Towards multi-terabit per second single wavelength links using multilevel modulation

Master Thesis

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Abstract

Orthogonal Time Division Multiplexing (OTDM) and advanced modulation formats represent the future in development of high speeds communications. In this text the feasibility of OTDM systems combining multilevel modulation is studied. The report is focused in Differential Quadrature Shift Keying (DQPSK) modulation. Computer simulations study the performance of an OTDM systems using DQPSK modulation at 1.28 Tb/s and a 640Gb/s OTDM system combined with DPSK modulation. Two solutions for channel demultiplexing are shown: Nonlinear Optical Mirror (NOLM) and Four Wave Mixing (FWM). These techniques need an optical control signal. Optimums values for set the signals and for the bandwidth of the filter are evaluated. Furthermore numerical simulations are performed to evaluate the transmission performances of a 1.28Tb/s OTDM-DQPSK system and a transmission through 561.6km is demonstrated.
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Chapter 1

Introduction

Today, since we are living in the information age, the demand for bandwidth is rapidly increasing. In order to solve this capacity of the communication networks has to increase. Optical techniques can cope with these high capacities due to their high bandwidth [1]. For example, the wavelength range from 1.3 to 1.6 µm which can be transmit in an optical fiber, correspond to a bandwidth larger than 40 THz, which is a larger magnitude than of any electrical field. However, a typical channel capacity for long haul transmission is limited nowadays to 10 Gb/s even 40 Gb/s [1], because of dispersion and nonlinear effects and for electronic limitations. More advanced systems can increase these transmission capacities of one single fiber by transmitting over this single fiber multiples channels. The fiber bandwidth is divided in lower data rate channels. This can be done in two different ways, using Wavelength Division Multiplexing (WDM) or Optical Time Division Multiplexing (OTDM). WDM systems increase the capacity by using multiples different wavelength channels over one fiber. Transmission bit rates over 25 Tb/s are achieved in one single fiber using WDM [2]. The other approach is OTDM. This technique combines the lower data rate channels in the time domain. Electrical data streams are modulated into short optical pulses at lower bit rates which are delay and combined to create a single optical data signal at higher bit rate. WDM systems at high bit rates over Tb/s entail a large number of transmitters and receivers. It leads to complexity in the system and physical and cost limitations [3]. It is preferred a fewer channels with a higher bit rate, rather than a large number of channels with a lower bit rate. That means that WDM systems are limited by electronics speed because WDM single channel capacity is limited by electronic data rates. Thus, the alternative to achieve a higher data speed per wavelength using OTDM technology increase the attention for research work. This technique requires new devices based in all-optical multiplexing and demultiplexing techniques such as high speed switches, optical delays or pulses compression techniques.

On the other hand, advanced modulation formats also attract much attention recently for high bit rate fiber transmission. Advance modulations can be used to improve the transmission performance and achieve high spectral efficiency. These modulations have been investigated in the early 90’s but then for a long time the effort
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researches were focus in the basic OOK modulation because, due the immature of semiconductor laser sources, optical phase was not stable enough to be modulated[5]. However in the recent days due to easier implementation of a stable receiver advance modulations is easier to implement at high bit rates and the research attention in these modulations reappears [6]. This master thesis focuses the attention in DPSK and DQPSK modulations. The use of Differential Phase Keying (DPSK) modulation combined with a balanced receiver offer a 3dB improvement in sensitivity compared with the common On-Off Keying (OOK) modulation [7]. This can be used to extend the transmission distance or increase the bit rate. Also DPSK has been shown to be quite robust to fiber nonlineaties [8-9].One step more to increase the bit rate is use a multilevel modulation such Differential Quadrature Phase Shift Keying (DQPSK). DQPSK results not only in an increase of the spectral efficiency as well in a improvement of polarization mode dispersion tolerance [10].

Thus, by using OTDM technique with advanced modulation good results have been achieved for high bit rates. Bit rates of 640Gb/s are obtained combining DPSK with OTDM [11-12]. This Master Thesis project is focused on study the feasibility of combining OTDM with the multilevel modulation DQPSK. Numerical simulations an OTDM-DQPSK system is carried out. These numerical simulations are used to understand the main issues involves in the system performance.

In chapter 2 a depth view of OTDM technique is presented. Section 2.1 deals with the principle of an OTDM system. In section 2.2. the main requirements and characteristics of an OTDM multiplexer are given. In 2.3 various demultiplexing schemes are presented. The attention is given to fiber based demultiplexing techniques and Non-Linear Optical Loop Mirror (NOLM) and Four Wave Mixing (FWM) are depth explained. In chapter 3 an overview of basic aspects for digital modulation are presented. Then the chapter focuses on advance modulations. In section 3.2.1 the main theory of DPSK modulation is given and in section 3.2.2 DQPSK modulation is presented. In chapter 4 results from computer simulations are reported. The tool used for the simulations is VPItransmissionMaker™/ VPIcomponentMaker™. After a previous introduction of some simulation issues, a performance of DPSK and DQPSK 10Gb/s systems are carry out in order to study the performance of the modulations. In sections 4.4 and 4.5 two channel demultiplexing solutions for OTDM system has been analyze. In section 4.4 a 640 Gb/s DPSK-OTDM system and a 1.28Tb/s DQPSK-OTDM system using a NOLM as demultiplexer are described and simulated. NOLM issues such frequency allocation and filter bandwidth are optimized in order to obtain less impact in phase modulation. Section 4.5 illustrates the results for transmission simulations of a 1.28 Tb/s DQPSK-OTDM signal. In section 4 the results of 1.28 Tb/s DQPSK-OTDM system using a NOLM are reported. This section is focus on the optimization of the control and data signals involved in FWM demultiplexing technique.

Chapter 5 concludes this thesis and gives an overview of the results obtained.
Chapter 2

Fundamental in OTDM system

Over the last years most of the research efforts in optical systems are focused on increasing the bit rate transmission. There are two ways on which bit rate increase can be done. The first consist on transmitting several channels spaced apart in the frequency domain. In the optical domain this approach is called wavelength division multiplexing (WDM). The second consists in launching the signal over the same wavelength but channel interleaving is done in time domain. It is called Optical Time Division Multiplexing (OTDM). Research work is giving increasing attention to the second technique and promising results have been achieved in the resent years. In this Master thesis the attention is focused in this multiplexing technique.

In section 2.1 and 2.2 is given introduction to OTDM system and a general requirements and performing for channel multiplexing. Then, section 2.3 is focused in channel demultiplexing. First a quickly overview of one electro optic demultiplexing technique (EAM) is carried out in section 2.3.1. Then, two different performances for an all-optical demultiplexing are presented in section 2.3.3: Non linear optical-loop mirror (NOLM) and four wave mixing technique (FWM).

2.1 OTDM principle

OTDM system consists on transmitting several optical channels over the same wavelength. N channels at bit rate $b_s$ are multiplexed in time generating an OTDM signal whose bit rate $B$ is equal to the single channel bit rate $b_s$ multiplied by the number of channels $N$, $B=NB_s$.

Figure 2.1 shows a general implementation of an OTDM multiplexer and demultiplexer. An OTDM signals is created interleaving in a single time slot $(1/b_s)$, $N$ channels.
Figure 2.1. General structure of an OTDM system: transmitter and receiver.

As it is seen in figure 2.1, an optical source is needed to generate a train of pulses without information at single bit rate \( b \). This stream of pulses is split in different branches. Every channel uses one branch to encode their information. Each channel can use an independent modulator. Before recombine the channels, each of them is delayed by an appropriated time delay. Since the width of the optical pulses is less than time slot \( 1/b \), all channels have to be allocated in the time intervals between pulses using appropriates delays. In the receiver to recover the single channel a clock recovery is needed. This devise obtained a signal at the bit rate of the channel. This signal is needed to synchronize the devise which acts as a gate to select the desirable channel. Demultiplexing channel is an important point on OTDM systems. To select the channel and removed the contribution of the neighbor channels and accurate temporal open gate is needed. Different devises for channel demultiplexing are widely explained in the following section 2.3.

### 2.2 Channel multiplexing

Channel multiplexing is needed to interleave the channels in time and generate the OTDM signal. Several devises are studied for this purpose. However, the most common technique for channel delaying is to use single model fiber which gives the desirable delay controlling the fiber length.

In the order to avoid overlapping of the channels the width of the optical pulses has to be less than time slot \( 1/B \), where \( B \) is the bit rate of the OTDM signal. So, the most important requirement for channel multiplexing is narrow pulses. The power from neighboring pulses can result in power crosstalk that can diminish the data channel. So in order to avoid crosstalk, pulses have to fulfill some requirement concerning full width at half maximum (FWHM) and pulse tail extension ratio. PTER
is the parameter that characterizes the power in the pulse pedestal and FWHM is given by the width between points on the pulse at which the power reaches half its maximum value. FWHM of pulses should be <40% of the time slot 1/B [13]. So, to achieve high bit rates higher than 320Gb/s a pulses with FWHM<0.6ps are needed. Several challenges for source pulse and compression techniques are achieved. Laser obtains in the recent years the improvement and the stability necessary to reach pulses with near 1ps [14-15]. Since this is not enough to reach some high bit rates a pulses compression technique is needed. The best compression techniques are based on fibers [16-17].

2.3 Channel demultiplexing

Channel demultiplexing is needed for selection of a single channel from the total OTDM high bit rate signal. It acts as a gate that selects the desirable single channel. The switching window is the main characteristic of this gate. To achieve error free in the system this switching window has to fulfill some temporal requirements and extension ratio requirements to be capable cross-talk suppression of the neighbor channels and keep the desirable channel without distortion.

Several demultiplexers have already been demonstrated for OTDM signal. Because of electronics limits, demultipled has to be done in electro-optical or all-optical domain. All channel demultiplexing techniques have in common the necessity of a control signal. This control signal consists in a periodic signal at the channel bit rate. In the case of electro optic modulators an electric signal is needed as control signal. However an optical signal is used in all-optical demultiplexing techniques.

2.3.1 Electro optical technique

Electro optical techniques use an electric sinus as control signal and the phase of this signal has to be adapted to choose the desirable channel. Electro optical devices are based on semiconductor devices. One common choice is electro absorption modulators (EAM). EAM is a semiconductor device which is usually used for intensity modulation. Its principle of operation is based on Franz-Keldysh effect. Franz-Keldysh effect consists on a change in the optical absorption of a direct semiconductor in the presence of an electrical field. These modulators are made as a single waveguide which consist in a doped p-i-n structure [18-20]. The optical loss change due the applied electric voltage produces the optical modulation. By increasing the electrical field the probability of the absorption of the semiconductor is increased and better extension ratio, defined as the difference of power level associated to a 0’s bits and 1’s bits, is obtained for optical modulation [19]

EAM can be used like optical gates using the appropriate sinusoidal driving voltage. Several designs for EAM as OTDM demultiplexer are possible. The basic
design is a single stage with one EAM modulator driven with an electrical sinusoid at the bit rate of the desirable OTDM channel. [20-21] A second structure is based in a fiber couple pair of separated EAM. This choice decreases the temporal width and increase the extension ratio of the switching window. Using this design a demultiplexing of a 10 Gb/s signal in a 160 Gb/s aggregated signal is achieve [22-23].

Other common electro-optical techniques is use Mach Zehnder Modulators (MZM)[1]. The structure of this gate consist on several MZM allocated in serie [24]. In [25] a 10.5 Gb/s channel is demultiplexed from a 336 Gb/s OTDM signal. Although EAM has a easier configuration, a switch based on MZM has a suitable switching window characteristics: narrower and rectangular. [26]

### 2.3.2 All-optical technique

The majority of the all-optical switching techniques are based on nonlinearities in fibers or in semiconductor devises. This technique also needs a control signal but in this case it consists in an optical pulse train. The two techniques presented are based on High Nonlinear Fibers (HNLF).

#### Nonlinear Optical Loop Mirror

Nonlinear Optical-loop Mirror (NOLM) is one of the most popular and promising method used in several OTDM studies and experiments. Demultiplexing 10 Gb/s data channels of 640 Gb/s OTDM signals have been achieved using this technique [11].

NOLM technique uses nonlinearities optical fibers. These nonlinearities consist on changes in the refractive index of the fiber depending on the light intensity. This dependence is called Optical Kerr effect and is governed by:

\[
 n_j' = n_j + n_2 P / A_{\text{eff}}, \quad j = 1,2
\]  

(2.1)

Where \( n_1 \), \( n_2 \) refer to the refractive index of the core and cladding of the optical fiber respectively, \( P \) is the optical power, \( n_2 \) is the nonlinear index coefficient and it is measured in \( \text{m}^2/\text{W} \). and \( A_{\text{eff}} \) is the effective area. \( A_{\text{eff}} \) is defined as the functionally equivalent area from which the core of the fiber absorbs the incident light and is defined as:

\[
 A_{\text{eff}} = \pi w^2
\]  

(2.2)

Where \( w \) is the field radius and is called spot size. The nonlinear effects are maximized in HNLF. So, it is possible to obtain nonlinear effects with short fiber lengths and lower power levels. The two main phenomenon arise from this nonlinearities are: Self Phase Modulation (SPM) and Cross Phase modulation (XPM). The first one leads a phase shift in the optical field induced by itself. This phase shift is:
\[ \phi_{\text{SPM}} = \gamma P_{\text{in}} L_{\text{eff}} \]  

(2.3)

Where \( L_{\text{eff}} \) is the effective length and is defined by

\[ L_{\text{eff}} = \frac{(1 - e^{\alpha L})}{\alpha} \]  

(2.4)

And \( \gamma \) is the non linear parameter measured in \( \text{W}^{-1}/\text{km} \) and is defined by:

\[ \gamma = \frac{2\pi n_2}{A_{\text{eff}} \lambda} \]  

(2.5)

The second phenomenon, XPM, occurs when two or more optical channels are co-propagated in the fiber. A phase shift is produced on channel \( j \)th due to the effect of the other channels power and the total phase shift is govern by:

\[ \phi_{\text{NL, } j} = \gamma L_{\text{eff}} P_j + \gamma L_{\text{eff}} \sum_{m \neq j}^M P_m \]  

(2.6)

Where \( P_j \) is the power of the \( j \)th channel, \( P_m \) is the power of the overlapping channels and \( M \) is the number of channels. The first term is the phase shift induced by the SPM and the second term is the term induced by the XPM. As equation 2.6 shows the phase shift induced by XPM is double the phase shift of SPM for the same average power.

In the concrete situation of two optical fields \( (E_1, E_2) \) at different frequencies \( (\omega_1, \omega_2) \) co-propagated in the same polarization direction optical field at \( \omega_1 \) suffers a phase shift described by:

\[ \phi^{1}_{\text{NL}} = \gamma L_{\text{eff}} (P_1 + 2P_2) \]  

(2.7)

Where \( P_1 \) and \( P_2 \) are the power associated at \( E_1 \) and \( E_2 \) respectively.

NOLM uses the XPM between control signal and data signal to switch the desirable channel. Control signal in this device is an optical pulse train at bit rate of the single channel. Optical control pulses has to be synchronized to the channel pulses which extraction is desirable. Delaying the control train pulse through the time slot enables selection of the channel. A NOLM is formed by a fiber loop with two 3dB’s couplers and a HNLF. Figure 2.2 show the schematic of a NOLM. Data signal \( (E_{\text{data}}) \) is launched in coupler A and is split in two signals that travel through the loop in different directions. Meanwhile, control signal \( (E_{\text{counter, in}}) \) is launched in coupler B. The data signal is travelling in the same direction as control signal is called co-propagated signal \( (E_{\text{co}}) \) and the one is travelling in the other direction is called counter-propagated signal \( (E_{\text{counter}}) \). The transmitted signal is signal at the output 1 of the coupler A and the reflected signal is the signal obtained at the output 2 of the same coupler.

In the absence of control signal, counter-propagate signal and co-propagate signal experience the same fiber effects. In the transmitted output counter-propagated
signal experience a phase shift of $\pi$ due to the coupler A, $\pi/2$ when data signal is split and $\pi/2$ when is recombined again. While co-propagated signal do not experience any phase shift due to the couplers. Thus, the two waves interfere destructively and no signal is transmitted. However, in coupler A output 1, the whole input signal is reflected because both signals, counter-propagated signal and co-propagated signal, have the same phase shift due to couplers and fiber. They interfere constructively. Due to this whole reflection this structure is called mirror.

To use the NOLM as a switch for an OTDM system, a control signal has to be injected in coupler B and it propagates only in one direction with the co-propagated signal. Due XPM induced by the control signal, the co-propagate signal suffer a phase shift in the specific channel that coincide with the control signal. To obtain a successful channel demultiplexing the phase shift induced has to be equal to $\pi$. Thus, the counter-propagated signal and the co-propagated signal interference constructive for desirable single channel in the output 2 because co-propagated signal experience a phase shift of $\pi$ due to XPM and counter-propagated signal suffer the same phase shift due to the couplers. Input single channel is transmitted entirely and the rest of the channels interference destructive and are not transmitted, they are reflected. A band pass filter at the output is needed to select the data signal and filter the control signal. To achieve suitable $\pi$ phase shift to switch on off the OTDM signal, the power of the control signal and the length of the fiber have to be chosen correctly.

Figure 2.2. Demultiplexing scheme for OTDM signal based on XPM in a nonlinear optical-loop Mirror.
In order to well understand the demultiplexing operation by a NOLM a studied of the experience of an optical field is carried out. Optical data field launched in the NOLM is $E_{\text{data}}$ and $E_{\text{control, in}}$ is control field. The principle of operation of a 3dB coupler is defined by equation 2.8 and a coupler model is illustrated by:

\[
\begin{bmatrix}
  E_3 \\
  E_4
\end{bmatrix} = \begin{bmatrix}
  1 & j \frac{1}{\sqrt{2}} \\
  j \frac{1}{\sqrt{2}} & 1
\end{bmatrix} \begin{bmatrix}
  E_1 \\
  E_2
\end{bmatrix}
\]

(2.8)

Where $E_1$ and $E_2$ are the two input fields in the coupler and $E_3$ and $E_4$ are the two output fields. Coupler A split data field in co-propagate field ($E_{\text{co}}$) and counter-propagate field ($E_{\text{count}}$).

\[
E_{\text{count}} = j \frac{1}{\sqrt{2}} E_{\text{data}}
\]

(2.9)

\[
E_{\text{co}} = \frac{1}{\sqrt{2}} E_{\text{data}}
\]

(2.10)

The transmitted data signal ($E_{\text{tx}}$) according equation is the sum at the output 3 in coupler A of both fields after their propagation in the loop.

\[
E_{\text{tx}} = j \frac{1}{\sqrt{2}} E_{\text{count, tx}} + \frac{1}{\sqrt{2}} E_{\text{co, tx}}
\]

(2.11)

Where, $E_{\text{tx, count}}$ and $E_{\text{tx, co}}$ are the counter-propagated field and co-propagated field after the HNLF.

\[
E_{\text{tx, count}} = e^{-\alpha L} E_{\text{count}} e^{j \phi_{\text{count}}} = j \frac{e^{-\alpha L}}{\sqrt{2}} E_{\text{data}} e^{j \phi_{\text{count}}}
\]

(2.12)

\[
E_{\text{tx, co}} = e^{-\alpha L} E_{\text{co}} e^{j \phi_{\text{co}}} = \frac{e^{-\alpha L}}{\sqrt{2}} E_{\text{data}} e^{j \phi_{\text{co}}}
\]

(2.13)

Where,

\[
\phi_{\text{count}} = \gamma P_{\text{count}} L_{\text{eff}}
\]

(2.14)

\[
\phi_{\text{co}} = \gamma [P_{\text{co}} + 2P_{\text{control}}] L_{\text{eff}}
\]

(2.15)

$\phi_{\text{co}}$ is the phase shift inducing in co-propagated signal due XPM and SPM that take place in the fiber. And $\phi_{\text{count}}$ is the phase shift for counter-propagated signal due SPM. $\alpha$ is the attenuation in the fiber, $\gamma$ and $L_{\text{eff}}$ is the effective length of the fiber. The dispersion in the fiber is not taken into account, $P_{\text{co}}$, $P_{\text{count}}$ and $P_{\text{control}}$ is the optical power associated to the co-propagated, count-propagated signal and control signals travelling through the fiber:
\[ P_{\text{count}} = |E_{\text{count}}|^2 = \frac{1}{2} |E_{\text{data}}|^2 = \frac{1}{2} P_{\text{data}} \quad (2.16) \]

\[ P_{\text{co}} = |E_{\text{co}}|^2 = \frac{1}{4} |E_{\text{data}}|^2 = \frac{1}{4} P_{\text{data}} \quad (2.17) \]

\[ P_{\text{control}} = |E_{\text{control, out}}|^2 = \frac{1}{2} |E_{\text{control, in}}|^2 = \frac{1}{2} P_{\text{control, in}} \quad (2.18) \]

Where \( E_{\text{control, in}} \) is the control field at the input of coupler B and \( E_{\text{control, out}} \) is the control field at the output of this port and \( P_{\text{data}} \) is the power of the data input signal.

According with these definitions it is possible to re-write the phase shifts as:

\[ \phi_{\text{count}} = \gamma \frac{P_{\text{data}}}{2} L_{\text{eff}} \quad (2.19) \]

\[ \phi_{\text{co}} = \gamma \left( \frac{P_{\text{data}}}{4} + 2 \frac{P_{\text{control, in}}}{2} \right) L_{\text{eff}} \quad (2.20) \]

The difference between phases shifts 2.19 and 2.20 can be approach by:

\[ \Delta \phi = \phi_{\text{count}} - \phi_{\text{co}} \approx \gamma P_{\text{control, in}} L_{\text{eff}} \quad (2.21) \]

Combining equations 2.11, 2.12 and 2.13 and replacing \( \phi_{\text{co}} \) and \( \phi_{\text{count}} \) for their real values.(equations 2.14,2.15). The data transmitter data field is:

\[ E_{\text{tx}} = \frac{e^{-\alpha L}}{\sqrt{2}} E_{\text{data}} \left[ e^{j \phi_{\text{data}} L_{\text{eff}}} - e^{j \phi_{\text{data}} + \phi_{\text{control}} L_{\text{eff}}} \right] \quad (2.22) \]

Developing the equation 2.22 and some mathematics, the result for the transmitted field is:

\[ E_{\text{tx}} = \frac{e^{-\alpha L}}{\sqrt{2}} E_{\text{data}} e^{-j \Delta \phi} \left[ 1 - e^{j \phi_{\text{control}} L_{\text{eff}}} \right] = -\frac{e^{-\alpha L}}{\sqrt{2}} E_{\text{data}} e^{j \Delta \phi} \left( e^{j (\phi_{\text{control}} L_{\text{eff}})} \right)^{1/2} \left[ \text{sen} \left( \frac{P_{\text{control, in}} L_{\text{eff}}}{2} \gamma \right) \right] \quad (2.23) \]

And doing the module square the optic power of the transmitted signal is found.

\[ P_{\text{tx}} = |E_{\text{tx}}|^2 = \frac{e^{-2 \alpha L}}{2} P_{\text{data}} \text{sen}^2 \left( \frac{P_{\text{control, in}} L_{\text{eff}} \gamma}{2} \right) \quad (2.24) \]

According this equation, no power is transmitted when \( P_{\text{control, in}} = 0 \). And \( E_{\text{data}} \) is transmitted when:

\[ \text{sen} \left( \frac{P_{\text{control, in}} L_{\text{eff}} \gamma}{2} \right) \neq 0 \quad (2.25) \]
So, whether control signal is transmitted with suitable power at the same time of the select channel according equation 2.24 the single channel will be demultiplexed. In order to achieve a successful demultiplexing channel the power difference between the channel and the rest of OTDM signal has to be as high as possible. Thus, the maximum transmitted power is required. This is obtained when:

\[ P_{\text{control,in}} L_{\text{eff}} \gamma = \pi \]  

Thus, when:

\[ P_{\text{control,in}} = \frac{\pi}{L_{\text{eff}} \gamma} \]  

The peak power of the train control pulses and fiber parameters has to fulfill equation 2.27 to be efficient as demultiplexer.

Another issue to take into account is the signal and control frequency located due to the impairment of walk-off. Walk-off is produced because data and control signals experience different groups velocities. So if it creates a delay between data and control signal that exceeds a time slot of the OTDM signal, the control signal interact with other channels and the demultiplexing of the single channel cannot be possible. Therefore control and data signal should be set symmetrically respect zero dispersion frequency to achieve nearly the same group velocity [27].

**Four wave mixing**

Four Wave Mixing (FWM) scheme presented is also based on fiber nonlinearities. In [28] this technique is used to a success demultiplexing 10G/s data channel from a 500Gb/s transmission signal.

FWM occurs when two or more frequencies are propagated together in an optical fiber. Due intensity dependence of refractive index, this co-propagation of several waves generates a number of new frequencies which frequencies depend on the frequencies of the input signals. This generation occurs under phase matching requirement based on the requirement of momentum conservation [29].

Demultiplexing technique is based on the concrete situation of two frequencies propagating together. An OTDM signal launched into a nonlinearity fiber together with a clock signal. Control signal is an optical pulse train synchronized to desirable single channel set at frequency \( \omega_{\text{control}} \). Clock signal is an intense power signal meanwhile OTDM signal is a weak signal set at \( \omega_{\text{signal}} \). Another signas appears at frequencies:

\[ \omega_{\text{FWM}} = 2 \omega_{\text{control}} \pm \omega_{\text{signal}} \]  

Figure 2.3 illustrates the schematic implementation of a demultiplexer based on FWM. The data signal and control signal situated at different wavelengths are coupler in a HNLF. The FWM in the fiber creates a replica of the single channel that wants to be demultiplexed at frequency \( \omega_{\text{FWM}} \). This is because the control signal is
synchronized with the desirable channel needs to be demultiplexed and every time that control pulse overlap with this channel creates a replica of this pulse at frequency $\omega_{FWM}$.

![Diagram of demultiplexing scheme for OTDM signal based on FWM in a HNLF.](image)

Figure 2.3. Demultiplexing scheme for OTDM signal based on FWM in a HNLF.

A band pass filter at the FWM wavelength is needed at the output of the fiber in order to separate the OTDM signal and control signal. The new frequency created is generated from the intensity of the two input powers. Power of the signal created can be written as:

$$P_{FWM} = \gamma^2 P_{control}^2 P_{signal} e^{-\alpha L} L_{eff}^2 \eta$$

(1.29)

Where $L$ is fiber length, $L_{eff}$ is the effective length, $\gamma$ the nonlinear coefficient, $\alpha$ is the fiber loss and $P_{control}$ and $P_{signal}$ are the power of control and data signal respectively. Efficiency is defined as the ratio between the demultiplexed power at $\omega_{FWM}$ and the input data frequency at $\omega_{signal}$. It is defined as [30]:

$$\eta = \frac{P_{FWM}}{P_{signal}} = \frac{\alpha^2}{\alpha^2 + \Delta \beta^2} [1 + \frac{4e^{-\alpha} \sin^2 (\frac{\Delta \beta L}{2})}{(1 - e^{-\alpha})^2}]$$

(1.30)

Where $\Delta \beta$ is the phase mismatch and is the difference of the propagation of control and data signal and can be approximated by[30]:

$$\Delta \beta = -\frac{\lambda^2 \pi}{c^2} S 2(f_{control} - f_{signal}) (f_{control} - f_o)$$

(1.31)

Where $c$ is the velocity in the vacuum, $S$ is the slope dispersion and $f_o$ is zero dispersion frequency. As equation 1.29 shows localization of control and data signal
is an important issue efficient generation of $P_{\text{FWM}}$. Efficiency is maximal when the phase matching condition is satisfied. It happens when $\Delta \beta$ is equal to zero, that means, according with equation 1.31 that control frequency should be set at zero dispersion frequency [30].

### 2.4 Summary

In this chapter description of an OTDM, the optical solution for the time division multiplexing, is given. Requirements to an implementation of OTDM system and alternatives for demultiplexer devises are discussed.

In the transmitter, narrow pulses source are required to avoid the overlapping channel. Challenges in this pulses source and several compression techniques are recently investigated achieving pulses width narrow than 460 fs.

In the receiver, several techniques for channel demultiplexing exist. High demultiplexing bit rates are given by all optical techniques based on HNLF. The faster results reported are based on NOLM schemes which can demultiplex channels at bit rates of 640Gb/s.
Chapter 3

Fundamental in advanced modulation formats

When the design of an optical system is carrying out, the process to convert the electrical binary data into an optical signal entails important issues to take into account. The election of a modulator, the signal format and the modulation format are one of them. The approach gives to this process has an influence in the performance of a system.

The goal of this chapter is to give an overview of digital modulation and demodulation schemes in optical communication for high bit rates. In the first part of the chapter a general introduction to digital modulation is given where different modulation formats are summarized. Then, in the second part of the chapter, the attention is focus on advanced modulation formats where Differential Phase Shift Key modulation and Differential Quadrature Phase Shift modulation is wide explained. For each modulation a theoretical principle definition is given followed by a detail description of the way to performance it.

3.1 Signal Formats

Non-Return to Zero (NRZ) and Return to Zero (RZ) are the most used signal formats. In the first format the pulse take up whole time interval reserved for a bit. In the other hand, in RZ format, pulses occupy a fraction of the bit slot and return to zero in every bit slot. Figure 3.1 shows an example of implementation of NRZ format and RZ format for a specific binary sequence where B is bit rate of the sequence, T is the time slot 1/B and A is the amplitude of the signal.
The biggest advantage associated at NRZ is better bandwidth efficiency; the bandwidth associated to the NRZ format is smaller than a RZ signal. However, Personick,[32],1973, was the first to report better sensitivity for RZ formats than NRZ for the same average power. Recent theory and experimental papers support this sensibility improvement. Advantage of RZ format stems in the fact that the field does not entail the time slot 1/B. So NRZ pulses have less peak power than RZ pulses for the same average power. It implies a wider open eye that is associated with a BER improvement. In addition, it was also demonstrated that for a RZ format improvement in nolinearity tolerance is achieved [33-34].

### 3.2 Modulations Formats

Different alternatives are possible in order to modulate the optical signal. This alternatives depend on the physical variable of the optical signal that is modulated. The general expression for the optical field is:

\[
E(t) = \sqrt{P(t)} \cos(\omega_o(t)t - \phi(t)) \hat{x}(t) \tag{3.1}
\]

Where \( P \) is the power amplitude, \( \omega_o \) is the carrier frequency, \( \phi \) is the phase and \( \hat{x} \) is the polarization vector. These four optical variables are possible alternatives for optical modulation. Depending on if the amplitude, the frequency or the phase of the signal is modulated the modulation techniques is called respectively Amplitude Shift Keying (ASK), Frequency-Shifted Keying(FSK), Phase Shifted Keying (PSK) or Polarization Shift Keying (POLSK). In the case of ASK \( P(t) \) is varied between two levels of power, one defines 0s bit and the other 1s bit. In this way, in FSK the frequency \( \omega_o \) is shifted between two values, in PSK is \( \Phi \) the shifted variable and polarization vector for POLSK.

On-off keying (OOK) modulation based in RZ or NRZ format is the most common ASK modulation schemes. In OOK, the amplitude, \( P(t) \), of the optical signal varies between two levels, one of them is 0. For a long time OOK has been the dominant modulation format in fiber-optical communication systems because of the
problems of generation and detection for a phase modulated signal. PSK modulations carry the information in the phase of the signal. This leads into a necessity of an accurate phase modulator, a stable laser and a synchronized laser for coherent receivers. In the recent years, improvements in phase noise resilience and easier implementation of stable receivers are obtained and it leads in an increase attention in advance modulations based on phase such DPSK.

### 3.2.1 DPSK

PSK modulation carries the information in the optical phase and for this reason a coherent receiver is needed to recover the phase information. Coherent receivers involve a narrow linewidth laser that needs synchronization with the signal and leads in more complicated receivers. Furthermore, since information in PSK modulation reside in the phase value, PSK is very sensitive to any changes in phase due to noise in the channel or linewidth of transmitter laser. Due to these facts an advance modulation, variant of PSK, called Differential Phase Shift Keyed (DPSK) appears. In DPSK modulation the direct detection is possible because the information reside in the difference phase of neighbor bits. Furthermore, signal is successful recovered as long as the phase difference between two neighbor’s bits is stable. This relaxes the requirements of phase stability in transmitter.

In DPSK modulation, the difference between consecutive bits of transmitter binary sequence is represented by the optical phase difference $[0, \pi]$ . The phase of the transmitted symbol is:

$$
\theta_i = \theta_{i-1} + \Delta \theta
$$

Where the $\theta_i$ is the phase of the transmitted symbol i, $\theta_{i-1}$ is the phase of the previous bit and $\Delta \theta$ is the difference of consecutive bits that is defined by:

$$
\Delta \theta = \pi x_i
$$

$x_i$ is the bit information. So the sequence is encoded in the difference between consecutive phases. In table 3.1, a concrete transmitted bit stream $x_i$ is differential encode in sequence $y_i$ and the phase transmitted $\theta_i$ is shown, $\pi$ when differential encode sequence is one and zero when is 0.

<table>
<thead>
<tr>
<th>$x_i$</th>
<th>1</th>
<th>1</th>
<th>0</th>
<th>0</th>
<th>1</th>
<th>1</th>
<th>1</th>
<th>0</th>
<th>0</th>
</tr>
</thead>
<tbody>
<tr>
<td>$y_i$</td>
<td>1</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>1</td>
<td>0</td>
<td>1</td>
<td>1</td>
<td>1</td>
</tr>
<tr>
<td>$\theta_i$</td>
<td>$\pi$</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>$\pi$</td>
<td>0</td>
<td>$\pi$</td>
<td>$\pi$</td>
<td>$\pi$</td>
</tr>
<tr>
<td>$\Delta \theta_i$</td>
<td>$\pi$</td>
<td>$\pi$</td>
<td>0</td>
<td>0</td>
<td>$\pi$</td>
<td>$\pi$</td>
<td>$\pi$</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>$r_k$</td>
<td>1</td>
<td>1</td>
<td>0</td>
<td>0</td>
<td>1</td>
<td>1</td>
<td>1</td>
<td>0</td>
<td>0</td>
</tr>
</tbody>
</table>

Table 3.1. Example of coding and decoding DQPSK for a concrete binary sequence.
The implementation of the transmitter requires a precoder and a modulator. In the receiver, it is possible to use direct detection to recover the sequence by using one delay interferometer. Figure shows the implementation of a DPSK system based on the implementation of [7].

Figure 3.2. Schematic representation of a DPSK transmitter and receiver.

Previous codification of the information is required by using difference between neighboring bits. This codification is done by logical circuit. The bit sequence \( x_i \) is encoded by a 1 bit delay feedback an XOR gate. The encoded bit \( y_i \) is equal to:

\[
y_i = y_{i-1} \oplus x_i
\]  

The generation of a DPSK signal requires an external modulator to change the phase of an optical signal according to the precoder electrical signal. This phase modulator imposes a phase shift of \( \pi \) in the signal when a bit “1” appears in the encoded sequence.

An important issue is the choice of modulator used in the transmitter. The simplest choice is use an optical phase modulator which induces a linear optical phase shift. The common phase modulator is electro optic modulator based on an electro optical crystal but it is also possible to exploit thermally induce refractive index changes or length changes. Electro optic modulators are based on the electro optic effect created in this crystal. Electro optic effect consist in the modification of the refractive index of a medium as a result of apply an electrical field. The most common electro optic crystal used for modulator is LiNbO\(_3\) crystal. Phase modulators are composed by a simple waveguide of LiNbO\(_3\) crystal through the optical signal is propagated. The refractive index of this crystal suffers a variation proportional to the electrical applied voltage [1]. This refractive index variation induces a phase shift in the optical field when is propagating through the waveguide. This phase shift is proportional to the electrical voltage used to drive the modulator. Equation 3.5 shows the relation between these two factors.

\[
\phi(t) = \pi \frac{V(t)}{V_\pi}
\]  

(3.5)
\[ \Phi(t) \text{ is phase shift induced by electro optic effect, } V(t) \text{ is the voltage applied and } V_{1/2} \text{ is the half-wave voltage. It is the voltage needed to induce a phase shift of } \pi. \]

That is why it is required an accurate voltage to drive phase modulator. Phase shift induce in the optical field is very sensitive to the electric field fluctuations. Phase modulator only modulates the phase keeping constant the amplitude of the optical signal. Figure 3.3 shows the principle of operation of a linear phase modulator.

\[ P_{\text{out}} = P_{\text{in}} \cos^2 \frac{\Delta \phi}{2}, \quad (3.6) \]
Where \( \Delta \phi = \phi_{\text{upper}}(t) - \phi_{\text{lower}}(t) \). Upper and lower phase shifts are \( \phi_{\text{upper}}(t) \) and \( \phi_{\text{lower}}(t) \), respectively. These phase shifts are related with the applied voltage following equation 3.5. The optical power function as function of applied voltage is governed by equation

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

(3.7)

An important issue is the way to drive the Mach-Zehnder to achieve phase modulation. The suitable operation point has to be found. To achieve a phase shift of \( \pi \) the electric voltage is applied in one of the arms and an amplitude of 2V is needed. This driving voltage enable a phase shift of \( \pi \) between 0 applied voltage and 2V applied voltage[7]. Figure 3.4 shows the transfer function and operation condition to obtain phase modulation. However if Mach-Zehnder modulator is used a part of phase modulation a residual amplitude modulation appears. This amplitude modulation is due to the way to drive the modulator. One change in the sign of the optical field involves a transition through the minimum of the power MZM transfer function and this is the reason for the residual amplitude modulation. This MZM method of performing phase modulation results in opposite optical phases for different bits that guarantee an accurate 180 phase shift on the signal even when the driving voltage is not exactly the required voltage.

Figure 3.4. Principle of operation of phase modulation using MZM. The inputs are a CW optical signal and the electrical signal wanted to modulate. Phase and amplitude of the resulting output is show.
The typical choice for a DPSK receiver is the balanced receiver. Figure 3.2 illustrates a typical configuration of a DPSK balanced receiver. It receiver consist in a pre process of the signal and direct detection. This receiver is made from one Mach-Zehnder interferometer and two photodiodes. The outputs of the two photodiodes are subtracted and sequence $r_i$ is obtained that corresponds to the sequence $x_i$. Table 4.1 illustrates $r_i$ for a concrete example. Direct detection using only a photodiode is inadequate to detect the optical phase of the receiver signal. Photodiodes only convert the incident optical power ($P_{in}$) into electric current ($I_p$) following the equation $I_p = R P_{out}$, where $R$ is the responsivity of the photodetector (units of A/W). Thus photodiode is insensitive to the optical phase. A process to provide the differential optical phase shift before the photodiode is needed for DPSK direct detection. It is performed using a Mach-Zehnder one delay interferometer. This device consists in 3dB couplers connected and one bit delay in one of arm. The receiver signal is launched in the interferometer. This signal is propagated through the two different arms of the interferometer. One arm delays a bit slot the signal. At the outputs two adjacent bit slots interfere. Due to this fields interference and the fact that the data is differential encode it is possible to recover the data stream ($r_i$). Because of the couplers a constructive interference is produced in the upper arm and a destructive interference take place in the lower arm. The same data sequence is recovered in both arms but they are inverted. If only one arm is photodetected the receiver is called single detection. However if the two outputs of the interferometer are photodetected and subtracted is called balanced receiver. The great benefit of balanced detection is that a 3dB lower optical signal-to-noise (OSNR) is required to achieve a particular BER.

3.2.2 Multilevel modulation : DQPSK

The main technique used in digital modulations to achieve greater spectral efficiencies is multilevel modulations. Using multilevel modulation is possible increase the bit rate keeping the same symbol rate as in binary modulations and the same spectral properties. These modulations consist in increase the number of bits per symbol, increasing the number of possible symbols. For example, in a 4-ASK modulation, the amplitude, $P$, of the signal is shifted between 4 values of power. Each value represent two bits (00,01,10,11). The optical multilevel modulation is one of the attractive candidates to significantly increase the channel bit rate and total capacity of future optical fiber communications.

DQPSK

DQPSK is a multilevel modulation where the information is encode in the optical differences between consecutive symbols. The phase shift can take 4 different values: $[0,\pi/2, \pi, 3\pi/2]$. Figure 3.5 illustrates the symbol constellation of a DQPSK signal.
Figure 3.5. Phase constellation of a DQPSK signal using Gray code.

Each symbol consists in two bits of information. The symbols per second rate is the half of the bit rate, bits per second.

Figure 3.6 shows the configuration of a typical DQPSK system based on [35]. The transmitter consists of a previous precorder and two parallel DPSK transmitters with a phase shift of $\pi/2$ between them. The receiver is formed by two balanced receivers using a phase shift in the interferometer equal to $\pi/4$ and $-\pi/4$.

Figure 3.6. Schematic model of the transmitter and receiver DQPSK system.

The encode sequence, $I_i$, $Q_i$, are generated by a precoder formed by logical gates. The precoder has 4 inputs: two inputs are data streams ($u_i, v_i$) and the other two are feedback of the outputs of the precoder of the previous bit ($I_{i-1}, Q_{i-1}$). Equations 3.8, 3.9 describe the precoding function.

$$I_i = v_i \oplus (I_{i-1}Q_{i-1}) + u_i \oplus (I_{i-1}Q_{i-1})$$  
$$Q_i = u_i \oplus (I_{i-1}Q_{i-1}) + v_i \oplus (Q_{i-1}I_{i-1})$$  

(3.8)  
(3.9)
Optical signal is split in the two arms of the transmitter using an optical power splitter. The two outputs ($Q_i$, $I_i$) of the precoder are used are independent phase modulated by two MZMs, one of them is $\pi/2$ shifted and both of them are combined forming the phase constellation illustrated in figure 3.5.

The DQPSK receiver is formed by two MZ-interferometers employed by balanced receivers. Thus, it is composed by two arms and in each arm a DPSK receiver is performed. The delays used for each MZ-interferometer is the double of the bit period, the symbol period. And different phase shift between interferometers arms are needed for each one, one is set to $\pi/4$ and the other to $-\pi/4$. It is possible to recover the sequence after the balanced receiver. Table 4.2 illustrates an example of a DQPSK signal generation and decoding. It can be seen that each symbol, $u_i v_i$, is represent by a value of the optical phase difference $[0, \pi/2, \pi, 3\pi/2]$.

<table>
<thead>
<tr>
<th>Transmitted sequence</th>
<th>$u_i$</th>
<th>0</th>
<th>1</th>
<th>1</th>
<th>1</th>
<th>0</th>
<th>0</th>
</tr>
</thead>
<tbody>
<tr>
<td>$v_i$</td>
<td>1</td>
<td>1</td>
<td>0</td>
<td>1</td>
<td>0</td>
<td>0</td>
<td>1</td>
</tr>
<tr>
<td>Encode sequence</td>
<td>$I_i$</td>
<td>0</td>
<td>0</td>
<td>1</td>
<td>1</td>
<td>1</td>
<td>0</td>
</tr>
<tr>
<td></td>
<td>$Q_i$</td>
<td>1</td>
<td>0</td>
<td>1</td>
<td>1</td>
<td>1</td>
<td>1</td>
</tr>
<tr>
<td>Phase of transmitted signal</td>
<td>$\Phi_i$</td>
<td>$7\pi/4$</td>
<td>$\pi/4$</td>
<td>$5\pi/4$</td>
<td>$5\pi/4$</td>
<td>$7\pi/4$</td>
<td></td>
</tr>
<tr>
<td>Optical phase difference</td>
<td>$\Delta\Phi_i$</td>
<td>$\pi/2$</td>
<td>$\pi/2$</td>
<td>0</td>
<td>$3\pi/2$</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Received sequence</td>
<td>$r_k$</td>
<td>1</td>
<td>1</td>
<td>0</td>
<td>0</td>
<td>1</td>
<td></td>
</tr>
<tr>
<td></td>
<td>$s_k$</td>
<td>0</td>
<td>1</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td></td>
</tr>
</tbody>
</table>

Table 3.1. Example of coding and decoding DQPSK for a concrete binary sequence.

**3.3 Summary**

Several years ago most of the efforts and attention research lied in the OOK modulations but since the development of the laser stability increase attention on PSK modulations. This attention is focusing in PSK advance modulations such DPSK. These modulations are based on differential phase between symbols. It makes modulation more resistant to phase noise. The receiver is a balanced received based on direct detection but requires a pre-process stage based on interferometers. This modulation also introduces a 3dB of improvement in received sensitivity. Multilevel modulations such DQPSK have the advantage over binary modulations of increase the bit rate of the signal keeping symbol rate and spectrum properties. A transmitter and receiver configuration is introduced for DPSK and DQPSK systems in this chapter and theory about these modulations is presented.
Chapter 4

Simulation results

This Chapter reports the simulation work carried out. The tool used to perform all this numerical simulation is VPItransmissionMarker [41].

Section 4.1 gives an introduction of the main figures of merit to evaluate the performance of simulations and how they are calculated. A brief discussion on general simulation parameters are also introduced in this section. Section 4.4 discusses the results obtained for simulations of an OTDM system using NOLM as a demultiplexer. Two different systems are reported: a 640 Gb/s OTDM system combined with DPSK modulation and a 1.28 Tb/s OTDM system using DQPSK modulation. Finally, section 4.4.4 is dedicated to fiber transmission for a 1.24 Tb/s DQPSK-OTDM system. In section a 1.28 Tb/s system DQPSK-OTDM system are reported using a FWM as a channel demultiplexing technique.

4.1 Simulation issues

4.1.1 Global parameters

The parameters that are common to all modules used in a simulation are called global parameters. These parameters are important to the correct operation of the simulation. Global parameters are set in this Master Thesis with the values given in Table 4.1.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Time window</td>
<td>$128/(10 \times 10^{-9})$</td>
</tr>
<tr>
<td>Bit Rate Default</td>
<td>$10 \times 10^{-9}$</td>
</tr>
<tr>
<td>Sample Rate Default</td>
<td>$20480 \times 10^9$</td>
</tr>
<tr>
<td>Sample Mode Bandwidth</td>
<td>$20480 \times 10^9$</td>
</tr>
</tbody>
</table>

Table 4.1. Global parameters of VPI simulation used in this Master Thesis.

The “Time Window” parameter sets the period of time of the simulated data. It is important because gives the resolution of the spectrum signals and the BER accuracy...
estimations. It is defined as $2^m/ \text{BitRate}$, where $m$ is the order of the PRBS used to generate data information. In all the simulations the PRBS order is 7. “Sample Rate” and “Bit Rate Default” defines the number of samples per second and the number of bits per second

### 4.1.2 Figures of merit

In this Master Thesis some figures of merit are used to describe the performance of the simulated systems and present the results. An overview of these figures and an explanation of how they are calculated are introduced in this section.

BER is the most common figure of merit to evaluate a system. The BER is understood as the probability of error for a received bit. The criterion used in this Master Thesis to consider a system error free is $\text{BER} < 10^{-9}$. To calculate the BER of the system a DPSK_BER module of VPI is used. This module estimated the error probability in DPSK systems with direct detection balance receivers. BER estimation is done using the optimum threshold ($D$). The received sample is compared with $D$ and depending on if it is below or over the threshold $D$ it is assigned one symbol or other. Equation 4.1 is used to estimate the BER.

$$BER = P_0 P_{0/1} + P_1 P_{1/0} \tag{4.1}$$

Where $P_0$ is the probability of transmitting a zero, and $P_1$ is the probability of transmitting a 1. Both of them are equal to 0.5. $P_{0/1}$ is the probability of error when 1 is transmitted and $P_{1/0}$ is the probability of error when a 0 is transmitted. These probabilities are governed by:

$$P_{0/1} = \int_0^\infty W_1(x)dx \tag{4.2}$$

$$P_{1/0} = \int_0^\infty W_0(x)dx \tag{4.3}$$

Where $W_1$ and $W_0$ are the density functions of one and zero levels respectively. To calculate this density functions the module takes into account optical beating noise, and the receiver thermal and shot noise. ASE-signal noise and ASE-ASE noise is calculated by using signal samples and Chi-square assumption. This module makes two different statistical analysis for input signal samples, one for the 1’s received where noise and optical signal are combined and another for 0’s received where only are noise. The samples for the negative detector and the positive detector are both used. This statistical analysis is used to compute the parameters [38] of the defined Chi density function.

Receiver noise is specified in the module using parameters showed in Table 4.2.
CHAPTER 4. SIMULATION RESULTS

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Electrical filter bandwidth</td>
<td>7.5 GHz</td>
</tr>
<tr>
<td>Shot noise</td>
<td>Included</td>
</tr>
<tr>
<td>Thermal noise</td>
<td>$10 \times 10^{-12}$ $A\sqrt{Hz}$</td>
</tr>
<tr>
<td>Dark current</td>
<td>0 A</td>
</tr>
</tbody>
</table>

Table 4.2. Receiver noise parameters used in VPI module BER_DPSK for BER calculations.

Gaussian assumption is used on thermal and shot noise. Parameters of Table 4.2 are used to calculated the variances as usual estimators. The resulting distribution is a combination of Gaussian distribution and chi-distribution. From the BER it is possible to derivate the receiver sensitivity. The sensitivity is defined as the minimum power required to obtain an error free system (BER=1·10^{-9}). In this Master Thesis to determine the power sensitivity an algorithm is implemented in Tcl (Tool Control Language). A power penalty metric is also used in this Master’s Thesis for common evaluation of the present results. Power penalty is defined as the increment of power required to achieve error free in the system due to some impairment introduce in the system. In systems which no carried out transmission his penalty is calculated respect the 10Gb/s system without channel multiplexing and channel demultiplexing stages. In the section which focuses on transmission issue penalty is calculated with respect to the whole OTDM system where transmitter and receiver are directly connect and no impairments due transmission take place.

Another tool used in this chapter to present some of the simulation results is the eye diagram of the signals. An eye diagram is the signal repeatedly plotted over a time interval which is a fraction of bit rate. In the case that eye diagram is performed for an optical train of pulses which not carrying amplitude information this plot is useful to see an overview of the pulses shape. If the case for eye diagram of an electrical signal obtained at the output of the receiver it is useful also to see the transition between symbols. The eye closure is defined as the maximum difference between the highest level (bit 1) and the lowest level (bit 0).

The last figure of merit presented to characterize the modulation is the constellation diagram. It shows the symbol localization in complex space. The horizontal axis is the real part of the optical field and the vertical axis is the imaginary part. The constellation diagram is achieved by sampling the signals in the appropriated sample time and plotting these samples in the complex space. In This Master Thesis a Matlab program has been used for illustrated constellation diagrams. The programs samples the optical field of the signal at the center of the pulses and plot it in a complex space.
4.2 DPSK and DQPSK

The following simulations are carried out to check that the DPSK and DQPSK system is operating properly following the characteristics discuss in theory in chapter 2. The performance of a DPSK and DQPSK systems are given in this section.

4.2.1 Model implementation.

DPSK

Figure 4.1 shows the implementation model that is used in this Master Thesis to simulate a DPSK transmitter and receiver. The implemented system has the same structure as the one explain in Chapter 2.

Pseudo random sequence is a binary sequence of length $2^7 - 1$. This sequence does not use pre-spaces, leading zeros, or post-spacing, trailing zeros. The bit rate of the generated sequence is 10 Gb/s. The pre-coder consists on an ideal logical gate XOR driven by the PRBS sequence and a feedback of one bit-delay. Once DPSK data is generated, module NRZ is the one that encode DPSK information in a NRZ rectangular signal with 1 volt of voltage for ones and 0V for zeros. A Rise Time module is used after the creation of DPSK sequence to convert the electrical NRZ rectangular pulses into smoother pulses with a rise time of 25 ps, that is a quarter of the time slot which is the inverse of the bit rate (100ps). This block is important for the correct operation of the modulator. Mach-Zehnder intensity external modulator is chosen to modulate the phase of the optical signal. MZM is driven by the NRZ PRBS sequence. Table 4.3 shows the parameters for this modulator.

The receiver is implemented as a balanced receiver. It is formed by an interferometer consisted in 3 dB coupler, two ideal PIN photodiodes, two electric filters, an ideal clock recovery and an ideal electric inverter. Electric filter are Bessel filter of fourth order with a bandwidth of 7.5 GHz.
Table 4.3. Parameters used in Mach-Zehnder modulator in DPSK transmitter.

**DQPSK**

The DQPSK system is based on the system studied in chapter 3, figure 4.2 shows the schematic.

Two independent modules generate a pseudo random data sequence. These modules are identical to the ones used previously for DPSK. The pre-coder VPI module is performed by ideal Boolean gates: NOT gates, AND gates and OR gates also and defined by:

\[
I_k = \overline{I_{k_1}} \overline{Q_{k_1}} u_k + I_{k_1} \overline{Q_{k_1}} v_k + I_{k_1} Q_{k_1} v_k + I_{k_1} Q_{k_1} u_k + I_{k_1} Q_{k_1} v_k
\]
\[
Q_k = I_{k_1} \overline{Q_{k_1}} v_k + I_{k_1} \overline{Q_{k_1}} u_k + I_{k_1} Q_{k_1} v_k + I_{k_1} Q_{k_1} u_k
\]  

(4.4)  
(4.5)

Where \(u_k\) and \(v_k\) are the original information bits and \(Q_k\) and \(I_k\) are codec bits. This logical expression is equivalent to the equation present in Chapter 3. The two branches of the transmitter perform as a DPSK transmitter, having the same parameters for the rise time and MZM. An ideal phase shift module is introduced in the upper branch of 90 degrees. The receiver is composed of two branches with a DPSK balanced receiver in each one. A DPSK BER module, introduce in section 4.1, calculate the BER in each branch and the total BER for DQPSK system is calculated as the average of the two values.
4.2.2 Simulation results

First a simulation of a 10 GHz DPSK system is carried out. An ideal continuous wave laser with a power level equal to 100mW is used. Then, a DQPSK 10Gb/s system is simulated using the same optical source.

Figure 4.3. a) Optical field at the output of the 10 Gb/s DPSK transmitter using CW laser. b) Optical field at the output of the 20Gb/s DQPSK transmitter using CW laser.

Figure 4.3a) illustrates the output optical field in the Mach-Zehnder of the DPSK transmitter. As discussed in Chapter 3, every transition between 0 and 1 bits leads a power dip in the optical field. Figure 4.3b) shows the optical field at the output of the DQPSK modulator. Three levels of power appear at the field. This appears when the two arms of DQPSK transmitter are combined. The highest levels take place when no transition takes place in any of the branches, and the lowest level power appears when a transition between a 0 and 1 bit occurs simultaneously in both arms. Finally, if there is a transition in only one arm (but not in the other), we obtain the middle level power.

Since OTDM is just possible using RZ signals, the response of the DPSK system for RZ signals is proved. Gaussian first order pulse without chirp are launched in the system. FWHM of pulses is swept in a range of [0.5-100]ps. Figure 4.4 shows the results for the sensitivity as a function of FWHM.
We observe that sensitivity is improved as the FWHM is decreasing. This is because as the pulses narrow the peak power increases for the same average power. It leads to a wider open eye of the electrical received signal that decreases the BER. Also the 20Gb/s DQPSK results are plotted in figure 4.4. Sensitivity of DQPSK has the same behavior as a function of FWHM. In this Master’s thesis we are working with OTDM systems at data rates of 640 Gb/s. As it is explained in chapter 2, this data rates require pulses with FWHM equal to 0.5ps. Sensitivity for DPSK systems using pulses with FWHM equal to 0.5ps is 24.58 dBm and for DQPSK system is -23.05 dBm.

PRBS length is another parameter that could influence the following results in this master thesis because of the influence in the calculation of the BER in DQPSK systems[42]. To be sure about stability of the sensitivity with PRBS length of data sequence, system sensibility is simulated for different PRBS lengths. Figure 4.5 shows the sensitivity versus several PRBS order. A penalty of 0.001 dBm appears between PRBS length, which is not significant.
Figure 4.5. Sensitivity calculated for a 20 Gb/s DQPSK system for different PRBS lengths.

The plot shows no important dependence on PRBS length. So a PRBS order equal to 7 is used in all the simulations.

### 4.3 OTDM Multiplexer

The performed of the channel multiplexing is a little different from the one shown in Figure 2.1 of Chapter 2. This structure is based on the structure used in previous COM lab experiments to implement an OTDM multiplexer. This implementation consists of generating an OTDM signal, but using only a single modulator instead of an independent modulator for each channel. Only one optical signal modulated at a bit rate $B$ is necessary to form an optical signal at a higher bit rate $NxB$, where $N$ is the number of channels that compose an OTDM signal and are formed by appropriately delaying the single data signal. This multiplexer consist in $n$ stages in series, where $n$ is related to the number of channels by $N=2^n$. Each stage is formed by one splitter, one combiner and one delay. The incoming signal is split in two, one part is delayed and then both parts are recombined. The delays used have to ensure that if the bit stream at the input of the multiplexer is a PRBS sequence with a given length the output also has to be a pseudorandom sequence. This is could also have been ensured using independent modulators. But in this case is achieved by using delays in each stage given by:

$$\tau_n = \left(\frac{PRBS\_length}{2^n}\right) \left(\frac{1}{B}\right)$$

(4.6)

Where $n$ is the number of the stage, $PRBS\_length$ is the length of the PRBS sequence used to generate input pulses and $B$ is the bit rate of the input pulses.
Figure 4.6 shows the model of a transmitter using this multiplexer that creates a 640 Gb/s OTDM signal composed of 64 channels. A train of pulses with a bit rate of 10 Gb/s is modulated and launched into the multiplexer. The required FWHM for the pulses is <0.6ps. An OTDM signal at 640Gb/s is achieved at the output at the demultiplexer. It has 6 stages and delays used in each one are ideal and are set with values showed in Table 4.4.

<table>
<thead>
<tr>
<th>Delay</th>
<th>Value [ns]</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\tau_1$</td>
<td>6.35</td>
</tr>
<tr>
<td>$\tau_2$</td>
<td>3.175</td>
</tr>
<tr>
<td>$\tau_3$</td>
<td>1.5875</td>
</tr>
<tr>
<td>$\tau_4$</td>
<td>0.79375</td>
</tr>
<tr>
<td>$\tau_5$</td>
<td>0.396875</td>
</tr>
<tr>
<td>$\tau_6$</td>
<td>0.1984375</td>
</tr>
</tbody>
</table>

Table 4.4. Delays values used in a 640Gb/s OTDM multiplexer using a single modulator.

4.4 NOLM

In this section a previous description of the simulated model of NOLM and a characterization of this device performance is given. Later an optimization of frequency allocation and filter bandwidth of NOLM is done for a 640Gb/s OTDM-DPSK system and 640 Gbaud/s OTDM-DQPSK system. Finally transmission issues of an OTDM-DQPSK system are presented.

4.4.1 Model implementation

In this Master Thesis simulations, in order to simplify the implementation in VPI of the NOLM presented in chapter 2 a new model based on Master project [43] is used. This model is shown in Figure 4.7.
Figure 4.7. Model used to simulate a demultiplexer for an OTDM signal using XPM in a nonlinear optical loop mirror.

For these master project simulations, the HNLF has the parameter described in Table 4.5.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Loss</td>
<td>0.3 dB/km</td>
</tr>
<tr>
<td>Length</td>
<td>500 m</td>
</tr>
<tr>
<td>Dispersion</td>
<td>0 ps/nm/km</td>
</tr>
<tr>
<td>Zero dispersion frequency</td>
<td>193.4 THz</td>
</tr>
<tr>
<td>Slope</td>
<td>0.019 ps/nm2/km</td>
</tr>
<tr>
<td>Nonlinear Coefficient</td>
<td>3.33x10^-20 m2/w</td>
</tr>
<tr>
<td>Effective Area</td>
<td>12x10^-12 m2</td>
</tr>
</tbody>
</table>

Table 4.5. HNLF parameters

It consists in four 3dB couplers and two HNLF with the same parameters. This model acts as the NOLM discussed in Chapter 2.

This model is formed by two arms, the upper arm acts as the path experienced by the co-propagated signal and the lower arm is equivalent to the path of the counter-signal. The data signal is injected in coupler A. This signal is split in two signals: co-propagated signal and counter-propagated signal. The control signal is injected in the upper arm using 3dB coupler B. It propagates through the fiber together with the co-propagated signal inducing the same phase shift as the real NOLM due to XPM and couplers B,D. In the same way, the effects that are experienced by the counter-propagated signal, due coupler A,C and SPM, travelling through this model lower arm are the same effects that counter-propagated signal suffer travelling through the NOLM. As in the NOLM structure this models needs a band pass filter to recover only data wavelength.
First, a static simulation with a CW signal is carried out in order to compare the analytical result found for NOLM output power in Chapter 2 with the results obtained for the simulation model. The parameters of fibers are given in Table 4.5, the data signal and the control signal are CW signal setting at 194.9 THz and 191.9 THz respectively. The band pass filter used for the simulation is a Gaussian second order filter with a bandwidth of 10 GHz. Power of data signal power is 100mW and the control power is swept between 0 and 2 W.

![Graph](image)

Figure 4.8. Output power of the data signal at the output of the NOLM as a function of input control peak power under static conditions

The graph in Figure 4.8 shows the output data power as a function of control input power. Analytical results (Chapter 2, equation 2.24), using values of table 4.5 for the fiber, are plotted together with the simulation results. As is expected, both plots have similar behavior but they do not match perfectly because in analytical expression issues such four mixing and dispersion are not taken into account.

Figure 4.8 is useful to set the suitable control input power to use NOLM as a switch. This value is the one that gives the maximum. This value for power control is 0.7 W.

### 4.4.2 OTDM-DPSK

Once the NOLM model is characterized, it is used to demultiplex an OTDM signal. There are some issues in fiber-based demultiplexing techniques that have to be taken into account when the parameters used in the NOLM are setting. The first one comes from the chromatic dispersion of the HNLF of the NOLM. Data signal and control pulses are sent via different wavelengths. Fiber dispersion produces different group velocities for these different wavelengths. Because of these different velocities, control and data pulses lose their temporal overlap after some propagation distance. To decrease
walk-off effect the control signal frequency has to be located symmetrically between zero dispersion frequency. This assures similar group velocities and allows a control signal synchronize with the desirable channel to demultiplex. To avoid pulse broadening due to chromatic dispersion signal and control frequencies should be closer to zero dispersion. But this fact leads to another problem due overlapping of the spectrum between these signals. High bit rates for an OTDM signals require narrow pulses and these narrow pulses have a broad spectrum. Closer localization of the wavelengths leads to a distortion of the data signal due to the overlapping of the control and data signal spectrum. The bandwidth of the filter at the output of the NOLM is another parameter that must be carefully select for the same reason. It has to be broad enough to conserve the shape of the signal and avoid distortion, but narrow enough to act with overlapping of control spectrum. In this section the effects of these two parameters in a 640Gb/s OTDM system combined with the DPSK and DQPSK modulations are studied.

A system combining DPSK modulation and OTDM is performed in this section. A 640bit/s rate is achieve by multiplexing 64 channels modulated using DPSK at 10 Gb/s. Figure 4.9 shows the schematic structure of the system used in the following simulations.

![Schematic model used to simulate a 640Gb/s OTDM system using DPSK modulation.](image)

This model implements DPSK modulator with parameters explained in previous section 4.2.1 and transmission pulses are Gaussian first order pulses with a FWHM of 0.5ps and a peak power of 900W. The higher value for the peak power is done to compensate the insertion loss in the multiplexer. So, the power of the OTDM signal at the output of the multiplexer is 100mW. The model NOLM used is the one explained above. The control signal consisted of a train of optical Gaussian pulses at 10Gb/s and with a FWHM of
0.5ps. As discussed in the previous section, the suitable peak power is 0.7 W. The DPSK receiver is set with the same parameters as in section.

First the detuning of data and control frequencies is studied. The filter bandwidth of the NOLM is set at 10 THz for these simulations. The DPSK-OTDM system is simulated by sweeping the detuning of the control and data frequencies from a 2 THz of separation and 16 THz of separation in steps of 0.1 THz. The penalty in the system is calculated as a reference of the penalty show in Figure 4.4 for a 10Gb/s DPSK system, -24.58dBm. The simulation results show a dependence on the system sensitivity with detuning of the frequencies. Figure 4.10 shows the results of the penalty as a function of the detuning control and data frequencies. The eye diagram at the output of the NOLM after the filter in three different detuning frequencies: 2.4 THz (figure), 3.5 THz (figure), 14 THz (figure) are also illustrated in the graph.

Figure 4.10. Power penalty in a 640Gb/s DPSK-OTDM system versus the detuning between the control signal and the data signal. Demultiplexed eye diagram for frequency separation of : a) 2.4 THz. b)3.5 THz c)14 THz.
CHAPTER 4. SIMULATION RESULTS

For detuning frequencies below 2.4 THz no error free system results are achieved. A penalty of less than 1 dB is obtained for separation frequencies higher than 2.8 THz and lower than 7 THz.

The optimum value for the penalty in the system is when the signals are located at 3.5 THz separated. In this case a penalty of 0.03 dB is achieved and the figure shows an open and clear eye. Some traces still appear because of the overlapping of the spectrum, but without high impact. Figure 4.11 shows the constellation diagram and the eye diagram of the output of the DPSK receiver. The constellation 4.11a) show that with this detuning no high impact appears in the phase domain, which also results in a wide open eye in figure 4.11b).

![Figure 4.11. Simulated performance of a 640 Gb/s DPSK-OTDM system using 2.4 THz of separation between data signal and control signal. a) Constellation diagram of 10 Gb/s demultiplexed signal b) Electrically converted received eye diagram.](image)

Higher penalty values are obtained for closer settings of the data and control signal frequencies. As Figure 4.10 shows, the eye diagram of the demultiplexed signal at the output of the NOLM, in the concrete case that frequency difference is 1.2 THz, present a double trace pulse. This double trace comes from the control signal that is closely located to the data signal and goes into the filter bandwidth of the NOLM. This overlapping spectrum has a severe impact in the phase of the demultiplexed signal as the constellation of Figure 4.12a) shows. The receiver signal is composed of the demultiplex data signal, but also of a contribution of the control signal that is also filter by the NOLM filter. This contribution of the control signal adds a phase that shift the data phase samples of the constellation diagram (figure 4.12a)) and decreases the differential phase between symbols. This leads a penalty in the eye opening of the receiver DPSK signal (figure 4.12b)).

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Figure 4.12. Simulated performance of a 640Gb/s DPSK-OTDM system using 3.5THz of separation between data signal and control signal. a) Constellation diagram of 10Gb/s demultiplexed signal channel b) Electrically converted received eye diagram.

When separation between signals increases over the optimum value, the penalty starts increasing as a function of frequency separation. This is because chromatic dispersion effect appears during propagation through the HNLF. Dispersion is linear with the wavelength and has a slope that is the slope dispersion. Since data and control frequencies are symmetrical to zero dispersion frequency, the walk off effect is avoided and the effect of dispersion is the same for data and control signal. However, as the separation between them increases, there is an increase in the separation from zero dispersion and an increase in the effect of dispersion. Dispersion effect is the result of different group velocities of the different frequencies of the pulse. This causes broadening in the pulses. During propagation through the fiber, pulses of data signal broader and overlapping neighbor channels resulting in a degrading and distorted demultiplex signal. Figure 4.13 illustrates the results of a detuning between control and data signals equal to 14THz.

Figure 4.13. Simulated performance of a 640Gb/s DPSK-OTDM system using 14THz of separation between data signal and control signal. a) Constellation diagram of 10Gb/s demultiplexed signal channel b) Electrically converted received eye diagram.

The constellation of the demultiplexed signal at figure 4.13a) shows phase and amplitude noise, the variance of the sample phase increases. However it maintains the
variance phase difference between symbols. Because the data reside in the different phase between symbols is still possible to obtain sensitivity. Figure 4.13 shows the eye diagram of the receiver signal and it is obvious the eye opening penalty due the phase noise introduced by dispersion.

Since the overlapping spectrum has this huge impact on the performance of the signal, it follows that the bandwidth of the filter used to filter demultiplexed signal is an important parameter in an OTDM system. In order to optimized this parameter the same system is simulated but keeping constant separation of data and control frequencies to 3.5 THz and sweeping the bandwidth of the NOLM filter. The penalty is understood in the same way as before, taking as a reference 10Gb/s DPSK sensitivity of -24.58 dBm. The penalty results are plotted as a function of filter bandwidth in figure. 4.14. It is also illustrated in the eye diagram of demultiplex signal at the output of the filter NOLM for three concrete situations: 0.2 THz, 2THz and 15THz.

The best sensitivity result is -24dBm (0.5dB of power penalty) at 2 THz bandwidth. Figure 4.14 shows that using a bandwidth narrow than 1 THz and higher bandwidth than 3.5 THz increase the penalty in more than 1 dB.

![Figure 4.14](image)

Figure 4.14. Power penalty for a 640Gb/s DPSK-OTDM system as a function of the filter bandwidth used in the NOLM. Visualization of eye diagram of demultiplexed10Gb/s signal using filter bandwidth of: a)0.2 THz, b) 2THz, c)15 THz
In the case of a narrow filter bandwidth as 0.2 THz a degradation of penalty is obtained. The filter induces phase noise, but because the correlation between phase noise suffer for different symbols the difference of the variance of the phase between symbol is remains close to $\pi$. (see figure 4.15). So it is possible to obtain error free system because an open eye is obtained. Figure 4.15 shows electric received eye diagram for the case of 0.2 THz of separation.

![Figure 4.15. Results of a 640Gb/s DPSK-OTDM system using 0.2 THz of bandwidth for the NOLM filter simulation. a) Constellation diagram of 10Gb/s demultiplexed signal channel b) Electrically converted received eye diagram.](image1)

For bandwidth values below this, the shape of the signal becomes more distorted and no error free transmission is achieved in the system.

The optimum value for the sensitivity is at 2 THz bandwidth. Since the control and data frequency are separated by 3THz, it is possible to avoid the spectrum overlap and this bandwidth has less impact in the demultiplex signal. Figure 4.16 shows a good result for the phase constellation and a wide eye opening for the case of filter bandwidth equal to 2THz.

![Figure 4.16. Results of a 640Gb/s DPSK-OTDM system using 2 THz of bandwidth for the NOLM filter simulation. a) Constellation diagram of 10Gb/s demultiplexed signal channel b) Electrically converted received eye diagram.](image2)

As increase of filter bandwidth the filter allows the spectrum of the control signal to pass, and a contribution from control signal is then added to data signal. This...
distorts the signal and decreases the penalty. For filter bandwidth broader than 10THz the penalty remains constant approximately 11 dB because both signal spectrums are within the filter bandwidth and there is no other source of degradation. If the simulation includes ASE noise or laser noise bandwidth, the degradation of the sensitivity will increase as a function of bandwidth filter, because the broader the bandwidth the more the noise bandwidth affects the signal. The filter signal in this case is compose of the data signal and also by the control signals. This explains the double trace and degradation in the eye diagram of the demultiplexed signal. So it is possible to approximate (not taking into account the impact of the shape of the filter) the received signal as:

$$\sqrt{P_{rec}(t)} = \sqrt{P_{sig}(t)}e^{j\phi_{sig}(t)} e^{j\omega_{sig}t} + \sqrt{P_{con}(t)}e^{j\phi_{con}e^{j\omega_{con}t}} = \sqrt{P_{sig}} e^{j\phi_{sig}e^{j\omega_{sig}t}} + \sqrt{P_{sig}} e^{j\phi_{con}e^{j\omega_{con}t}}$$

Where $P_{rec}$, $P_{sig}$ and $P_{con}$ are the power of demultiplexed signal, the power of data signal and the power of control signal respectively, are the phases of data and control signal and are the frequencies of data and control signal. The contribution of the control signal adds a phase offset to the clean data phase at the received signal. The phase of the control signal is set to 0 at the transmission source and the differential between frequencies is $\omega_{con} - \omega_{sig} = 2 \cdot \pi \cdot (3 \text{THz})$. The sample phase that forms the constellation at the output of the NOLM has a contribution from control signal that can be approximated by:

$$\frac{\sqrt{P_{sig}(t)}}{\sqrt{P_{con}(t)}} e^{j(\omega_{con} - \omega_{sig})} = \sqrt{70} e^{j(2 \cdot \pi \cdot 3e12)} = \sqrt{70} \cos(2\pi3 \cdot 10^{12}) + j\sqrt{70} \sin(2\pi3 \cdot 10^{12})$$

This can explain shift samples in the constellation of Figure 4.17. It is the constellation of the demultiplexed signal in the case of filter bandwidth equal to 15 THz. It can be seen that the linear phase shift induce by the control signal that has a very high penalty in the closure of the eye as show in Figure 4.17.

![Figure 4.17 Constellation diagram of 10Gb/s demultiplexed signal channel of a 640Gb/s DPSK-OTDM system using 15 THz of bandwidth for the NOLM filter simulation.](image)
4.4.3 OTDM-DQPSK

The optimization for filter bandwidth and detuning frequencies is carried out for a 640Gbaud/s OTDM system using DQPSK modulation. Since DQPSK carried two bits for symbol, the bit rate of the system increases to 1.24 Tb/s with respect use DPSK modulation. Since the spectrum width is the same for both modulation formats, the same effects for filter bandwidth and detuning control and data frequencies concerning overlapping spectrum are expect.

Figure 4.18 shows the system model used to simulate a 160Gbaud/s OTDM system with DPQSK modulation. The DPQSK transmitter and receiver follow the model explained in Chapter 3. The multiplexer is implemented as in section 4.3 Implementation of the NOLM is the one discuss above with the same HNLF parameters (Table 4.5). Data pulses have a FWHM of 0.5ps and a peak power of 900 W. FWHM of control pulses is 0.5ps and peak power is the suitable value of 0.7W.

Figure 4.18. Schematic model used to simulate a 1.28 Gb/s OTDM system using DQPSK modulation and NOLM channel demultiplexer.

The first issue to simulate is the detuning between control and data frequency. As in DPSK simulation, the detuning of data and control frequencies is swept keeping constant bandwidth filter of the output of NOLM at 10THz. Referenced sensitivity for penalty calculation is the sensitivity achieved with a 10Gbaud/s (20Gb/s) DQPSK system without multiplexing demultiplexing system. This is the sensitivity simulated at section and it is -23.05 dBm.

Figure 4.19 shows the penalty result as a function of detuning the control and data frequencies. Three different eyes diagrams for demultiplexed signal for different
frequencies values are illustrated in the graph. As expected the behavior of the graph is the same as in DPSK simulations.

Figure 4.19. Power penalty in a 640Gb/s DPSK-OTDM system versus the detuning between the control signal and the data signal. Demultiplexed eye diagram for frequency separation of: a) 2.8 THz. b)4 THz c)14 THz.

The optimum separation frequency for control and data signal is 4.5 THz, at which value the system penalty achieves its lowest value. For this case a clear DQPSK constellation (figure 4.20a) is achieved, which has a 4 phase samples separated by $\pi$ and an open eye diagram (figure4.20b).
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Figure 4.20. Simulated performance of a 1.28 Tb/s DQPSK-OTDM system using 4THz of separation between data signal and control signal. a) Constellation diagram of 10Gb/s demultiplexed signal b) Electrically converted received eye diagram.

For signal localization closer than 3.4 THz, we obtained penalty results greater than 1dB. Demultiplexed eye in the case of separation frequencies equal to 2.8THz (figure 4.19a) presents the same double trace as DPSK system modulated previously. This is due to the overlap caused for data signal by the control signal spectrum. This spectrum overlapping leads a severe degradation of the eye diagram in the demodulated DQPSK signal and the phase diagram due contribution of control signal as in the case of DPSK. Figure 4.21 shows the constellation diagram and received electrical eye diagram for the case of 2.8 THz. A phase shift is illustrated in the constellation diagram in figure 4.21a). This phase shift is due to constant phase contribution of control signal that results in a penalty in the closure of the received eye diagram that decrease the eye opening.

Figure 4.21. Simulated performance of a 1.28 Tb/s DQPSK-OTDM system using 2.8THz of separation between data signal and control signal. a) Constellation diagram of 10Gb/s demultiplexed signal b) Electrically converted received eye diagram

Dispersion for higher separation signals causes higher penalty values in the system. As it can be seen in figure 4.22a), in the constellation of the demultiplexed signal after NOLM filter for the case of 14THz of separation. As in the case of DPSK signals, phase noise are introduced in the constellation diagram due the overlapping of
data pulses produces for broadening of the pulses because of dispersion. It shows some noise phase that leads to penalty in the eye opening of the electrical received eye diagram (figure 4.22b).

Figure 4.22. Simulated performance of a 1.28 Tb/s DQPSK-OTDM system using 14THz of separation between data signal and control signal. a) Constellation diagram of 10Gb/s demultiplexed signal b) Electrically converted received eye diagram

The second parameter to discuss is the NOLM bandwidth filter. The system used in this simulation is the same as in Figure 4.23 but in this case control and data frequencies are 3.5 THz separated. Meanwhile, filter NOLM bandwidth is swept from 0 to 15 THz in steps of 0.1THz. The results are presented in figure 4.23. The graph plots the penalty as a function of the filter bandwidth. This graph also illustrated the eye diagram of the demultiplexed signal after the filter when the bandwidth is 0.4 THz, 2THz and 15THz.
Figure 4.23. Power penalty for a 1.28Gb/s DPQSK-OTDM system as a function of the filter bandwidth used in the NOLM. Eye diagram of demultiplexed 10Gb/s signal using filter bandwidth of: a) 0.4 THz, b) 2 THz, c) 15 THz

It is not possible to achieve error free performance with bandwidth narrow than 0.4 THz. Figure 4.24a) shows the impact of a bandwidth of 0.4 THz in the constellation of the demultiplexed signal and Figure 4.24b) in the electrical received eye diagram. The impact of this bandwidth is no very severe. It can be seen some noise in the constellation diagram that leads to a penalty in the open aye diagram. However it is still a clear eye diagram.
The optimum value for penalty is 1.2 dB, which is achieved using a filter bandwidth of 2.2 THz. The constellation at this is close to ideal constellation. Dispersion of the phase samples is small and the phase difference between the four symbols is $\pi$. However, the penalty increases as the bandwidth of the filter increase. This is because the filter selects an increasing part of control signal spectrum. At filter bandwidth higher than 10 THz, the penalty remains constant around 11 dBs because the whole control spectrum is within the width of the filter. Since no other source of distortion like ASE noise is taken into account the penalty stop increasing. As explained for DPSK, the contribution of the control signal adds a constant phase shift to the constellation diagram. This results in a closed eye diagram for the demodulated signal. Figure 4.25 shows the constellation and the eye diagram of the signal when a filter bandwidth equal to 15 THz is used. The shows the behavior expected. A phase shift is reproduced in the constellation diagram and a severe penalty can be seen in the eye diagram.

As can be deduced, dispersion in the fiber that makes up the NOLM is an important parameter for the quality of the demultiplexed signal. In previous simulation
we discussed how allocation of frequency signal far from zero dispersion frequency induces pulse broadening that causes an increase in receiver sensitivity. In the next simulation a study of the impact of the dispersion slope parameter of the NOLM fiber in demultiplexed signal is carried out.

The set up showed in Figure 4.18 is used. All the parameters remain the same as in previous simulations. However, frequency separation between control and data frequencies is set at 3THz and filter bandwidth is set at 20 THz. Meanwhile, the sensitivity is measured for different values of the dispersion slope.

Figure 4.26 shows the penalty results as a function of the slope of the HNLFs of the NOLM. In this figure the eye diagram of demultiplexed signal for three different values of slope, [0.01,0.028,0.08] ps/(nm²·km) fiber is also shown.

![Figure 4.26. Power penalty in a 1.28 Tb/s DPQSK-OTDM system as a function of the dispersion slope of HNLF used in the NOLM. Eye diagram of demultiplexed 10Gb/s signal using slopes of: a)0.01 ps/(nm²·km) b) 0.028 ps/(nm²·km) c) 0.05 ps/(nm²·km)](image)

As it can be appreciated in figure 4.26, the values slope below 0.019 ps/(nm²·km) gives high penalty values. Increasing the slope improves the penalty. The optimum penalty is obtained for a 0.028 ps/nm²·km slope and above this value the penalty increase as a function of the slope. These high values for reduced slopes are due to the spectrum shape. Figure 4.27a) shows the spectrum of the signal before the NOLM
filter using 0.01 ps/(nm\(^2\)·km) slope in HNLF and figure 4.27b) shows the spectrum for the same signal but using 0.028 ps/(nm\(^2\)·km) slope fibers. In the case of higher slope the signals spectrum are more defined and narrower than in the case of 0.01 ps/(nm\(^2\)·km). This shape spectrum results in less overlapping of the signals.

Figure 4.27. Spectra of data and control signals at the output of the NOLM after filter the signal in a DQPSK-OTDM system for a slope of: a) 0.01 ps/(nm\(^2\)·km) b) 0.028 ps/(nm\(^2\)·km)

This overlap in the spectrum introduces phase shift in the demultiplex signal. Figure 4.28a) shows the constellation of demultiplexed signal using fibers dispersion slope of 0.01 ps/(nm\(^2\)·km).

Figure 4.28. Simulated performance of a 1.28 Tb/s DQPSK-OTDM system using a HNLF of 0.01 ps/(nm\(^2\)·km. a) Constellation diagram of 10Gb/s demultiplexed signal b) Electrically converted received eye diagram

It is show in the constellation the phase shift induced in phase samples by the overlapping of control spectrum, which also leads in a closer electric received eye diagram as seen in Figure 4.28b). So dispersion in this case result in the beneficial avoidance of the overlapping spectrum of the control spectrum.
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For higher values of dispersion slope the penalty start to increase because the impact of the dispersion on the pulses is more severe than the beneficial impact in the spectrum. Figure 4.29a) shows the constellation diagram for signals demultiplex by HNLFs with a 0.05 ps/(nm$^2$-km) dispersion slopes. As Figure 4.29a) shows, phase noise appears in the signal. It is due the dispersion effect and not due for overlapping of the control spectrum.

![Constellation diagram](image)

**Figure 4.29.** Simulated performance of a 1.28 Tb/s DQPSK-OTDM system using a HNLF of 0.05 ps/(nm$^2$-km). a) Constellation diagram of 10Gb/s demultiplexed signal b) Electrically converted received eye diagram

### 4.4.4 Transmission span

Once the optimization of some of the parameters of the system model is carried out, this Master Thesis goes through the issue of transmission. Transmission through optical fiber has some important impairment that leads to degradation in the receiver signal and diminish the penalty in the system. Dispersion is one of these issues. Dispersion is a fiber effect originating of the fact that different frequencies associated with a pulse propagated at different group velocities. This causes a spreading of the transmission pulse as it travels through the fiber. Intuitively this can be explained as every frequency arriving at different times at the output of the fiber. Since OTDM signals need narrow pulses to avoid the overlapping between the neighbor pulses, the effect of dispersion is one of the most critical effects in fiber transmission. So it is necessary to compensate for dispersion effect. A possible technique for dispersion compensation slope is dispersion compensation fibers (DCF). Dispersion can also be negative in silica fibers at the range of frequencies below zero dispersion frequency. Typical DCF has negative dispersion and produces a negative slope, with dispersion per unit per length values that are typically ten times higher than those of the dispersion to compensate. Although the material component is fixed in silica fibers, it is also possible design a fiber with negative dispersion by varying parameters such core radius cladding radius. So a pulses that has experienced a certain amount of positive dispersion can be easily correct itself by traveling through a fiber that has an equivalent amount of negative dispersion. Usually dispersion compensation fibers are placed in line in the path of the signal every few kilometers of fiber.
Loss and nonlinearities that, as was demonstrated previously are useful to demultiplex the OTDM signal are the cause of the limitation on the transmission length. The most common method to transmit signals for long distances is to divide the total distance in different stages called span and repeat the signal. In each stage the signal is repeated and is compensated for different fiber impairments such as loss and dispersion.

In this section a 1.24 DQPSK-OTDM signal is transmitting through spans in order to discuss the distance limitation of this signals. Figure 4.30 illustrates the model system simulated. OTDM multiplexer and DQPSK modulation are generated and received with the same model system as is present in previous simulations. The model of DQPSK system back to back is illustrated in Figure 4.18. The data signal frequency is set at 194.9THz and data pulses have a FWHM of 0.5ps. The peak power of data pulses is 90 W, so 10 mW (-20 dB) are obtained at the output of the multiplexer. The technique used for channel multiplexing is the NOLM. The model introduced in the NOLM section is the one used in this system. The control signal has a power of 0.7 W and the frequency is set at 191.9THz, 3THz from the data signal and symmetrical to zero dispersion frequency. It is found the optimum allocation for these signals in previous sections. The control pulses have a FWHM of 0.5 ps and the first order Gaussian filter used has a bandwidth of 20 THz. A pre-amplifier is placed at the output of the transmitter, and there are also several amplifier stages installed along the link. Transmission over several number of spans are simulated. Each span is composed of 80 km of Single Mode Fiber fiber, an amplifier to compensate the loss of the fiber and DCF to compensate the dispersion. Finally an amplifier is at the end of the span to compensate the attenuation of DCF. Parameters of the SMF and DCF are in table 4.6.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>SMF</th>
<th>DCF</th>
</tr>
</thead>
<tbody>
<tr>
<td>Attenuation [dB/km]</td>
<td>0.2</td>
<td>0.5</td>
</tr>
<tr>
<td>Dispersion [ps/nm-km]</td>
<td>17</td>
<td>-100</td>
</tr>
<tr>
<td>Slope [ps/nm$^2$-km]</td>
<td>0.0578</td>
<td>-0.34</td>
</tr>
<tr>
<td>Effective Area [$\mu$m$^2$]</td>
<td>80</td>
<td>20</td>
</tr>
<tr>
<td>Nonlinear index [m$^2$/W]</td>
<td>2.6·10$^{-20}$</td>
<td>3.7·10$^{-20}$</td>
</tr>
<tr>
<td>Length [km]</td>
<td>80</td>
<td>13.6</td>
</tr>
</tbody>
</table>

Table 4.6. Parameters used for SMF and DCF in a transmission span.

The first amplifier used in a span is an ideal amplifier with a gain of 16 dB to compensate for the loss introduce by SMF and a noise figure of 4dB. The second amplifier has a gain of to compensate the loss insert for DCF and a figure of noise of 4 dB. The gain of the last amplifier of the last span, the one before the receiver, has to compensate the loss for the last DCF but as well vary for every different of launched power in order to achieve a fix peak power of 100mW (-10dB) in the receiver in every case. This is done in order to have the same power for every cases of launched power as in the case of the back-to back system. This transmission configuration is show in figure 4.30.
Figure 4.30. Block diagram of transmission scheme over several spans for a OTDM-DQPSK signal with a bit rate of 1.24 Tb/s.

The launched power into the fiber is swept by varying the gain in the pre-amplifier. The sensitivity of the system is calculated at the input of the DQPSK receiver. The penalty calculations are performance as the different between sensitivity after transmission and OTDM-DQPSK back-to back system, where transmitter and receiver are directly connected. Figure 4.31 shows the results for different number of spans.

Figure 4.31. Power penalty introduced in a 1.24Tb/s DQPSK-OTDM system due transmission over several spans versus launched signal power. Visualization phase constellation of demultiplexed signal after 6spans transmission at a) -20 dB power launched. b) -11 dB power launched.
In the graph, the penalty in the system is plotted as a function of the launch power. The constellations in the case of transmission of 6 spans are illustrated also in the figure. Figure 4.31a) is the constellation when data peak power is -20dB, and figure 4.31b) is constellation for data peak power of -11 dB, after 6 spans of transmission.

The number of spans increases the power penalty in the system. The impairment of fiber effect increase the penalty in the system, if the signal is transmitted through a high distance the impact of fiber effects increase and the penalty also increases. However, fiber is not the only detrimental factor in transmission but amplifiers also increase the penalty in the system adding ASE noise to the signal.

As it can be expect, Figure 4.31 shows that penalty depends on the launched power into the fiber. The penalty is limited by high values of power and lower values of power. This can be explained by analyzing the sources of signal degradation. The most important impairments due to the fiber are dispersion, loss and nonlinearities. Dispersion and loss are compensating by using amplifiers and DCF, but the nonlinearities still degrade the signal. This nonlinear effect appears because of the dependence of the refractive index of the fiber with the signal intensity (equation 2.3) and this is modulated according instantaneous power. Whether this index change is due to the pulse’s own power the phenomenon is called self phase modulation and, as it is discuss in Chapter 2, section 2.3.2 results in a phase shift of:

$$\varphi_{SPM} = \gamma P L_{\text{eff}}$$  \hspace{1cm} (4.9)

Where $L_{\text{eff}}$ is the effective length and $\gamma$ is the non linear parameter and P is the power of the signal. High signal power levels cause a high phase shift. It is one of the most significant limitations for highest values of power.

Amplifiers are also a cause of the limitation in transmission. Amplifiers add noise to the signal during amplification because spontaneous emission [1]. The metric for noise performance of a circuit is the Noise Figure (NF). When the gain needed for an amplification increase then the noise added at the signal also increases. So, although it is possible to launch a low power signal into a transmission link to avoid fiber nonlinearities the necessity of large amplification in the receiver to obtain enough power to achieve error free introduces noise that increase the penalty in the system. Figure 4.31 follows this behavior. For lower values of power the penalty in the system increase due the necessity of higher values of gain and self phase modulation increases the penalty in the cases of higher values of power.

Dynamic range is defined as the range between the smallest and largest useful input power levels. The input useful levels are called to the power levels launched in the transmission link that gives penalty values lowers than 2 dB. As it is seen in Figure 4.31 as the number of span increases the amplification dynamic range decrease. This is logical because more spans mean more amplifiers and more fiber length. A successful transmission over a two spans is achieved by a dynamic range of .3dBs, from -16 dB to -12dB of input power.
4.5 Four Wave Mixing

Four Wave Mixing is the second demultiplexing technique presented based on nonlinearities of the fiber. In this section a brief characterization of this technique is given. Then a performance of a DQPSK-OTDM system using FWM as a demultiplexer is then presented.

4.5.1 Model implementation

As is discuss in chapter 2, the demultiplexing technique based on FWM consist of launching the OTDM data signal together with the control signal. The localization of control and data signal is an important issue, as well as the power of these signals. Efficiency is maximized when the frequency control is set at zero dispersion frequency [30].

To confirm the difference wavelength dependence of the efficiency and to optimize the values of the parameters of the FWM demultiplexed signal the efficiency is calculated for different power levels and different range of frequencies. The implemented model is shown in figure 4.32.

![Figure 4.32. Schematic model used to simulate the FWM demultiplexing scheme.](image)

Data and control signals are launched into a HNLF using a 3 dB couple. The FWM effect induced in the fiber creates another the frequency that is filtered by a first order Gaussian filter. Data signal and control signal are created by an ideal CW laser. The fiber used is a HNLF and the parameters are show in Table 4.5 they are the same fiber used for the NOLM. The frequency of the control signal is set close zero dispersion frequency, meanwhile the frequency data signal is swept around this value. The control frequency is not set at the exactly zero dispersion frequency because dispersion function used for VPI is linear and not take into account the third order dispersion and the phase matching condition is always true. The zero dispersion frequency of the fiber is 193.4 THz and the data signal is swept in the range of 200.4-185.4 THz. The efficiency is
calculated for two different values of control and data power. First the power of the control is set at 0.7 W and the power of data signal at 0.1W. In the second case the power data signal is set at 10mW and the pump at 100mW.

Figure 4.33 shows the efficiency as a function of the detuning frequencies between the data signal frequency and the control frequency. Efficiency is calculated as the ratio of the power of the generated signal between the power of the data signal. An efficiency bandwidth is defined for the half of the maximum efficiency achieved.

As closer as the control frequency is set from the signal frequency as high efficiency is achieved. The bandwidth efficiency for a control power equal to 700 mW and 100mW is 5.2THz In the bandwidth efficiency presents some irregularities. This is due to the parametric gain. Parametric gain is an effect in optical fibers that attenuates or amplifies the frequency at some frequencies induced by an intense pump power [ref]. So the higher Pcontrol launched into the fiber, the higher impact of parametric amplification. Efficiency for a control pump of 100mW and a data signal power of 10mW is 9 dB lower than the first efficiency simulated. The bandwidth efficiency is wide and is equal to 7.2THz. For lower values of power the efficiency is flat. It confirms that the efficiency depends on the power signals and on the frequency signals as it is seen in Chapter 2.
4.5.2 OTDM-DQPSK

FWM demultiplexing technique is used in this section to simulate an 1.24 Tb/s OTDM-DQPSK system. The issue about detuning between control and signal frequencies is discussed.

Figure 4.34 shows the schematic of the model used to simulate the system. First the 20Gb/s DPQSK signal is modulated. The issues involved in this modulation strategy are the same as used in previous sections for modulate DQPSK (section 4.2). The data pulses peak power at the transmission source is set at 90W to obtains after multiplexer a peak power of 10mW (-20 dB). The multiplexer used is presented in section 4.3 and as it is explained in this section introduces 6dBs of attenuation for stage. The FWM is the scheme used to demultiplex data signal in the receiver. The control signal used is a 10Gb/s pulses train with FWHM equal to 0.5ps and peak power equal to 100mW. Because the highest bandwidth efficiency and the constant behavior of the efficiency, the peak power associated to control and data signals are chosen equal to 100mW and 10mW respectively. The bandwidth of the second order gaussian filter used to filter demultiplex signal after the HNLF is equal to 20 THz. This bandwidth is the optimum bandwidth as found for NOLM demultiplexing technique in section 4.4.3. The DQPSK receiver is the same used in previous sections.

The control frequency is set at 193.4 THz, which is the frequency zero dispersion of the fiber. Meanwhile the frequency of the control signal is swept between a 1THz of separation and a 15 THz of separation from control signal. As the demultiplexed signal frequency is function of the control and data frequencies, the center frequency of the NOLM filter has to be swept as well by:

\[ \omega_{FWM} = 2\omega_{control} - \omega_{signal} \]  

(4.10)
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Where $\omega_{FWM}$ is the demultiplexed frequency, $\omega_{\text{control}}$ is the control frequency and $\omega_{\text{signal}}$ is data signal frequency.

The first simulation of this system does not give error free for any of the detuning frequencies between control and data signal. The reason for the low values of BER is the low values for the demultiplexed signals generated by four wave mixing. Even for detuning frequencies between control and data frequencies in the efficiency bandwidth, the power generated of demultiplexed signal is below -30 dBm. Since the sensitivity calculated in section 2.2 for 10Gb/s DQPSK system is equal to -23.05, the low power of demultiplex signal makes it impossible to achieve error free performance in the system. For this reason a pre-amplifier before the DQPSK receiver is included in the system. This amplifier has a gain of 20 dB and a noise of figure of 4 dB. Although the amplifier adds noise to the system, it is needed to achieve error free in the system.

The system is simulated using an amplifier and the power penalty introduced by demultiplexer is calculated in the system for different separation frequencies for control and data signal. The results are illustrated in figure 4.35, where the power penalty is plotted versus the detuning frequencies. Also the eye diagram of the demultiplex signal at 3 THz and 4.5THz separation frequencies are illustrated.

![Figure 4.35. Power penalty introduced as a function of the separation between the control signal and the data signal in a 1.28Tb/s DQPSK-OTDM system using a FWM demultiplexing technique. The eye diagram for detuning frequencies of: a) 3 THz, b) 4.5 THz.](image-url)
As it is expected, the system using FWM demultiplexing technique has dependence on the detuning between control and data signal. The behavior for closer data and control signal location is the same as in NOLM. Closer frequencies lead to an increase of power penalty because of control contribution. This contribution leads to a phase shift in demultiplexed signal. This phase shift is appreciated in Figure 4.36a) where the constellation diagram of the demultiplex signal is illustrated. Figure 4.36b) illustrates the eye diagram of the receiver signal. A closer eye diagram is observed than in the optimum case that is plotted in Figure 4.37b).

![Figure 4.36. Simulated performance of a 1.28 Tb/s DQPSK-OTDM system using 3THz of separation between data signal and control signal. Control signal is situated in zero dispersion frequency. a) Constellation diagram of 10Gb/s demultiplexed signal b) Electrical converted received eye diagram.](image)

The optimum value for frequency is at 4.5THz of separation. At this point the penalty in the system is 2.4 dB. This value of penalty is higher than in NOLM simulations because of the impact of the amplifier. As it can be seen in Figure 4.37 some phase noise appears in the constellation diagram. The different of the variance of symbols is still near pi but this phase noise introduce a penalty in the eye closure as Figure 4.37b shows.

![Figure 4.37. Simulated performance of a 1.28 Tb/s DQPSK-OTDM system using 4.5THz of separation between data signal and control signal. Control frequency is set at zero dispersion frequency. a) Constellation diagram of 10Gb/s demultiplexed signal b) Electrical converted received eye diagram.](image)
However for far localization of the frequency the power penalty increase is not as rapid as in the NOLM demultiplexing scheme. The impact of this high separation is not as severe as in the previous case. In the concrete case of 12 THz separation, an increment of 0.1 dB is used instead of the increment of 3 dB for the case of the NOLM. This is because in FWM case the control signal is set at the frequency zero dispersion, so the control pulses are not broader for dispersion. Although data pulses are broader due to dispersion, the overlapping contribution of neighbor pulses can be avoided because control pulses keep narrow bandwidth and it defines the switching window. In order to see the influence of control pulse broadening the same system is simulated with 12THz of detuning frequency between control and data signal, but symmetrical to zero dispersion frequency. That is a data frequency equal to 199.4 THz and a control frequency equal to 185.4 THz. The penalty obtained in this case is 3.3 dB over the optimum penalty. Figure 4.37 shows the constellation diagram and the received eye diagram in this case.

Figure 4.38. Simulated performance of a 1.28 Tb/s DQPSK-OTDM system using 4 THz of separation between data signal and control signal. Frequencies are symmetrically to zero dispersion frequency. a) Constellation diagram of 10 Gb/s demultiplexed signal b) Electrical converted received eye diagram.

### 4.6 Summary

In this chapter main characterizes of performance for OTDM systems combined with advanced modulations at high bit rates are presented. To study the operation of these systems, a models of these systems are designed using VPI modules. Two solutions for channel demultiplexing are evaluated: NOLM technique and FWM technique. In section 4.4 are reported the results of computer simulations of an 640 Gb/s OTDM-DPSK system and 1.28 Tb/s OTDM-DPQSK system using a NOLM channel demultiplexing technique. In order to reach the final simulations setup, optimization of the two signals and filter parameters involved in demultiplexer are carried out. First a characterization of the response of a NOLM under static conditions is done in order to set the best peak power for the control signal used in the demultiplexer. The peak power
found is of 0.7W, which is the value that gives the maximum output power for
 demultiplexed signal. The frequency location issue is studied in this chapter for both
 systems. The results provide evidence of the importance of detuning frequency in the
 case of 640Gb/s DPSK signal as well as in the case of 1.28Tb/s DQPSK-OTDM signal.
 It is well known that the control frequency and data signal frequency have to be
 allocated symmetrical to zero dispersion frequency to avoid walk off. However, the
 requirements for separation of data signal from control signal are studied. The optimum
 value for these separations is 3.5 THz in both systems using a 1THz bandwidth second
 order Gaussian filter. The optimization of the bandwidth for this separation is then
 performed. The behavior for OTDM-DQPSK system and OTDM-DPSK system as a
 function of the bandwidth of the filter is also the same. The optimum value found is a
 bandwidth of 2THz. The influence for the slope used in the HNLF involved in NOLM
 for a 1.28Tb/s OTDM-DQPSK system is also show in section 4.4.3. The penalty
 introduced by a NOLM using this optimum values for 640Gb/s OTDM-DPSK system
 is 0.32 dB and for the 1.28Tb/s OTDM-DQPSK system is 1.2dB.

 Once signals and filter requirements have been defined, 1.28b/s DQPSK-OTDM
 signal is transmitted through different number of spans using 80km of SMF and 13.6
 km of DCF. The achieved distance with less of 2dBs of penalty is 561.6 km. And the
 dynamic range for this distance and for penalties under 2 dBs is 4dBs.

 In section 4.5 a FWM demultiplexing technique for a 1.28Tb/s OTDM-DQPSK
 system is performed. A low power for control and data signal is chosen because of the
 dependence between power signals and efficiency. The control frequency is set at zero
 frequency dispersion and the optimum value of separation for data signal is 4.5 THz .
 At this point the power penalty is 2.4 dB. This increase in the penalty for FWM is due the
 needed of an amplifier.
Chapter 5

Conclusions

This Master Thesis has investigated the feasibility of OTDM systems combining advance modulations at bit rates of 640Gb/s and 1.28Tb/s. Over the recent years OTDM and advanced modulations are techniques depth study by laboratories research and also are the issue addressed in this work. The focus of the study is on use DPSK and DQPSK modulations combined with this multiplexing techniques.

Demultiplexing technique is one of the main issues in the optimization of OTDM systems. At high bit rates all optical switches are required. The channel demultiplexing technique based on fiber nonlinearities is the common solution. Two schemes have been analyzed by means of computer simulations: Non linear optical mirror and Four Wave Mixing. The impact of these demultiplexing schemes at high bit rates on phase modulations has been reported.

The first demultiplexing scheme simulated is base on NOLM. A model for a NOLM based on 3dB couplers and HNLFs is used. Static conditions are used to study the behavior of the model. The results for the output power as a function of the input control power involve in the NOLM is obtained. An optimum level 0.7 W peak power is required to obtain the maximum output power. Then, simulation work is focused in NOLM as a demultiplexer. Two systems are implemented: 640Gb/s DPSK-OTDM system and a 1.28Tb/s DQPSK-OTDM system. NOLM is based on XPM effect on HNLF and two waves are involved: control signal and data signal. An OTDM signal at high bit rates such 640Gb/s requires narrow pulses with a FWHM equal to 0.5ps. This leads to problems in overlapping spectrum between control and data signals. This overlap spectrum has severe impact in the phase of the signal, the main variable for DPSK and DQPSK modulations. Detuning between these frequencies and the filter bandwidth used to select the desirable data channel are studied since are the main parameters involve which can minimize this problem. The same behavior for DPSK and DQPSK OTDM systems are obtained. The optimum detuning frequencies are achieved in the range from 3.5THz to 4.5THz when frequencies are symmetrical to zero dispersion frequency. In order to optimize the frequency bandwidth 3.5THz separation between data and control frequencies is the better choice. The bandwidth of the filter which gives the better sensitivity to the system is 20THz. This parameters avoid
overlapping spectrum and not distort the signal due dispersion for high separation frequencies or due narrow filters.

The issue of transmission is also simulated for the case of 1.28Tb/s DQPSK-OTDM signal. The signal is transmitted over spans composed by 80 km of SMF and 13.6 km of DCF, in order to compensate the impairment of dispersion. A transmission over 561.6km is achieved by a dynamic range of .3dBs, from -17dB to -12dB of input power.

FWM demultiplexing technique has also been the object of simulation work. This technique also involves a data and control signal. The efficiency of FWM generation is analyzed under static conditions in order to studied power of data and control signal requirements. Lower level powers for these signals give a high efficiency bandwidth. Then simulation set up has studied the possibility of use FWM technique to demultiplex an 1.28Gb/s OTDM-DQPSK signal. Detuning separation between these frequencies is studied in the case that control frequency correspond zero dispersion frequency. Lower levels are achieved for demultiplex signal that put into evidence the necessity of amplify the signal. These leads into higher values of sensitivity because the noise introduced by the amplifier. Optimum detuning frequency of 4.5THz is reported.

The present work has studied the performance of an OTDM system at high bit rates using DQPSK modulations. A possible way to set some of the main parameters involve in demultiplexing scheme based in HNLF to achieve the less impact in this phase modulation is present. These results present a good performance for DQPSK modulations combined with OTDM systems. OTDM systems double the bit rate using DQPSK instead DPSK. The results reported are based on simulations. Simulations are a useful tool to understand the main requirements and performance of a system. However do not give real results and real limitations for a system as experimental results where real devices are involved. For this, as a future work of this Master thesis a experimental studied is proposed.

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Bibliography


Spanish Summary: Towards multi-terabit per second single wavelength links using multilevel modulation

Actualmente, como vivimos en la era de la información, la demanda de ancho de banda está creciendo rápidamente. Para paliar esta demanda la capacidad de las redes de comunicación tiene que ser incrementada. Las comunicaciones ópticas pueden sostener estas altas capacidades gracias a su gran ancho de banda [1]. Por ejemplo el ancho de banda que puede ser transmitido en una fibra óptica es de 40 THz, que es muy superior al de cualquier campo eléctrico. Sin embargo actualmente cualquier canal óptico es limitado a 40Gb/s. Esta limitación es causada por el efecto de dispersión y los efectos no-lineales que se producen durante la transmisión a través de fibras ópticas. Sistemas de transmisión más elaborados son capaces de incrementar la capacidad de una línea óptica. Uno de estos sistemas consiste en transmitir sobre múltiples canales ópticos. El ancho de banda óptico es dividido canales de menor velocidad de transmisión y enviados a través de la misma fibra óptica. Esto puede hacerse de dos maneras diferentes, usando Wavelength Division Multiplexing (WDM) o Optical Time Division Multiplexing (OTDM).

Los sistemas WDM incrementan la capacidad usando diferentes longitudes de onda sobre una misma fibra óptica. Usando esta tecnología tasas de transmisión alrededor de 25 Tb/s son conseguidas [2].La segunda técnica mencionada, OTDM, combina múltiples canales en el tiempo. Las secuencias de información son modulados en estrechos pulsos ópticos que son retrasados y combinados en tiempo creando un único canal con una tasa de transmisión elevada.

Los sistemas WDM a velocidades alrededor de Tb/s conllevan un número de transmisores y receptores muy elevado que complica y encarece el sistema [3]. Es preferible menos canales con una tasa de transmisión elevada a un número mayor de canales con menos tasa de transmisión. Por eso, en los laboratorios ha incrementado la atención a la alternativa de aumentar la velocidad de transmisión por longitud de onda usando la tecnología OTDM. Esta tecnología necesita nuevos componentes de multiplexado y demultiplexado basados en tecnología solamente óptica como retardadores ópticos, interruptores de alta velocidad o tecnologías para la compresión de pulsos ópticos. Estas tecnologías son las utilizadas en este proyecto de fin de carrera.

Por otro lado, el formato de la modulación utilizada también puede ser utilizado para aumentar la tasa de transmisión. Modulaciones avanzadas consiguen aumenta la velocidad de transmisión, mejorar el comportamiento de la transmisión y consigue alta
eficiencia espectral. Estas modulaciones fueron investigadas a principios de los 90 pero después por largo tiempo los estudios de investigación se centraron en la modulación básica OOK porque modular la fase de una señal óptica resultaba ser un proceso poco estable [5]. Sin embargo actualmente se ha conseguido la implementación de moduladores y receptores estables que facilitan la implementación de estas modulaciones [6]. Este proyecto se centra en la modulación multinivel Differential Quadrature Phase Keying (DQPSK). Esta modulación incrementa la eficiencia espectral de la transmisión incrementando el número de bits por símbolo.

La tarea de esta tesis es caracterizar el comportamiento de un sistema OTDM combinado con la modulación multinivel DQPSK a altas velocidades de transmisión mediante simulaciones de dicho sistema. Para este estudio la herramienta de simulación utilizada es VPItransmissionMarker [41].

Dos realizaciones de demultiplexado del canal del mismo sistema OTDM son estudiadas. Ambas son implementadas con módulos del VPI. La primera usa como demultiplexador la técnica Non lineal Loop Mirror (NOLM) y la segunda la técnica de Four Wave Mixing (FWM).

Primero se simula un sistema OTDM de 640Gb/s que trabaja con una modulación DPSK y se optimizan algunos de los principales parámetros. Los parámetros que se han trabajado en esta primera optimización tenían que ver con las dos señales (la señal de información y la señal de control) y el filtro que se ven envueltos en el proceso de demultiplexación de la señal OTDM. En la primera parte de la simulación se trabaja con el demultiplexador NOLM y en la segunda se repiten las mismas optimizaciones pero usando un demultiplexado basado en FWM.

La primera comprobación es la fuerte dependencia de la localización en frecuencia de las dos señales, la de información y la de control, que forman parte en el proceso de demultiplexación tanto para el basado en NOLM como para FWM. Colocar en frecuencias muy próximas estas dos señales conlleva un solapamiento del espectro que degrada considerablemente la calidad de la señal. Por otro lado al colocar las señales simétricas a la frecuencia de cero dispersión para evitar el efecto de walk-off, una separación grande provoca un aumento del efecto de dispersión en ambas señales que destruye también la calidad de la señal. El otro aspecto que hemos tenido en cuenta a la hora de optimizar el sistema es el ancho de banda del filtro óptico usado a la salida del proceso de demultiplexación. Este filtro es el encargado de filtrar la señal donde está la información de la señal de control. Un ancho de banda demasiado corto destruiría la señal y un ancho de banda muy ancho provocaría interferencias entre estas dos señales. Una vez que se ha comprobado que un sistema OTDM es totalmente compatible con una modulación de fase como DPSK se duplica la velocidad de transmisión usando una modulación multinivel DQPSK. Se pretende hacer el mismo estudio de optimización que en el caso anterior y comprobando la viabilidad del sistema. Se consigue un sistema libre de errores optimizando los parámetros del sistema. De la misma forma que en el sistema anterior, se optimiza la separación frecuencial de las señales y el ancho de banda de filtro. Se comprueba que la dependencia del sistema respecto la localización
en frecuencia de las señales y el ancho de banda del filtro tiene el mismo comportamiento que el sistema combinada con DPSK.

Una vez las señales y los restricciones en el ancho de banda del filtro están definidas, se comprueba el funcionamiento del sistema 1.28 Tb/s DQPSK-OTDM después de transmitirlo a través de diversos spans de 80 km de SMF y 13.6 km de DCF. Se consigue transmitir la señal a una distancia de 561.6 km.