi(t)$, at the output of a noise free modulator:

$$E(t) = A \cos (\omega t + \phi_S(t) + \varphi(t)) \quad [W]$$

(2.35)

where $\phi_S(t) [rad]$ is the modulated phase and $A$ is the signal amplitude. In order to be able to extract any phase information we need that $\varphi(t) [rad]$ fulfill some specific requirements that will be explained in chapter 4.
Chapter 3 – Design of a Digital Coherent System

3.1 Digital System Overview

In order to implement an optical coherent communication system first we need to study and design the digital transmitter and digital receiver architecture, as well as the methods that will allow us to study the system performance. Based on this principle, we will generate a digital system that contains:

- Digital transmitter, responsible for data and signal generation.
- Digital receiver, responsible for impairment compensation and data demodulation.
- Digital control system, responsible for error counting and performance evaluation.
3.2. Digital transmitter

The digital transmitter is the block responsible for data and signal generation. First, a pseudo-random sequence of data is generated and it will act as the information message to be transmitted. Then a unique word preamble is added. This preamble is needed in order to detect the beginning of the message and to design an ambiguity resolution circuitry at the receiver that will help us with the demodulation (see 3.3). The data bits are grouped depending on the modulation order \((M)\). Each group of data generates a particular signal, called symbol, which depends on the modulation format chosen. In our project complex modulations are of special interest, so at the output of the signal generator two signals will be produced: the in-phase and quadrature signals.

In the homodyne case these signals will be sent through different output channels due to their baseband nature (Figure 18). If not, an overlap would be produced with the consequent lost of information.

For the heterodyne case the same procedure for data generation is applied, but instead of sending the data through different channels, the I and Q branches are upconverted by multiplying each one of them by the sine and cosine of the selected
carrier frequency $w_c \text{[rad/s]}$, respectively. Both branches are then added and the resulting signal is sent through the channel to the receiver (Figure 19).

![Figure 19. Heterodyne Digital transmitter. Signals are generated using the same procedure than in the homodyne case. When the in-phase $S_i[n]$ and quadrature $S_q[n]$ signals are generated, they are multiplied by the cosine and sine of the carrier frequency $w_c$. Both branches are then added and the resulting signal is sent through the channel to the receiver.]

In our project we will use no return to zero (NRZ) signaling, which implies that the generated waveform is suppressed carrier in nature [22]. The power spectrum of the baseband signals generated for each branch in the homodyne case is shown in Figure 20.

![Figure 20. Power spectrum of NRZ signaling. $T_b$ is the symbol time and $L$ is the pattern length.]

In the heterodyne case the signal spectrum is exactly the same, but centered around $w_c$ instead of around the zero frequency.

### 3.3. Digital Receiver

The demodulation process of a coherent system can be divided into several major sections (Figure 21). First, since the incoming waveform is suppressed carrier in nature coherent detection is required. This means that the first action must be recover the intermediate frequency (heterodyne case) in order to translate the modulated data
into the baseband frequencies or compensate any frequency drift between the central frequency of the signal and the central frequency of the local oscillator (homodyne case). Next, the input waveform is multiplied by the estimated carrier frequency, which allows us to derive the clock-synchronization information that will indicate us the optimal sampling points.

![Figure 21. Digital coherent receiver block system.](image)

Once we have the sampled data, we can work with the obtained points to estimate the phase shift introduced by the constant phase difference between the transmitter source and the optical and electrical sources at the receiver. After preamble decoding, the demodulator is able to translate the data into bits, which will be compared to the transmitted bits in order to get the number of error, obtaining this way the bit error rate of the system.

3.3.1. Carrier Frequency Estimator

Ideally, the carrier frequency of a system could be parameterized and this block would be unnecessary. The receiver would know the exact frequency that the transmitter is using and no error would be produced. In practice, we find that the received frequency doesn’t match exactly the transmitted frequency. This offset is generated by the difference between the transmitter and local oscillator laser, as well as differences in the clock rates between the two systems.

A frequency offset produces a phase shift on the constellation and it introduces errors, leading in most of the times to a complete loss of the data (Figure 22).
The demodulated signal for both heterodyne and homodyne cases, can be explained through the expressions obtained in the previous chapter modeling the frequency mismatch as $\Delta \omega$:

$$I_i(kT_S) = A_i \cdot \cos[\Delta w k T_S + (\phi_{s,k} - \Delta \phi_T(kT_S))] + I_n(kT_S) \ [A]$$

(3.1)

where $\phi_s = n \frac{2\pi}{M} + \phi_0 \ [rad]$ is the modulated phase corresponding to a M-PSK system, $A_i$ is the intensity amplitude, and $\Delta \phi_T(kT_S) \ [rad]$ is the total phase offset, which will be covered in the next section. Here we have to introduce the assumption that the line widths of the lasers are very narrow, so we can consider $\Delta \phi_T \ [rad]$ constant over a number of symbol periods so it becomes time independent. In that expression $k$ indicates the number of samples taken by the sampler and $T_S$ indicates the sampling time.

We can easily notice that the phase drift corresponds to a constant frequency offset, which can be expressed as:

$$\Delta \varphi = 2\pi \Delta f T_S \ [rad]$$

(3.2)

This expression helps us to express the frequency drift in terms of the phase variation of consecutive samples. We can calculate the frequency offset as an average of $K-1$ data points containing the phase difference between samples. It has to be noticed that each symbol contains a phase jump inherent to the phase modulation, which will have to be discard in order to neglect the modulation effect. For that purpose we will use the phase increment estimation algorithm proposed in [24], which provide a good performance for real-time communication systems.

This estimator is based on the multiply-filter-divide concept, which takes advantage of the self-multiplication in order to remove the modulation components. It works following the scheme shown in Figure 23.
Figure 23. Multiply-Filter-Divide estimator \((M=4)\). The input signal phase is multiplied by the modulation order \((M)\) to neglect the modulation factor. Then it is band-pass filtered around the frequency \(M^*w_c\). The frequency obtained is divided by the modulation order in order to obtain the carrier frequency.

The input signal is power to the \(M\)-order, which deletes the modulation phase \(\phi = \frac{2\pi}{M}\) [rad] by trigonometrical identity, turning the modulation phase into a \(2\pi\) modulus phase. Then, a band pass filter is applied around \(f_{BP} = M \times f_{IP} \) [Hz], which contains the pure carrier component multiplied by \(M\). Then we will have just to divide the frequency in order to extract the exact carrier frequency and use it to demodulate the incoming signal.

3.3.2. Clock Recovery

The common receivers are configured to select a concrete number of samples depending on their sampling frequency and frequency modulation in order to find the correct sampling points. A reference clock difference between transmitter and receiver can produce a misalignment on the symbol duration having:

\[
T_{Symbol,TX} = N \times T_{s,TX} \ [s] \quad (3.3)
\]
\[
T_{Symbol,RX} = M \times T_{s,RX} \ [s] \quad (3.4)
\]

where \(T_{Symbol,TX} \neq T_{Symbol,RX}\). Many solutions have been proposed and studied, many of them based on edge detectors that look for transitions between zero and one [25]. Generally, these methods are based on statistical parameters, which introduce some uncertainty, especially in noisy scenarios [25]. In our project we will consider an easy and robust method, based basically on the previous characterization of the clock difference of transmitter and receiver. In the other hand, we will have no flexibility to change the transmitter or receiver without changing the system parameters.

Our method is based on the previous calculation of the coefficient \(T_{s,TX}/T_{s,RX}\), which is used in the receiver to resample the incoming waveform:
This implies that the incoming waveform will have to be re-sampled in order to have the same number of samples that the transmitter. This way, the symbol durations in both places are matched. We can neglect the noise added by the re-sampling process if we consider that the sampling frequency is much higher than the data rate. Finally we obtain the re-sampling frequency as:

\[
f_{s, RX} = K \cdot f_{s,TX} \cdot \frac{M}{N} \text{ [Hz]}
\]  \tag{3.6}

3.3.3. Phase offset compensation

Since the line widths of the lasers used in this project are small (10 KHz) in compare to the modulation bandwidth, the variations of the carrier phase are much slower than the modulated phase. Therefore, as we saw in the section 3.2, by averaging the carrier phase over many symbol intervals lead us to an accurate demodulation, where we obtain the baseband signals for the heterodyne and homodyne case. In this section we will consider that the frequency estimation is ideal, leading to the general expressions:

\[
I_i(t) = R \sqrt{P_S} \sqrt{P_{LO}} \cdot \cos[\varphi_S - \Delta \varphi_T] + I_{n,i}(t) \tag{3.7}
\]

\[
I_Q(t) = R \sqrt{P_S} \sqrt{P_{LO}} \cdot \sin[\varphi_S - \Delta \varphi_T] + I_{n,Q}(t) \tag{3.8}
\]

where \(\varphi_S\) is the modulated carrier phase and \(\Delta \varphi_T\) is the phase offset introduced by the total phase difference of the system sources. This offset is calculated differently for homodyne and heterodyne case due to the fact that, as we already mentioned, the heterodyne case impose an additional downconversion, which adds an additional phase offset coming from the electrical oscillator:

\[
\Delta \varphi_{T,het} = \varphi_0 - \varphi_{LO} - \varphi_{RF} \tag{3.9}
\]

\[
\Delta \varphi_{T, hom} = \varphi_0 - \varphi_{LO} \tag{3.10}
\]

where \(\varphi_0 [\text{rad}]\) is the reference phase of the laser source in transmission, \(\varphi_{LO} [\text{rad}]\) is the phase reference of the laser source at the local oscillator, and \(\varphi_{RF} [\text{rad}]\) is the reference phase of the recovered intermediate carrier, which only makes sense in the heterodyne case. So the objective of this block is to calculate \(\Delta \varphi_T [\text{rad}]\) in order to be
able to extract the absolute modulated phase $\Phi_S$ [rad], which contains the information. For that purpose, we can digitally reconstruct the signal samples as:

$$Z(t) = I_p(t) + jI_Q(t) = R\sqrt{P_S}\sqrt{P_{LO}} e^{j(\Phi_S - \Delta\Phi_T)} + n(t) \quad (3.11)$$

The random process $n(t)$ models the shot noise in the system and it is associated with a complex zero mean Gaussian distribution characterized by $N(0,\sigma^2)$ [26]. In order to obtain the phase offset, the M-PSK data modulation can be deleted by calculating $Z(t)^M$, since $M \cdot \Phi_S = 2\pi n + \pi$ [rad]. If we define $\Phi_{est}$ [rad] as the phase offset calculated to compensate $\Delta\Phi_T$ and considering no noise effect we would have that:

$$\Phi_{est} = \frac{1}{M} arg\{Z(t)^M\}^* \quad (3.12)$$

Here, all the information required to calculating the phase offset is at hand. Unfortunately, there are two sources of error to consider here. The factor $1/M$ in this operation introduces an M-fold phase ambiguity [23], which can be eliminated using a correction scheme that uses a known sequence that acts as a data preamble. A second error source is the mentioned shot noise present in each sample, which distorts the phase estimation. To mitigate this effect, a filter is employed [26]. The filter implements a phase averaging over the entire block of symbols selected, acting as a low pass filter. The phase estimate is, thus, given as:

$$\Phi_{est,LP} = \frac{1}{M} arg\left\{\sum Z(t)^M\right\}^* [rad] \quad (3.13)$$

This shot noise can be compensated by the filter, but this estimation will inherently introduce some error on the phase estimation. It is obvious that the higher number of samples we consider, the better estimation we will get. But in practical systems we cannot consider as many samples as we want due basically to two reasons. First, it is the line width laser limitations, which constrains the time when the phase offset $\Delta\Phi_T$ [rad] can be considered constant in time. If the acquisition time is higher than the correlation time of the laser, the estimation will reduce the phase accuracy of each sample instead of increasing it. Second, real-time systems have limited memory and calculation power, which impose a physical limitation to achieve the calculation rate needed.
3.3. System performance evaluation block and error counting

A digital control between transmitter and receiver is needed in order to evaluate the performance of the system. After the signal is demodulated by the receiver, the data obtained is compared to the data sent by the transmitter. The performance of the system is calculated based on the bit error rate (BER) calculation. This parameter is calculated as:

\[
BER = \frac{\# \text{bit errors}}{\# \text{transmitted bit}} \tag{3.14}
\]
Chapter 4 – Experimental Set Up

On the previous chapter we have defined the necessary blocks for the reception and demodulation on an optical coherent system. In this chapter we will define our experimental set up, describe different devices compounding the system and characterize several effects introduced by hardware limitations.

4.1. System architecture

The idea is to build a complete coherent system using the devices that are commercially available in order to study the feasibility of a real coherent communication system. For that purpose we will use a mix of digital, electrical and optical devices. The digital devices are the front-end blocks, which correspond to the digital transmitter and digital receiver.
4.1.1. Coherent optical transmitter

As we mentioned on the previous chapter the digital transmitter is responsible for the signal generation and all the subsystems included on it: modulation format selection, and system parameters configuration. An Arbitrary Waveform Generator (AWG) is use to conform the electrical signals. We need to take into account that two channels are needed for the homodyne case and only one for the heterodyne case. This electrical signal is used to modulate the optical signal, which is generated on the laser source. This light coming from the laser at 1550nm is transmitted through an optical isolator, which protects the source of possible back reflections, and through an optical beam splitter 90/10. The function of the beam splitter is to divide the incoming signal on two: the optical signal that will be modulated and the optical signal that will act as local oscillator. The electrical signal generated on the AWG modulates the optical signal using a Mach-Zehnder optical modulator.

In the heterodyne case, the output electrical signal coming from the AWG is used to modulate the optical signal using a Mach-Zehnder optical modulator, as shown in Figure 24.

![Figure 24. Heterodyne optical modulation system. The light beam coming from the laser is sent through an optical isolator. The resulting signal in divided by a beam splitter 90/10. The high power signal is sent through a circulator $E_{in}(t)$ to the one-single-port Mach Zehnder modulator where the modulated optical signal $E_{out}(t)$ is generated. The modulator is driven by the heterodyne current $I_{het}(t)$ generated at the arbitrary waveform generator (AWGN).](image)

In the homodyne case, the two baseband signals coming from the AWG shape the optical signal using a Mach Zehnder modulator of two branches, one shifted $90^\circ$ from the other. The scheme is shown in Figure 25.
Figure 25. Homodyne optical modulation system. The light beam coming from the laser is sent through an optical isolator. The resulting signal is divided by a beam splitter 90/10. The high power signal is sent through a circulator to the two channel Mach-Zehnder modulator $E_{in}(t)$. The modulator is driven by the two baseband currents $I_{hom,1}(t)$ and $I_{hom,q}(t)$ generated by an arbitrary waveform generator (AWGN). The output modulated signal is $E_{o}(t)$.

Using Labview we are able to control the digital modulator as well as the AWG. The control panel embeds all the parameters needed on the transmitter. The Labview software responsible for the transmitter is shown in Figure 26.

Figure 26. Transmitter software design using Labview

4.1.2. Optical coherent receiver

We have already presented in previous chapters the basic scheme of heterodyne and homodyne receivers. In order to implement it we will reproduce exactly the same designs, but substituting each block for commercial devices. One important difference to mention is that we will use part of the power sent by the transmitter as a local oscillator on the receiver. This way we will only need one laser source on the
complete system, which makes the system more affordable. This is called self-homodyne and self-heterodyne detection.

At the receiver a variable attenuator is used to control the power of the light that act as local oscillator. In order to down convert the optical signal we will use the two different configurations that have already been described in chapter 2. For the heterodyne case, a 180° hybrid is used and its outputs will be transformed into an electrical signal using one balanced photo detector (Figure 28).

![Figure 27. Self-Heterodyne optical receiver set-up. The signal coming from the channel $E_s(t)$ is mixed in the hybrid with the one sent from the splitter 90/10 on the transmitter, which is passed through an attenuator. The hybrid outputs are connected to a balanced photo detector, which generates the intermediate frequency current $I_{het}(t)$. This current is translated to the digital domain by an analog to digital converter in order to apply digital compensation methods and demodulation techniques.](image1)

For the homodyne case, as we have shown in chapter 2, a 90° Hybrid will be used instead and two balanced photo detectors are needed (Figure 29).

![Figure 28. Self-homodyne optical receiver set-up. The signal coming from the channel $E_s(t)$ is mixed in the hybrid with the one sent from the splitter 90/10 on the transmitter, which is passed through an attenuator to obtain $E_{lo}(t)$. The hybrid outputs are connected to two balanced photo detectors, which generate the baseband frequency currents $I_{hom,1}(t)$ and $I_{hom,2}(t)$. This currents are translated to the digital domain by an analog to digital converter in order to apply digital compensation methods and demodulation techniques.](image2)
The electrical signals coming from the balanced photo detectors are converted into digital signals using an Analog to Digital Converter (ADC). At this point we will be able to apply the digital processing algorithms explained in chapter 3 in order to recover the transmitted data. Also, we will be able to introduce any channel error, as channel impairments or additive or multiplicative noises. The system parameters defined for the transmitter will be used in the receiver, and we will be able to extract the system throughput and the system BER.

Using the same procedure as in the transmitter, the receiver is completely controlled using Labview. The software is responsible for the data acquisition, impairment compensation and system BER performance evaluation. The software design is shown in Figure 29.

![Figure 29. Receiver software design using Labview.](image)

### 4.2. Device Characterization

In order to completely characterize and define our system we need to examine the characteristics of our devices. We will obtain their optical and electrical parameters and we will calibrate the performance of each component.
4.2.1. Clock Deviation

As we mentioned before, a reference deviation is usually present when we construct a communication system in which the digital transmitter and digital receiver use different clocks. This yields an electrical time/frequency deviation which can destruct the communication if no correction is applied. In our system, the objective is to measure this clock deviation in order to apply a correction parameter at the receiver.

The objective is to calculate and calibrate the clock deviation that exists between our digital transmitter and digital receiver. For that purpose we need an external reference that, independently of its own deviation, will measure the central frequency generated by each device at a single frequency.

The first objective is to see if this clock deviation generates a constant frequency deviation through the working frequency range. For that, we change the central frequency of the generated sine and measure the deviation for each frequency and each channel (Figure 30).

![Figure 30. Clock frequency deviation between transmitter and receiver for both transmitting channels](image)

In this figure we observe that the deviation increases as the central frequency grows up. The results are almost identical for both channels. From this graph we can...
extract that the deviation is linear and depends on the working frequency. Another way to look at it is by plotting the percentual deviation present for both channels.

From this graph we can extract that the deviation can be modeled as a percentual deviation around 0.0015%. With this calibration we will be able to adjust the clock of both systems in order to compensate this effect.

Other parameters that could influence this deviation were studied, as output power, pre-distortion filters or output filters, but no dependences were found.

![Graph showing relative clock frequency deviation between transmitter and receiver for both transmitting channels.](image)

**Figure 31.** Relative clock frequency deviation between transmitter and receiver for both transmitting channels

### 4.2.2. Isolator – Thorlabs: IO-H-1550-APC

This device, which allows the transmission of light in only one direction, is placed at the output of the laser in order to prevent the entry of any light beam into it, which could damage our source.

![Isolator diagram. Transmission of light is only in one direction, from port 1 to port 2.](image)

**Figure 32.** Isolator diagram. Transmission of light is only in one direction, from port 1 to port 2.
The key characteristics of this device are obtained using a power source and a power meter while changing the input and output.

<table>
<thead>
<tr>
<th>Insertion Loss (dB)</th>
<th>Return Loss (dB)</th>
<th>Isolation (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.81</td>
<td>55.32</td>
<td>36.4</td>
</tr>
</tbody>
</table>

*Table 1. Isolator characteristics*

### 4.2.3. Circulator – Thorlabs: 6015-3-APC

This is a three-port device designed such that light entering any port exits from the next. Circulators can be used to achieve bi-directional transmission over a single fiber. In our system it is used to monitor the back reflections and control any possible failure on the set-up.

![Optical circulator scheme](image)

*Figure 33. Optical circulator scheme*

The measured characteristics are:

<table>
<thead>
<tr>
<th>Direction</th>
<th>Insertion Loss (dB)</th>
<th>Return Loss (dB)</th>
<th>Isolation (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1 -&gt; 2</td>
<td>1.21</td>
<td>53.6</td>
<td></td>
</tr>
<tr>
<td>2 -&gt; 3</td>
<td>1.36</td>
<td>53.76</td>
<td></td>
</tr>
<tr>
<td>2 -&gt; 1</td>
<td></td>
<td>53.76</td>
<td>49.61</td>
</tr>
<tr>
<td>3 -&gt; 2</td>
<td></td>
<td>53.24</td>
<td>54.703</td>
</tr>
</tbody>
</table>

*Table 2. Circulator characteristics*

### 4.2.4. Polarization controller – FPC030

This device allows us to modify the polarization state of the incoming light from the laser. In our system, the modulator works only with one polarization, so we have to adjust the polarization state at their entry on order to maximize the modulated power. In this case it is a three freedom degree, manually controlled polarizer.
### Polarizator Controller characteristics

<table>
<thead>
<tr>
<th>Direction</th>
<th>Insertion Loss (dB)</th>
<th>Return Loss (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input -&gt; Output</td>
<td>0.293</td>
<td>49.76</td>
</tr>
<tr>
<td>Output -&gt; Input</td>
<td>0.502</td>
<td>51.707</td>
</tr>
</tbody>
</table>

*Table 3.*

#### 4.2.5 Mach Zenhder Modulator – Covega 056-003

This device is used to modulate the optical intensity by using optical interferometry. The incoming light is divided in two beams. One of these branches is delayed by the radiofrequency signal coming from the AWG. At the output, these two beams are mixed, generating a constructive or destructive wave. This way we are able to obtain an intensity modulated signal. The main characteristic of the device is the output power as a function of the applied voltage. The results are presented on figure 34.

*Figure 34. Transmission function of MZ modulator Covega 056*
By default, this device has been “bias trimmed” such that it’s zero-volt operating point is near negative-slope quadrature. The implication of operating on the negative slope quadrature point is that the modulator will cause data inversion. It can also be corrected by applying a bias voltage to achieve operation on the positive slope quadrature point. This is not a preferred solution as the required bias voltage will be quite large. In our system we have the advantage that the phase offset will be corrected by software. This means that an inversion of the data will not affect our system and that we are able to work at the optimal bias voltage. Another important data extracted from Figure 34. is that the maximum input voltage is ±2 V. If not, we would introduce modulation errors and undesired effects.

The chirp parameter is a key characteristic of any optical modulator. In our case the device is zero-chirp designed, so we can neglect its effects. Other parameters are shown in table 4.

<table>
<thead>
<tr>
<th>Insertion Loss (dB)</th>
<th>Return Loss(dB)</th>
<th>E/O Bandwidth (GHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>4.0</td>
<td>40</td>
<td>10.2 (data sheet)</td>
</tr>
</tbody>
</table>

Table 4. MZ modulator Covega 056 characteristics

4.2.6. Optical attenuator – Thorlabs: CWD-AL-10H

An optical attenuator is a device used to reduce the power level of an optical signal. In our system we use it as a local oscillator power controller.
This way we can modify several parameters as the working regime or the SNR.

<table>
<thead>
<tr>
<th>Insertion Loss (dB)</th>
<th>Return Loss (dB)</th>
<th>Maximum attenuation (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.18</td>
<td>55.32</td>
<td>160 (data sheet)</td>
</tr>
</tbody>
</table>

*Table 5. Attenuator characteristics*

### 4.2.7. Phase shifter – Phoenix photonics VPS 15

The phase shifter is a compact, simple to operate, all-fiber device for wideband operation. Applying a voltage to the pins it gives a controlled modification of the phase shift through the device. The phase shifter provides phase shifts up to $50\pi$ over a broad wavelength range.

![Phase shifter diagram](image)

*Figure 37. Phoenix photonics phase shifter*

The main characteristic we are interested in is the frequency response, shown in Figure 38. We will use it to change the phase of the local oscillator signal in order to prevent the power variations at the mixer output, which are due to the variation of the laser central frequency and the path difference.

![Phase shifter frequency response graph](image)

*Figure 38. Phase shifter frequency response*
Also, the input voltage needed to perform a phase shift of $\pi$ is calculated to be 0.05V. Other important parameters are shown in table 6.

<table>
<thead>
<tr>
<th>Insertion Loss (dB)</th>
<th>Return Loss(dB)</th>
<th>E/O Bandwidth (KHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>3.2</td>
<td>47</td>
<td>18.25</td>
</tr>
</tbody>
</table>

*Table 6. Phase shifter characteristics*

**4.2.8. Balanced photo detector – U2T - BPDV2020R**

The balanced photo detection has been widely covered on chapter 2 and its importance has been demonstrated. In table 7 the characterization of our detector is shown.

<table>
<thead>
<tr>
<th>Return Loss (dB)</th>
<th>CMRR (dB)</th>
<th>E/O Bandwidth (GHz)</th>
<th>DC Responsivity (A/W)</th>
</tr>
</thead>
<tbody>
<tr>
<td>24.7</td>
<td>15</td>
<td>40 (data sheet)</td>
<td>0.51</td>
</tr>
</tbody>
</table>

*Table 7. Balanced photo detector characteristics*

**4.3. Implementation limitations and impairment effects**

The main problems that exist in coherent communications have been explained theoretically in chapter 3, but in a real system we have to take in account a greater number of non-idealities that can influence in different manners to the obtained results. In order to address these effects, in this chapter we will try to define this impairments, how they influence our system, and the means to minimize its consequences.

**4.3.1. Bandwidth and transmitter limitations**

Considering that every device has a well-defined frequency range, where it can work properly, we would have to define the maximum bandwidth that we can use assuming a correct frequency response. In our case, as we can see in Table 8, the most limited device is the arbitrary waveform generator. This means that the maximum bandwidth that we can use to transmit data is 625MHz.

<table>
<thead>
<tr>
<th>Device</th>
<th>3dB Bandwidth</th>
</tr>
</thead>
<tbody>
<tr>
<td>Arbitrary Waveform Generator N82341A</td>
<td>625 MHz</td>
</tr>
<tr>
<td>Mach-Zehnder Covega 056</td>
<td>10GHz</td>
</tr>
<tr>
<td>Mach-Zehnder Covega 060</td>
<td>10 GHz</td>
</tr>
<tr>
<td>Balanced Photodetector BPVD-2020R</td>
<td>40 GHz</td>
</tr>
</tbody>
</table>
The arbitrary waveform generator presents several parameters providing different advantages or disadvantages with regards to the working and filtering regime at its output. There are three key tools that control our signal generator.

The first tool is an integrated pre-distortion filter, which compensates for the variation in the magnitude response of the output response as a function of frequency. This process creates a linear phase response and attenuates the lower frequency signals. The consequence is both a small dynamic range and a reduced output voltage at all frequencies.

Another integrated tool is the reconstruction filter at 500MHz realized as a 7-pole elliptical filter plus thru-line output. The filter purpose is to attenuate the harmonics generated by the DAC and reduce the noise floor. In the other hand, they cause a power loss around 2 dB and reduces the bandwidth to 500MHz.

The last parameter we can configure in the signal generator is the amplification level. It consists on an analog amplifier, placed after the DAC, which amplifies the RF signal. Unfortunately, it diminishes the signal purity and increase the noise floor.

In order to decide which parameters would provide us a better solution, we present the AWG results in the table 9. Here we can see that the best noise floor level and an acceptable harmonic generation are obtained with the pre-distortion filter active. In this scenario, introducing the reconstruction filter doesn’t improve the performance of the noise floor, but it would mean a high power loss.

<table>
<thead>
<tr>
<th>Predist. Filter</th>
<th>Amplification</th>
<th>Reconstr. Filter</th>
<th>Max. Power(dBm)</th>
<th>Noise Floor (dB)</th>
<th>2nd Harmonic (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>x</td>
<td>✓</td>
<td>x</td>
<td>0</td>
<td>-82</td>
<td>-53</td>
</tr>
<tr>
<td>x</td>
<td>✓</td>
<td>✓</td>
<td>-0.82</td>
<td>-80</td>
<td>-71</td>
</tr>
<tr>
<td>x</td>
<td>x</td>
<td>✓</td>
<td>-5.84</td>
<td>-96</td>
<td>-75.19</td>
</tr>
<tr>
<td>x</td>
<td>x</td>
<td>✓</td>
<td>-6.11</td>
<td>-96</td>
<td>-75</td>
</tr>
<tr>
<td>✓</td>
<td>✓</td>
<td>x</td>
<td>-7.2</td>
<td>-100</td>
<td>-77</td>
</tr>
<tr>
<td>✓</td>
<td>✓</td>
<td>✓</td>
<td>-10.1</td>
<td>-100.1</td>
<td>-80.3</td>
</tr>
<tr>
<td>✓</td>
<td>x</td>
<td>✓</td>
<td>-11.4</td>
<td>-100.1</td>
<td>-82.2</td>
</tr>
<tr>
<td>✓</td>
<td>x</td>
<td>✓</td>
<td>-16.27</td>
<td>-105</td>
<td>-87.11</td>
</tr>
</tbody>
</table>

Table 9. AWG performance with different configuration tools
4.3.2. Laser source: line width effects.

As we saw in chapter 3, there is a phase offset $\phi_s \ [\text{rad}]$ that must be calculated in order to demodulate any M-PSK signal. Since usually the line frequency width is much smaller than the symbol rate, it is reasonable to assume that $\phi_s \ [\text{rad}]$ is constant over each symbol duration. In real systems, we have to consider how this line width influences the bit error rate of the system. For that we use the expression obtained in [27]:

$$BER = \int \frac{1}{\sqrt{2\pi\sigma_\phi}} e^{-\phi^2/(2\sigma_\phi^2)} d\phi$$

(4.1)

where $\sigma_\phi^2 = 2\pi \tau \Delta \theta \ [W], \tau [s]$ is the coherence time, and $\Delta \theta [Hz]$ is the laser line width. In Figure 39 we can see the influence of the variance on the bit error of an ideal PSK demodulator.

![Figure 39. BER performance for synchronous PSK (homodyne and heterodyne) in presence of phase noise with a variance $\sigma^2$](image)

Our laser specifications indicate that its Lorentzian line width is less than 0.1 KHz. Also, as we see in Figure 39, if the laser variance $\sigma_\phi^2$ is lower than 0.01 there is no effect on the system BER. This means that we can calculate the maximum coherence time so we can avoid this effect:

$$\tau_{\text{max}} = \frac{\sigma_{\phi, \text{max}}^2}{2\pi\Delta \theta} = \frac{0.01}{2\pi \times 100} = 1.59155 \times 10^{-5} [s]$$

(4.2)

This time corresponds to a symbol frequency of 62.831 KHz. The maximum bandwidth in our system is 625 MHz, which means that the phase offset can be calculated calculated over a maximum number of symbols $N_{\text{max}}$ calculated as:
\[
N_{\text{max}} = \frac{BW}{f_s} = \frac{625 \times 10^6}{62.831 \times 10^3} \approx 10^4 \text{ [symbols]}
\]

where BW is our system bandwidth [Hz] and \( f_s \) is the symbol frequency [Hz]. This means that using a maximum of \( 10^4 \) symbols we will avoid the laser phase noise influence over the phase offset estimation.

### 4.3.3. Central frequency deviation and path difference fading

Another effect that is present in self-heterodyne systems is the low frequency beat due to the combination of two effects: the variation of the central frequency of the laser and the path difference. At the receiver, the two optical signals are mixed in the hybrid. These two signals have travelled through different paths which may differ on the length (Figure 40).

![Figure 40](image.png)

**Figure 40.** Optical path difference in self-homodyne and heterodyne systems. The light beam generated at the source \( E_{\text{la}}(t) \) is separated into two beams. One will act as signal beam, \( E_{\text{s1}}(t) \) and is sent through the system travelling a distance \( d_{\text{si}} \). \( E_{\text{o2}}(t) \) is sent to the coherent detector travelling a distance \( d_{\text{LO}} \), where it acts as local oscillator \( E_{\text{LO}}(t) \).

If we consider that the central frequency of the laser is time-varying, at the receiver we would have a time-dependent phase factor due to the path difference:

\[
\Delta \phi(t) = \phi_{\text{SI}}(t) - \phi_{\text{LO}}(t) = (d_{\text{SI}} - d_{\text{LO}}) \cdot \frac{2\pi}{\lambda(t)} \ [\text{rad}]
\]  

(4.3)

where \( d_{\text{SI}} \) is the signal path, \( d_{\text{LO}} \) is the local oscillator path, and \( \lambda(t) \) is the instantaneous wavelength of the laser at a concrete time \( t \). We can express this phase offset as a sum of two terms, a constant one and a time-dependent one:

\[
\Delta \phi(t) = (d_{\text{SI}} - d_{\text{LO}}) \frac{2\pi}{c} f_0 + (d_{\text{SI}} - d_{\text{LO}}) \frac{2\pi}{c} \Delta f(t) = \phi_0 + \phi_d(t) \ [\text{rad}]
\]  

(4.4)
The term $\Delta f(t)$ [Hz] introduces a time-varying phase, which self-modulates the signal at the output of the hybrid and reduces the power detected at the balanced photo detectors, having:

$$ I(t) = K \cdot \cos[w_{FI}t + \theta_0 - \theta_{LO}] = K \cdot \cos[w_{FI}t + \theta_0 + \theta_d(t)] \quad [A] \quad (4.5) $$

To clarify the effect of this varying phase we can use the standard identity:

$$ \cos(A + B) = \sin A \cos B + \cos A \sin B \quad (4.6) $$

This way we obtain an expression that relates directly the output intensity and the time-varying phase:

$$ I(t) = K \{\cos[w_{FI}t + \theta_0] \cos[\theta_d(t)] - \sin[w_{FI}t + \theta_0] \sin[\theta_d(t)]\} \quad [A] \quad (4.7) $$

This effect can result in destructive signal interference. In order to compensate this effect, we have to introduce a new block to control the phase of the local oscillator. For that purpose we introduce a phase shifter after the variable attenuator. A digitally controlled voltage source applies a voltage that controls the phase shifter. The scheme is shown in Figure 41.

![Figure 41. Local Oscillator phase controller system](image)

The algorithm that controls the voltage source is based on the second derivative algorithm. The objective of this algorithm is to apply phase shift $\theta_d(t)$ [rad] that compensates the time dependent term $\theta_d(t)$ [rad] so we get:

$$ \theta_T(t) = \theta_d(t) + \theta_A(t) = n \frac{\pi}{2} \quad [rad]; \quad n = 0,1,2,3 \quad (4.8) $$
Theoretically, if this term is well estimated we would have only one sinusoidal term in equation 4.7, which is our objective. In order to estimate this phase term $\Phi_A(t)$ [rad] we will focus on maximize the term $|I(t)| |A|$ that we get in the receiver. By maximizing the power obtained we ensure that the resulting signal is only over one projection of the plane. The algorithm is based on minimize the second derivative of the intensity power by applying a phase-step with the phase shifter. Depending on the new derivative result we will apply again the same phase-step or the negative.

In order to optimize the algorithm we need to calculate two basic parameters. The first one is the phase step introduced by our phase shifter. Here, the voltage source imposes a lower limit due to its minimum resolution (0.001 [V]) that corresponds to a minimum phase step of $3^\circ$. The second parameter is the sampling time used to update the phase shifter voltage. The combination of these two parameters describes the system response. A higher phase step at a high sampling rate could be able to correct faster changes but it would introduce a higher ripple on stationary conditions. On the other hand, by using a low phase step we would accurately obtain a phase correction on stationary conditions but it won’t be able to correct fast laser changes.

For that purpose we analyze and measure the effect of these two parameters over the losses on the average received power and on the relative fading variance $\sigma^2_P/P_m$, where $\sigma^2_P$ is the power variance and $P_m$ is the average power.

*Figure 42. Fading variance and power losses in the presence of different phase steps and sampling periods. The ripple and power losses are minimized for a phase step of 10 degrees with an update every 50 ms.*
As we see in Figure 42 the ripple and power losses are minimized for a phase step of 10 degrees with an update every 50 ms. The power loss at this point is 0.208 dB and the fading variance is -29.8 dB. With these results we demonstrate the functionality of the closed loop compensation technique.

4.3.4. Modulation imbalance factors

When characterizing modulation measurements, it is important to note that the phase, amplitude, and frequency of the RF signal is obtained based on the amplitude of each I and Q vector. Thus, errors in the baseband signal translate to errors in the RF signal. When we work in the heterodyne scenario this effect is negligible, because the RF signal is generated digitally, which provide us a high accuracy. In the other hand, when we work with homodyne systems, these parameters are defined by the optical modulator at the transmitter. Actually, this factor can increase the bit error rate. In this section we will explain how they influence the system quality and what is the error introduced by our own modulator. In this case we will describe the effect of the two main factors in our system: the gain imbalance factor and the quadrature skew. Mathematically, we can describe these effects as [28]:

\[ I' = a \times I \] 
\[ Q' = a \times \sin(\varphi) \times I + b \times \cos(\varphi) \times Q \]

Here \( \varphi \) is the quadrature skew, \( a \) is the modulation gain at the I branch and \( b \) is modulation gain in the Q branch. The gain imbalance factor is defined as:

\[ \gamma = 20 \log \left( \frac{a}{b} \right) [\text{dB}] \] 

The gain imbalance indicates the ratio of I gain to Q gain in dB. This effect is present in many modulators due to the possible differences between the two modulation channels. Also, it is usually caused by amplitude errors in the DAC’s or because of inconsistencies between each of the analog mixers. This effect, even with a few dB, can prevent proper demodulation of the signal. In order to see how it affects our receiver we simulate this error in the transmitter. The results can be seen in Figure 43.
This source of error is typically observed in the optical modulator or in the local oscillator splitter. The problem here is that the phase introduced between the I and Q channels is not exactly 90°. Some systems specify up to 3 degrees or more of skew [29]. For lower order modulation schemes, quadrature skew has relatively small effect on the system throughput. However, like other sources of error, higher order modulation schemes are significantly impaired. In the figure 4.12, we show the performance of our system for different magnitudes of the quadrature skew.

Figure 43. System BER in the presence of different gain imbalance factors

Figure 44. System BER in presence of quadrature skew imbalance
4.4.4. Effect of asymmetric hybrids

In chapter 2 we defined the demodulation schemes and their equations considering ideal hybrids. Now we will analyze the same equations using a generic description of the coupler considering all the non-ideal parameters. For that, we will model the basic coupler as [30]:

\[
\begin{bmatrix}
E_3 \\
E_4
\end{bmatrix} = \begin{bmatrix}
\sqrt{1 - \varepsilon} & \sqrt{\varepsilon} \\
\sqrt{\varepsilon} & \sqrt{1 - \varepsilon}
\end{bmatrix} \begin{bmatrix}
E_1 \\
E_2
\end{bmatrix} e^{j\phi}
\]

(4.12)

where \( \phi \) [rad] is the phase introduced between the two branches and \( \rho \) [rad] is the phase shift due to the coupler. Following the same procedure as in chapter 2, we have that:

\[
E_3(t) = (\sqrt{1 - \varepsilon} \cdot E_1 + \sqrt{\varepsilon} \cdot E_2 e^{j\phi}) e^{j\rho}
\]

(4.13)

from there we can obtain the correspondent intensity.

\[
I_3(t) = R \cdot |E_3(t)|^2 = R \cdot (E_3(t) \cdot E_3^*(t))
\]

(4.14)

Developing the expression we find the intensity as:

\[
I_3(t) = R \cdot (1 - \varepsilon) |E_1|^2 + R \cdot \varepsilon |E_2|^2 + 2R(1 - \varepsilon) \varepsilon
\]

(4.15)

\[
* |E_1|^2 |E_2| \cos((w_1 - w_2)t + \theta_1 - \theta_2 + \theta)
\]

Applying the same procedure for \( I_4(t) \) we have that:

\[
I_4(t) = R \cdot (1 - \varepsilon) |E_2|^2 + R \cdot \varepsilon |E_1|^2 + 2R(1 - \varepsilon) \varepsilon
\]

(4.16)

\[
* |E_1|^2 |E_2| \cos((w_1 - w_2)t + \theta_1 - \theta_2 - \theta)
\]

Using balanced detection and after applying a band pass filter, the resulting intensity is obtained as:

\[
I_{Het}(t) = I_3(t) - I_4(t)
\]

(4.17)

\[
= K \{\cos((w_1 - w_2)t + \theta_1 - \theta_2 + \theta)
\]

\[
- \cos((w_1 - w_2)t + \theta_1 - \theta_2 - \theta) \}
\]

where \( K = 2R(1 - \varepsilon) \varepsilon \cdot |E_1|/|E_2|^2 \cdot |A| \). In the ideal case \( \theta = \pi/2 \) [rad], which would simplify the equations. Applying the trigonometric identity \( \cos(\alpha + \beta) = \cos(\alpha) \cos(\beta) - \sin(\alpha) \sin(\beta) \) and substituting \( \alpha = (w_1 - w_2)t + \theta_1 - \theta_2 \) and \( \beta = \theta \) we have that the resulting intensity is:

\[
I_{Het}(t) = -2 \cdot K \cdot \sin[\theta] \cdot \sin((w_1 - w_2)t + \theta_1 - \theta_2)
\]

(4.18)
Comparing this result to the one obtained in Chapter 2, we observe that if \( \epsilon \neq 1/2 \) and \( \theta \neq \pi/2 \) [rad] they will produce a power loss on the detected intensity, which will imply a lower SNR at the receiver.

We can obtain the parameters of our concrete hybrid. The calculated coupling ratio of our coupler is \( \epsilon = 0.4936 \) which implies a loss of 8.34e-5 dB. So the effect due to the transmission mismatch can be neglected. The phase term is obtained using the balanced photo detection set up, based on the electrical DC level when no modulation is present and having one of the hybrid inputs empty. In this case, the calculated phase is \( \theta = \left( \frac{\pi}{2} \right) \times 0.912 \) [rad]. This implies a power loss of 0.004 dB. As we see, this effect is more representative than the transmission parameter but, still, it is very small and can be neglected.

For the homodyne case the calculations are more complicated because of the presence of four couplers and an intermediate shifter. In order to solve this problem we will model it as a 4x4 matrix based on the notation of Figure 45. To facilitate the task, we will assume that all the couplers are identical and we will neglect any noise, which have already been discussed on chapter 2.

![Figure 45. Homodyne receiver non-ideal model. Each port of the hybrid contains a field described by the letter it contains. Equations are described below.](image)

\[
1) \text{First Coupler System} \quad \begin{bmatrix} A' \\ B' \\ C' \\ D' \end{bmatrix} = \begin{bmatrix} T_1 & T_2 e^{j\phi} & 0 & 0 \\ T_1 e^{j\phi} & T_2 & 0 & 0 \\ 0 & 0 & T_1 & T_2 e^{j\phi} \\ 0 & 0 & T_1 e^{j\phi} & T_2 \end{bmatrix} \begin{bmatrix} A \\ B \\ C \\ D \end{bmatrix} \quad (4. 19)
\]
2) Transition Stage

\[
\begin{bmatrix} E' \\ F' \\ G' \\ H' \end{bmatrix} = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 0 & 1 & 0 \\ 0 & e^{j\phi} & 0 & 0 \\ 0 & 0 & 0 & 1 \end{bmatrix} \begin{bmatrix} A' \\ B' \\ C' \\ D' \end{bmatrix}
\] (4.20)

3) Second Coupler System

\[
\begin{bmatrix} E \\ F \\ G \\ H \end{bmatrix} = \begin{bmatrix} T_1 & T_2 e^{j\phi} & 0 & 0 \\ T_1 e^{j\phi} & T_2 & 0 & 0 \\ 0 & 0 & T_1 & T_2 e^{j\phi} \\ 0 & 0 & T_1 e^{j\phi} & T_2 \end{bmatrix} \begin{bmatrix} E' \\ F' \\ G' \\ H' \end{bmatrix}
\] (4.21)

Substituting 2 in 3 we obtain:

\[
\begin{bmatrix} E \\ F \\ G \\ H \end{bmatrix} = \begin{bmatrix} T_1 & T_1 e^{j\phi} & 0 & 0 \\ T_1 e^{j\phi} & T_1 & 0 & 0 \\ 0 & 0 & T_1 & T_1 e^{j\phi} \\ 0 & 0 & T_1 e^{j\phi} & T_1 \end{bmatrix} \begin{bmatrix} A' \\ B' \\ C' \\ D' \end{bmatrix}
\] (4.22)

Where using 1 we obtain:

\[
\begin{bmatrix} E \\ F \\ G \\ H \end{bmatrix} = \begin{bmatrix} T_1 & 0 & T_1 e^{j\phi} & 0 \\ T_1 e^{j\phi} & T_1 & 0 & 0 \\ 0 & 0 & T_1 & T_1 e^{j\phi} \\ 0 & 0 & T_1 e^{j\phi} & T_1 \end{bmatrix} \begin{bmatrix} A \\ B \\ C \\ D \end{bmatrix}
\] (4.23)

Considering that the input B and D are zero, we can simplify the previous transmission coupler matrix as:

\[
\begin{bmatrix} E \\ F \\ G \\ H \end{bmatrix} = \begin{bmatrix} T_1^2 & T_1 e^{j\phi} \\ T_1 e^{j\phi} & T_1^2 \\ T_1 T_2 e^{j\phi} & T_1 T_2 e^{j2\phi} \\ T_1 T_2 e^{j2\phi} & T_1 T_2 e^{j\phi} \end{bmatrix} \begin{bmatrix} A \\ B \\ C \\ D \end{bmatrix}
\] (4.24)

Now, we can separate this matrix in smaller ones, one for each balanced photo detector. This way, we could apply the simpler results obtained in the heterodyne case.
If we observe these equations we can notice that both are very similar in concept. The expression (4.25) is the same as eq. (4.12), but having the entry A shifted \( \phi_d \), which ideally is 90º. Considering perfect frequency match we can derive the expression for the output currents of the photo detectors:

\[
I_i(t) = I_e(t) - I_p(t) = K_1 \cdot \{ \cos[\theta_1 - \theta_2 + \theta] - \cos[\theta_1 - \theta_2 - \theta] \} \\
= 2K_1 \sin[\theta] \cdot \sin[\theta_1 - \theta_2] 
\]

\[
I_q(t) = I_c(t) - I_H(t) \\
= K_2 \cdot \{ \cos[\theta_1 - \theta_2 + \phi_d + \theta] - \cos[\theta_1 - \theta_2 + \phi_d - \theta] \} \\
= 2K_2 \sin[\theta] \cdot \sin[\theta_1 - \theta_2 + \phi_d] 
\]

where \( K_1 = 2RT_4|E_s||E_{LO}| \cdot [A] \) and \( K_2 = 2RT_1^2T_2^2|E_s||E_{LO}| \cdot [A] \). So we have a very similar scenario than in the heterodyne case. The influence of the transmission coefficients for each branch is exactly the same than in the heterodyne case, so we can assume that their influence on the power loss is negligible. To analyze the phase term we have to separate the two branches: for the I branch, the effect is the same as in the heterodyne case \( \sin[\theta] \cdot \text{factor} \), but in the Q branch the sum and difference of \( \phi_d \text{ [rad]} \) and \( \theta \text{ [rad]} \) may increase the error introduced. Let’s assume that:

\[
\phi_d = \frac{\pi}{2} + \delta; 
\]

Substituting in (4.28) we have:

\[
I_q(t) = 2K_2 \sin[\theta] \cdot \sin[\theta_1 - \theta_2 + \frac{\pi}{2} + \delta] = 2K_2 \sin[\theta] \cos[\theta_1 - \theta_2 + \delta] \\
= 2K_2 \sin[\theta] \left[ \cos(\theta_1 - \theta_2) \cos(\delta) - \sin(\theta_1 - \theta_2) \sin(\delta) \right] [A] 
\]

This parameter cannot be neglected. As we see, it affects to the orthogonality of I and Q branches, which may cause an additional error in compare to the heterodyne case. The phase deviation of our hybrid is 3º, which would imply a power penalty of 0.12 dB in the worst scenario.

The effect of the transmissions parameters \( T_1 \) and \( T_2 \) is demonstrated to diminish the output intensity of each branch. Assuming the model used before, where \( T_1 = \sqrt{1 - \varepsilon} \) and \( T_2 = \sqrt{\varepsilon} \), we have that the penalization on the in-phase branch is
0.00223 dB and the penalization on the quadrature-branch is $1.423 \times 10^{-4}$ dB. We observe that the penalization increases in compare to the heterodyne case. This is due to the fact that the signal passes through twice the couplers. Still, the penalization is small enough to be neglected.

### 4.4. System parameters

In order to test the behavior of the transmitter, channel and receiver implementation, LabView software is used. The whole transmission system is developed using LabView, which performs the digital signal processing and controls the surrounding equipment using standard drivers and functions.

![Transmitter control panel using Labview](image)

*Figure 46. Transmitter control panel using Labview. From this panel the digital transmission parameters are configured as well as the control over the arbitrary waveform generator.*

The prototype is designed with the main objective of validating the transceiver of an optical coherent system that uses complex modulation formats as well as direct
methods at a maximum symbol rate of 625 MHz on single polarization. All the devices used are commercial equipment with real and non-ideal behavior.

The signal in the heterodyne case is modulated by an electrical signal which contains data at 312.5 Msymbols/s using a carrier at 312.5 MHz. The number of symbols sent is based on the laser line width characterization derived on section 3.4.2., which implies that, at this data rate, we are able to send a maximum of $5 \times 10^3$ symbols assuming a constant phase offset. For the homodyne case we can use the whole bandwidth available, having a data rate of 625 Msymbols/s, which implies that the maximum number of symbols while the phase offset is constant is $10^3$. In both systems, the data rate is chosen to be the maximum allowed by the AWG, which implies that the transmitter will be using 2 samples per symbol in the homodyne case and 4 in the heterodyne case. The AWG works in the optimal scenario shown in chapter 3.4.1., using a pre-distortion filter, maximum amplification and no filter at 500MHz.

![Figure 47. Receiver control panel using Labview. From this panel we are able to control the receiver parameters, the A/D acquisition configuration and the local oscillator phase control. At the same time it implements the demodulation process, presenting the IQ diagram and demodulated I and Q signals.](image)
The shape filter used can be modified by parameterization, being able to use raised cosine, root raised cosine, Gaussian filter or none. The fixed preamble used to solve the phase ambiguity at the receiver is chosen to be eleven symbols and it is configured in order to be unique. In the transmitter it is also possible to configure other parameters as frequency deviation, white noise addition or IQ impairments.

At the receiver, a variable attenuator controls the local oscillator power that arrives to the hybrids of 90° and 180°. In the heterodyne case a phase shifter controlled by a low bandwidth AWG is introduced. Balanced photo detectors translate the optical signal into electrical signals, which are sampled by a digital oscilloscope working at a sampling rate of 4 GHz, which uses 8 and 16 samples per symbol for the homodyne and heterodyne case. Digital filtering is applied by re-sampling and channel impairments are optionally applied using additive and multiplicative noises. The digital signal processing algorithms explained in chapter 2 are used applying the system parameters defined in the transmitter.

After demodulation in burst mode the system bit error rate can be calculated under different simulation characteristics and system parameters, which allow us a high flexibility on the system definition.
Chapter 5

Chapter 5. System Validation

In this Chapter the robustness of the whole coherent system is tested. First of all, the functionality of the receiver in absence of impairments is checked to demonstrate the system validation. We will calculate the signal to noise ratio for the homodyne and heterodyne cases. We will compare the throughput of the system to the ideal receiver and theoretic limit. Then, we continue testing the behavior of our system in presence of shot noise (white Gaussian additive noise, AWGN).

From our tests, it is demonstrated that the system is able to send and receive information using a single polarization complex modulation, achieving different data rate as a function of the system parameters used. We will also test the BER performance of the system in absence of each compensation block, which will indicate the performance improvement and penalization that introduces each block. From this study it is shown that the phase and
frequency estimator achieves a good practical performance by averaging the information along
the symbols sent.

In order to test the different compensation block we analyze the system bit error rate
without any kind of compensation. The phase compensation must be always present, because
the optical phase drift between the optical signal and the local oscillator must be corrected in
order to have some correct data in a complex system.

Figure 48. IQ diagram and eye diagram in absence of frequency correction

For that purpose we will assume the most simple phase corrector, which averages the
phase offset over only one symbol period. As we see in Figure 48, when no frequency
correction is applied the demodulated symbols are completely random. This means that a
small frequency deviation, if it is not compensated, can destroy completely the transmitted
data. Even when no additive noise is present, the demodulated phase is not correct due to the
phase variation of the baseband signal, which would correspond to a low frequency
modulation of the baseband signal. This frequency offset present in our system is due to a
frequency drift between the transmitter and receiver carrier frequencies. This drift is caused by the deviation between the AWG reference clock and the receiver reference clock, which needs to be compensated.

When the frequency offset estimator is introduced, we observe that we can demodulate correctly the transmitted data. In this case we have a fixed system, where we can compensate the frequency deviation through parameterization without any loss of information. Still, the difference with the theoretical limit is large. This is because the sampling rate is not matched with the one at the transmitter. The effect is that, at the receiver, the shape filter is not optimal and it introduces an error due to inter-symbol-interference (ISI).

![Figure 49](image)

*Figure 49. System performance in the presence of compensation algorithms. System bit error rate is obtained for different SNR scenarios and compensation techniques. When compensation techniques are not applied, data demodulation becomes random. By introducing a frequency offset estimator the system is able to demodulate data correctly. The clock recovery block improves the receiver performance, but the system becomes optimal when the phase offset estimator is applied. In this situation, for SNRs higher than 8 dB the sensitivity penalization is around 0.1 dB in compare to the theoretical limit.*

Furthermore, the sampled point for each symbol is deviated from the optimal sampling point, which introduces an additional error. There is a threshold where, if the transmitted data is large enough, the synchronization is completely lost. This threshold depends on the clock deviation. In our system, the deviation between transmitter and receiver was calculated to be 0.00159%. This sets the limit on:
The BER shown in Figure 49 is calculated using 500 symbols (N) in order to avoid this situation for the scenario without clock recovery.

When the clock recovery block is included, the behavior of the receiver improves up to 6dB, obtaining a closer curve to the theoretic results. At this point we still have a penalization close to 1.5 dB, which can be improved by estimating the phase offset over a higher number of symbols. In this case we use the same number of symbols that in the no-clock recovery scenario. The resulting BER after averaging over 500 symbols almost matches the theoretical limit when the SNR is higher than 8dB. The small deviation present at higher SNR may be caused by some of the impairments explained in chapter 4 as asymmetric hybrids, limited bandwidth, or by small deviations on the frequency clock parameter estimation.

\[
N_{\text{max no clock}} = \frac{1}{0.00159} = 628 \text{ [symbols]}; \quad (4.31)
\]

Figure 50. Receiver performance in low SNR scenarios in the presence of different compensation algorithms. Working with a SNR of 8 dB the penalty introduced by the system over the SNR is 1.02 dB. In the worst scenario, when SNR per bit is 1, the maximum penalization introduced by our receiver is 2dB.

Working with a SNR of 8 dB the penalty introduced over the SNR is 1.02 dB (Figure 50). For lower SNRs the phase estimator loses some accuracy and the differences are higher. The maximum penalization introduced by this block is around 2dB in the worst scenario considered. This additional error is caused by the phase unwrapping stage, which transform a continuous phase into a circular one. This block is needed for averaging and it uses a LMS
algorithm to implement it. At low SNRs, phase jumps are not detected and contribute significantly to the calculated average. When the SNR is good enough to avoid this jumps the penalty introduced is less than 0.1dB (Figure 51).

![Figure 51. Receiver performance in high SNR scenarios in the presence of different compensation algorithms. The system performance working with SNR per bit higher than 8 dB almost matches the theoretical limit. The maximum penalization introduced is 0.1dB, which demonstrates the viability of the system](image)

From this study, the functionality of the system and blocks has been validated. The considered case is the optimal one, where the frequency offset is calculated through deterministic methods instead of using statistical ones. The results demonstrate the viability of the system and the performance of the blocks.

By taking a look to the IQ diagrams we can visualize the effects already mentioned. When clock recovery is not applied the sampled point are not optimally chosen so we have a high dispersion around each ideal symbol. By adding the phase recovery at this point we see that the performance improves but the phase offset is not correctly estimated. This is due to the fact that the phase offset is calculated by using the sampled points. If these points contain an error, it will be transmitted to the phase offset estimator.

When clock recovery is applied, the sampled points are chosen to be optimal and they are more compact around the constellation point. Still, the IQ diagram is phase shifted and the phase estimator is needed in order to avoid errors. We get the optimum scenario by applying both algorithms. Here, the IQ diagram presents the demodulated symbols around the ideal
constellation points (Figure 52). The distance to the ideal point is mainly due to the additive
noise and bandwidth limitations.

![No Clock Recovery & No Phase Estimation](image1)

![No Clock Recovery & Phase Estimation](image2)

![Clock Recovery & No Phase Estimation](image3)

![System Performance](image4)

*Figure 52. Demodulated IQ diagram working with an SNR of 10 dB and (a) No clock recovery and no phase estimation (b) no clock recovery and phase estimation (c) clock recovery and no phase estimation (d) frequency and phase compensation and clock recovery.*

The same conclusions can be obtained from the eye diagrams (Figure 53). Here the eye
opening is clearly maximized by using the whole chain of algorithm already described.

Another parameter to be studied is the number of symbols used for the phase offset
estimation. As we mentioned, there is a maximum limit of $10^4$ symbols imposed by the
coherence time of the laser. The number of symbols to be chosen must ensure that the phase
offset is constant over that time. As we saw in previous chapters, other effects may influence
this coherent time. The central frequency deviation of the laser introduces a phase variation,
which depends on the magnitude of this variation and the path difference between signal and
local oscillator. This effect reduces the maximum limit imposed by the laser line width.
We proved that the maximum frequency variations produced at the receiver are around 1 Hz, so to ensure a constant offset the maximum coherence time must be much lower than this quantity.

![Figure 53. Eye diagram at the receiver using (a) No clock recovery and no phase estimation (b) no clock recovery and phase estimation (c) clock recovery and no phase estimation (d) frequency and phase compensation and clock recovery.](image)

In order to measure this effect in our receiver we calculate the system BER for a different number of symbols. The results are shown in Figure. The objective is to find an optimal number of symbols which minimizes the penalization. In the graph we observe that by choosing N=50 or N= 200 we introduce a penalization around 0.5dB. This is because the phase estimator has not enough data to completely estimate the phase in the presence of noise. If we increase N up to 500 symbols the optimum BER is obtained. When we keep increasing the number of symbols we get a slightly higher penalization, which is approximately 0.3dB. The reason of this penalization is that, despite the better theoretical performance in presence of noise, we reach the phase coherence time.
Figure 54. System BER performance as a function of the number of symbols (N) selected to estimate the phase offset. The receiver performance is optimal using N=500.

This means that the optimal receiver must implement the phase offset estimation by dividing the incoming signal on smaller data containers of 500 symbols, and apply each estimated offset over the same group of data.
Chapter 6

6. Conclusions and future work

In this work a study and implementation of an optical coherent transmission system has been made and demonstrated. The theory and basics of this technology are presented, as well as the digital system processing needed to obtain a robust communication system. Afterwards, a practical system has been designed, implemented and tested. The coherent system implementation has shown the viability of optical communications using complex modulation formats, which improves the spectral efficiency and provides better sensitivity at the receiver.

The following objectives and conclusions have been accomplished in this project:

- An optical coherent communication system has been designed and implemented.
- Its robustness has been demonstrated by reaching the theoretical performance limit in the presence of AWGN for SNR higher than 10 dB.
• For SNRs lower than 10 dB we cannot model the system as ideal, but still the system performance is 1.5 dB from the theoretical limit.

• Digital signal processing has been proved to be advantageous and essential in a real optical coherent receiver. A frequency offset estimation algorithm is mandatory in any system. Also, phase estimation and clock recovery strongly improve the receiver performance.

• The implementation limits have been studied. The performance penalization introduced by the parameters of physical devices has been studied and quantified.

• A feed-back control system was implemented in order to compensate the power fading due to destructive interferometry at the receiver.

With coherent communications we are able to introduce information in the phase, frequency and amplitude signal by using the front-end architectures described in this project. Balanced photo detection is today a must in coherent detection due to two main reasons: it provides twice the power than non-balanced detection and minimizes the effect of the noise present in the photodiodes. Once the received signal is translated to the digital domain we are able to implement real-time algorithms that are able to compensate the main impairments that affect optical communications. This work is focused on the receiver design and also on the system requirements that a prototype must have in order to provide a good practical performance.

In this project we have shown the different coherent systems that are present nowadays considering the advantages and disadvantages of each detection system as well as the different sensitivities needed for each modulation format. Complex modulation formats have been chosen due to their high spectral efficiency, their low sensitivity and the capacity of build a practical system based on commercial devices that exist in the present.

The algorithms needed in any coherent system have been presented, studied and implemented. Our system is based on three compensation techniques specially implemented for coherent systems. First, a clock recovery algorithm has been designed and implemented. Due to the fact that other existing algorithms based on statistical methods impose a penalization on the system BER, we have decided to apply a more robust compensation. By taking advantage of our system parameters and calibration, we are able to compensate the clock deviation, assuming negligible the penalization introduced. Phase and frequency estimation are implemented by using a multiply-filter-divide method, which eliminates the
modulation factor on phase-shift-keying communications and allow us to use statistical algorithms to estimate these factors.

An experimental set-up has been implemented using commercial devices. We have accomplished a detailed device characterization in order to study any possible influence of these devices on the system. Some of these devices present some limitations that generate differences from the ideal scenario. The penalization introduced by each one of these limiting devices has been calculated or compensated. The bandwidth limitations impose us a transmission upper limit of data rate. The laser source deviation introduces a maximum coherence time that limits the number of symbols that can be used for the frequency and phase estimation. The penalization associated to this deviation was found to be under 0.001 dB, so it can be neglected.

We have shown how the modulation imbalance factors may affect the communications and the receiver performance. Using this theoretical approach we have been able to calculate the penalization introduced by our modulator. For the heterodyne case, this modulation imbalance factors can be digitally controlled and minimized, which allow us to consider a penalization close to zero. In the homodyne case these factors could be electrically compensated at the modulator by introducing an offset voltage or by modifying the transmitted signal at the digital transmitter.

Also the effect and penalization of asymmetric hybrids have been studied in detail. We have obtained the equations that show their behavior and the penalization introduced. In the heterodyne case, the penalization introduced by our modulator is very small: 8.34e-5 dB due to the transmission factor and 0.004 dB due to the phase deviation. In the homodyne case, our modulator would introduce a penalization of 0.2 dB, which cannot be neglected.

We have demonstrated the system and receiver implementation viability. Our system is able to compensate the most influential effects that appear in practical systems. It is shown that the phase estimator achieves an excellent performance by averaging the estimation along enough number of symbols (N). The optimal N is calculated to be around 500 samples. In this scenario and under good SNR conditions, higher than 10 dB, we are able to maintain a system BER very close to the theoretical limit.
Under these conditions our system only adds a penalization of 0.01 dB, which demonstrate the full functionality of our prototype. Under low SNR conditions, below 8 dB, our phase estimator introduces higher penalization due to the phase noise and phase jumps that cannot be compensated. Still, the maximum penalization in the worst possible scenario (1dB of SNR) is only 2.5 dB.

![Figure 51. System BER on presence of AWGN. System bit error rate is obtained for different SNR scenarios and compensation techniques. When compensation techniques are not applied, data demodulation becomes random. By introducing a frequency offset estimator the system is able to demodulate data correctly but introducing many errors. The clock recovery block improves the receiver performance, but the system becomes optimal when the phase offset estimator is applied. In this situation, for SNRs higher than 8 dB the sensitivity penalization is around 0.1dB in compare to the theoretical limit.](image)

This project is part of a research that, based on the implemented prototype, will study the viability of optical coherent technology in free-space communications using compensation techniques. The objective is to increase the flexibility of the transceiver by adding different modulation formats as well as new optical techniques that will try to compensate the effects introduced by the atmosphere. The results presented on this project will be the base of further studies about the coherent receiver performance on free space communications.
Bibliography


