Master’s Thesis

Wavelength Tunable Short Pulses Sources Based on Frequency Comb Generation in a Mach-Zehnder Modulator

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Abstract

In this thesis we present the study of the implementation of ultrashort pulses using frequency comb generation in a Mach-Zehnder modulator and chirp compensation in a standard single-mode fiber. Two possible configurations have been developed and optimized to obtain a picosecond pulse train. An additional stage of pulse compression was necessary to arrive to femtosecond pulses, and a nonlinear fiber was employed. An analysis of a complete optical communication system using the generated pulses with OOK modulation and OTDM was devoted and a relation between pulse parameters and the BER sensitivity was evaluated. A numerical model with Matlab® software has been developed for the required analysis. Finally, an experimental solution of the pulse source has been devoted in the Fotonik’s laboratory, obtaining results that we compared with the simulated ones.
I would like to express my gratitude to Christophe Peucheret and Michael Galili, the supervisors of this thesis. Thanks for motivating and guiding me along my thesis, for having always time to answer my questions, and for all the help and advises during the time in the laboratory too.

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Gracias.
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<td>ASK</td>
<td>Amplitude Shift Keying</td>
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<td>AWGN</td>
<td>Additive White Gaussian Noise</td>
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<td>BER</td>
<td>Bit Error Rate</td>
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<td>BW</td>
<td>Bandwidth</td>
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<td>CW</td>
<td>Continuous Wave</td>
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<td>DD</td>
<td>Direct Detection</td>
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<tr>
<td>DEMUX</td>
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<tr>
<td>DF-HNLF</td>
<td>Dispersion-Flattened High NonLinear Fiber</td>
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<tr>
<td>EO</td>
<td>Electro-Optic</td>
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<tr>
<td>ER</td>
<td>Extinction Ratio</td>
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<tr>
<td>FROG</td>
<td>Frequency-Resolved Optical Gating</td>
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<tr>
<td>fs</td>
<td>femtosecond</td>
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<tr>
<td>FT</td>
<td>Fourier Transform</td>
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<td>FWHM</td>
<td>Full Width at HalfMaximum</td>
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<td>GVD</td>
<td>Group-Velocity Dispersion</td>
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<td>IFT</td>
<td>Inverse Fourier Transform</td>
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<tr>
<td>MMF</td>
<td>Multi Mode Fiber</td>
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<tr>
<td>Abbreviation</td>
<td>Full Form</td>
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<tr>
<td>MUX</td>
<td>Multiplexer</td>
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<td>MZM</td>
<td>Mach-Zehnder Modulator</td>
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<tr>
<td>NLSE</td>
<td>NonLinear Schrödinger Equation</td>
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<tr>
<td>NRZ</td>
<td>Non-Return-to-Zero</td>
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<tr>
<td>OBPF</td>
<td>Optical BandPass Filter</td>
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<td>OOK</td>
<td>On-Off Keying</td>
</tr>
<tr>
<td>OSA</td>
<td>Optical Spectrum Analyzer</td>
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<td>OSNR</td>
<td>Optical Signal to Noise Ratio</td>
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<td>OSO</td>
<td>Optical Sampling Oscilloscope</td>
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<tr>
<td>OTDM</td>
<td>Optical Time Division Multiplexing</td>
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<tr>
<td>PP</td>
<td>Power Penalty</td>
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<td>PRBS</td>
<td>Pseudo Random Bit Sequence</td>
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<tr>
<td>ps</td>
<td>picosecond</td>
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<td>PSD</td>
<td>Power Spectral Density</td>
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<td>RZ</td>
<td>Return-to-Zero</td>
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<td>RMS</td>
<td>Root Mean Square</td>
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<td>SMF</td>
<td>Single Mode Fiber</td>
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<td>Signal to Noise Ratio</td>
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1.1 Motivations and State of the art

Nowadays, most of the people use Internet for finding information, sending and receiving emails, downloading and sharing data and use online applications such as video calls, online TV or games. This growth demands the usage of networks with higher capacities to have a faster and easier access to information.

Most commercial optical communication systems employ on-off keying (OOK) modulation and direct detection (DD). OOK is a binary modulation where the presence of light represents a bit 1, while its absence represents a bit 0. The main advantage of this modulation is that it is the simplest modulation format, and the receivers have low complexity.

There are two ways to increase the capacity employing the OOK modulation: the wavelength division multiplexing (WDM), and the time division multiplexing (TDM). Optical networks use a combination of WDM and TDM to optimize the transmission capacity. TDM may be realized by electrical multiplexing (ETDM) or by optical multiplexing (OTDM)[25].

As the bit rate used in optical communications keeps increasing, optical networks will evolve into photonic networks, in which ultrafast optical signals of any bit rate and modulation format will be transmitted and processed without optical-electrical-optical conversion. OTDM technology get us the chance to investigate and develop high-speed optical signal processing and explore the ultimate capacity for fiber transmission in a single wavelength channel.

One of the key elements of ultrafast optical communication systems is optical short pulse generation, where stable pulse sources are required to generate short pulses at a high repetition frequency. Pulse source requirements are especially critical in OTDM systems where bit rates above 640 Gbit/s, require sub-picosecond pulses to be generated at a repetition rate of 10 or 40 GHz. [12].
Currently, various methods for high-repetition rate optical pulse generation have been studied and developed using mode locked lasers [22][19], fiber ring lasers [6] or gain-switched laser diodes [11]. Such pulse sources might suffer from long term stability and overall reliability issues. Furthermore, tunability in terms of emission wavelength and pulse repetition frequency, which is a desirable feature in real systems, is often difficult to achieve in practice.

A novel short pulse generation method using frequency comb generation in a Mach-Zehnder modulator has recently been proposed and demonstrated [13][18]. In this technique, light emitted by a continuous wave (CW) laser is modulated in an external Mach-Zehnder modulator (MZM). Following spectral shaping in an optical bandpass filter (OBPF) and chirp compensation in a piece of standard single mode fiber (SMF), is obtained a periodic train of short pulses at a repetition frequency corresponding to that of the electrical sinusoidal driving signals.

1.2 Goals and Thesis Structure

In this thesis, the implementation of pulse sources using frequency comb generation in a Mach-Zehnder modulator will be investigated, and the pulse source will be optimised to achievable pulse width, pulse repetition frequency and wavelength tuning range. An additional pulse compression stage will be employed to obtain subpicosecond pulses generated at a repetition rate of 10 GHz.

This thesis is divided into five chapters. In chapter 2 we introduce the theoretical background related to the aspects developed in this thesis. In chapter 3 we explain how we generate the different blocks that we need to simulate our optical communication system, giving a brief introduction in the commands used in Matlab® software. In chapter 4 is described the generation of ultrashort pulses using frequency comb generation in a Mach-Zehnder modulator. It is also explained the pulse compression stage and the system performance analysis. In chapter 5 the results of the experimentally frequency comb generator using Mach Zehnder modulator tested in the laboratory are collected. Finally, the conclusions of this thesis are given in chapter 6.
This chapter wants to provide the necessary theoretical basis at the reader for understanding both our model of an optical communication system and its implementation. It is just a brief introduction to handle the following topics with confidence.

The first section gives a short opening about signal description; then the fiber optics effects and information about modulating signals, BER and OSNR are detailed; as well as an introduction to OTDM and related concepts to optical transmitters and receivers.

2.1 Signal Description

In this section we will develop an expression for the electric field of the optical signal [15]. We can describe a plane wave of the type

$$E(r, t) = \text{Re}\{A(r, t)e^{j(\omega_0 t - k \cdot r)}\}$$  \hspace{1cm} (2.1)

where $E(r, t)$ is the electric field, which depends on space and time; $A(r, t)$ is a complex vector that is related to the complex envelope of the field; and $\omega_0$ is the angular frequency and $k$ is the wave vector that characterises the light.

Defining an orthonormal basis $(x, y, z)$ where $z$ is the direction of propagation, and where $r = r_x x + r_y y + r_z z$, then (2.1) can be expressed as

$$E(z, t) = \text{Re}\{A(z, t)e^{j(\omega_0 t - k \cdot z)}\}$$  \hspace{1cm} (2.2)

$A(z, t)$ can be decomposed on $(x, y)$ according to

$$A(z, t) = \kappa[u_x(z, t)x + u_y(z, t)y]$$  \hspace{1cm} (2.3)

where $u_x(z, t)$ and $u_y(z, t)$ are complex functions and $\kappa$ is a normalization factor. That
expression allows for intensity, phase and polarisation of the light wave to be time dependent. The magnitude of $A(z, t)$ can be calculated

$$|A|^2 = A \cdot A^* = \kappa^2[|u_x(z, t)|^2 + |u_y(z, t)|^2]$$

hence the light intensity

$$I(z, t) = \frac{1}{2} \epsilon_0 cn \kappa^2 [|u_x(z, t)|^2 + |u_y(z, t)|^2]$$

and the optical power

$$P(z, t) = \frac{1}{2} \epsilon_0 cn \kappa^2 A_{eff} [|u_x(z, t)|^2 + |u_y(z, t)|^2]$$

where $\epsilon_0$ is the vacuum permittivity, $c$ the speed of light; $n$ the refractive index of the medium; and $A_{eff}$ corresponds to the area over which the intensity is integrated.

The parameter $\kappa$ can be chosen to have a simple expression of optical power as

$$P(z, t) = |u_x(z, t)|^2 + |u_y(z, t)|^2$$

consequently,

$$\kappa = \sqrt{\frac{2}{\epsilon_0 cn A_{eff}}}$$

whose dimension is in $\sqrt{\Omega/m^2}$.

The coordinates of $A(z, t)$ along $x$ and $y$ ($u_x(z, t)$ and $u_y(z, t)$) can be expressed in terms of their magnitude and phase components, so that

$$A(z, t) = \kappa [A_x(z, t)e^{-j\phi_x(z, t)}x + A_y(z, t)e^{-j\phi_y(z, t)}y]$$

which immediately leads to

$$A(z, t) = \kappa e^{-j\phi(z, t)} [A_x(z, t)e^{-j\frac{\delta(z, t)}{2}}x + A_y(z, t)e^{j\frac{\delta(z, t)}{2}}y]$$

where

$$\phi(z, t) = \frac{\phi_x(z, t) + \phi_y(z, t)}{2}$$

is the common phase, and

$$\delta(z, t) = \phi_x(z, t) - \phi_y(z, t)$$
is the differential phase term. Introducing the normalized Jones vector,

\[ e(z, t) = \frac{1}{\sqrt{A_x^2(z, t) + A_y^2(z, t)}} [A_x(z, t)e^{-j\phi_z(t)} \mathbf{x} + A_y(z, t)e^{j\phi_z(t)} \mathbf{y}] \] (2.13)

we can express \( A(z, t) \) as

\[ A(z, t) = \kappa e^{-j\phi(z, t)} \sqrt{A_x^2(z, t) + A_y^2(z, t)} e(z, t) \] (2.14)

and therefore

\[ A(z, t) = \kappa \sqrt{P(z, t)} e^{-j\phi(z, t)} e(z, t) \] (2.15)

\( A(z, t) \) is proportional to the complex envelope of the field, multiplied by an unit vector describing the state of polarisation. Now we can go to find a simply expression for the electric field considering that in the study of optical modulation formats the space dependence is not always necessary, and \( \kappa = 1 \text{ m}^{-1} (s/F)^{1/2} \).

Let us conclude that we will concentrate our study on phase modulation, \( \phi(t) \), although also \( P(t), w_0 \) and \( e(t) \) can be modulated. So we will refer to the electric field, forgetting the real part, as

\[ E(t) = \sqrt{P(t)}e^{-j\phi(t)}e^{-jw_0(t)} = A(t)e^{-jw_0(t)} \] (2.16)

### 2.1.1 Baseband transformation

The electric field with respect to the reference angular frequency \( w_{ref} = 2\pi f_{ref} \) is defined as

\[ E(t) = \sqrt{P(t)}e^{-j\phi(t)}e^{-j(w_0 - w_{ref})t} \] (2.17)

In case of single channel operation, a good choice of the reference frequency is \( w_{ref} = w_0 \), in which case

\[ E(t) = \sqrt{P(t)}e^{-j\phi(t)} \] (2.18)

In chapter 3 we will explain the reasons why we will work with a baseband signal.

### 2.2 Optical Fiber effects

The medium we are interested in is the optical fiber, which means that our optical signals will be transmitted over long distances. We encourage the reader to [3] and [4] if is interested in more information about Maxwell’s equations and fiber geometrical description to understand the guiding mechanism because we are going to concentrate...
2.2. Optical Fiber effects

our efforts in two important effects in optical fibers, such as fiber dispersion and nonlinear effects, and also the advantages of a single-mode fiber (SMF) over a multi-mode fiber (MMF).

2.2.1 Nonlinear Schrödinger Equation

Considering the function $A(z,t)$, presented in section 2.1, the nonlinear Schrödinger equation (NLSE)\cite{3} describes the propagation through a nonlinear medium as an optical fiber

$$\frac{\delta A}{\delta z} = -\frac{\alpha}{2} A - j\beta_2 \frac{\delta^2 A}{\delta t^2} + \frac{\beta_3}{6} \frac{\delta^3 A}{\delta t^3} + j\gamma |A|^2 A$$  \hspace{1cm} (2.19)

The first term relates to the losses, included through the $\alpha$ parameter; the second and third ones to chromatic dispersion ($\beta_2$ and $\beta_3$); and the last one to the nonlinear kerr effect. These aspects will be discussed in the following three subsections.

2.2.2 Losses

Light propagating in an optical fiber will incur in losses. Under general conditions, changes in the average optical power in a fiber follows \cite{20}

$$\frac{dP}{dz} = -\alpha P$$  \hspace{1cm} (2.20)

where $\alpha$ is the attenuation coefficient, measured in $m^{-1}$. If we define $L$ as the fiber length, we can relate the power to the input and the output of the fiber as

$$P_{out} = P_{in} e^{-\alpha L}$$  \hspace{1cm} (2.21)

We used to express $\alpha$ in [dB/Km] by using the relation

$$\alpha (dB/Km) = -10 \log_{10} \frac{P_{out}}{P_{in}}$$  \hspace{1cm} (2.22)

and refer to it as the fiber-loss parameter.

Several factors contributes to losses in a fiber, the two most important are material absorption and Rayleigh scattering. All the sources of losses are included in $\alpha$ parameter. Standard value for SMF is $\alpha=0.2$ dB/Km (see \cite{3}) at 1.55 $\mu$m. One of the reasons why in optical communications usually works at 1.55 $\mu$m is because here losses are the lowest for standard SMFs.

2.2.3 Dispersion

Dispersion causes pulse broadening when travelling over an optical fiber. The phenomenon of dispersion has two contributions, intermodal dispersion and intramodal
2.2. Optical Fiber effects

dispersion. The main advantage of SMF is that intermodal dispersion is absent. However, intramodal dispersion still takes place and pulse broadening does not completely disappear.

The group velocity associated with the fundamental mode is frequency dependent because of chromatic dispersion. Each frequency travels at different group velocities, this phenomenon is related to group-velocity dispersion (GVD).

A specific spectral component travel through a fiber of length $L$ in a time $T = L/v_g$, where $v_g$ is the group velocity, defined as

$$v_g = \left( \frac{d\beta(w)}{dw} \right)^{-1}$$  \hspace{1cm} (2.23)

where $\beta$ is the propagation constant. By using $\beta = \bar{n}w/c$, where $\bar{n}_g$ is the group index given by

$$\bar{n}_g = \bar{n} + w \left( \frac{d\bar{n}}{dw} \right)$$  \hspace{1cm} (2.24)

Group velocity causes pulse broadening because different spectral components of the pulses are dispersed during propagation and do not arrive simultaneously at the fiber output.

GVD could be expressed expanding $\beta(w)$ in a Taylor series as

$$\beta(w) = \beta_0 + \beta_1(\Delta w) + \frac{1}{2}\beta_2(\Delta w)^2 + \frac{1}{6}\beta_3(\Delta w)^3 + \ldots$$  \hspace{1cm} (2.25)

where $\beta_0$ and $\beta_i = \left( \frac{d^i \beta(w)}{dw^i} \right)|_{w = w_0}$. The parameter $\beta_2 = \left( \frac{d^2 \beta}{dw^2} \right)$ is known as the GVD parameter. Generally, fiber dispersion is defined by two parameters,

$$D = \frac{d}{d\lambda} \left( \frac{1}{v_g} \right) = -\frac{2\pi c}{\lambda^2} \beta_2$$  \hspace{1cm} (2.26)

$$S = \frac{d^2}{d\lambda^2} \left( \frac{1}{v_g} \right) = \left( \frac{2\pi c}{\lambda^2} \right)^2 \beta_3 + \frac{4\pi c}{\lambda^3} \beta_2$$  \hspace{1cm} (2.27)

where $D$ is the dispersion parameter, in units of $[ps/(nm\cdot km)]$ and $S$ is the dispersion slope, measured in $[ps/(nm^2\cdot km)]$.

It is possible to find fibers with $D=0$ at 1.55 $\mu$m, but we are not interested in it, because dispersion is an useful phenomenon that contains the effects of nonlinearities. Therefore, an important parameter is the dispersion length. It is the length that Gaussian pulse takes to get broadened of a factor of $\sqrt{2}$, and is defined as,

$$L_D = \frac{T_0^2}{|\beta_2|}$$  \hspace{1cm} (2.28)

where $T_0$ is the full width of the pulse at 1/e of the peak value.
2.2. Optical Fiber effects

For SMF with $\beta_2 < 0$, that means that high frequencies travel faster than lower frequencies. As a consequence, a positive chirped pulse will experience compression before broadening. This is a characteristic that we exploit in our system, and we will give more details in chapter 4.

2.2.4 Nonlinear Kerr Effect

The *kerr effect* is a change in the refractive index of a material in response to an applied electric field. The dependence of the refractive index can be expressed as

$$n = n_0 + n_2 \frac{|A(z,t)|^2}{A_{eff}}$$

where $n_0$ is the linear refractive index, $n_2$ is the nonlinear-index coefficient, $|A(z,t)|^2$ is the optical signal power and $A_{eff}$ the fiber effective area.

The nonlinear propagation coefficient can be expressed in terms of $n_2$, the optical wavelength $\lambda$ and $A_{eff}$ as

$$\gamma = \frac{2\pi n_2}{\lambda A_{eff}}$$

with values ranging from 1 to 5 $W^{-1}$/Km, depending on $\lambda$ and $A_{eff}$.[3]

The nonlinear length can be defined as

$$L_{NL} = \frac{1}{\gamma P_p}$$

where $P_p$ is the peak power, $L_{NL}$ is the length over which a Gaussian pulse experiences a phase shift of 1 rad in correspondence of its peak when there are neither losses nor dispersion. Assuming that we only have nonlinearities, a solution of (2.19) is

$$A(z,t) = A(0,t)e^{-j\gamma |A(0,t)|^2 z}$$

here, the dependence of the refractive index leads to a phase modulation of the signal. This phenomena is called *self-phase modulation* (SPM), and induces a frequency chirp in optical pulses. The effect is broadening spectrum, but with the presence of dispersion, this spectral broadening is converted into pulse broadening.

To conclude, we would like to explain the soliton systems. A soliton is a particular pulse with a balance between GVD and SPM that verify

$$N^2 = \frac{\gamma P_p T_0^2}{|\beta_2|} = \frac{L_D}{L_{NL}}$$

where $N$ is the soliton order. GVD broadens optical pulses during their propagation and if SPM-induced chirp satisfies (2.33), cancel the GVD-contribution, then the optical pulse would propagate undistorted in the form of a soliton. In [3] is possible to find a complete study of solitons.
2.3 Modulating signal

In this thesis we will work with the on-off keying (OOK) modulation, but first of all let us introduce some previously notation and concepts.

The number of symbols depends on the modulation and we use the modulation order to define it, and we will refer to it as $M$. If one symbol contains $m$ bits, then

\[ M = 2^m \quad (2.34) \]

The symbol rate is the rate which we transmit a symbol, we will label it as $Rs$. If one symbol contains $m$ bits, then

\[ R_b = m \cdot R_s \quad (2.35) \]

and it is called bit rate. We can define now the symbol slot ($T_s$) and bit slot ($T_b$) as

\[ T_s = \frac{1}{R_s}; \quad T_b = \frac{1}{R_b} = \frac{1}{m \cdot R_s} = \frac{T_s}{m} \quad (2.36) \]

We represent the symbols in a complex system, called constellation, and each modulation format has an specific constellation. So we need to define the bit mapping, where each symbol is defined as a complex number as

\[ S_k = S_{ik} + jS_{qk} \quad (2.37) \]

where $S_{ik}$ is the real part of the symbol $S_k$; and $S_{qk}$ is the imaginary part of $S_k$.

In this work we will focus only on the study of OOK modulation, to check the system performance when using the generated train of pulses to transmit information.

2.3.1 OOK modulation

OOK is the simplest modulation format. It is also called amplitude-shift keying (ASK) because the signal is amplitude modulated and we can assume just two different values. In presence of carrier represents a bit 1, while its absence represents a 0, and we do not modify the phase of the signal.

We use $m=1$ bit per symbol, so the modulation order is $M=2$, and the symbol rate parameter correspond to the bit rate. The OOK constellation is shown in figure 2.1.

![Figure 2.1: Constellation of OOK Modulation.](image)
2.3. Modulating signal

2.3.2 RZ and NRZ signals

We can find all modulation formats in two different ways, return to zero (RZ) or nonreturn to zero (NRZ). If we use NRZ, the signal occupies the whole symbol slot (in case that we transmit "1"), whereas in the RZ format the signal return to zero at the symbol slot limits. Here we show an example using OOK modulation (see figure 2.2).

![RZ format and NRZ format](image)

**Figure 2.2:** RZ – NRZ format OOK modulation.

2.3.2.1 TFWHM, Trms and ER

In this section we want to clarify some concepts that will be important later in our study. All of them are related to the pulse shape.

First of all, when referring to the width of a pulse, we will always relate to its full width at half maximum (TFWHM), that is the difference between time instants when the pulse power is equal to half of maximum peak power (see figure 2.3). But sometimes if we have sidelobes in our pulse shape, TFWHM does not take care about it and pulse shape can not be completely characterized by TFWHM. Then we need to introduce the root mean square (RMS) width (Trms) of the pulse [3] as

\[
T_{\text{rms}} = \left[\langle t^2 \rangle - \langle t \rangle^2 \right]^{1/2}
\]

(2.38)

where the angle brackets denote averaging with respect to intensity profile, i.e.,

\[
\langle t^m \rangle = \frac{\int_{-\infty}^{+\infty} t^m |A(z,t)|^2 dt}{\int_{-\infty}^{+\infty} |A(z,t)|^2 dt}
\]

(2.39)

\(|A(z,t)|^2 = P(t)| is the pulse power. Trms measures how much the pulse is spread in time with respect to the middle point. Notice that the smaller Trms is, the more energy of the pulse is confined close to the peak.

The last concept introduced is the extinction ratio (ER). Considering a pulse, ER is the ratio between the peak power and minimum power (see figure 2.3). We desire an ER as high as possible and we used to measure it in dB.

\[
ER = \frac{P_{\text{max}}}{P_{\text{min}}}
\]

(2.40)
2.4 Fourier Transform

In telecommunications we often refer to the signal description both in time and frequency domain. We want to clarify the concept of Fourier Transform (FT) (and inverse Fourier Transform (IFT)) to explain how we move from one domain to the other. For this reason we define FT as

\[ X(f) = F(x(t)) = \int_{-\infty}^{+\infty} x(t) e^{-j2\pi ft} dt \]  

(2.41)

and IFT as

\[ x(t) = F^{-1}(X(f)) = \int_{-\infty}^{+\infty} X(f) e^{j2\pi ft} df \]  

(2.42)

2.4.1 Frequency Domain Representation

When we represent a signal in the frequency domain we used to talk about Power Spectral Density (PSD), measured in [W/Hz] and also known as spectrum of the signal. PSD is defined as the FT of the autocorrelation\(^1\) signal as

\[ S(f) = F(R(t)) = \int_{-\infty}^{+\infty} R(t) e^{-j2\pi ft} dt \]  

(2.43)

2.4.2 Filter

We use a filter to remove the undesired frequency components of a signal. A filter is characterized by an impulse response, usually called \( h(t) \), which is the output of the

\(^1\)The autocorrelation function of a signal \( x(t) \) is calculated as: \( R_{x,x}(t) = x(t) * x^*(t) = \int_{-\infty}^{\infty} x(\tau) x^*(t-\tau) d\tau \)
2.5. BER and OSNR

filter when the input signal is a \( \delta(t) \) (delta Dirac's funcion). The Fourier transform of the filter impulse response is called *frequency response*, denoted with \( H(f) \).

For an input signal \( x(t) \), the output of the filter is given by the convolution of \( x(t) \) with \( h(t) \)

\[
y(t) = x(t) * h(t) = \int_{-\infty}^{+\infty} x(\tau) h^*(t - \tau) d\tau
\]

Applying Fourier transform, and using the property that convolution converts to product when using \( FT \), we obtain

\[
Y(f) = F(y(t)) = X(f) \cdot H(f)
\]

An important parameter related to filters is the *filter bandwidth* (BW). To measure it we need to evaluate the lower and the upper cut-off frequencies, which we define as the frequencies where \( |H(f)|^2 \) is half of the maximum value, and in that case we talk about 3-dB BW (because 3-dB corresponds to divide the maximum by 2).

We used in our system both optical and electrical filters, and their implementation will be explained in following sections.

2.5 BER and OSNR

Nowadays, we want to transmit as many information and as fast as possible, but it should be in a proper way. Then we need to talk about how accurate is our systems in terms of detecting the signal, so we used to refer to *Bit Error Rate* (BER), *Optical Signal to Noise Ratio* (OSNR) and *Signal to Noise Ratio* (SNR), so it is also important to find a relationship between them.

2.5.1 Bit Error Rate (BER)

*BER* is a very important parameter in the design of a digital communication system. The *BER* is defined as the number of incorrect received bits over a total number of transmitted bits, when the signal is sent through an *additive white gaussian noise channel* (AWGN). So normally errors are not systematic and they are due to noise.

\[
BER = \frac{\text{no. of incorrect received bits}}{\text{no. of transmitted bits}}
\]

2.5.2 OSNR

*OSNR* is important because it suggests a degree of impairment when the optical signal is carried by an optical transmission system that includes optical amplifiers. It is defined as the ratio between the signal power and the noise power in an optical channel. In optical communications we used to draw the relationship between *BER* and *OSNR* to evaluate the system behaviour.
2.6. OTDM

It would be important to specify a precise definition of OSNR as:

$$\text{OSNR} = \frac{<P(t)>}{p \cdot \text{No} \cdot Bref} \quad (2.47)$$

where at the numerator we found the mean signal power ($<P(t)>$), while at the denominator we have the number of polarizations in which noise is present ($p$), the single-sided noise spectral density per polarization ($\text{No}$), and a reference bandwidth ($Bref$), which is 0.1 nm (or 12.5 GHz)$^1$ in our model, as generally found in literature [20]. The value of $p$ can be either 1 or 2, depending if the noise occurs only in one or in both polarizations ($x$ and $y$).

2.5.3 Relation OSNR and SNR

The $\text{SNR}$ can be defined in different ways. In our model it is defined as one of the most used ways,

$$\text{SNR} = \frac{E_b}{p \cdot \text{No}} \quad (2.48)$$

where at the numerator $E_b$ is the energy per bit, and at the denominator, as previously in (2.47), $p$ is the number of polarizations in which we have noise, and $\text{No}$ is the single-sided noise spectral density per polarization.

It is important to find a relationship between both $\text{SNR}$ and $\text{OSNR}$ terms, because in some papers [8] the results associate BER with $\text{SNR}$, whereas in others [24] the relationship is between BER and $\text{OSNR}$. In chapter 4 we will also compare our results with some papers [8] and books [3] for validation and for being able to change safely both relationships. If we define $m$ as the number of bits in one symbol, the relationship between $\text{SNR}$ and $\text{OSNR}$ could be define as

$$\text{OSNR} = \frac{\text{SNR} \cdot Rs \cdot m}{Bref} \quad (2.49)$$

and also in dB as

$$\text{OSNR(dB)} = \text{SNR(dB)} + 10 \log_{10} \left( \frac{Rs \cdot m}{Bref} \right) \quad (2.50)$$

2.6 OTDM

Multiplexing is a process where several optical signals (called channels) are combined and transmitted together over the same medium. The technique that we use in our system to multiplex information is optical time-division multiplexing (OTDM).

The use of OTDM allows to increase the overall data transmission capacities without increasing the data rate of a single channel. However, every channel has a fraction of

$^1$To convert from $\Delta \lambda = 0.1 \text{nm}$ to $\Delta f = 12.5 \text{GHz}$ we apply the relation $\Delta f = \frac{c}{\lambda^2} \cdot \Delta \lambda$
the symbol slot. If we have $N$ channels in our OTDM system, every channel will have a time slot of $T_{sch} = T_s / N$, while the data rate at the output will be $R' = R_s \cdot N$.

When multiplexing different channels, we expect to have our pulse completely included in $T_{sch}$, but this is never realized in practice and some of the energy of the pulse interacts with the adjacent channels. We call this phenomenon crosstalk, and it is an effect that decrease the quality of transmission introducing errors in the information we want to transmit. We measure it in two different ways, Energy crosstalk ($E_{crosstalk}$) and Power crosstalk ($P_{crosstalk}$).

As we can see in the figure 2.5, $E_{crosstalk}$ measures how much energy of the pulse is included in the adjacent channel, while $P_{crosstalk}$ measures the difference, in dB, between the peak value of the main pulse and the power value in the middle of the adjacent channel. Both contributions will be studied in chapter 4.
2.7 Optical Transmitters

In this section we just want to define some notations about two important components that we use in our transmission system, such as light sources and modulator.

Optical transmitters need a light source and we will concentrate in lasers, particularly in continuous-wave (CW) lasers. We transmit the information in a determined modulation, and we combine the laser with an external modulator to obtain it. In our study we will focus on Mach-Zehnder modulator (MZM).

2.7.1 CW laser

A laser is a light source that usually has two parts: the laser cavity, in which light can circulate (e.g. between two mirrors), and an active medium. which serves to amplify the light.

We have to define the radiative processes occurring in a laser when a light is generated. In a laser we have several energy levels, \( E_i = h \nu_i \), but for simplicity we just consider two levels to explain the processes[9]:

- **Stimulated absorption**: It is a process by which an incident photon is absorbed by an electron which increments its energy level.

- **Spontaneous emission**: It is a process by which an excited electron in \( E_2 \) releases energy in the form of a photon, with energy \( h \nu_i = E_2 - E_1 \) and random frequency, phase and direction.

- **Stimulated emission**: It is a process by which an incident photon forces an excited electron to release its energy in the form of a new photon with exactly the same frequency, phase and direction.

The main phenomenon during the generation of a lightwave in a laser is stimulated emission. In our system we can adapt (2.16) for the output of a CW laser as (deleting the phase modulation terms)

\[
E(t) = \sqrt{P_0} e^{-j \omega_0 t} \quad (2.51)
\]

where \( P_0 \) is the output power and, if \( f_0 \) is the central frequency, \( \omega_0 = 2\pi f_0 \), then the spectrum will have a single peak at \( f_0 \) considering the laser as mono-mode. In the case of real lasers, we should modify (2.51) to take into account both phase and intensity noise.

\[
E(t) = \sqrt{P_0 + \delta P} e^{-j (\omega_0 t + \delta \omega)} \quad (2.52)
\]

The phase noise is due to the presence of spontaneous emission in the laser, when the generated photon has random phase and direction; whereas the intensity noise is due to the quantization of photons and electrons.
2.7. Optical Transmitters

2.7.2 Mach-Zehnder Modulator

MZM is based in the electro-optic effect (EO) [3]. In the Mach-Zehnder an electric field is applied in the two different arms of the modulator, that allow us to control the phase, and it makes Mach-Zehnder a phase modulator, and if the contribution in each arm is properly controlled, to obtain an intensity modulator. Let us introduce the model for the MZM, the different operation modes and the expression with Bessels functions.

2.7.2.1 Model

The structure of the MZM is depicted in figure 2.6. It consists of a waveguide where the input signal is divided into two different arms, with a contribution of $\sqrt{\alpha}$ and $\sqrt{1-\alpha}$ respectively (splitting ratio $\alpha$ in the input), where in each arm the optical signal has a different phase shift equal to $\phi_1$ and $\phi_2$, which is also dependent on the applied voltage $V_1(t)$ and $V_2(t)$, and finally both signals are combined again, with a contribution of $\sqrt{\beta}$ and $\sqrt{1-\beta}$ (splitting ratio $\beta$ in the output).

![Figure 2.6: Model of Mach-Zehnder Modulator.](image)

Where $V_i(t) = V_{dci} + V_{aci}(t)$, $V_{aci}(t) = V_{ppi} \ast d_i(t)$ and $d_i(t) = \frac{1}{2}\sin(2\pi f_s t)$, for $i=1,2$.

And the phase shift is defined as [14]:

$$\phi_i(t) = \frac{\pi V_i(t)}{V_\pi}; \quad i = 1, 2;$$  \hspace{1cm} (2.53)

The output field can be expressed as:

$$E_{out}(t) = \left(\sqrt{\alpha}e^{-j\phi_1(t)} + \sqrt{1-\alpha}e^{-j\phi_2(t)}\right) E_{in}(t)$$  \hspace{1cm} (2.54)

For simplicity, we consider that the input and output Y-junctions are $\alpha=\beta=1/2$. We can re-write the expression (2.54) as:

$$E_{out}(t) = \left(\frac{1}{2}e^{-j\phi_1(t)} + \frac{1}{2}e^{-j\phi_2(t)}\right) E_{in}(t)$$  \hspace{1cm} (2.55)

and defining the next parameters:

$$\Delta\phi(t) = \frac{\phi_1(t) - \phi_2(t)}{2}$$  \hspace{1cm} (2.56)

$$\Theta\phi(t) = \frac{\phi_1(t) + \phi_2(t)}{2}$$  \hspace{1cm} (2.57)
2.7. Optical Transmitters

combining (2.55) with (2.56) and (2.57) we obtain the expression of \( E_{out}(t) \) as:

\[
E_{out}(t) = \frac{E_{in}(t)}{2} e^{-j\phi(t)} \left( e^{-j\Delta \phi(t)} + e^{j\Delta \phi(t)} \right) = E_{in}(t) e^{-j\frac{\phi_1(t) + \phi_2(t)}{2}} e^{-j\phi(t)}
\]

(2.58)

\[
E_{out}(t) = E_{in}(t) \cos \left( \frac{\phi_1(t) - \phi_2(t)}{2} \right) e^{-j\frac{\phi_1(t) + \phi_2(t)}{2}}
\]

(2.59)

and finally the expression of \( P_{out}(t) \) as:

\[
P_{out}(t) = |E_{out}(t)|^2 = P_{in}(t) \cos^2 \left( \frac{\phi_1(t) - \phi_2(t)}{2} \right)
\]

(2.60)

2.7.2.2 MZM Transfer function

We are going to represent the transfer function of the modulator \( P_{out}/P_{in} \) as a function of the difference between the voltages applied to each arm.

Introducing (2.53) in (2.60) we obtain

\[
\frac{P_{out}(t)}{P_{in}(t)} = \cos^2 \left( \frac{\pi}{2V_{\pi}} (V_1 - V_2) \right)
\]

(2.61)

\[
V_1 = V_{dc1} + \frac{V_{pp1}}{2} \sin(2\pi f_s t) \quad \text{and} \quad V_2 = V_{dc2} + \frac{V_{pp2}}{2} \sin(2\pi f_s t) \quad \text{and} \quad V_1 - V_2 \text{ could be expressed as}
\]

\[
\Delta V = V_1 - V_2 = \Delta V_{bias} + \frac{\Delta V_{pp}}{2} \sin(2\pi f_s t)
\]

(2.62)

where \( \Delta V_{bias} = V_{dc1} - V_{dc2} \) known as modulator bias; and \( \Delta V_{pp} = V_{pp1} - V_{pp2} \) is the peak-to-peak difference between the driving signals.

In figure 2.7 is shown the transfer function of the MZM.

![Figure 2.7: Transfer function of MZM.](image)
2.7. Optical Transmitters

2.7.2.3 Operation modes

Usually, there are two operation modes in a MZM, the push-pull mode and the asimetric mode.

- **Push-Pull mode**: In this operation mode, we apply in both arms of MZM the same signal but with an opposite phase in each one of them. The benefit of this operation mode is that the resulting modulation is an amplitude modulation and the output signal is unchirped.

The condition is:

$$\phi_1(t) = -\phi_2(t) = \phi(t) \quad (2.63)$$

Under this condition applied in (2.59) is possible to define :

$$E_{out}(t) = E_{in}(t) \cos(\phi(t)) \quad (2.64)$$

which is an amplitude modulation where the expression of $P_{out}(t)$ is :

$$P_{out}(t) = |E_{out}(t)|^2 = P_{in}(t) \cos^2(\phi(t)) \quad (2.65)$$

- **Asimetric mode**: In the asymetric mode operation of MZM, the input signals in both arms of the modulator are different, and the output signal is chirped. So in that mode the expression of the electric field and power are (2.59) and (2.60).

We will concentrate our efforts working in this operation mode because it allows us to work with phase modulation, and we are interested in chirped signals to use pulse compression.

2.7.2.4 Frequency chirp and MZM

Let us clarify the concept of **chirp**. Equation (2.53) shows that the phase shift is time dependent. In communications, a time varying phase is equivalent to a change in the instantaneous frequency of the signal

$$f(t) = f_0 + \delta f(t) \quad (2.66)$$

where

$$\delta f(t) = -\frac{1}{2\pi} \frac{\delta \phi(t)}{dt} \quad (2.67)$$

The frequency experiences instantaneous shifts depending on the time derivative of the phase, and this phenomenon is known as chirp.

It is generally considered that frequency chirp has to be avoided in optical transmitters, since it will broaden the spectrum. However, it is also known that frequency chirp can modify the evolution of a pulse in a fiber link. As explained in section 2.7.2.3, the chirp effect can be controlled in a MZM applying the proper voltage in each arm of the modulator.
2.7. Optical Transmitters

2.7.2.5 MZM and Bessel function

We expressed the optical field at the output of the MZM as (2.55) but it is also important for us to find an expression using Bessel functions.

Combining the expression (2.53) and (2.55) and $V_i(t) = V_{dc} + \frac{V_{pp}}{2} \sin(\omega t)$ we obtain

$$E_{out}(t) = \frac{E_{in}(t)}{2} \left( e^{-j\frac{\pi V_{dc}}{V_\pi}} e^{-j\frac{\pi V_{pp}}{2} \sin(\omega t)} + e^{-j\frac{\pi V_{dc}}{V_\pi}} e^{-j\frac{\pi V_{pp}}{2} \sin(\omega t)} \right)$$ (2.68)

and using

$$\sum_{n=-\infty}^{+\infty} J_n(A) e^{jn\phi} = e^{jAsin(\phi)}$$ (2.69)

where $J_n(x)$ is the $n$-th order Bessel function, and using the property $J_n(x) = J_n(-x)$, then

$$A_i = \frac{\pi}{V_\pi} V_{pp}; \quad \text{and} \quad \phi_i = \omega t; \quad i = 1, 2;$$ (2.70)

Introducing it on (2.68) we define

$$E_{out}(t) = \frac{E_{in}(t)}{2} \left( e^{-j\frac{\pi V_{dc}}{V_\pi}} \sum_{k=-\infty}^{+\infty} J_k(A_1) e^{jk\omega t} + e^{-j\frac{\pi V_{dc}}{V_\pi}} \sum_{k=-\infty}^{+\infty} J_k(A_2) e^{jk\omega t} \right)$$ (2.71)

Developing the previous expression

$$E_{out}(t) = \frac{E_{in}(t)}{2} \sum_{k=-\infty}^{+\infty} \left( J_k(A_1) e^{j(k\omega t - \frac{\pi V_{dc}}{V_\pi})} + J_k(A_2) e^{j(k\omega t - \frac{\pi V_{dc}}{V_\pi})} \right)$$ (2.72)

Now, we declare

$$\theta_i = -\frac{\pi V_{dc}}{V_\pi}; \quad i = 1, 2;$$ (2.73)

and we obtain the optical field at the output of MZM as

$$E_{out}(t) = \frac{E_{in}(t)}{2} \sum_{k=-\infty}^{+\infty} \left( J_k(A_1) e^{j(k\omega t + \theta_1)} + J_k(A_2) e^{j(k\omega t + \theta_1)} \right)$$ (2.74)

as we can find in [17] and [18].
Let us conclude with two important definitions that we will use later on chapter 4. We define $2\Delta A$ as the peak-to-peak phase difference induced in each arm; and $2\Delta \theta$ is a dc bias difference between the arms:

\[
\Delta A = \frac{A_1 - A_2}{2} \quad (2.75)
\]

\[
\Delta \theta = \frac{\theta_1 - \theta_2}{2} \quad (2.76)
\]

### 2.7.2.6 Frequency comb generation and MZM

One of the applications that offer an optical frequency comb generator is ultrashort pulse generation. The principle of the optical frequency comb generation using a MZM [17] is shown in figure 2.8.

[Figure 2.8: Concept of ultraflat optical frequency comb generation using a MZM.]

As we seen in figure 2.8, spectral unflatness can be canceled if the dual arms of the MZM are driven by in-phase sinusoidal signals, \textit{rf-a} and \textit{rf-b} in figure 2.8, with a specific amplitude difference. To make the comb flat in the optical frequency domain, the driving condition becomes [13]

\[
\Delta A \pm \Delta \theta = \frac{\pi}{2} \quad (2.77)
\]

This condition is called the flat spectrum condition of the EO-modulated lightwave, and $\Delta A$ and $\Delta \theta$ are described in the previous section.

### 2.8 Optical Receivers

In this section we introduce some notations about optical receivers in an optical communication system. We show a general block diagram of an optical receiver, as well as information about photodetectors and the decision circuit.
2.8. Optical Receivers

2.8.1 Block Diagram

A general block diagram for a direct detection (DD) receiver is shown in figure 2.9.

![Figure 2.9: Block diagram for direct detection receiver.](image)

First of all we introduce an optical bandpass filter (OBPF) to reduce the ASE noise (affecting frequencies at the output of the spectrum of signal) and improve the receiver sensitivity. Then there is a photodetector (in our case a photodiode) to convert the signal from the optical domain to the electrical one (further explanation section 2.8.2). At the output of photodetector appears shot and thermal noise, and an electrical filter is used to reduce it. Finally, we employ a decision circuit to determine the received data.

2.8.2 Photodetectors

Photodetector devices are used to detect the optical power of an incident light. There are many types of photodetectors, and we will use a photodiode.

Photodiodes are semiconductors devices which contain a p-n junction, and often an intrinsic layer between the n and p layers. Devices with an intrinsic layer are called PIN photodiodes [1].

In a photodiode, the photocurrent is

\[ I = R \cdot P_{in} \]  

(2.78)

where \( R \) is the responsivity, measured in \([A/W]\), and it is directly related to the quantum efficiency, which is the fraction of the incident photon which contributes to the photocurrent

\[ R = \frac{\eta e}{h\nu} \]  

(2.79)

with the quantum efficiency \( \eta \), the electron charge \( e \) and the energy \( h\nu \) of the incident photons.

The noise performance of a photodiode can be determined by shot noise, thermal noise and dark current:

- **Shot noise:** It is related to the discreteness of photons and electrons, and can be observed as oscillations in the photocurrent when a constant optical input power is applied to the photodiode.
2.8. Optical Receivers

- **Thermal noise**: It is the noise generated by thermal agitation of electrons in a conductor, which is reflected in a fluctuation of the current through a resistor even in absence of applied power.

- **Dark current**: It is the amount of output current measured when there is no power applied to the photodiode input.

2.8.3 Decision circuit

The decision changes the analog photocurrent to the digital domain, and makes an output signal based on an specific instant of time, called **sampling time** \( t_{\text{sample}} \), and a **decision threshold** \( I_{\text{th}} \).

The method to decide a "1" or a "0" is simple. Considering a single threshold, if the photocurrent at \( t_{\text{sample}} \) is higher than the threshold, we decide we have a "1". Otherwise, the output is a "0".

It is important to clarify that the received signal at the input of the decision circuit is affected by noise (caused by noise sources explained previously), which introduces errors in our detection, so we will measure at the output of the decision circuit the **BER** of the system.
Analysis and Design

In the following sections we will explain how we generate the different parts of the optical communication system model developed on this project.

For the simulation we use Matlab® software that allows us to implement the different components of the system. Most of the functions of the model derive from the theory described in chapter 2. We are not going to detail all the single functions and operations used for the simulations, we will try to define just the most important parts.

This chapter is divided into seven parts. The first part presents the scheme of our optical communication system. The second introduces the model setup for the simulations. The third describes our transmitter model. In the fourth one, we model the noise in the optical channel. In the fifth, we develop our model of an optical receiver. The sixth traces how we determine the BER in our system. Finally, in the last section we explain the spectrum analysis.

3.1 Optical communication system

The first section presents the scheme of our optical communication system (see figure 3.1), where we show the connections of the different parts that will be described in this chapter.

The upper part corresponds to the optical transmitter, while in the lower part we add noise to the signal before going to the optical receiver.

3.2 Model setup

This section explains how we generate the time and the frequency arrays used in the different simulations on Matlab® over the time and the frequency domains.! The second part describes a Pseudo-random bit sequence (PRBS) used to transmit information.
3.2. Model setup

Analysis and Design

3.2.1 Time and Frequency Array

In our model we represent signals in the digital domain, so we need to define the time and frequency array.

The time array is an array of temporal values to describe a $T$ duration time window. The parameters that determine the time (and frequency) array are: number of symbols ($N_{\text{symbols}}$), the number of samples used in each symbol ($N_{\text{SamplesPerSymbol}}$) and the symbol rate ($N_{\text{SymbolRate}}$).

We have a total number of points ($N_{\text{Samples}} = N_{\text{SamplesPerSymbol}} \times N_{\text{symbols}}$) to define the time array, and the distance between two consecutive points (time resolution) is $\Delta t$, defined as $\Delta t = 1/fs$, where $fs$ is the sample rate $SampleRate = N_{\text{SamplesPerSymbol}} \times N_{\text{SymbolRate}}$, so the total duration $T$ is $T = N_{\text{Samples}} \times \Delta t$.

Figure 3.1: Scheme of our optical communication system. The upper part corresponds to the transmitter, that continues in the lower part with the receiver.

Figure 3.2: Time and Frequency array definitions.
When we refer to the spectrum of the signal (and also for the filtering operations) we need to work in the frequency domain. Both time and frequency array have the same length, and as in the time domain, we define the frequency resolution, which in that case depends on the time window $T$ of the simulation, as $\Delta f = 1/T$.

Let us clarify that $\text{SamplesPerSymbols}$ is defined as a power of 2, because when we need to go from one domain to the other, the discrete Fourier Transform operation in Matlab® software (and also the inverse Fourier Transform) is faster if the number of samples is a power of 2.

Finally, the code is:

\[
\begin{align*}
\text{TimeArray} &= (0: \text{Nsamples}-1) \times \text{delta}_t. \\
\text{FrequencyArray} &= (-\text{Nsamples}/2: \text{Nsamples}/2 - 1) \times \text{delta}_f.
\end{align*}
\]

### 3.2.2 Pseudo Random Bit Sequence (PRBS)

In optical communications, a PRBS sequence is widely used to evaluate the performance of a system. A PRBS is a semi-random sequence because it appears approximately random data, but actually the entire sequence is deterministic and repeated indefinitely [26]. In our implemented model, we use a PRBS sequence for BER testing because it is easy to synchronize a transmitted PRBS pattern to a received stream for bit by bit comparison.

A $n$-bit PRBS generator consists of a $n$-bit shift register and a feedback [20], where typically the feedback is a XOR gate, which inputs are the outputs of the shift register and the output is fed back to the first stage of the register.

If the two inputs bits are identical, the XOR produces a "0" at the output, otherwise produces a "1". Selecting the appropriate location for the XOR gate, a so-called maximum length sequence with a period of $2^n-1$ contains all possible $n$ pattern, except for the one with $n$ zeros can be produced.

In figure 3.3 is shown the PRBS sequence of length $2^7 - 1 = 127$ bits implemented in our model. In this case, the XOR gate is situated between the output and the sixth register.

![Figure 3.3: PRBS generator scheme.](image)


3.3 Transmitter Model

In the next section we will explain the different blocks that are included in our optical communication transmitter system. We divide the section into four parts. The first contains the different blocks for the generation of a picosecond pulses. The second part analyzes the elements for pulse compression to femtosecond pulses. The third part takes into account the structure of OOK transmitter. Finally, we describe the structure of the OTDM multiplexer.

3.3.1 Ultrashort pulses using Frequency Comb Generation in a Mach-Zehnder modulator

Here we are going to describe the different blocks that we will need for the generation of a train of ps-pulses, and all of them are based on the theory explained in chapter 2. As we can see in figure 3.4, it is composed by a CW laser, an external modulator (MZM), an optical band-pass filter (OBPF) and a single-mode fiber (SMF).

![Figure 3.4: Scheme for a ps-pulses train generation.]

This is the complete scheme that we will employ in our thesis, although in chapter 4 we will also analyze a configuration without the OBPF between the MZM and the SMF.

3.3.1.1 CW laser

In our system we work with a CW laser as a light source, and it operates at a certain wavelength that is \( \lambda = 1550 \) nm, so the central spectrum frequency is \( f_0 = c/\lambda = 193.12 \) THz, and it is too high compared with the modulation frequency \( f_s = 10 \) GHz and we will need a high quantity of samples to describe the signal. Moreover, the spectrum until \( f_0 \) will be unused. To solve this problem, we are going to represent a baseband version of the signal (see section 2.1.1), and for that reason we will select \( f_{ref} = f_0 \).

We want to clarify that in our implementation of a CW laser source, we avoid the effects related to the phase and intensity noise of the laser, so our model is based on equation (2.51).

3.3.1.2 Mach-Zehnder Modulator

Following the CW laser there is a MZM. We will base our implementation of the MZM model using Matlab® software in the theory explained in section 2.7.2.
3.3. Transmitter Model

In the part of generation the ps-pulse train, we use a MZM working on the asymmetric mode operation; whereas we use a MZM operating in the push-pull mode for the implementation of the OOK transmitter.

Working in one or another mode operation, the MZM output depends on the phase shift of each arm of the modulator, which at the same time depends on the DC voltage applied in each arm ($V_{dc}$) and on the electrical signal generated ($d_i = 1/2 \sin(2\pi f_s t)$) multiplied by $V_{pp}$ ($V_{aci} = V_{pp}\cdot d_i(t)$), where $f_s$ correspond to the SymbolRate.

Moreover, we define the splitting ratios $\alpha$ and $\beta$ equal to 1/2 in both MZM.

3.3.1.3 Optical filter

Generally, we use a filter to reduce the noise affecting the signal. In that case, at the output of the MZM we have a chirped signal (because the MZM operates in asymmetric mode) and we utilize the filter for spectral shaping and to reduce the amount of chirp.

As we explain in section 2.4.2, the operation of filtering is done in the frequency domain, and to transform from the time domain we use the Fourier and inverse Fourier transform (equations (2.41) and (2.42)). In Matlab® software, we use the discrete Fourier transform ($X = \text{fft}(x)$) and inverse Fourier transform ($x = \text{ifft}(X)$) between two vectors of length $N$ defined as

$$X(k) = \sum_{i=1}^{N} x(i) e^{-\frac{2\pi j}{N}(k-1)(i-1)}$$

(3.1)

and

$$x(i) = \frac{1}{N} \sum_{k=1}^{N} X(k) e^{\frac{2\pi j}{N}(k-1)(i-1)}$$

(3.2)

As a consequence, we use the function `fft` and `ifft` to calculate the FT and IFT transformations. Another important function used is "fftshift", that moves the zero frequency components to the center of the array. Finally, the operation of filtering in Matlab® is

$$x_f = \text{ifft} (\text{fft}(x) \cdot \text{fftshift}(H_f)),$$

where $H_f$ is the transfer function of the optical filter.

In our model, optical filters are band-pass filters modeled as Gaussian filters of $m$-th order. The transfer function is

$$H(f) = \exp\left(-\frac{1}{2}\left(\frac{f-f_c}{f_0}\right)^{2m}\right)$$

(3.3)

where $m$ is the order of the filter, $f_c$ is the central frequency and $f_0$ is related to the $f_{3dB}$ as

$$f_0 = \frac{f_{3dB}}{2 \sqrt{\ln(2)}}$$

(3.4)

where $f_{3dB}$ is the 3-dB BW of the filter.
3.3. Transmitter Model

3.3.1.4 SMF

As we mentioned previously, at the output of the MZM we have a chirped signal and we employ a piece of SMF as a dispersive element to compensate the frequency chirp in the optical signal, which converts the flat optical comb signal to a picosecond pulse train.

The SMF is defined by the fiber length \( L \) and the dispersion parameter \( D \) (see equation (2.26)). We use a SMF with positive \( D = 17 \text{ ps/(nm·km)} \), but the length of the fiber will depend on the amount of chirp of the signal.

As we explain in section 2.2.3, the GVD is frequency dependent, so we need to operate in the frequency domain. The different spectral components of an optical pulse propagate inside the fiber according to the relation

\[
A(L, w) = A(0, w) e^{iβL} \tag{3.5}
\]

where \( A(0, w) \) is the FT of the pulse at the input of the fiber, \( A(L, w) \) is the FT of the pulse at the output of the fiber, \( L \) is the SMF length and \( β(w) \) is the GVD (see equation (2.25)).

In our implementation of SMF, we just consider the effect of \( β_2 \) (not considering high order dispersion \( β_3 \)) and avoid the effect of losses. In figure 3.5 is shown the process for a SMF simulation, where \( Δw = w - w_0 \), and \( w_0 \) is the frequency at which pulse spectrum is centered (remember we are working in base-band, so we consider \( w_0 = 0 \)), so \( Δw = w = WArray \) and \( WArray = 2πFrequencyArray \).

![Figure 3.5: Scheme for SMF diagram](image)

The code is:

\[
\begin{align*}
X &= \text{fft}(\text{Einput}); \\
H &= \text{fftshift}(\exp(1i*0.5*\text{beta}_2*z*WArray.^2)); \\
Xz &= X.*H; \\
Eoutput &= \text{ifft}(Xz);
\end{align*}
\]

3.3.2 Pulse compression with Split-Step Fourier method

In this section, we are going to describe the different blocks that we use for the pulse compression of the generated ps-pulses to a fs-pulses. As we can see in figure 3.6,
3.3. Transmitter Model

it is composed by an amplifier, a Dispersion Flattened High NonLinear Fiber (DF – HNLF), an OBPF and a SMF.

![Diagram of transmitter model]

**Figure 3.6:** Scheme for a pulse compression stage.

3.3.2.1 Amplifier

We introduce an amplifier because as we will explain later in the next section, the nonlinearities of the DF – HNLF depend on the value of $|P(t)|^2$ at the input.

The implementation of an amplifier is quite simple, just need to multiply the input signal by $\sqrt{G}$ to obtain a gain $G$ in the power of the signal.

$$E_{\text{out}} = \sqrt{G} \times E_{\text{in}};$$

3.3.2.2 Dispersion Flattened High NonLinear Fiber (DF-HNLF)

We obtain a train of ps-pulses but we still need a pulse compression stage to arrive to fs-pulses. For this reason we use a DF – HNLF, where the proper combination of dispersion and nonlinearities can lead to pulse compression.

In the DF – HNLF we need to simulate together the effects of losses, dispersion and nonlinearities, so we apply to numerical solutions to solve the general NLSE, and in our model we implement the commonly used Split-Step Fourier method (SSMF) [16]. In this section we avoid the effect of losses, because the length of the fiber is not too long, and the contribution of losses is not relevant, so the equation of NLSE

$$\frac{\delta A}{\delta z} = -j \frac{\beta_2}{2} \frac{\delta^2 A}{\delta t^2} + \frac{\beta_3}{6} \frac{\delta^3 A}{\delta t^3} + j \gamma |A|^2 A$$

(3.6)

can be written as

$$\frac{\delta A}{\delta z} = [\tilde{D} + \tilde{N}] A$$

(3.7)

where

$$\tilde{D} = -j \frac{\beta_2}{2} \frac{\delta^2 A}{\delta t^2} + \frac{\beta_3}{6} \frac{\delta^3 A}{\delta t^3}$$

(3.8)

$$\tilde{N} = j \gamma |A|^2$$

(3.9)
3.3. Transmitter Model

In general, dispersion and nonlinearities act together along the fiber length. The SSMF obtains an approximate solution by assuming that in propagating the optical field over a small distance \( \Delta z \), called step size, the dispersive and nonlinear effects act independently. Moreover, the propagation from \( z \) to \( z + \Delta z \) is carried out in two steps. The first step considers the dispersion, and in the second the nonlinearity acts alone.

The linear part \( \tilde{D} \) is easily solved in the frequency domain

\[
\frac{\delta \tilde{A}}{\delta z} = \tilde{D}(w) \tilde{A}
\]  (3.10)

the FT \( \tilde{D}(w) \) of \( \tilde{D} \) is obtained by using the Fourier transformation replacement property \( \delta/\delta t \rightarrow -jw \)

\[
\tilde{D}(w) = j \frac{\beta_2}{2} |w|^2 - j \frac{\beta_3}{6} |w|^3
\]  (3.11)

\[
\tilde{A}(z + \Delta z, w) = \tilde{A}(z, w) \exp[j \frac{\beta_2}{2} |w|^2 - j \frac{\beta_3}{6} |w|^3]
\]  (3.12)

The nonlinear part is implemented in the time domain

\[
A(z + \Delta z, t) = A(z, t) \exp[j \gamma |A(z, t)|^2 \Delta z].
\]  (3.13)

Considering a step of length \( \Delta z \), we follow the next process (see figure 3.7):

1. Apply FT to the input signal.
2. Introduce dispersion over a total length of \( \Delta z/2 \).
3. Go back to time domain of optical signal applying IFT.
4. Calculate nonlinear effect over a total length of \( \Delta z \).
5. Repeat the steps 1, 2 and 3.

![Figure 3.7: Process for Split-step Fourier method.](image)

Now it is important to determine the step size \( \Delta z \), because there are several techniques to define \( \Delta z \). In our model we study three different ones:
3.3. Transmitter Model

- **Constant step size**: Using this method we just divide the total length of the fiber in $N$ small parts of length $\Delta z$ ($\Delta z = L/N$). This is the simplest method of the three, although the global accuracy can be improved increasing the number of steps, and it needs more number of fft too [21].

- **Adapative step size**: For a step size of $\Delta z$, the effect of the nonlinear operator $\tilde{N}$ is to increment the phase by an argument of $\phi_{NL} = |A(z,t)|^2 \Delta z$. If we impose an upper limit $\phi_{NL}^{max}$ on the nonlinear phase, we obtain the step size

  $$
  \Delta z = \frac{\phi_{NL}^{max}}{\gamma |A(z,t)|^2}
  $$

  (3.14)

  and then the value of $\Delta z$ will change depending on the value of $|A(z,t)|^2$ at the input.

- **Error method step size**: Given the pulse $|A(z,t)|^2$, our aim is to compute the pulse at $2\Delta z$, where $\Delta z$ is the step size (the first $\Delta z$ defined using "adapative step-size method"). Then we apply the following process:

  1. We obtain the solution $A(z + \Delta z)$ using one step of $l = 2\Delta z$, known as *coarse solution* ($A_c$).
  2. We obtain the solution $A(z + \Delta z)$ using two steps of $l = \Delta z$ each one, known as *fine solution* ($A_f$).

  We define the relative local error $\delta$ of the step, where

  $$
  \delta = \frac{||A_f - A_c||}{||A_f||}
  $$

  (3.15)

  $$
  ||A|| = \sqrt{\int |A(t)|^2 dt}
  $$

  (3.16)

  and to evaluate the integral we use the *Simpson’s rule*\(^1\) [2].

  Then the step size is chosen by keeping the relative local error within a specified range $(1/2\delta_G, \delta_G)$, where $\delta_G$ is a *goal relative error*.

  1. If $\delta > 2\delta_G$, the solution is discarded and the step size is divided by 2.
  2. If $\delta \in (\delta_G, 2\delta_G)$, the step size is divided by a factor $2^{1/3}$ for the next step.
  3. If $\delta < 1/2\delta_G$, the step size is multiplied by a factor $2^{1/3}$ for the next step.

  Then, the final solution at the output of $2\Delta z$ is given by

  $$
  A(z + 2\Delta z) = \frac{4}{3}A_f - \frac{1}{3}A_c
  $$

  (3.17)

  It is the method that we will use in our simulation, although it requires more time to decide $\Delta z$, it needs less number of fft [21], which consumes a lot of computational calculation.

\(^1\)The Simpson’s rule is a method for numerical integration and define the following approximation: $\int_a^b f(x)dx = \frac{b-a}{6} [f(a) + 4f(\frac{a+b}{2}) + f(b)]$. 
3.3. Transmitter Model

3.3.2.3 OBPF and SMF

The implementation of the OBPF and SMF in the pulse compression stage is the same than in section 3.3.1.3 and 3.3.1.4 respectively.

3.3.3 OOK Transmitter

For the transmission of data using an OOK modulation, we implement in our model an OOK Transmitter (see figure 3.8).

As the OOK is an amplitude modulation, we use a MZM working in push-pull mode operation (see section 2.7.2.3), with a modulation bias $\Delta V_{bias} = V\pi$; and a peak-to-peak difference between driving voltage $\Delta V_{pp} = 2V\pi$. In both arms of the MZM we apply the same electrical signal, but it depends if the input bit is 1 or 0.

\[
\phi_1(t) = -\phi_2(t) = \phi(t) = \frac{\pi}{V\pi} V_{0,1}(t); \quad V_{0,1}(t) = \frac{V\pi}{2} + V\pi d_{0,1}(t)
\] (3.18)

The output signal is

\[
P_{out}(t) = P_{in}(t) \cos^2(\phi(t)) = P_{in}(t) \cos^2\left(\frac{\pi}{V\pi} \left(\frac{V\pi}{2} + V\pi d_{0,1}(t)\right)\right).
\] (3.19)

for bit 0, $d_0=0$

\[
P_{out}(t) = P_{in}(t) \cos^2\left(\frac{\pi}{2}\right) = 0;
\] (3.20)

for bit 1, $d_0 = 1/2$

\[
P_{out}(t) = P_{in}(t) \cos^2\left(\frac{\pi}{2} + \frac{\pi}{2}\right) = P_{in}(t) \cos^2(\pi) = P_{in}(t);
\] (3.21)

3.3.4 OTDM Multiplexer

Once we have the signal at the output of the OOK Transmitter, we need to implement the OTDM-Multiplexer (OTDM – MUX). The input signal at the OTDM – MUX is a PRBS sequence and we would like that the output signal will be a PRBS sequence too.
3.3. Transmitter Model

In an OTDM system, the multiplexing in the transmitter should be carried out as shown in figure 2.4, where once we have the train of pulses with a bit rate SymbolRate, the signal is split into a number of channels $N$, where in every channel is applied a different delay, and then with the MUX the signal is interleaved to obtain the OTDM signal with an output bit rate of $N \cdot$ SymbolRate. This is a parallel-structure multiplexer.

The structure of the MUX in the laboratory is implemented using a serial-structure multiplexer (see figure 3.9). In that case, once we have the train of pulses, with bit rate SymbolRate, the signal is split into two, and in one of the arms is applied a certain delay, then the both signals are interleaved and the output signal has a bit rate of $2 \cdot$ SymbolRate.

To maintain the PRBS pattern in the output of the MUX, the relative delay in every stage should be precisely controlled following the next relation [26]:

$$d_m = (2^n - 1)(p - 1) + \frac{2^n - 1}{2^m}; \quad m, p = 1, 2, 3... \text{bits} \quad (3.22)$$

where:

- $d_m$ is the delayed bits in $m$-th stage.
- $n$ is the $n$-bit register for reproducing the $2^n-1$ PRBS.
- $p$ is the number of periods considering the $2^n-1$ bits per period.

![Figure 3.9: Scheme of the serial multiplexer implemented.](image)

In our thesis we use $p=1$ and $n=7$, and the number of stages will depend on the number of channels that we want to multiplex to obtain a bit rate of $N \cdot$ SymbolRate at the output.

For the implementation with Matlab® software we use the Matlab® function circshift that allows us to shift the data signal and to simulate the delay effect. The code is:

```matlab
number_stages = log2(N);
Eout_array = zeros(N,length(TimeArray));
Eout_array_aux = zeros(length(TimeArray),1);
Eout_array_total = zeros(1,length(TimeArray));
Eout_array_total(1,:) = Eout;
```
3.4 Optical channel noise

for i=1:number_stages
    delay = length(info)/2^(i);
    Eout_array_aux(:,1) = circshift(Eout_array_total(1,:),
        samples_period*delay);
    Eout_array(i,:) = Eout_array_aux(:,1)';
    Eout_array_total(1,:) = (1/sqrt(2))*(Eout_array_total(1,:)+
        Eout_array(i,:));
end

3.4 Optical channel noise

Once we have modeled the signal at the output of the OTDM Transmitter, we will send the signal over an optical channel, usually a fiber link that use to introduce losses, and we will use an amplifier to compensate it. Most optical amplifiers amplify light through stimulated emission and are susceptible to amplified spontaneous emission (ASE) noise.

In our thesis, we work in back-to-back configuration (where the receiver structure is situated just after the transmitted without using a fiber) to make the BER measurements, and we add noise to the signal before going to the receiver. Remember the relation OSNR (see section 2.5.2)

\[
OSNR = \frac{< P(t) >}{p \cdot No \cdot Bref} \tag{3.23}
\]

where in our model we work with \( p=1 \) (consider only 1 polarization) and \( Bref = 0.1nm \) (or 12.5GHz).

We model the ASE noise as AWGN over the electric field \( E(t) \), so both real and complex part of \( E(t) \) are noisy. We need to define the value of \( No \) depending on the OSNR we would like to obtain [7], we calculate the standard deviation of the Gaussian noise as

\[
\sigma_N = \sqrt{No \cdot \text{SampleRate}} / 2 \tag{3.24}
\]

because \( No \) is the single-sided power spectral density, and \( \text{SampleRate} \) is the width of our baseband span. We need to generate different instances of noise for the real and imaginary part of the electric field, and that explains the factor 2.

The implemented code is:

\[
Bref = 12.5e+009;
\]

\[
\%Average power of input signal
Pav = sum(abs(sig).^2)/length(sig);
OSNR_lin = 10^(OSNR_dB/10);
\]

\[
\%Noise power
No = Pav/(Bref*OSNR_lin);
\]
Here we are going to describe the blocks of our OTDM-Receiver structure. As we can see in figure 3.10, the OOK Receiver has a simple structure, and the main elements are the photodiode and the decision circuit.

![Diagram of OOK receiver structure](image)

**Figure 3.10:** Scheme of the OOK receiver structure.

### 3.5.1 OTDM Demultiplexer

For the demultiplexing operation we implement an optical gate, and then we select for the following studies only one OTDM channel of the multiplexed data signal\(^1\), although multiple-channel output operation can be achieved by a "serial" or "parallel" configuration of several of these gates.

The operation of demultiplexing is shown in figure 3.11

![Diagram of demultiplexing operation](image)

**Figure 3.11:** Demultiplexing operation.

For the optical gate we use a Gaussian-shape gate, whose time-response is defined as

\[
h(t) = \exp \left( -\frac{1}{2} \frac{(t - t_0)}{T_0} \right)^{2m}
\]  

(3.25)

where \(m\) is the Gaussian-order, \(t_0\) the time response delay to select the appropriate

\(^1\)We assume that all the channels have similar performance to reduce the amount of computational calculation for the analysis of the receiver structure.
3.5. Receiver Model

channel, and \( T_0 = \frac{T_{FWHM}}{1.665} \) [3]. In our case we define a 2-nd order Gaussian-gate, with a \( T_{FWHM} = \frac{T_{sch}}{4} \).

3.5.2 Optical Filter

The operation of filtering in the receiver is the same than the previously explained in section 3.3.1.3. In that case, the filter allows us to reduce the effect of the noise introduced by the channel, and we apply a Gaussian filter of 1-st order in our received model with an optical bandwidth of \( b_o = 2.5 \text{ SymbolRate} \).

3.5.3 Photodiode

We detailed in section 2.8.2 the theory related to the photodiodes, so in that case we just need to employ the relation

\[
I = R \cdot P_{in}
\]  

(3.26)

In our system, we neglect the noise effects of shot and thermal noise, and we just consider the contribution of Gaussian noise.

The parameter of responsivity \( R \) is set to \( R=1 \), because it is only an amplitude term that is not going to affect the BER measurements. The Matlab® code is quite simple:

\[
P_{in} = \text{power}(\text{abs}(E_{in}),2);
I = R*P_{in}.
\]

3.5.4 Electrical filter

The performance of an electrical filter is quite similar to the one of an optical filter, because it is also realized in the frequency domain (see section 3.3.1.3).

Electrical filters are low-pass filters, and the implosional response could be Gaussian, Bessel or Butterworth. The transfer function for the Bessel and Butterworth filters depends on Bessel and Butterworth polinomial functions respectively (can be found in some notes as [10]).

\[
H_{BE} = \frac{BE_n(0)}{BE_n(s)} \quad H_{BW} = \frac{BW_n(0)}{BW_n(s)}
\]  

(3.27)

while the Gaussian transfer function for a low-pass filter is

\[
H_G = \exp\left(-\frac{1}{2}\left(\frac{f}{f_0}\right)^{2m}\right)
\]  

(3.28)

where \( m \) is the order of the filter; and \( f_0 \) is the cut-off frequency (see equation (3.4)).

In our model we implement a 4-th order Bessel filter, with an electrical bandwith of \( b_e = 0.75 \text{ SymbolRate} \).
However, the filter introduces a pulse shift, and we need to compensate it before going to the decision circuit. For that reason, we apply the cross-correlation\(^1\) between the pulse at the input and at the output of the filter, which returns a maximum for a certain \(t_d\) that corresponds with the delay. It is compensated with the circhift function. The code is:

\[
\text{% Compensate the delay of the I_filter.} \\
cr_corr = xcorr(I,I_filter); \\
m = \max(cr_corr); \quad \text{%m: maximum of the crosscorrelation.} \\
pm = \text{find}(cr_corr==m); \quad \text{%pm: position of the maximum->delay.} \\
delay_filter = Nsamples-pm; \\
I_aux = circhift(I_filter',-delay_filter); \\
I_filter_delay = I_aux'; \\
\]

### 3.5.5 Decision

In the decision circuit stage, once we obtain the signal at the output of the electrical filter, we must decide if the received information corresponds to a "1" or a "0". For this purpose we will take an array of samples from the output signal and we will make a decision over two parameters, a sampling time \((t_{\text{sample}})\) and a threshold \((I_{th})\) (see figure 3.12).

The decision rule is quite simple, if the value of the sample is higher that \(I_{th}\), we consider a "1"; otherwise a "0".

![Figure 3.12](image)

**Figure 3.12:** Example of decision with a threshold current \(I_{th}\), and three decision points at \(t_{\text{sample}}\), \(t_{\text{sample}} + T\) and \(t_{\text{sample}} + 2T\).

The code is:

\[
\text{% Decision circuit:} \\
\text{sample = zeros(1,length(I_filter));} \\
\text{received_info = zeros(1,length(I_filter));} \\
\text{for k=1:length(I_filter)} \\
\]

\(^1\)The cross-correlation function is calculated as: \(R_{x,y}(t) = x(t) \ast y^*(t) = \int_{-\infty}^{\infty} x(\tau) y^*(t-\tau)d\tau\)
3.6 BER analysis

```matlab
sample(k) = I_filter_delay(tsample + k*T);
if sample(k) > Ith
    received_info(k) = 1;
else
    received_info(k) = 0;
end
```

3.6 BER analysis

Once we have the "received_info" signal, we compare it with a reference signal (the PRBS sequence transmitted) and we evaluate the number of errors (see figure 3.13), that allow us to measure the BER performance of the system.

![Figure 3.13: Example of reference and received signal with a threshold current Ith, and three decision points at tsample, tsample + T and tsample + 2T, where an error occurs at tsample + 2T.](image)

To estimate the BER in our system, we will apply the Monte-Carlo algorithm \[5\], where the error probability is

\[
BER = \frac{N_e}{N}
\]  

(3.29)

where \(N_e\) is the number of errors occurred; and \(N\) is the total number of bits processed.

In our case we define \(N_e = 100 \cdot m\), where \(m = \log_2 M\) is the number of bits per symbol (in OOK \(M = 2 \Rightarrow m = 1\)), so the total number of errors is \(N_e = 100\).

The problem is that our transmitted signal is a \(2^7 - 1 = 127\) bits PRBS sequence, that is short for the total number of errors that we want to measure. To solve that problem, we will run different simulations\(^1\) (measuring in each simulation the number of errors \(n_e\) and adding the result to a variable total_errors) until we have a total number of errors higher or equal to \(N_e = 100\). Then the total number of bits processed is determined with the total number of simulations \((N_{sim})\) transmitted as \(N = 127 \cdot N_{sim}\), and the BER is evaluated as (3.29).

\(^1\) We maintain the transmitted signal fixed, to reduce the simulation time, and in every simulation we apply a different performance of the noise at the output of the OTDM – MUX.
3.7 Spectrum analysis

As explained in section 2.4.1, the spectrum of a signal is its PSD. The PSD is equivalent to calculate the power of the FT of the signal. So, for a signal $x(t)$, we just apply the FT of the signal and then take the square of its absolute value.

$$S(f) = |X(f)|^2.$$ (3.30)

We represent it in the frequency domain. The code is:

```matlab
\% Spectrum of a signal x(t):  
X = fft(x)/length(x);  
XX = power(abs(fftshift(X)),2);  
\% Represent Spectrum:  
plot(FrequencyArray,10*log10(XX));
```
3.7. Spectrum analysis
In this chapter we will show the results of our study after many simulations realized with our implemented Matlab® model. We will explain the analysis for the generation of ultrashort pulses using frequency comb generation in a MZM. Then we will continue with a stage of pulse compression using nonlinear fiber. Finally, we will study the BER performance of the complete system employing OTDM and OOK modulation.

Let us explain that we will introduce some validation test along the chapter just to verify the implementation of our blocks, explained in the previous chapter, with the theory and with other results from the literature.

4.1 Ultrashort pulses using Frequency Comb Generation in a Mach-Zehnder modulator

We will study the different parameters of the elements that we used for the ultrashort pulse generation using optical frequency comb. Firstly, we will develop the flat spectrum condition in terms of the applied voltage in each arm of the MZM. Secondly, we will study the behaviour of the MZM working at different modulator bias point. Thirdly, we will explain the technique of chirp compensation using SMF. Finally, two possible configurations will be studied for the optical frequency comb in a MZM.

4.1.1 Set-up values

First of all we defined a few values that we used for the analysis of this section. The CW laser output power was $P_0=1$W (30dBm) and the central frequency was 193.12 THz, so we set the reference frequency to the same value to work in base-band configuration (we explained the reasons in section 3.3.1.1) and we did not simulate the effects of phase and intensity noise of the laser. The number of symbols was set to two ($N_{\text{symbols}}=2$) with $\text{SamplesPerSymbol}=4096$, and the $\text{SymbolRate}=10$ Gbaud/s, because the pulses were produced at a frequency of 10 GHz. We set the value of $V_\pi=2.8$V.
4.1. Ultrashort pulses using Frequency Comb Generation in a Mach-Zehnder Modulator Simulations

4.1.2 Optical frequency comb generation using MZM

As we explained in section 2.7.2.6, one of the applications that offer an optical frequency comb is ultrashort pulses. We worked with the MZM operating in the asymmetric mode. In section 2.7.2.6 we found the expression for flat spectrum condition as

\[ \Delta A \pm \Delta \theta = \frac{\pi}{2} \]  

but we needed to express this condition in terms of the applied voltage \( V(t) \) in both arms of the modulator. We defined \( \Delta A \) using (2.70) as

\[ \Delta A = \frac{A_1 - A_2}{2} = \frac{\pi}{2V_\pi} \left( \frac{V_{pp1} - V_{pp2}}{2} \right) = \frac{\pi}{4V_\pi} \Delta V_{pp} \]  

and \( \Delta \theta \) using (2.73) as

\[ \Delta \theta = \frac{\theta_1 - \theta_2}{2} = -\frac{\pi}{2V_\pi} (V_{dc1} - V_{dc2}) = -\frac{\pi}{2V_\pi} \Delta V_{bias} \]  

Finally, we obtained the flat spectrum condition in terms of the differential voltage applied to the arms of the MZM as

\[ \frac{\Delta V_{pp}}{2} \pm \Delta V_{bias} = V_\pi \]  

In our analysis we employed the positive sign of (4.4).

Let us clarify that we concentrated our study of the MZM parameters (\( \Delta V_{bias} \) and \( \Delta V_{pp} \)) selecting \( \Delta V_{bias} \) between \( \frac{V_\pi}{2} \) and \( V_\pi \), and then calculating the appropriate value of \( \Delta V_{pp} \) to accomplish the condition (4.4). The reason for this range of \( \Delta V_{bias} \) was because if the modulator bias was smaller than \( \frac{V_\pi}{2} \), then to fulfill (4.4), the value of \( \Delta V_{pp} \) should be bigger than \( V_\pi \) and the output pulse had two peak values (see figure 4.1(a)), whereas between \( \frac{V_\pi}{2} \) and \( V_\pi \) it had only one peak (see figure 4.1(b)).

![Figure 4.1: Example of pulse at the output of MZM using: a) \( \Delta V_{bias} < \frac{V_\pi}{2} \) and b) \( \Delta V_{bias} > \frac{V_\pi}{2} \).](image)
4.1. Ultrashort pulses using Frequency Comb Generation in a Mach-Zehnder modulator

The values of $\Delta V_{bias}$ and $\Delta V_{pp}$ were always related to $V_{\pi}$, because then a change in the value of $V_{\pi}$ would not affect our pulse analysis. Once we decided the value of $\Delta V_{pp}$, we determined the values of $V_{pp1}$ and $V_{pp2}$ to have a difference between the two in-phase sinusoidal modulating signals of the $MZM$ of approximately 1dB.

As a parameter to decide which configuration was better, we used the $T_{rms}$ value because it takes into account the pulse shape. The $T_{FWHM}$ measurements do not consider the existence of small sidelobes near the central lobe, and they increase the crosstalk with adjacent channels. However, we also measured the $T_{FWHM}$ for the different $\Delta V_{bias}$.

4.1.3 Chirp compensation using SMF

We used a piece of $SMF$ to compensate the frequency chirp ($C$) in the optical comb signal with the following parameters (see table 4.1). Before finding the necessary length of $SMF$, we had to check our model of optical fiber.

<table>
<thead>
<tr>
<th>SMF Parameters</th>
<th>D</th>
<th>17 ps/(nm·km)</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\lambda$</td>
<td></td>
<td>1550 nm</td>
</tr>
</tbody>
</table>

**Table 4.1:** Parameters of $SMF$

4.1.3.1 Measure of the chirp of the signal

The pulse shape was approximately Gaussian, so we could express the frequency chirp of the signal (2.67) as [3]

$$\delta w(t) = 2\pi \delta f(t) = -\frac{\delta \phi(t)}{dt} = \frac{C}{T^2_0} t$$ \hspace{1cm} (4.5)

and the second derivate as

$$-\frac{\delta^2 \phi(t)}{dt^2} = \frac{C}{T^2_0}$$ \hspace{1cm} (4.6)

As we can see in **figure 4.2**, we could approximate the frequency chirp as lineal ($y = mx$) in the area of the peak value of the pulse, and related it to (4.5), then taking the maximum (or minimum) of the 2nd derivate, value $m$, the chirp value was

$$C = m \times T^2_0$$ \hspace{1cm} (4.7)

and we measured the value of $T_{FWHM}$ from the pulse (and applying the relation $T_{FWHM} = 1.665 T_0$ for a Gaussian pulse) we obtained the chirp $C$ of the signal.
4.1. Ultrashort pulses using Frequency Comb Generation in a Mach-Zehnder modulator

Simulations

4.1.3.2 Test of SMF

As we explained in section 2.2.3, a pulse can be broaden or compressed while propagating in a fiber (depending of the initial value of $C$ and the value of $\beta_2$). In that case we tested the evolution of a chirped Gaussian pulse propagating along a SMF with the parameters of table 4.1.

The evolution of the pulse width along the fiber for a chirped Gaussian pulse (see [3]) can be calculated as

$$\frac{T(z)}{T_0} = \sqrt{\left(1 + \frac{C\beta_2 z}{T_0^2}\right)^2 + \left(\frac{\beta_2 z}{T_0^2}\right)^2}$$  \hspace{1cm} (4.8)

where $T_0$ is the half-width at 1/e of the intensity point at the input, and $T(z)$ is the half-width after a fiber length of $z$. In figure 4.3 we show the relationship $\frac{T(z)}{T_0}$ as a function of the propagation distance $\frac{z}{L_D}$, where $L_D = \frac{T_0^2}{|\beta_2|}$ is the dispersion length.

The pulse can be broaden or compressed depending on the value of the product $C\beta_2$. We were interested in $C\beta_2<0$, because the pulse initially decreases and has a minimum at $z_{\min}$ (see figure 4.3), and then starts to increase. It means that at the beginning the induced dispersion compensate the initial chirp until it is completely compensated at $z_{\min}$.

Then we could calculate the necessary length of the fiber to compensate the chirp as

$$L_{SMF} = z_{\min} \ast L_D.$$  \hspace{1cm} (4.9)
4.1. Ultrashort pulses using Frequency Comb Generation in a Mach-Zehnder modulator

The first option that we analyzed for the frequency comb generation was using a MZM and then a piece of SMF (with the parameters of table 4.1) to compensate the chirp, without introducing an OBPF between them (see figure 4.5).
4.1. Ultrashort pulses using Frequency Comb Generation in a Mach-Zehnder modulator

Simulations

![Diagram of CW Laser, MZM, and SMF](image)

**Figure 4.5:** Flat frequency comb using a MZM and SMF.

We analyzed for different values of $\Delta V_{\text{bias}}$ the values of $T_{\text{rms}}$, $T_{\text{FWHM}}$, $C$ and the necessary length of SMF ($L_{\text{SMF}}$) and the results are in figure 4.6. Looking the graphs, we observed that the smaller the value of $\Delta V_{\text{bias}}$, the higher the values of $T_{\text{rms}}$, $T_{\text{FWHM}}$ and chirp of the pulse, but the necessary length of SMF was smaller. This was due to the contribution $C T_0^2$ in the evolution of the pulse width (see equation (4.8)), because the quadratic exponent in the denominator made that for a bigger $T_0$, then the smaller $L_{\text{SMF}}$ was required.

![Graphs of a) $T_{\text{rms}}$, b) $T_{\text{FWHM}}$, c) Chirp, and d) $L_{\text{SMF}}$.](image)

**Figure 4.6:** Values at the output of MZM with several $\Delta V_{\text{bias}}$ for: a) $T_{\text{rms}}$; b) $T_{\text{FWHM}}$; c) Chirp and d) $L_{\text{SMF}}$.

Subsequently, for each one of the values of $\Delta V_{\text{bias}}$, and applying the calculated $L_{\text{SMF}}$, we measured the $T_{\text{rms}}$, $T_{\text{FWHM}}$ and $C$ at the output of the SMF (see figure 4.7). Observing the graphs of figure 4.7, we could conclude that the best option for $\Delta V_{\text{bias}}$ was 0.72 V$_\pi$ considering $T_{\text{rms}}$, although in terms of $T_{\text{FWHM}}$ the best option was 0.5 V$_\pi$ (note that in both cases the chirp at the output was 0).
4.1. Ultrashort pulses using Frequency Comb Generation in a Mach-Zehnder Modulator

Simulations

Figure 4.7: Values of $T_{\text{rms}}$, $T_{\text{FWHM}}$ and Chirp at the output of SMF for several values of $\Delta V_{\text{bias}}$.

If we observed the pulse shape for both options of $\Delta V_{\text{bias}}$ (see figure 4.8), we discovered that there were two big sidelobes limiting the number of channels that we could multiplex in the OTDM system, because they introduced high crosstalk with respect to the adjacent channels. In figure 4.9 is shown the spectrum at the output of SMF, that in both cases had a high BW, then we decided to introduce an OBPF between the MZM and the SMF to reduce the spectral width and the amount of chirp.

Figure 4.8: Pulses at the output of SMF for a) $\Delta V_{\text{bias}} = 0.5V\pi$; b) $\Delta V_{\text{bias}} = 0.72V\pi$. 

Figure 4.7:

- $T_{\text{rms}}$ vs. $V_{\text{bias}}$
- $T_{\text{FWHM}}$ vs. $V_{\text{bias}}$
- Chirp vs. $V_{\text{bias}}$

Values:
- $a) T_{\text{FWHM}} = 2.64 \text{ ps}$
- $b) T_{\text{FWHM}} = 4.64 \text{ ps}$. 

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4.1. Ultrashort pulses using Frequency Comb Generation in a Mach-Zehnder modulator

Simulations

4.1.5 Configuration introducing an OBPF between MZM and SMF

We introduced an OBPF between the MZM and the SMF (see figure 4.10), and we analyzed the effects of this filter in the pulse shape. The OBPF was a first order Gaussian filter.

We studied which one was the most suitable value of $\Delta V_{bias}$ together with the study of the appropriate $BW$ of the filter, using as a parameter of decision the $T_{rms}$ value of the pulse at the output of the filter. We also checked the value of $T_{FWHM}$ for the different combinations of $\Delta V_{bias}$ and $BW$ of the filter.

<table>
<thead>
<tr>
<th>MZM Parameters</th>
<th>$\Delta V_{bias}$ = 0.5$\pi$ V</th>
<th>$\Delta V_{bias}$ = 0.72$\pi$ V</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_\pi$</td>
<td>2.80 V</td>
<td>2.80 V</td>
</tr>
<tr>
<td>$V_{pp1}$</td>
<td>25.76 V (33.20 dBm)</td>
<td>14.42 V (28.16 dBm)</td>
</tr>
<tr>
<td>$V_{pp2}$</td>
<td>22.96 V (32.20 dBm)</td>
<td>12.85 V (27.16 dBm)</td>
</tr>
<tr>
<td>$V_{dc1}$</td>
<td>0.70 V</td>
<td>1.01 V</td>
</tr>
<tr>
<td>$V_{dc2}$</td>
<td>-0.70 V</td>
<td>-1.01 V</td>
</tr>
</tbody>
</table>

Table 4.2: Parameters of MZM.

Figure 4.9: Spectrum of the signal at the output of SMF for a) $\Delta V_{bias} = 0.5V\pi$; b) $\Delta V_{bias} = 0.72V\pi$.

In table 4.2 are shown the parameters employed for the MZM in both cases.
4.1. Ultrashort pulses using Frequency Comb Generation in a Mach-Zehnder modulator

Simulations

**Figure 4.11:** Analysis of $T_{rms}$ and $T_{FWHM}$ for different values of $BW$ of the OBPF and for different values of $\Delta V_{bias}$ between $V_{\pi}/2$ and $V_{\pi}$.

As we can see in figure 4.11(a), introducing the OBPF the best combination using $T_{rms}$ criteria was $\Delta V_{bias}=V_{\pi}/2$ (then $\Delta V_{pp}=V_{\pi}$), and the appropriate $BW$ was set to 55GHz. This combination was also optimum for the value of $T_{FWHM}$ (see figure 4.11(b)).

In table 4.3 are shown the parameters employed for the MZM and the OBPF.

<table>
<thead>
<tr>
<th><strong>MZM Parameters</strong></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_{\pi}$</td>
<td>2.8 V</td>
</tr>
<tr>
<td>$V_{pp1}$</td>
<td>25.76 V (33.20 dBm)</td>
</tr>
<tr>
<td>$V_{pp2}$</td>
<td>22.96 V (32.20 dBm)</td>
</tr>
<tr>
<td>$V_{dc1}$</td>
<td>0.70 V</td>
</tr>
<tr>
<td>$V_{dc2}$</td>
<td>-0.70 V</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th><strong>OBPF Parameters</strong></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Type</td>
<td>Gaussian</td>
</tr>
<tr>
<td>Order</td>
<td>1</td>
</tr>
<tr>
<td>BW</td>
<td>55 GHz</td>
</tr>
</tbody>
</table>

**Table 4.3:** Parameters of MZM and OBPF.

The spectrum of the signal at the output of the MZM and the OBPF is shown in figure 4.12, where we can see the reduction of bandwidth due to the filter.

Once we had decided the optimum value of $\Delta V_{bias}$ and $BW$ of the filter, we studied the evolution of the pulse along the fiber (see equation 4.8) to find the necessary value of $z_{min}$ to compensate the chirp.

Considering a SMF with the parameters of table 4.1, in figure 4.13 is shown the evolution of the $T(z)$ as a function of the length of the fiber. It is possible to see that for the minimum value of $T_{rms}$ the chirp was also reduced to 0, and we also checked the value of $T_{FWHM}$ at the output of the SMF and it had the minimum value.
4.1. Ultrashort pulses using Frequency Comb Generation in a Mach-Zehnder modulator

Simulations

![Graph](image)

Figure 4.12: Spectrum of signal with $\Delta V_{bias}=0.5V_{p}$ at the output of a) MZM b) OBPF.

The appropriate length of the SMF was 885m, which corresponded to a total dispersion value of $D=15 \text{ ps/nm}$. Finally, as a result of the chirp compensation using SMF, a picosecond pulse with a pulse width of 7.8ps was obtained.

![Graph](image)

Figure 4.13: Evolution of 1)$T_{rms}$; 2)$T_{FWHM}$ and 3) Chirp, as a function of the length of the SMF.

To conclude this section, we show in figure 4.14 the pulse at the output of the MZM and at the output of SMF. This pulse was wider with the OBPF than without it, but we had the advantage that the sidelobes dissapeared in the pedestal of the pulse.
4.2 Pulse compression using DF-HNLF

An OTDM system transmission with a bit-rate above 640Gbit/s requires sub-picosecond pulses generated at a repetition rate of 10GHz (or 40GHz). So we needed to add a pulse compression stage in our system to arrive to femtosecond pulses, and for this reason we employed a nonlinear fiber combined with an OBPF and a piece of SMF (see figure 4.15).

![Figure 4.15: Scheme for a pulse compression stage.](image)

4.2.1 Set-up values

There are several types of nonlinear fiber, and for our pulse compression stage we employed a dispersion-flattened high nonlinear fiber (DF − HNLF). The parameters of the DF − HNLF are shown in table 4.4.

<table>
<thead>
<tr>
<th>DF-HNLF Parameters</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>D</strong></td>
</tr>
<tr>
<td><strong>S</strong></td>
</tr>
<tr>
<td><strong>γ</strong></td>
</tr>
<tr>
<td><strong>λ</strong></td>
</tr>
</tbody>
</table>

Table 4.4: Parameters of DF − HNLF.

As an OBPF we used a first order Gaussian filter with a BW of 14 nm, and the length of SMF was the necessary for chirp compensation.
4.2.2 Test of Split-step Fourier Method

We simulated the effects of the $DF - HNLF$ applying the $SSFM$, but we had to test our implemented method previously. In that case we checked dispersion (second order) together with nonlinearities using the error-method step size with a relative goal error $\delta_G=0.001$.

The test was done transmitting a first-order soliton pulse (see section 2.2.4), with a dispersion parameter $D=17$ ps/(nm-km) and $\gamma=2.5$ (W·km)$^{-1}$, and we set the pulse width to $T_0=10$ ps. Setting $N=1$ in equation (2.33), the peak-power of the pulse was $P_0=86.7$ mW. We propagated the pulse over a total length of 1000km, and as seen in figure 4.16, the soliton pulse was preserved along that distance. We could conclude that our implemented model of $SSFM$ was correct.

![Figure 4.16: Evolution of a soliton pulse over a total distance of 1000km.](image)

4.2.3 Analysis of pulse compression stage

We analyzed different lengths for the $DF - HNLF$ together with different values of optical pulse power ($P(t) = |E(t)|^2$) at the output of the amplifier, and we modified this values as $P'(t) = G\cdot P(t)$.

We made the analysis as a function of the average power of the pulse ($P_{av_{in}}$) at the input of the $DF - HNLF$. Then, a possible change of the $CW$ laser output power ($P_0$) that modify $P_{av_{in}}$, could be compensated modifying the value of $G$. Subsequently, the analysis of the effects in the nonlinear fiber will not change.

We could not use an infinite gain for the amplifier. As a parameter to limit the range of values of $G$ we imposed that $P_{av_{in}}$ was between 1W and 2W. For each one of the values of $P_{av_{in}}$, we studied several values of $L_{DF-HNLF}$. The analysis of the pulse parameters ($T_{rms}$ and $T_{FWHM}$) were done after the $OBPF$ and the $SMF$. 
4.2. Pulse compression using DF-HNLF

Our goal was to obtain a pulse with a $T_{FWHM}$ around 350fs. The higher the $P_{av\_in}$, the greater the effect of the nonlinearities of the fiber. Therefore, if we employed a fiber with the same length, the pulse with the higher $P_{av\_in}$ would be more compressed. In figure 4.17 we checked that to obtain a $T_{FWHM}=342$fs pulse (with a time resolution of 24.40fs), the higher the value of $P_{av\_in}$, the lower the total length of $DF-HNLF$ was necessary to obtain the pulse.

![Figure 4.17: $T_{FWHM}$ for different $P_{av\_in}$ and $L_{DF-HNLF}$](image)

Then we made the analysis of $T_{rms}$ (see figure 4.18). We observed that a high value of $P_{av\_in}$ corresponds to a high value of $T_{rms}$. We selected a pulse with $P_{av\_in}=1.12$W, because as seen in figure 4.18, we obtained a pulse with approximately the same $T_{rms}$ than with the minimum $P_{av\_in}=1.00$W, but we needed around 50m less of fiber to acquire it.

![Figure 4.18: $T_{rms}$ for different $P_{av\_in}$ and $L_{DF-HNLF}$](image)
4.2. Pulse compression using DF-HNLF

If we observed the pulse shape for $P_{\text{av}_{\text{in}}}=1.12\,\text{W}$, which had a $T_{\text{FWHM}}=342\,\text{fs}$ (see figure 4.19(a)), we could visualize that a pedestal existed and it seemed that far from the center the pulse tended to 0, although if we plotted the pulse in logarithmic scale (see figure 4.19(b)), we could see that the pulse was not completely 0 and it made that when we increased the $P_{\text{av}_{\text{in}}}$, due to the increment at the extremes, the $T_{\text{rms}}$ was also higher.

![Figure 4.19: Generated pulse with $P_{\text{av}_{\text{in}}}=1.12\,\text{W}$ and $T_{\text{FWHM}}=342\,\text{fs}$ represented in a) lineal scale b) logarithmic scale.](image)

In figure 4.20 is shown the final pulse (zoom of the previous figure 4.19(a)), with a $T_{\text{FWHM}}=342\,\text{fs}$.

![Figure 4.20: Zoom of the pulse with $P_{\text{av}_{\text{in}}}=1.12\,\text{W}$ and $T_{\text{FWHM}}=342\,\text{fs}$.](image)

For each value of $P_{\text{av}_{\text{in}}}$ we could obtain a pulse of 342fs with a different pedestal, and we analyzed the $E_{\text{crosstalk}}$ (measured as the percentage over the total energy of the pulse $E_{\text{total}}$) and $P_{\text{crosstalk}}$ of the generated pulses\(^1\), considering that we would multiplex $N=32$ and $N=64$ channels in the OTDM-MUX. We observed in figure 4.21 that the higher the $P_{\text{av}_{\text{in}}}$, the lower the value of $E_{\text{crosstalk}}$ and the higher the $P_{\text{crosstalk}}$ was.

\(^1\)We only measure the $E_{\text{crosstalk}}$ and $P_{\text{crosstalk}}$ in one of the sides of the pulse.
4.3. System performance with OOK modulation and Direct Detection

In this section we would like to evaluate how the crosstalk measured previously affects the OTDM system. Crosstalk produces power penalty (PP) in sensitivity, and we would like to relate the crosstalk with this penalty.

4.3.1 Test BER measurement

We previously tested the implementation of SMF and the SSFM method, and we checked our model of decision circuit to measure the BER performance of our system. We compared the results to ones found in [8].

We transmitted a square signal with a NRZ–OOK modulation and we applied ASE noise to the signal (as explained in section 3.4 to only 1 polarization $p=1$) and the OSNR and SNR defined as in section 2.5.

The receiver model consists in a matched-optical filter, which optimizes the SNR of the system, and a photodiode (no thermal noise nor shot noise), without an electrical filter at the output. The system scheme is shown in figure 4.22.

![Figure 4.22: Scheme of system with square NRZ–OOK modulation, Matched optical filter and photodiode.](image)

**Figure 4.22:** Scheme of system with square NRZ–OOK modulation, Matched optical filter and photodiode.
4.3. System performance with OOK modulation and Direct Detection

The correspondence between our numerical results and the theoretical ones were almost perfect (see figure 4.23).

![Figure 4.23: BER – SNR curve for NRZ – OOK modulation with our model and reference, with Matched optical filter and photodiode.](image)

4.3.2 Set-up values

For the BER performance of the system we employed \textit{SamplesPerSymbol} = 4096, a \textit{2}^{7-1}=127 PRBS sequence and we used a \textit{NRZ – OOK} modulation. We added \textit{ASE} noise only in one polarization. In the receiver structure, we employed the \textit{OTDM – DEMUX} explained in section 3.5.1. A first order Gaussian optical bandpass filter with \textit{b}_o=2.5\cdot\textit{SymbolRate} was used to suppress out of band \textit{ASE} noise. A 4th order Bessel low pass filter with cutoff frequency equal to 0.75 times the bit rate was then used after photodetection in an ideal photodetector (no thermal noise nor shot noise) with responsivity equal to 1 A/W. (Scheme of the receiver structure in figure 4.24).

![Figure 4.24: Scheme of the receiver structure.](image)

4.3.3 OSNR-BER measurements

We measured the crosstalk \textit{PP} produced by channel multiplexing in an \textit{OTDM} system with a back-to-back configuration. We also measured the \textit{OSNR} corresponding to a \textit{BER} of 10^{-3}. The \textit{sampling time} (\textit{t}$_{\text{sample}}$) was chosen at the centre of the bit slot and the \textit{threshold} (\textit{I}$_{th}$) was optimized over 100 threshold values.

Firstly, we transmitted the generated signal without multiplexing (\textit{R}$_b$=10Gb/s) and we measured the necessary \textit{OSNR} for \textit{BER}=10^{-3}. Then, we multiplexed \textit{N}=32 and \textit{N}=64 channels (using the method explained in section 3.3.4), and we measured the \textit{PP} (see figure 4.25) result of the difference between both sensitivities. For each pulse generated in section 4.2.3 we had a \textit{P}_\text{crosstalk}, a \textit{E}_\text{crosstalk} and a \textit{PP} too.
4.3. System performance with OOK modulation and Direct Detection

![Figure 4.25: Example of $PP$ in a $BER - OSNR$ curve.](image)

We should expect that for a lower $\%$ of $E_{crosstalk}/E_{total}$ (and for a higher $P_{crosstalk}$), the lower the $PP$ occurred. But if we observe figure 4.26(a) and (c) for $E_{crosstalk}$; and figure 4.26(b) and (d) for $P_{crosstalk}$, the obtained points were dispersed in the evaluated range.

![Figure 4.26: Relation between the $PP$ for 1) $N=32$ a) $\%E_{crosstalk}/E_{total}$ - and b) $P_{crosstalk}$; and 2) $N=64$ c) $\%E_{crosstalk}/E_{total}$ - and d) $P_{crosstalk}$.](image)
4.4 Conclusion

A ps-pulse has been developed with the frequency comb in a MZM and using a standard piece of SMF. We can conclude that it is better to introduce an OBPF between the MZM and SMF, because the pulses at the output have no sidelobes, which are detrimental in terms of crosstalk for the OTDM transmission.

Additional pulse compression stage has been introduced to reduce from picosecond to femtosecond pulses, that allows us to work with bit rates of 320 Gbit/s or 640 Gbit/s (multiplexing $N=32$ and $N=64$ channels respectively) in an OTDM system. We check that to obtain the same pulse width, the higher the value of the optical pulse power is, the lower the necessary length of the nonlinear fiber.

In conclusion, there is no correlation between the terms of $P_{crosstalk}$ or $E_{crosstalk}$ and the BER sensitivity. Anyway, if we have to use one of them, we would prefer the $P_{crosstalk}$ because it is more easily measured from the pulse.
5

Laboratory experiments

All the previous results were realized using the Matlab® software to simulate the behaviour of the different blocks conforming the system. Finally, we had the opportunity to characterize the pulse source generator using flat frequency comb in a MZM in the Fotonik’s laboratory. This chapter is divided into two parts. Firstly, we will give a description of the studied system, and later we will offer the obtained results.

5.1 System description

Let us begin giving a brief introduction of the system analyzed. We modulated the signal emitted by a CW laser in an external MZM, which was driven by two 10GHz sinusoidal electrical signals (setting both signals in phase using a delay line) and then we realized chip compensation using a piece of SMF. In figure 5.1 is shown the experimental system.

![Figure 5.1: Experimental setup for frequency comb pulse generation using MZM.](image)

Notice that in the experimental system we did not use an OBPF between the MZM and the piece of SMF, because the spectrum of the signal at the output of the MZM was not too much wide (see section 5.2).
5.1. System description

Laboratory experiments

The pulse width was measured with an oscilloscope and the spectrum of the signal with an optical spectrum analyzer (OSA).

5.1.1 CW laser

The laser employed in the experiment was a CW laser, working at $\lambda=1551.28$nm, and having a light-current curve (measured) which is shown in figure 5.2. As we can see, the laser had a threshold current of $I_{th}=25$mA. The measurements were made with an output power of 10mW ($I=70$mA).

![Figure 5.2: Light-current curve of CW laser.](image)

5.1.2 MZM

As a MZM we used a LiNbO$_3$ modulator, which is one of the most frequently used in OTDM experiments. It was used to modulate the generate pulse train.

First of all, we had to know the half-wave voltage $V_\pi$ because we needed it to adjust the rest of MZM parameters to have a flat frequency comb signal. We measured the relation $P_{out}/P_{in}$ as a function of the voltage (see figure 5.3) and the measured value of $V_\pi=4$V.

![Figure 5.3: Transfer function of MZM.](image)
5.2. Results

5.1.3 Optical coupler and Delay line

We generated a 10 GHz sinusoidal electrical signal to drive the two arms of the modulator. So we needed to use an optical coupler to split the signal in two branches, and in one of them we introduced a delay line that we adjusted to have both signals in phase (see figure 5.4).

![Optical Coupler and Delay Line Diagram]

**Figure 5.4:** Setup of driving signals for MZM.

The optical coupler had two inputs and two outputs. The optical signal in one of the inputs was divided and, as a consequence, the half of the power (ideally) went into each one of the outputs; and the same occurred with the other input. In our case, we only used one of the inputs to divide the 10 GHz signal that we needed for both arms of the modulator.

5.2 Results

First of all, we measured the peak-to-peak voltage at the output of both amplifiers, obtaining

\[ V_{pp1} = 7.8 \text{ V} \]
\[ V_{pp2} = 11 \text{ V} \]

We selected \( \Delta V_{bias} \) to fulfill the condition for the frequency comb generation

\[ \frac{\Delta V_{pp}}{2} + \Delta V_{bias} = V_{\pi} \]  \hspace{1cm} (5.1)

In that case \( \Delta V_{pp} = -3.20 \text{V} \) (remember \( V_{\pi} = 4 \text{V} \)), so the theoretical \( \Delta V_{bias} \) should be 5.60V. Setting this value at the power supply, we visualized the output of the MZM in the Oscilloscope (see figure 5.5), and we also introduced the parameters of \( V_{\pi}, V_{pp1}, V_{pp2} \) and \( \Delta V_{bias} \) in our Matlab® model to check the results.

As we can see in figure 5.5, both signals were quite similar, and in table 5.1 is shown the \( T_{FWHM} \) value at the output of the MZM without SMF.

For the rest of measurements we did not use the oscilloscope previously mentioned because it has a time resolution of 8ps, which was too much for the expected pulses; Instead, we used an optical sampling oscilloscope(OSO), that has a time resolution of 1ps, which was much better for our pulses.

Firstly, we measured again the pulse at the output of the MZM without introducing SMF, and we obtained a pulse that was a little bit wider than with the previous oscil-
5.2. Results

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Measured</th>
<th>Simulated</th>
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<tbody>
<tr>
<td>FWHM</td>
<td>45.5 ps</td>
<td>44.4 ps</td>
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Table 5.1: Obtained $T_{FWHM}$ results without SMF.

Figure 5.5: Output signal of the MZM using a) Matlab® implemented model b) Oscilloscope

loscope, with a $T_{FWHM} = 51.4 \text{ps}$, and with some ripples on the top. We shifted the bias to $\Delta V_{bias} = 4.50 \text{V}$, where the pulse in OSO was optimum with a $T_{FWHM} = 49.6 \text{ps}$ (see figure 5.6a). We did not know why appeared this difference between the oscilloscope and OSO for the same setup, but we made the rest of the experiments with the OSO because it was more precise.

The spectrum of the signal at the output of the MZM is shown in figure 5.6b, and as we can see, the spectrum was not too wide, so we decided not to use an optical filter and we connected at the output of MZM a piece of SMF to compensate the chirp.

Figure 5.6: Signal at the output of the MZM without SMF using a) OSO b) OSA.

With our Matlab® implemented model we measured the chirp of the signal and the necessary length of the SMF to compensate it, obtaining a $L_{SMF} = 3.2 \text{ km}$ (using a fiber with $D = 17 \text{ ps/(nm-km)}$) and a total dispersion of 54.40 ps/nm.
Then we observed the output of the pulse connecting different pieces of SMF until a total length of 5000m. The results of the OSO and the spectrum analyzer are shown in figures 5.8 and 5.9.

At the same time we measured with the OSO the pulse width for every configuration, and the results are shown in table 5.2. As we can see in the table and in the graphs, the higher the length of SMF, the lower the pulse width and the higher the peak power of the pulse was until a length of 3.5km. It was approximately the value of the necessary length of SMF to compensate the chirp that we determined with our Matlab® model. Later the pulse width started to broaden and the peak value decreased again.

<table>
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<tr>
<th>L_SMF</th>
<th>Measured</th>
<th>Model</th>
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<tbody>
<tr>
<td>500m</td>
<td>39.4 ps</td>
<td>36.3 ps</td>
</tr>
<tr>
<td>1000m</td>
<td>27.8 ps</td>
<td>28.1 ps</td>
</tr>
<tr>
<td>1500m</td>
<td>21.4 ps</td>
<td>20.1 ps</td>
</tr>
<tr>
<td>2000m</td>
<td>14.8 ps</td>
<td>13.6 ps</td>
</tr>
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<td>2500m</td>
<td>10.3 ps</td>
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<td>3500m</td>
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<td>4000m</td>
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<td>4500m</td>
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</tr>
<tr>
<td>5000m</td>
<td>8.84 ps</td>
<td>9.75 ps</td>
</tr>
</tbody>
</table>

Table 5.2: Results of $T_{FWHM}$ for different SMF measured and simulated with Matlab® model.

Finally, in figure 5.7 we can see the evolution of the pulse width for different values of SMF measured in the laboratory and the evolution with Matlab® model too.
5.2. Results

Figure 5.8: Signal output of SMF for different lengths a) OSO and b) OSA.

- a1) OSO SMF = 500m
- b1) OSA SMF = 500m
- a2) OSO SMF = 1000m
- b2) OSA SMF = 1000m
- a3) OSO SMF = 1500m
- b3) OSA SMF = 1500m
- a4) OSO SMF = 2000m
- b4) OSA SMF = 2000m
- a5) OSO SMF = 2500m
- b5) OSA SMF = 2500m
5.2. Results

Figure 5.9: Signal output of SMF for different lengths a) OSO and b) OSA
Let us conclude explaining that we tried to measure the phase of the signal to visualize the chirp using the Frequency-Resolved Optical Gating (FROG)[23] method in the laboratory, but we had problems because the pulse was too wide to make the analysis. Finally, we could not take the measurements, despite using the pulse with the minimum $T_{FWHM}$ (8.2 ps).

5.3 Conclusion

We could not arrive to a picosecond pulse train shorter than 8.2 ps due to the limit of peak-to-peak voltage induced in both arms of the $MZM$. It limits the amount of chirp generated that we used afterwards for pulse compression employing chirp compensation in a piece of $SMF$.

Therefore, we could conclude that observing the different figures, the obtained results in the laboratory were quite similar with the ones obtained with our implemented model, which gave reliability to it, although we just tested the part of frequency comb generation using the $MZM$.

Although most of the analysis for the generation of a ps-pulse train using a $MZM$ were completed, it would be also important to test the pulse compression stage and BER measurements. Unfortunately, due to a time constraint, they could not be performed.
Chapter 6

Conclusion

The main objective of this thesis was the implementation of pulses sources using frequency comb generation in a Mach-Zehnder modulator.

We have developed a numerical model using Matlab® software for a complete optical communication system, not only for the frequency comb generated pulses. The implemented model has been validated in chapter 4 introducing some tests that we compared with results from the literature.

We investigated the realization of ultrashort pulses using two possible combinations. Firstly, we designed a short train of pulses using a frequency comb in the MZM without introducing a filter between the modulator and the SMF. We obtained picosecond pulses of around 3 ps, but they had sidelobes which were detrimental for the use of these pulses in an OTDM system. However, introducing an OBPF, whose parameters had been optimized together with the MZM, we were able to obtain wider pulses (around 8ps) but without sidelobes.

Secondly, it was necessary to add a stage for pulse compression to use the generated pulses with the frequency comb in a high speed optical communication system using OTDM. The method used in this thesis enabled us to get pulses of 350 fs, although far from the main lobe the pulse was not completely annulled. So that a future analysis with a different compression stage would be necessary for the optimization of the fs-pulse.

Thirdly, an analysis of the pulse using two different pulse parameters as $P_{crosstalk}$ and $E_{crosstalk}$ was also carried out. Moreover, we analyzed the system performance and we tried to find a correlation between OSNR and the previous parameters, but we did not find a good relationship.

Finally, we had the chance to go to the laboratory and make some experimental analysis of the pulses using frequency comb generation. We were limited by the peak-to-peak voltage that we were able to induce in both arms of the MZM. The measured values were compared simulating the conditions in our implemented model, and we obtained good similarities that gave reliability to our results.
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