Master Erasmus Mundus in Photonics Engineering, Nanophotonics and Biophotonics.

Euro photonics

MASTER THESIS WORK

INTENSITY MODULATION AND DIRECT DETECTION ORTHOGONAL FREQUENCY DIVISION MULTIPLE ACCESS DIMENSIONING FOR NEXT GENERATION PASSIVE OPTICAL NETWORKS.

Bruno Alberto Calmels

Supervised by Dr. María Concepción Santos Blanco (Universitat Politècnica de Catalunya, UPC)

Presented in Barcelona, on August 26th, 2012.
Registered at

Escola Tècnica Superior d’Enginyeria de Telecomunicació de Barcelona
Contents

1 Introduction 5

2 Orthogonal Frequency Division Multiplexing for Passive Optical Networks 9
  2.1 Brief history of OFDM ........................................ 9
  2.2 Mathematical formulation of OFDM .......................... 10
  2.3 OFDM system description ..................................... 13
    2.3.1 OFDM transmitter ....................................... 15
    2.3.2 OFDM receiver ........................................... 23
  2.4 O-OFDM for next-generation passive optical networks .... 29

3 Architectures for optical OFDMA-PONs 33
  3.1 Optical carriers .............................................. 33
  3.2 RF up/downconversion and Hermitian symmetry ............... 34
  3.3 Optical double sideband .................................... 36
  3.4 Detection and reconstruction of the OFDM band .............. 36
  3.5 Proposed architecture ...................................... 36

4 Simulations 39
  4.1 Single-user scenario ......................................... 39
  4.2 Multi-user scenario .......................................... 44

5 Point-to-point O-OFDM 47
  5.1 Chromatic dispersion and equalization ....................... 47
  5.2 Direct detection and frequency guard band .................. 48
  5.3 Double sideband and amplitude fading ....................... 50
  5.4 Phase Noise .................................................. 56
  5.5 Frequency chirp ............................................... 58
  5.6 Drive amplitude .............................................. 60
  5.7 Nonlinear effects ............................................ 65

6 Multipoint-to-point O-OFDMA 67
  6.1 Electrical carrier frequencies ............................... 67
  6.2 Optical beat interference and optical carriers separation .. 68
  6.3 Drive amplitude .............................................. 72
  6.4 Dynamic bandwidth assignment and variable received power .. 73
<table>
<thead>
<tr>
<th>Chapter</th>
<th>Title</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>7</td>
<td>Conclusions</td>
<td>79</td>
</tr>
<tr>
<td>8</td>
<td>Future lines of work</td>
<td>81</td>
</tr>
</tbody>
</table>
Chapter 1

Introduction

In the last couple of decades, optical communications have become increasingly more important. The refinement of optical fiber production processes; the invention of Erbium-Doped Fiber Amplifiers (EDFAs) with which much longer distances without electronic regeneration could be reached; and the development of Wavelength Division Multiplexing (WDM) techniques were some of the fundamental facts that established lightwave systems as a mature technology [1]. Nowadays, a vast majority of core and a growing part of metropolitan and access networks are optical.

Driven by the exponentially increasing demand for broadband services, global IP traffic has increased eightfold over the past 5 years, and is expected to increase fourfold over the next 5 years. This means that by 2016, the equivalent of all movies ever made could be transmitted over global Internet every 3 minutes [2]. This phenomenal growth will put tremendous pressure on the underlying communication infrastructure.

In order to cope with the future requirements of optical networks, novel techniques are being studied and applied. A list of them would include advanced modulation techniques, line coding, enhanced forward error correction (FEC) and digital signal processing at the receiver and the transmitter [3]. However, a great majority of the currently existing lightwave systems work based on single-carrier modulation. This is the 'conventional' modulation format that has been the workhorse in optical communications for more than three decades [4]. These techniques include non-return-to-zero (NRZ), return-to-zero (RZ), duobinary, differential phase shift keying (DPSK) and coherent quaternary phase shift keying (QPSK). On the other hand there is multicarrier modulation (MCM), in which the data are carried through many closely spaced subcarriers.

One type of this is Orthogonal Frequency Division Multiplexing, that is fundamentally an approach to high-speed transmission in which high aggregate data rates are achieved by parallel transmission of partially overlapped (i.e. spectrum efficient) lower rate frequency-domain tributaries (or subcarriers)[5]. The subcarriers are so narrow that each, within itself, is practically unaffected by dispersion [6]. This means that dispersion affects each subcarrier only by changing its phase (and possibly the amplitude), which can be corrected by only a complex multiplication in what is called a 1-tap equalizer. In contrast, most single-carrier systems require time-domain equalizers consisting of a weighted sum of delayed versions of the input signal, using multiple taps of a delay line, being much more complex. OFDM has been used extensively in wired and wireless communication systems also because it is an effective solution to inter-symbol interference (ISI) caused by a dispersive channel [7]. The addition of a form of guard band called Cyclic Prefix, avoids this type of interference.

Despite all of these advantages, OFDM did not start to be seriously considered as a modulation option for optical communications until 5 years ago. On the one hand, this happened because there were no important capacity problems that justified the use of complicated modulation and multiplexing formats. On the other hand, there were no fast enough digital signal processors (DSPs) to handle their speed requirements. Some years ago, both of these facts changed, making it possible and necessary to tackle research on more complex modulation formats.
By now, after half a decade of intense research and development, remarkable success has been achieved by many groups throughout the world. Most of them show the immense possibilities of OFDM when applied to long-haul links, i.e. for distances of thousands of kilometers. In the laboratory, capacities above 1 Tbps over 7200 km have been demonstrated employing several Dense WDM channels [6]. Recently, capacity above 100 Tbps over more than 350 km was as well achieved [8].

All these efforts aimed at applying OFDM for developing high-capacity backbone networks [9] are coherent with those of a great part of the R&D community, maybe using different modulation techniques. In order to accommodate the tremendous traffic volume expected to arise in the next few years, it is necessary to have core networks working at capacities never experienced before. This is how nowadays, WDM backbone links at 40 Gbps per channel are widely available and mature [4]. However, a telecom network does not consist of only its backbone, but of metropolitan and access networks as well. While the first is about to start offering terabit-per-second capacities, access networks appear as the major bottleneck for providing broadband services.

Until very recently, the two by far most deployed access technologies worldwide were Digital Subscriber Line (DSL), along with its variants, and broadband access through Cable Television (CATV). They offer end user rates of between 1.5 and 50 Mbps for downlink (from the central office to the users’ premises), and lower for uplink (from the users to the central office). This bandwidth restraint and the up/downlink asymmetry, together with stringent distance limitations, showed that these technologies would not be suitable for the forthcoming traffic growth motivated by emergent applications.

In this context, the trend in access networks is nowadays to bring high-capacity optical fiber closer to the residential homes and small business. The different FTTx models (Fiber to the Home, Fiber to the Curb, Fiber to the Premises, etc.) offer the potential for unprecedented access bandwidth to the users. Depending on which flavor is installed, fiber reaches either the end user’s home or very near to it, from where other technologies like VDSL take over, avoiding distance-related limitations. The number of worldwide FTTH subscribers is projected to reach 100 million by 2013 [10], and most FTTx solutions are based on Passive Optical Networks (PONs) standards. A PON is a point-to-multipoint network architecture that employs unpowered (passive) optical splitters to enable a single optical fiber to serve multiple premises, typically in the order of 16-128. By sharing the same physical medium among many users, the total network cost due to amount of fiber and central office equipment is reduced as compared to point-to-point architectures.

While the already installed PONs successfully provide the user with bitrates of some Gbps, because of utilizing time domain medium access (TDMA), they will certainly reach serious difficulties when moving to higher capacities [11]. Wavelength division multiple access (WDMA) cannot surpass these limitations either [12], besides of being scarcely cost-effective and offering little granularity. Moreover, the employed NRZ modulation format is prone to polarization mode dispersion (PMD) and chromatic dispersion (CD) for greater bandwidths. Therefore, there exists the need to find a novel scheme, able to provide access networks with the desired characteristics that techniques in use nowadays cannot. Recently, OFDM has been regarded as a promising solution for Next-Generation Passive Optical Network (NG-PON) due to its robust dispersion tolerance and the flexibility on both multiple services provisioning and bandwidth allocation [13, 14].

There exist several different options for transmitting and receiving an Optical OFDM signal [15]. Their differences in the electronic domain are based mainly on how they produce the OFDM baseband electrical signal, and on the modulation and detection techniques. While there are interesting ongoing projects creating and modulating the subcarriers in the analog domain [16], these techniques are still at a research stage. In contrast, doing so in
the digital domain is the preferred approach in a vast part of the R&D community. This is due to this leveraging the advances in FFT/IFFT electronics, and pre/post-processing in the digital domain thanks to DSP.

Another stage that brings about a myriad of possibilities is related to the optical domain, and how the electrical-to-optical and optical-to-electrical conversions are performed. The former can be realized by optical amplitude modulation, or by intensity modulation. For amplitude modulation generally a Mach-Zehnder modulator (MZM) is used, while intensity modulation can be as well obtained with MZM, electroabsorption modulators (EAM) or directly modulated lasers (DML). On the other hand, optical-to-electrical conversion can be executed by direct detection (DD) with a photodiode, or by coherent detection, which needs balanced photodiodes and optical hybrids. MZMs have proved to be effective devices, but at this point in time their cost is still one of their disadvantages, and EAM do not prove as cost-effective as DMLs either [17]. As cost-effectiveness is one of the key design requisites in this work, IM by DMLs is employed. Among detection possibilities, a similar argument supports the choice of direct detection, which has been shown to work properly for the bitrates sought in this work [18]. A major part of the last years’ publications on Optical OFDM produce an optical Single Sideband signal. For this purpose, optical filters which eliminate one of the two generated bands can be employed, as most of the modulation techniques yield double sideband signals which consist of a sideband at each side of the optical carrier. But these components are not convenient to have in transceivers thought for PONs, because they depend on the specific wavelength and electrical bandwidths. Therefore, double sideband modulation is the option left.

It would be expensive and complicated to build whole lightwave systems in order to put them or some of their parts under study. The system could cost hundreds of millions of dollars. Besides this, the setbacks related to setting up a real scenario are often intended to be left aside when a proof-of-principle is sought for. Due to this and to the increasing complexity and number of components involved in a photonic system, during the last decades specific simulation tools have been created [19]. The chosen one for the present work is VPItransmissionMaker™, a worldwide reference in the field. It is widely used as an R&D tool to evaluate novel component and subsystems designs in a systems context, investigate and optimize systems technologies (e.g. coding, modulation, monitoring, compensation, regeneration) for short-reach, access, aggregation, core and long-haul applications [20].

This work has the aim of investigating the suitability of OFDM as the modulation format for Next-Generation Passive Optical Networks, and of delivering dimensioning guidelines. As NG-PONs are expected to reach a considerable number of houses and buildings all over the world, cost-effectiveness is one of the primordial drivers of this research. Therefore, directly modulated lasers have been chosen for performing intensity modulation of a directly detected double sideband optical OFDM signal. The advantages, drawbacks, potentialities and dimensioning principles of such a scheme in the scope of NG-PON are to be studied in the present work.

Objectives of the present work:

- Review of Optical OFDM, studying its specific characteristics and design parameters. The advantages and disadvantages compared to its competing technologies should be highlighted and contribution to their understanding should be made, in a general point-to-point case, and particularly for Passive Optical Networks.

- Proposal of a next-generation passive optical network architecture. Based on the review of the basics of O-OFDM and a study of state-of-the-art research and avail-
able techniques, a system should be proposed according to requirements, expected performance and design guidelines.

- Creation of simulation scenarios suitable for studying the diverse aspects and characteristics of Optical OFDM. Flexibility with respect to the different design parameters should be regarded, ideally in a user-friendly manner. The different phenomena affecting the system ought to be distinguishable and separable for their study. If programming code were to be produced or modified, comments should be added, making it as worthy as possible for future modifications or use by people not involved in the present work. At least one point-to-point and one point-to-multipoint scenarios should be created, in order to study the characteristics and assess the performance of each environment.

- Investigation of the phenomena not completely covered by known research, in the specific conditions dealt with here. Aided by the simulation scenarios created, those effects arising in the transmitter, channel or receiver, that may bear special interest should be studied.

- Analysis of the suitability of the proposed Intensity Modulation and Direct Detection OFDM system for Next-Generation Passive Optical Networks. Novel effects stemming from multiple users sharing the communication channel ought to be studied, as well as the effect of different network configurations on system performance. Specific PON features like Dynamic Bandwidth Allocation could be essayed as well.

- Comparison between simulation and laboratory experiments of the worthiest obtained results, regarding the assessment of IM/DD OFDM for NG-PONs suitability.

With these objectives in mind, the rest of the report is divided as follows: Chapter 2 presents a summary of the history of Optical OFDM, introduces the basic concepts that will be needed in future chapters, explains the features and design parameters typical of an OFDM system, mentions some of its advantages compared to other techniques, analyzes the theoretical strengths of Optical OFDM for PONs; Chapter 3 provides insight into the NG-PON architecture proposed in this work; Chapter 4 covers issues concerning to the simulation software used, as well as the developed scenarios and code, and explains the design parameters chosen for our simulations; Chapter 5 analyzes (formally and through simulations) the phenomena affecting an intensity modulation and direct detection optical OFDM system like the proposed and suggests ways of avoiding or alleviating their effects, from a system design point of view; Chapter 6 researches into phenomena intervening due to the multipoint-to-point nature of a PON, explains ways to combat them, mentions the dimensioning principles taken in our proposal, and shows the results of PON techniques put into practice both in simulation and laboratory; Chapter 7 finally points out the conclusions of this work.
Chapter 2

Orthogonal Frequency Division Multiplexing for Passive Optical Networks

2.1 – Brief history of OFDM

Some of the most important milestones in the development of OFDM can be observed in Fig. 2.1. Its history goes back to 1966, when a patent by Chang of Bell Labs [21] suggested for the first time the utilization of a multiple-carrier modulation system, with the particularity that the subcarriers were orthogonal among them. After this, in 1969, the idea of using FFT to generate the subcarriers came along [22]. An important addition to the technique, the cyclic prefix, which provides with some of its key advantages, came about in 1980 [23]. During the 1980’s, OFDM got increasing attention, as applications in the radio broadcasting [25], mobile communications [26] and wireline [27] were put forward. This last meant the invention of digital subscriber loop (DSL), widely used in access networks. Since 1995, OFDM has been adopted as modulation technology of many standards and a broad range of applications. Nowadays, various standards like Wi-Fi (IEEE 802.3a/g), WiMAX (IEEE 802.16e), asymmetric digital subscriber line (ADSL; ITU G.992.1), Long Term Evolution (LTE) for mobiles and many of the European and Asian terrestrial television and radio broadcasting standards employ OFDM modulation.

Despite its success in other communication areas, OFDM did not attract much attention for optical communications until the beginning of this century. Although the first paper suggesting its use in the optical domain appeared in 1996 [28], it was not before 2001 that its main advantage of combating dispersion was noticed as a critical feature [29]. One optical implementation of OFDM had been reported in 2002 [30], but it apparently did not attract much interest, as not many publications followed it. However, some years after this, a few groups had already reported different digital-processing implementations of optical OFDM featuring dispersion compensation [31, 32, 33]. After these pioneering publications, and the work following them, which addressed some of the key questions about bringing OFDM...
to the optical domain, groups around the world working on the topic started multiplying. As a consequence, Optical OFDM is currently being considered as a serious contender for 100+ Gbps, and there have been demonstrations of extremely high speed and high spectral efficiency transmissions, reaching the Tbps regime [34].

Some years after the interest in OFDM for long-haul systems emerged, it started being seriously considered as a possible constituent of access networks, and the amount of research and publications on the topic massively increased. In 2007 some of the first demonstrations of such systems appeared [35]. Since then, OFDM for passive optical networks has become a very hot topic of research, being investigated by several groups of academia and industry across the world.

2.2– Mathematical formulation of OFDM

Multicarrier modulation techniques employ several carriers, each of which is modulated by signals containing information to be transmitted. If an analogy with music may be done, this resembles to playing several notes (frequencies) at the same time (this could be considered as a "chord"), as opposed to single-carrier modulation, which plays only one note at a time, as can be clarified by Fig. 2.2.

![Figure 2.2: Musical analogy of modulation types. Top: Single-carrier. Bottom: Multi-carrier. [16]](image)

A generic multicarrier modulation and demodulation diagram can be seen in Fig. 2.3.

![Figure 2.3: Conceptual diagram of a multicarrier modulation/demodulation system. [4]](image)
It also shows an IQ modulator/demodulator, which performs a complex multiplication of the incoming signals. The signal $s(t)$ input to the channel is represented as:

$$s(t) = \sum_{i=-\infty}^{+\infty} \sum_{k=1}^{N_{sc}} c_{ki} s_k(t - iT_s)$$ (2.1)

$$s_k(t) = \Pi(t) e^{j2\pi f_k t}$$ (2.2)

$$\Pi(t) = \begin{cases} 
1 & \text{if } 0 < t \leq T_s \\
0 & \text{if } t \leq 0 \text{ or } t > T_s 
\end{cases}$$ (2.3)

where $c_{ki}$ is the $i^{th}$ information symbol at the $k^{th}$ carrier, $s_k$ is the waveform for the $k^{th}$ carrier, $N_{sc}$ is the number of carriers, $f_k$ is the frequency of the carrier, $T_s$ is the symbol period, and $\Pi(t)$ is the pulse shaping function.

The reception part consists of a bank of matched filters, which perform a correlation between the incoming signal and the same set of carriers used for modulation. The output of each of these filters are the detected information symbols $c'_{ki}$, given by:

$$c'_{ki} = \frac{1}{T_s} \int_0^{T_s} r(t - iT_s) s_k^* dt = \frac{1}{T_s} \int_0^{T_s} r(t - iT_s) e^{-j2\pi f_k t} dt$$ (2.4)

where $r(t)$ is the received time domain signal, $c'_{ki}$ are the estimates of the information symbols $c_{ki}$ and the * operator means complex conjugation. If the channel does not produce changes on the signal, $r(t)$ is just the sum of all the modulated carriers.

Classical multicarrier systems generate the carriers in the analog domain, obtaining bandlimited signals for each of the carriers, whose spectra do not overlap. This implies the need for a great number of oscillators and filters at both the transmitter and the receiver sides. Cost-effective oscillators and filters need a frequency guard band between channels, which means low spectral efficiency, as can be observed in Fig. 2.4.

![Figure 2.4: Spectrum of classical multicarrier modulation. [5]](image)

In this aspect, OFDM introduces the idea of using overlapping yet orthogonal sets of carrier frequencies. The orthogonality property originates from the correlation between any pair of carriers, given by:

$$\delta_{kl} = \frac{1}{T_s} \int_{-\infty}^{+\infty} s_k s_l^* dt = \frac{1}{T_s} \int_0^{T_s} \exp(j2\pi f_k t) \exp(-j2\pi f_l t) dt$$

$$= \frac{1}{T_s} \int_0^{T_s} \exp(j2\pi(f_k - f_l) t) = \exp(j\pi(f_k - f_l)T_s) \frac{\sin(\pi(f_k - f_l)T_s)}{(\pi(f_k - f_l)T_s)}$$ (2.5)

So it can be seen from the exponential term that if the condition:

$$f_k - f_l = m \frac{1}{T_s}$$ (2.6)

being $m$ an integer, is fulfilled, for $k \neq l$ the correlation $\delta_{kl} = 0$, and then the two subcarriers are orthogonal to each other\(^1\). This implies that these orthogonal subcarriers sets, with their

\(^1\)From now on, the term 'subcarrier' is used for what have been called 'carrier' until this point. There is no implication behind this, except to follow OFDM bibliography, which majorly uses the term subcarrier.
frequencies spaced at multiples of the inverse of the symbol period, can be recovered with
the matched filters as seen in Fig. 2.3, without intercarrier interference (ICI), despite strong
signal spectral overlapping [4].

An OFDM signal of 4 orthogonal subcarriers ($N_{sc}=4$) can be observed in Fig. 2.5 for
a single symbol. The effect of the pulse shaping function $\Pi(t)$ in the frequency domain is
to give the subcarriers a shape of a Sinc ($\sin(x)/x$) function. Note that the orthogonality
property means that the subcarriers’ maxima coincide with the zeros of the rest. A trail of
symbols is sketched in the last picture of Fig. 2.5.

$$s_m = \sum_{k=1}^{N} c_k e^{j2\pi f_k (m-1)T_s/N}$$  \hspace{1cm} (2.7)
where \( t \) has been replaced by the sampling time \( \frac{(m-1)T_s}{N} \). In order to fulfill the orthogonality condition (Eq. 2.6) the chosen subcarriers frequency differences need to be integer multiples of \( 1/T_s \), which can be accomplished by setting that:

\[
f_k = \frac{k - 1}{T_s}
\]

and then substituting into Eq. 2.8:

\[
s_m = \sum_{k=1}^{N} c_k e^{j2\pi \frac{(k-1)(m-1)}{N}} = \mathcal{F}^{-1}\{c_k\}
\]

where \( \mathcal{F}^{-1} \) is the Inverse Fourier Transform and \( m \in [1, N] \). Performing alike for reception, we arrive at:

\[
c_k' = \mathcal{F}\{r_m\}
\]

where \( r_m \) are the received signal samples at every interval \( T_s/N \) and \( \mathcal{F} \) is the Fourier Transform. The result of Eqs. 2.9 and 2.10 is that the discrete values of the transmitted signal \( s(t) \) are given by the N-point IDFT of the information symbol \( c_k \). At the receiver, the information symbols \( c_k \) can be obtained as a N-point DFT of the received signal samples \( r_m \). This represents a fundamental advantage, since DFT/IDFT calculations can be performed through very efficient algorithms called fast fourier transform (FFT) and inverse fast fourier transform (IFFT) in digital signal processors (DSP) that are massively produced and are thus affordable. These processes can be observed as fundamental blocks in the transmitter and receiver sketched in Fig. 2.6 and 2.7 in next section. Moreover, this avoids the need for several complex filters or oscillators. Having presented the mathematical formulation of OFDM, let us move towards the explanation of the processes involved in an OFDM system.

2.3– OFDM system description

There are several different flavors of OFDM, that differ in many stages of the transmission and reception systems. As already mentioned, one of the source of differences is the way in which the subcarriers are produced. However, we do not discuss those systems that generate them in the analog domain, because of the advantages that digital techniques provide for this task nowadays, without discarding that at some point in the future analog techniques may present tough competition to the digital ones. Therefore, the main differences among different design principles that we deal with here stem from the electrical-to-optical and optical-to-electrical conversion stages. In consequence, before moving to these, we can explain the OFDM transmitter and receiver in the electrical domain, which are fairly constant regardless of what is done in the optical domain. In order to do this, we first present in Fig. 2.6 and 2.7 the block diagrams of the electrical parts of a transmitter and a receiver.

The transmitter is input with a bit stream, which is divided in sets of bits which are parallelized. These parallel bits are mapped onto symbols whose form depends precisely on the mapping scheme. After this, training symbols are added, which are entire OFDM symbols that are known by both transmitter and receiver, and are used for synchronization and channel estimation. After this, the IFFT is calculated, using the mapped symbols for the different subcarriers. Zeros may be added depending on the chosen design. After this, a parallel-to-serial conversion is performed, cyclic prefix is added in order to alleviate dispersive effects and avoid interference. Finally, the real and imaginary parts of the signal are usually separated and turned into analog signals with the digital-to-analog converters.
Orthogonal Frequency Division Multiplexing for Passive Optical Networks

(DACs). After unwanted aliasing components are filtered out from the analog signals, these may be treated in different ways, according to the optical modulation to be used.

In the receiver, most of the stages execute reverse processes of those in the transmitter. The first two stages are a low pass filter that may be needed to eliminate unwanted components from detection, and posterior analog-to-digital converters. In the case of the receiver, there appear some steps in the electrical domain that may vary depending on the optical demodulation. One example is the following block, phase noise compensation (carrier recovery), which is carried out when coherent detection is done, but not for direct detection. The next one, symbol synchronization, is a very important one, since the receiver has to detect and synchronize its FFT window with the symbol as the transmitter sent it, in order to perform correct decoding. After this, the cyclic prefix is removed, and the signal converted from serial to parallel in order to calculate the FFT and recover the subcarriers separately. The zeros that were added in the transmitter are next eliminated, as well as the training symbols, which are however used for synchronizing and estimating the channel response. The estimation is used later in the following step for equalization. Other type of phase noise compensation can be applied at this point, before demapping and parallel-to-serial conversion in order to finally get the demodulated bit stream.

In the following sections, these summarized processes are discussed in higher detail. This is done because great insight into most of them is needed. There are some which we need to understand in order to be able to design and dimension the related parameters, as well as to program code emulating them for the simulation environment. Others, we need to be able to compare, in order to decide which are the most suitable for our objectives.
2.3.1 OFDM transmitter

Serial-to-parallel conversion and mapping

A serial bit sequence enters first a serial-to-parallel converter that builds blocks of \( M \times N \) bits. Here, \( N \) is the number of subcarriers to be modulated with information, and \( M \) is the number of bits with which a single subcarrier is to be modulated in each symbol. \( M \) depends on the mapping process, which is the next step. It is in the mapping when information symbols \( c_{ki} \) (Eq. 2.1) are obtained.

Mapping represents the very step where information is encoded in each of the subcarriers. In OFDM, each subcarrier can be modulated in a variety of ways, which include amplitude shift keying (ASK), phase shift keying (PSK), quadrature amplitude modulation (QAM), and some others. Fig. 2.8 shows examples of different mappings that could be applied to the subcarriers.

![Figure 2.8: Different modulation formats that can be applied to the subcarriers in the mapping step. The carrier in red is modulated with ASK, the orange and green ones with certain PSK scheme, and the blue one is not changed by the modulation, which can happen in most modulation formats.][16]

There is no specific requirement for the mapping scheme to be the same for every subcarrier; in fact, state-of-the-art research has begun to use adaptive bit-loading algorithms that tend to maximize spectral efficiency by dynamically set the number of bits used for each subcarrier based on channel estimations \([38, 39, 40, 41]\). However, in a major part of the present work, QPSK modulation is used for all the subcarriers. The constellation of this modulation format, also known as 4QAM, can be seen in Fig. 2.9, together with other formats.

Depending on the chosen mapping scheme, the information symbol \( c_{ki} \) will take different forms. For instance, if ASK is utilized, \( c_{ki} \in (0, 1] \) and is a real number, with BPSK \( c_{ki} \in \{1; -1\} \), and in the case QPSK is used, \( c_{ki} \in \{1 + j; 1 - j; -1 + j; -1 - j\} \) being a complex phase with constant amplitude. This is shown in Fig. 2.10.

Training symbols

The next stage after mapping is the addition of training symbols. These are symbols that both the transmitter and the receiver know beforehand, which upon detection are used for synchronization and channel estimation. They are sent periodically, every a certain
number of 'payload' (useful, carrying information) symbols. This number depends mainly on channel dynamics. When the receiver notices that it has received a training symbol, as it is able to compare the received symbol with that it should have received, channel estimation is possible by calculating how it has altered the training symbol. A common practice is to use two OFDM symbols that are identical. Through correlation processes, this can be used for synchronization between the transmitter and the receiver.

**IFFT**

Already having a block of \( N \) parallel complex coefficients \( c_{ki} \), the **IFFT calculation** is done, following the results found in Eq. 2.9. Some of the IFFT module inputs may be set to zero. The reason for this to be done is that the DACs yield alias bands at both sides of the OFDM band generated by the IFFT. If all the input pins of the IFFT are used, it is impossible to separate the wanted OFDM band and the unwanted aliasing components using a real filter. This is illustrated in Fig. 2.11. In order for this process to be satisfactory, the unused inputs should correspond to the subcarriers located at the edges of the OFDM band, and the process is called **oversampling**. This is important to be understood at this point, hence some issues related to the IFFT calculation need to be addressed. The process of putting certain subcarriers to zero is done in the by a feature called **zero padding**, which can be defined as the vector of subcarriers set to 0. In this work it is closely related to the simulation environment, and we come back to this in Sec. 4.2 in order to explain some issues for multi-user scenarios.

IFFT modules’ inputs correspond to \( N \) equally spaced frequencies in the range \([-f_{Nyq}, +f_{Nyq}]\). \( f_{Nyq} \) is called **Nyquist frequency** and is the maximum frequency a DAC can modulate (see Fig. 2.12). Normally, the input vector of coefficients \( \{y_i\} \) in Fig. 2.12 is numbered...
2.3 OFDM system description

Figure 2.11: Aliasing components appearing separated from the OFDM band due to oversampling. In this case, an electrical filter can eliminate the alias. [16]

starting from the DC channel, i.e. the coefficient modulating the 0-frequency subcarrier. This channel is usually not used for modulation, since it coincides with RF carriers normally introduced before optical modulation, and because it would impose stringent requirements on the low-pass characteristics of optic and electronic components. The following input vector elements modulate subcarriers of increasing frequencies, until element number \( N/2 + 1 \), which corresponds to the Nyquist frequency and is located on both edges of the OFDM band. Subsequent input vector elements correspond to negative frequencies of decreasing absolute value. This means, for example, that the second input coefficient modulates the subcarrier at frequency \( f = 1/T_s \), which is the lowest positive frequency one. The last coefficient, similarly, modulates the subcarrier at the frequency immediately to the left of the DC one, i.e. \(-f = -1/T_s\), where \( T_s \) is the symbol duration. The subcarriers to be left in 0 are therefore the Nyquist frequency one (input \( N/2 + 1 \)) and, if needed, the ones immediately above and below it.

Figure 2.12: IFFT input vector elements and their corresponding frequencies. [16]

The output of the IFFT module is a vector of parallel values that correspond to samples of the signal to be transmitted, in the time domain. These parallel values are then input to a parallel-to-serial converter in order to get a proper signal to transmit. After this, the cyclic prefix is added. This is a process of great importance in an OFDM system.

Cyclic prefix

Due to chromatic dispersion in the fiber, different wavelengths have different group velocities. This means that there exists a temporal delay among the subcarriers. Let us for the sake of simplicity only consider two subcarriers suffering a delay spread \( t_d \) (sometimes also called walk-off). This is the same as considering several subcarriers with the two taken here being the "fast subcarrier" and the "slow subcarrier", whence these would be the
fastest and slowest respectively of all the subcarriers. Fig. 2.13.A shows two consecutive OFDM symbols upon transmission. Both subcarriers are then aligned. Fig. 2.13.B shows the situation at reception, where the delay spread $t_d$ has appeared. The FFT window (for the FFT calculation needed for demodulation) has been chosen to be aligned with the fast subcarrier. In this case, the slow subcarrier has crossed the symbol boundary, so leading to **inter-symbol interference** (ISI). Moreover, for the slow subcarrier, that has moved into the adjacent OFDM symbol, the fundamental orthogonality condition (Eq. 2.6) is not fulfilled anymore because it is not the same waveform during the complete symbol. This leads to **inter-carrier interference** (ICI), meaning that this subcarrier's zeros do not coincide with the other subcarriers' maxima.

![OFDM symbols without cyclic prefix](image)

Figure 2.13: OFDM symbols without cyclic prefix. A: At the transmitter. B: At the receiver. [4]

The solution to these two problems comes with the cyclic prefix. It consists on adding a cyclic extension, i.e. extending the same waveform, for a duration $\Delta G$, where the subscript $G$ is related to it being a form of 'guard interval'. It is usually performed by copying the last part of each OFDM symbol to the beginning of it, as done in Fig. 2.14.A for the slow subcarrier (although in reality it is done for the whole serial signal, thus for all subcarriers simultaneously). Fig. 2.14.B shows the situation at reception. Now, a complete OFDM symbol for the slow subcarrier is also maintained inside the same FFT window used before. This is achieved because a part of the cyclic prefix has moved inside the window to replace the part that has shifted out. As a consequence, the OFDM symbol for the slow subcarrier is almost identical to the transmitted one, except for an additional phase shift. This phase shift is dealt with by channel estimation and is subsequently removed before symbol decision. Notice that the FFT is not calculated over the entire OFDM symbol duration anymore, but over a shorter duration $t_s$. So, there is a portion of the symbol that is discarded.

The condition for ISI-free transmission can be then stated from analyzing Fig. 2.14. It is clear that the cyclic prefix allocated should be larger than the walk-off:

$$t_d < \Delta G$$

(2.11)
If the fiber dispersion coefficient is known, $t_d$ can be calculated. Another phenomenon that the cyclic prefix alleviates is differential group delay (DGD), that is the difference in propagation time between the two principal polarization states in the fiber, and can cause additional walk-off. If no polarizers are introduced in the system and direct detection with a single photodiode is used, the maximum DGD can be estimated by the PMD (polarization mode dispersion) fiber coefficient and the fiber length [1]. We can then calculate the cyclic prefix to be allocated as:

$$\Delta_G > \frac{DLBc}{f^2} + DGD_{\text{max}}$$

(2.12)

where $D$ is the fiber dispersion coefficient, $L$ its length, $B$ is the total bandwidth of the OFDM signal, $f$ its central frequency, $c$ is the speed of light and $DGD_{\text{max}}$ the maximum DGD [16]. The first term in the right hand side corresponds to chromatic dispersion, and the second to PMD. It is convenient at this point to anticipate that PMD is considered here for completeness, but it is not taken into account in following sections of the study, because it has been proved that PMD is negligible for lengths under 100 km, i.e. passive optical networks reach [12].

An elegant way to formulate the cyclic prefix is to use the transmitted signal $s(t)$ as stated in Eq. 2.1 but changing the pulse shape function as:

$$\Pi(t) = \begin{cases} 1 & -\Delta_G < t \leq t_s, \\ 0 & t \leq -\Delta_G \text{ or } t > t_s. \end{cases}$$

(2.13)

This represents an OFDM symbol as seen in Fig. 2.15.

After the addition of the cyclic prefix, digital-to-analog converters are used to get a continuous-time signal $s(t)$ from the coefficients $s_m$. As mentioned when oversampling was treated, DACs produce aliasing components at the sides of the OFDM band. Therefore, low-pass filters are used for eliminating them. The number of DACs and LPFs needed depend on the modulation scheme to be used.
Electrical-to-optical conversion

This process, which may be as well called simply optical modulation, is where the signal is taken from the electrical to the optical domain, and its details need to be considered carefully, because it has important effects on the performance of the whole system and the phenomena that should be taken into account. In Optical OFDM, there are two possible types of modulation. The two of them along with different options within them can be observed in the inverted tree diagram in Fig. 2.16. One is intensity modulation (IM) and the other amplitude modulation (AM). The former implies that the optical intensity is linearly modulated by the OFDM signal $s(t)$. There are several devices to perform this kind of modulation. The three most commonly used are Mach-Zehnder Modulators (MZM; as in [42]), Electroabsorption Modulators (EAM; as in [43]) and Directly Modulated Lasers (DML; as in [44]). In amplitude modulation, the optical field is linearly modulated by the OFDM electrical signal (this is why it is also usually called field modulation). In order to modulate the optical field, an IQ Mach-Zehnder Modulator (also called Complex Modulator, Super Mach-Zehnder or Cartesian Modulator) can be used. A simple MZM at null-point may be employed as well.
2.3 OFDM system description

Optical intensity is a real and strictly non-negative quantity. Consequently, if intensity modulation is to be performed, the signal with which it is modulated needs to be as well real and positive. Let us remind at this point that the output of the IFFT module are complex coefficients, which after parallelization and addition of a cyclic prefix remain being samples of a theoretical complex continuous-time signal to be transmitted. A way of obtaining a real and positive signal is needed then.

Two techniques have been put forward for getting a real-valued signal. One consists in enforcing Hermitian symmetry at the IFFT output. Hermitian symmetry means that the negative and the positive frequency components are complex conjugates of each other, as shown in Fig. 2.17. This causes the IFFT output (time-domain coefficients) to be real, and it is what is done e.g. in ADSL, so that the baseband signal can be transmitted. The other technique is to perform RF upconversion, also called electrical IQ (in-phase and quadrature) multiplexing. Normally, the real and imaginary parts of the IFFT output are accessible separately. The real part of \( s(t) \) (which is called in-phase component, \( I(t) \)) is multiplied by a cosine signal of certain frequency \( f_{RF} \), and it is done alike with the imaginary part (to be called quadrature, \( Q(t) \)) by a -sine signal, and then both summed, as follows:

\[
s'(t) = I(t) \cos(2\pi f_{RF} t) - Q(t) \sin(2\pi f_{RF} t)
\]

(2.14)

An OFDM transmitter featuring electrical IQ multiplexing can be observed in Fig. 2.18. The result of multiplying a signal by a cosine or a sine function, which is called frequency upconversion, is that two replicas of the signal appear centered at \( \pm f_{RF} \) and attenuated by half each, as proved in the following:

\[
y(t) = I(t) \times \cos(2\pi f_{RF} t) \quad \Rightarrow \quad Y(f) = \mathcal{F}(f) * \frac{1}{2} (\delta(f - f_{RF}) + \delta(f + f_{RF}))
\]

(2.15)

\[
Y(f) = \frac{1}{2} (\mathcal{F}(f - f_{RF}) + \mathcal{F}(f + f_{RF}))
\]

(2.16)

where \( * \) is the convolution operation, \( \mathcal{F}(f) \) is the Fourier Transform of \( I(t) \) and \( Y(f) \) is the Fourier Transform of the result of the upconversion. When multiplying by a sine function, something similar occurs:

\[
z(t) = Q(t) \times \sin(2\pi f_{RF} t) \quad \Rightarrow \quad Z(f) = \frac{1}{2j} (\mathcal{F}(f - f_{RF}) - \mathcal{F}(f + f_{RF}))
\]

(2.17)

where \( \mathcal{F}(f) \) is the Fourier Transform of \( Q(t) \) and \( Z(f) \) that of the product. \( f_{RF} \) has to be chosen such that the two copies of the OFDM band do not overlap. That is to say that \( f_{RF} \geq BW/2 \), being BW the whole OFDM band width.

Hermitian symmetry imposition has an important practical advantage, that is to avoid the usual problem of IQ imbalances, which occur when the cosine and sine signals are not perfectly orthogonal. This causes that at reception \( I(t) \) and \( Q(t) \) mix with each other, therefore it has to somehow be fixed. Besides its simplicity from an analog component point...
of view, Hermitian symmetry technique excels in needing only one DAC\(^2\). On the other hand, the condition imposed causes spectral efficiency and total data rate to be reduced by half, because one side of the OFDM band is completely determined by the values of the other side. Because of this also, 1/4 of the IFFT and 1/2 of the DAC bandwidth are used for data. As opposed to this, using electrical IQ multiplexing both the IFFT and the DACs are fully used. A drawback compared to the previous technique is that it needs two DACs\(^3\).

If amplitude modulation is to be exploited, there are two possible devices to profit from. A normal MZM biased in its null point can be used. In this case, as it can only employ a real modulating signal, one of the two techniques just explained needs to be used (either HS or electrical IQ prior to optical modulation). On the other hand, IQ MZMs\(^4\) can be fed with a complex modulating signal (modulating the real and imaginary parts with one MZM each, in quadrature between them), so no previous process needs to be applied.

When optical modulation is performed, the OFDM spectrum and its complex conjugate, which is mirrored with respect to the optical carrier, are observed in the optical spectrum (see Fig. 2.19). This implies, as seen in the right hand picture of Fig. 2.19, a real signal. This kind of optical modulation is called **optical double sideband modulation**, precisely due to the presence of these two bands, at each side of the carrier. A more detailed study of intensity modulation is to be done in Chap. 5.

Double sideband modulation entails some disadvantages that are discussed in Sec. 5.3 and would be desirable to avoid. In its place, **optical single sideband modulation** (OSSB) would be useful to obtain. What is done for this matter is to perform double sideband modulation and after it a band-pass filter used to eliminate one of the bands, as shown in Fig. 2.20. In order for this to be possible with real optical filters, the two sidebands need to be separated in frequency. If electrical IQ multiplexing is performed, usually the

\(^2\)Many reference authors in the field note this characteristic of HS; however, as it transmits half of the payload information, it is not so much a real advantage.

\(^3\)The same comment as before applies here, since in this case, double the information is being transferred.

\(^4\)Which is basically a 'super' MZM containing one MZM in each of its arms, and them being in phase quadrature.
bands would be separated from the optical carrier, thus between them as well. This can be achieved in the Hermitian symmetry approach by letting some low frequency subcarriers be zero. The drawback of this approach is that an optical filter is needed, which diminishes the transmitter flexibility since the filtering needs to be done for specific frequencies. This would mean that $f_{RF}$ must be constant if electrical upconversion is performed, as well as the bandwidth.

Although the mathematical formulation of OFDM done in Sec. 2.2 has not implied at any point the strict need of the subcarriers to be generated in an electronic DSP, we have only treated that case, due to its aforementioned advantages. However, there are other ways of generating the subcarriers by performing the IFFT analogically, entirely in the optical domain. This approach is commonly called all-optical OFDM [5] while it may also receive the name of coherent WDM [36]. A block diagram of a generic all-optical OFDM transceiver is displayed in Fig. 2.21. A comb of unmodulated phase-locked orthogonal subcarriers is first produced by a tone generator, such as a pulsed laser. Then, these phase-locked tones are separated by means of an array waveguide grating (AWG) after an optical IQ modulator modulates each individual subcarrier in a manner defined by the mapping scheme. After this parallelized modulation, the tones are rejoined by another AWG and transmitted over the fiber. At reception, the tones are again separated and processed parallely.

Recently, using all-optical OFDM, bit rates as high as 26 Tbps have been obtained [46]. Nevertheless, its overall complexity, mainly at the receiver side, makes it prohibitive for PONs for the time being. Consequently, and as has already been explained that it is our objective to propose the most convenient techniques for next-generation passive optical networks, we shall not come back to all-optical OFDM in the present work.

### 2.3.2 OFDM receiver

In order to follow the logic order of steps in the communication system as we have done so far, in this section we deal first with the optical-to-electrical conversion and its different possibilities, and then move on to giving a detailed explanation of the electrical OFDM
receiver.

**Optical-to-electrical conversion**

In optical OFDM there are basically two possible detection approaches. These are direct detection (DDO-OFDM) and coherent detection (CO-OFDM). The first one is the simplest detection technique, and consists in inputting the optical signal to a photodetector. A thorough mathematical formulation of direct detection is done in Sec. 5.2, so it is skipped here, but some of its requirements and characteristics are mentioned in order to compare it with the other approach. Fig. 2.22 depicts direct detection process, showing the signals at different points. The photodiode, which acts as an optical-to-electrical converter is a square-law detector, because its output current is proportional to the square of the incoming optical field amplitude modulus. This, as is proved in Sec. 5.2, imposes the need of a frequency guard band between the optical carrier and the OFDM band in order to avoid interference from mixing products.

In addition, direct detection requires an optical carrier. This happens because the detection of the signal is originated in the beating of the components with each other, because of the quadratic nature which means a convolution in the frequency domain. Therefore, the OFDM band as well as the optical carrier need to be present at detection. This is seen in the mathematical formulation in Sec. 5.2. Research studies have concluded that an optimum level of optical carrier power is as much as that of the OFDM band [47]. This fact increases the O-OFDM signal vulnerability to OSNR (optical signal-to-noise ratio) degradations, as there is less power in proportion in the "useful" band that contains information for a constant transmitted power.

---

**coherent detection** (CO-OFDM) is intrinsically more complex than direct detection since it needs more components, however yielding some advantages. For CO-OFDM, there is no need of an optical carrier to be sent together with the OFDM band from the transmitter, since at reception a local oscillator (LO) is used to produce an optical carrier that is mixed with the incoming signal. The frequency of this added carrier can be equal to the transmitter carrier, in which case the technique is called homodyne, or different in the heterodyne
version. Fig. 2.23 shows these two variants. Homodyne CO-OFDM can be thought of as the reverse process of modulating with an optical IQ MZM. By means of the 90° optical hybrid and balanced photodetectors, the OFDM band can be recovered completely, i.e. its amplitude and phase information can be retrieved. Heterodyne reception normally locates the LO carrier at one of the sides of the OFDM band, so after detection at the photodiode electrical IQ downconversion needs to be done in order to get the baseband OFDM signal.

![Diagram showing homodyne and heterodyne reception](image)

Figure 2.23: Two possibilities for coherent detection OFDM. [16]

As CO-OFDM does not require a transmitter-side optical carrier, all the power can be allocated in the OFDM sideband. This makes it more resilient to OSNR degradation than DDO-OFDM. Moreover, as a linear capture of the full optical field is done, the mixing products that cause direct detection techniques to need a frequency guard band no longer appear, which improves the spectral efficiency of coherent reception. Another advantage of this technique is that optical phase information is preserved, as opposed to direct detection. Moreover, a serious drawback of optical double sideband modulation as frequency selective amplitude fading (see Sec. 5.3) does not take place. Furthermore, this technique increases receiver sensitivity [5].

Many advantages of CO-OFDM have been mentioned, compared to DDO-OFDM. Nevertheless, it still is a much more complicated technique, which usually needs for three bias voltages to be adjusted, requires phase noise compensation and complex and costly components. From a PON point of view, as the receiver needs to employ very cost-effective hardware, coherent detection may not be the optimum choice. As it is explained in Chapter 3, another fundamental cause for opting for DD is that it enables detection at low frequencies of bands far away in the spectrum, by beating each with its corresponding optical carrier.

**Electrical domain**

Now we go on with the description of the electrical part of an OFDM receiver. Its block diagram is shown again in Fig. 2.24 for comfort of the reader. In an overall view, most of the steps are inverse to those in the transmitter. The photocurrent that is output from the detector is converted into a voltage signal by a transimpedance amplifier (sketched in Fig. 2.22 after the photodiode). After this, RF filtering is applied (seen as a capacitor in Fig. 2.22) in order to eliminate the DC component caused the detection of the optical carrier.

In some cases, depending on the kind of modulation and detection performed, RF downconversion may be needed next. If RF upconversion has been applied, this step is necessary after direct or heterodyne detection, in order to recover the baseband OFDM signal located in frequency as after the transmitter’s DAC. RF downconversion is an inverse
process to the upconversion, in the sense that the signal is multiplied by a cosine and a sine functions in an electrical IQ mixer, as shown in Fig. 2.25. By doing so, the real and imaginary components of the OFDM signal can be recovered independently. Though RF downconversion permits full use of FFT and ADC bandwidths for data treatment, it depends on an electrical IQ mixer and 2 analog-to-digital converters. This can bring about unwanted complexity at the receiver.

There is an alternative way of demodulating the OFDM band without employing RF downconversion. It is the inverse of leaving unmodulated subcarriers near DC in the transmitter. So, it consists in inputting the signal as it is detected to the DSP part, filtering the unwanted mixing products, and demodulating it with the FFT, as if the low frequency subcarriers have been left unused. This is shown in Fig. 2.26. As no IQ demultiplexing is done, only a real signal can be demodulated in this way, which means that the OFDM sideband possesses Hermitian symmetry, as seen in Fig. 2.26. As opposed to its simplicity from an analog component point of view, regarding digital component perspective it is not efficient, because only 1/4 of the FFT and 1/2 of the ADC bandwidths are exploited for data treatment.

ADC, PNC and symbol synchronization

The next steps to be done after having the OFDM sideband spectrally located where it is intended, in the electrical domain, analog-to-digital conversion is performed. If the method enables to have the real and imaginary parts separately, at this point they are...
2.3 OFDM system description

summed, and the subsequent digital processing modules deal with complex numbers.

Afterwards, phase noise compensation (PNC) can be performed. This is also known as carrier recovery, since its objective is to track the transmitter’s laser phase noise (random phase shifts) in order to correct for that phase difference in the LO. This is a process done in CO-OFDM systems, since in DDO-OFDM systems there does not exist a LO. Because of this, phase noise affects the latter in a different way than the former ones. Phase noise compensation is a complex task, and is currently under intense research, with several techniques being proposed and studied [48]. As CO-OFDM is not used in system design and dimensioning, this topic is not dwelt upon here. Phase noise effects on DDO-OFDM systems are studied in detail in Sec. 5.4.

The following procedure is symbol synchronization. It is mandatory that the receiver FFT window is properly synchronized with the OFDM symbols, because otherwise inter-symbol and inter-carrier interference take place, similarly as it happens when no cyclic prefix is added to the symbols in a dispersive link of considerable length (different symbols are mixed and the orthogonality condition is no longer fulfilled; see Sec. 2.3.1). The training symbols are usually used for this purpose. When working with simulation software, symbol synchronization is not obvious, since dispersion makes some subcarriers to arrive at the receiver first, as explained in Sec. 2.3.1. This is solved by the cyclic prefix in the majority of OFDM symbols, but due to the numerical methods underlying the software operation, the last transmitted symbols may mix with the first, and vice versa. This may be solved with a smart choice of reference frequency.

Cyclic prefix removal, S/P conversion and FFT

Once the receiver is synchronized to the OFDM symbols correctly, the next thing to do is cyclic prefix removal. Fig. 2.27 shows symbols in reception in back-to-back (without fiber) and after some length of dispersive fiber. Striped in gray are the the ranges in which the FFT window can start in order to completely avoid ISI and ICI, which means to only let into the window waves of the desired OFDM symbol. It is clear that in the first case there is a wider range, while as dispersion turns greater this range narrows, because the walk-off among subcarriers also grows (see Sec. 2.3.1). The ideal synchronization is then to start the window in the middle of the dashed range.

As the following process is to demodulate the already synchronized OFDM symbols with the FFT module, it is necessary to parallelize the serial sample sequence in a serial-to-parallel converter. At the output of the FFT, the received estimates $c_{ki}^r$ of the information symbols are obtained for each subcarrier $k$ and the symbol $i$, as studied in Eq. 2.9 in Sec. 2.2.
TS extraction, channel estimation and 1-tap equalization

The subsequent blocks in Fig. 2.24 are training symbols removal, channel estimation and 1-tap equalization. The three of them are related, since the training symbols are used for channel estimation, which is then used for compensating for channel distortions in the 1-tap equalizer. As commented before, the training symbols are known beforehand by the receiver. Hence, upon reception, the receiver can compare the received symbols $c_k'$ for each subcarrier $k$ to those sent by the transmitter, which are known because of it being a training symbol. The received symbol $c_k'$ can be expressed as a function of the sent symbol $c_k$ as:

$$c_k' = c_k H_k$$  \hspace{1cm} (2.18)

where $H_k$ is the system transfer function. It is virtually impossible to establish a general form for $H_k$ that takes into account all the phenomena affecting the received symbols and serves for every system architecture. This is due to this transfer function not only accounting for chromatic dispersion, PMD, nonlinear effects, channel noise, but also for modulation- and detection-related issues[4]. It is intrinsically different how phase noise, for example, affects CO-OFDM and DDO-OFDM [50]. Therefore, no general description is attempted here, because the phenomena affecting the systems under study are the objective of the forthcoming parts of this work.

The distortions introduced along the channel and upon detection are nevertheless corrected by the transmitter. This is done, as anticipated, using the received symbols at training sequences. One of the greatest advantages of OFDM is that as the subcarriers are within themselves so narrow that it can be considered that each of them is affected by only a complex constant, i.e. a phase shift and an amplitude change. This is of remarkable importance, because this can be corrected through the so-called 1-tap equalizers, which are computationally very convenient [5]. The transfer function estimates $H_k'$ can be calculated under this assumption as:

$$H_k' = \frac{c_k'_{TS}}{c_k'_{TS}}$$  \hspace{1cm} (2.19)

where the superscripts in the symbols show that they belong to the training sequences. The correction of the payload OFDM symbols is performed with these calculated coefficients as follows:

$$c_k = \frac{c_k'}{H_k}$$  \hspace{1cm} (2.20)

Actually, in practice this calculation is performed as [4]:

$$c_k = \frac{H_k}{|H_k|^2} c_k'$$  \hspace{1cm} (2.21)
Fig. 2.28 shows different subcarriers being corrected after channel estimation using the training symbols. Only phase shifts are shown, but amplitude variations can be alleviated in the same process as well.

Most of the phenomena affecting the transmission vary in times of the order of milliseconds [4], due to which training symbols need to be sent regularly over periods ideally shorter than this, in order to repeat the channel estimation process and recalculate the correction coefficients $H'_k$. Sometimes, several training sequences are sent consecutively, in which case these coefficients are the result of the averaging over all the training symbols.

The remaining step after what has been explained is demapping, which is basically the inverse of mapping. It implies a decision process in order to finally estimate the sent bits from the received symbols. After this, parallel-to-serial conversion is done, thus yielding a serial bit sequence.

2.4—O-OFDM for next-generation passive optical networks

Although novel services like video-on-demand (VoD), real-time network games, peer-to-peer applications, video file swapping, Internet Protocol TV (IPTV) and other emerging web applications are creating an increasing need for high end user bandwidth, the cost of high-speed access technologies remain prohibitively high for the average household. Until very recently, typical access technologies consisted of 1.5 Mbps downlink and 128 kbps uplink digital subscriber line (DSL) networks. Variations of this were available, stretching the velocities until 50 Mbps at most, but with serious limitations with respect to the reach. The other typical possibility was broadband access through cable television (CATV), which offers similar performance, but can only provide limited and asymmetric bandwidth access to the user, alike DSL and its variants [9].

Motivated by the limitations of current access network technologies, and empowered by the enormous bandwidth offered by optical fibers, passive optical networks are currently
being deployed, offering unprecedented bit rates and supporting various services. As commented in Chapter 1, a PON is a point-to-multipoint topology where all the components between the **optical line terminal** (OLT) and the **optical network units** (ONUs) are passive, which minimizes the overall system cost. Several services, such as video, voice and data traffic are delivered to the end users. Fig. 2.29 displays an example of a PON. The links are bidirectional, and it is common that different wavelengths are used for different services: upstream data, downlink data and RF video, for example.

![Figure 2.29: Generic example of a passive optical network.](image)

PON technologies that have been developed during the last years include Ethernet PON (EPON) [53], APON, BPON and GPON (all based on ATM) [54] and the non-standardized WDM-PON [55]. All of these PON flavors employ **time division multiple access** (TDMA). This consists in assigning time slots to the different ONUs, moment in which they are allowed to transmit or receive, while the other ONUs do not transmit. This proves convenient for 10 Gbps capacities. Nonetheless, at this velocity, they already mandate expensive components, complex scheduling algorithms and framing technologies, besides being highly sensitive to packet latency [12]. Moreover, with higher capacities in mind, design and implementation of 40\(^{+}\) Gbps burst-mode receivers is far from simple, from both the technical and economic perspectives.

Another possibility is **wavelength division multiple access** (WDMA) [56], in which each ONU is assigned its own wavelength. This technique has two main disadvantages as well. The first one is not being straightforwardly compatible with legacy distribution networks. In order to employ such technology, part of the network would have to be wavelength-aware. This could be done either by replacing the passive optical couplers with WDM multiplexers, or by introducing wavelength-tunable devices on the user side. Moreover, as the wavelength assigned to each ONU is fixed, all ONU upgrades mandate corresponding changes at the OLT. The second drawback is that WDMA is not capable nowadays of dynamically allocating bandwidth with sub-\(\lambda\) granularity. This means that the smallest unit of bandwidth that a user can be assigned is a whole WDM carrier [18]. This is far from the expected flexibility for a PON, besides being costlier, and thus a superior scheme is expected to be found.

Other medium access technologies have been put forward for PONs as well. Optical code-division multiplexing PON (OCDM-PON) different CDM codes are assigned to different users. Its disadvantages are expensive encoders and decoders and multiaccess interference and noise because of the codes becoming non-orthogonal after intensity modulation [4]. Another proposed technique is **subcarrier multiplexing** PON (SCM-PON),

---

\(^{5}\)Asynchronous Transfer Mode
where one dedicated electrical subcarrier is used for each ONU. This one has the weakness of not allowing for a bandwidth assignment scheme as dynamic as the one allowed by OFDMA.

**Orthogonal frequency division multiple access (OFDMA)** enables assigning different subcarriers to multiple users in a dynamic way. OFDM-PONs can be as well combined with TDM so that users can be assigned certain subcarriers in given time slots, which yields one additional dimension for resource partitioning compared to the previously commented multiple access methods, giving very fine granularity to the system. A sketch of a system exploiting this possibility is shown in Fig. 2.30. Such a system can offer heterogeneous services, as being analog baseband T1/E1 and analog wireless RF signals together with Ethernet packets. The frequency-time plot depicts how certain services that have stringent quality of service requirements can be statically assigned certain subcarriers, while the remaining can be used for allocating burst traffic, like that stemming from Internet use.

![Figure 2.30: TDM-OFDM-PON architecture for heterogeneous services delivery.](image)

A system like the one in Fig. 2.30 could be as well combined with WDMA, if several such systems work on different wavelengths. OFDM-PONs are also a cost effective solution, since it can serve more ONUs per OLT receiver than WDM-PONs (several users share each wavelength). Besides cost-efficiency, backward compatibility with already deployed systems (the vast majority of which employ intensity modulation and direct detection) would be an asset for next-generation PONs. Given that fiber may be already installed, investment costs would be critically reduced if it can be reutilized [57]. Optical paths without amplification or dispersion correction methods should be then expected. However, this makes the case for OOFDMA-based PONs even stronger, since as we have proved in the last chapter OFDM features a 1-tap equalization method to cope with these impairments.

In consequence, it can be stated that OFDM-PON systems have some advantages compared to other PON technologies, like being the flexibility and granularity it offers for resource sharing, its scalable architecture (it can be combined with TDMA and WDMA). Other strengths that should be mentioned are its protocol independence and transparency, since practically any kind of signal (analog or digital) can be put onto the orthogonal subcarriers, which act as transport pipes; it has an improved spectral efficiency (typical of OFDM systems); it is a cost effective solution; and it can operate with simple medium access control (MAC) with low overhead [4].
Chapter 3

Architectures for optical OFDMA-PONs

When it comes to designing an OFDMA-PON, several architectures appear as possible. The main differences among them are those related to modulation/detection combinations. The different possibilities’ basics have been discussed in Sec. 2.3.1 and 2.3.2. Among them all, intensity modulation and direct detection architectures offer great optical-domain simplicity. Moreover, if the IM choice is a directly modulated laser, and no optical filters or array waveguide gratings are utilized, the system’s cost-effectiveness is supreme among all possible O-OFDM systems. However, there are as well trade-offs implied, like higher transmitter-side power due to the mandate of a strong optical carrier (Sec. 5.6) [5], or a lower spectral efficiency, stemming from the need for a frequency guard band because of the direct detection (Sec. 5.2) [31].

The forthcoming study is centered in the upstream multipoint-to-point architecture, since it is generally more challenging than the downstream counterpart. This occurs because as in any OFDM system, the subcarriers (possibly assigned to different ONUs) need to remain orthogonal in order to enable proper data recovery (see Sec. 2.2) [58]. Such a problem does not appear in downstream because the entire O-OFDM signal is built by the OLT, which regardless of using one or several wavelengths can manage to maintain the orthogonality among the subcarriers. This task is not easy at all for upstream, since subcarriers modulated by different ONUs are spectrally located with respect to different laser sources, which in case of being shifted in frequency cause the subcarriers to lose their orthogonality.

3.1– Optical carriers

The main question to settle regarding upstream OFDMA-PON concerns the source and nature of the optical carriers to be modulated by each ONU. One of the simplest approaches may be to deploy ONUs with lasers at the same wavelength. If the ONUs transmit at a certain wavelength, determined by its local laser, they are called colored ONUs. However, in reality keeping the same exact wavelength at different lasers is rather impossible. This occurs because the link lengths are typically longer than the coherence length of the cost-effective lasers used at such transmitters, which have linewidths in the order of the MHz. So the laser sources are in the general case completely decorrelated. When optical signals from transmitters that modulate the same nominal wavelength (presenting phase shifts and frequency drift among them) are detected together, optical beat interference (OBI) occurs, appearing as unwanted mixing products that degrade the performance of the system (treated in more detail in Sec. 6.2). This is an issue that must be solved in order to enable successful transmission.

One possible approach to solve the OBI problem is that of colorless ONUs (or source-free ONUs). The term implies the ONUs being agnostic of the exact wavelength of the optical carrier; i.e., there are no light sources or filters that demand accurate wavelength control. The ONUs can remodulate an optical carrier sent by the OLT (using a reflective semiconductor optical amplifier (RSOA), for instance). Doing so, the same wavelength is guaranteed for every ONU, and the OFDM spectrum can be filled continuously, each
user locating its band on the frequency it has been assigned. This, nonetheless, requires coherent detection at the OLT. Also, due to path differentials between the OLT and the different end users, the broadcasted carrier components can become as decorrelated at the ONUs as if they had been originated in different lasers sources [5]. Furthermore, this approach is sensitive to Rayleigh backscattering, reflections and limited gain [58].

An alternative approach could be the so-called hybrid WDM-OFDMA, which imposes orthogonal wavelengths in order to solve the optical beating problem [59]. However, this requires multiple optical sources, that can be located at the OLT or the ONUs, as well as wavelength tunability. Single-wavelength optical carrier suppression with coherent OLT detection has as well been suggested [12]. By employing coherent detection at the OLT, the optical upstream carriers are no longer needed. Due to this, they can be eliminated at the transmitters, and a single optical carrier be reinserted at the receiver, which avoids the effects of having uncorrelated carriers at the same nominal wavelength. However, this technique has two disadvantages. The first is that it advocates the use of properly biased modulators at the transmitters in order to eliminate the carriers. This is impossible to do with DMLs, which always entail optical power at the carrier frequency. This task could be alternatively done by optical filtering, in which case the ONUs colorless feature is lost, thus minimizing some of its advantages. On the other hand, though suggesting the use of sophisticated modulators (e.g. MZMs) for carrier suppression, it does not guarantee that the subcarriers be orthogonal. Therefore, it needs frequency guard bands between different ONUs’ signals in order to avoid ICI, thus not being too spectrally efficient.

A colored ONU could be thought as well to transmit on a preselected wavelength, which could be different for every transmitter. This would result in large inventorying and complex provisioning for providers, nonetheless. Non-preselected random wavelengths do not have this last drawback. This option is based upon lasers that emit at a beforehand not known wavelength. Despite of it being random in a band, feedback techniques can be utilized for finely tuning it by thermal, electrical or mechanical means and minimizing the overlapping probability with other ONUs’ wavelengths [58]. This last one is the paradigm chosen for the present work.

In other words, for our PON architecture we advocate using optical sources at the ONUs whose wavelengths are not known before being installed. Upon deployment, it may happen that they overlap with the sources of other ONUs sharing the same optical distribution network (ODN). In this case, the new source can be tuned in order to locate it on a 'free' wavelength location. It has been shown that for lasers that can be thermally tuned by 0.1 nm/ºC, as DFB and VCSEL, over a range of 20 ºC, approximately 99% of 64 installed lasers can be successfully accommodated [58]. The discussion about the necessary carrier separation in order to avoid OBI is left for Sec. 6.2, but a schematic view of the spectrum of 3 users sharing the ODN can be seen in Fig. 3.1.

3.2 – RF up/downconversion and Hermitian symmetry

As it has been explained in Sec. 2.3, a must for applying intensity modulation is a modulating electrical signal which is real. There are two ways that have been discussed in order to get a real electrical OFDM signal. One is to use RF upconversion prior to the optical modulation in the transmitter, and RF downconversion after direct detection in the receiver. The alternative way is to generate a Hermitian-symmetrical (HS) signal to modulate the subcarriers with, such that its IFFT is a real time-domain signal. The decision in this regard is an important characteristic of the system, because it determines how each ONU exactly modulates the subcarriers that have been assigned to it. If the HS approach were to
be used, in order to locate the OFDM band exactly on the subcarriers it has been assigned, the ONU would only input data on the corresponding pins of its IFFT module, leaving the rest with 0’s. On the contrary, if the RF approach were preferred, the intermediate RF frequency $f_{RF}$ would be chosen in order to correctly locate the electrical band on the frequency corresponding to the assigned subcarriers.

These two techniques appear as valid in the design of a OFDMA-PON, both offering strengths and weaknesses. On the one hand, the RF-option demands extra circuitry on both transmitter and receiver that the other does not. However, RF electronics suitable for this task are similar to those needed for wireless OFDM, which are very mature, so cheap and reliable chips are available for this nowadays. On the other hand, HS-solution has the disadvantage of halving the other in spectral efficiency, because by imposing the symmetry condition half of the subcarriers are determined by the other half. Moreover, while in the RF approach the DACs and IFFTs are fully utilized for data modulation, with HS half of the DAC and a quarter of the IFFT are only used for this, at maximum. In addition, it implies that the ONUs should have an IFFT module of as many elements as the OLT’s FFT. Furthermore, the bandwidth of ONUs’ DAC needs to be as big as OLT’s ADC. This is a huge advantage for RF technique, which only imposes that ONUs’ IFFT and DAC need be sufficient for dealing with the maximum number of subcarriers that can be assigned to a single user.

Clarifying this with an example: if the OLT functions on a total bandwidth of 50 GHz divided into 256 subcarriers, using the HS technique, every ONU need to have an IFFT of 256 elements and a 50GHz-bandwidth DAC. As opposed to this, if the maximum capacity an ONU can get is one quarter of the total, RF-method asks for 64-element IFFT and 12.5GHz-bandwidth DAC at the ONUs. Additionally, if a considerable guard band needs to be put in place (Sec. 5.2), as happens for increasing the capacity, in the HS-case this means leaving extra unused subcarriers, while it is done by using a higher RF frequency in the other case. These are remarkable advantages that lead us to adopt the RF-technique for our design.
3.3– Optical double sideband

Since OFDM was started to be proposed for the optical domain, one of the main limitations that academia noticed consisted on frequency-selective fading stemming from direct detection of an optical double sideband signal. This is mathematically proved in Sec. 5.3. As early as 2006, papers appeared supporting the use of optical filters [31] or complex modulators in order to obtain an optical single sideband signal. Since then, extremely little research has again considered ODSB modulation, taking the elimination of one of the sidebands in the optical domain for granted. However, as far as passive optical networks are concerned, including an optical filter is not necessarily the wisest choice, because these function mostly for fixed frequencies, while resource allocation and dynamic bandwidth assignment processes in such networks connote in varying frequencies. The cost of the optical filter is another issue that would be optimum to avoid. Moreover, particularly for our architecture proposal, fixed optical filters cannot coexist naturally with random wavelength laser sources which would be tuned ad-hoc.

Clearly, frequency-selective amplitude fading must be avoided in extremely high bitrate systems seeking for uttermost spectral efficiency. However, for lower (yet nowadays still considered high) bitrate scenarios where mechanisms to vary the location of the ONUs’ OFDM bands through upconversion at variable \( f_{RF} \) are inevitably planned to exist, this phenomenon may not be too fatal. We may consider that if the frequencies that suffer amplitude fading the most can be avoided, such a system has chances to work satisfactorily, at the cost of spectral efficiency. As proved in Sec. 5.3, frequency ranges restricted for some ONUs can be utilized by others, due to amplitude fading’s dependency on link length. This means that a system is able to employ its full bandwidth if the subcarrier assignment takes into account the most affected frequencies for each individual end user.

3.4– Detection and reconstruction of the OFDM band

Upon detection in the OLT, as direct detection is employed, the OFDM band is properly reconstructed at low frequency in the electrical domain, since each one is detected with its corresponding carrier. Replicas of the different ONUs’ sidebands result as well, but they can be filtered out, because they are located at higher electrical frequencies, due to detection with other than their respective carriers. This may be more clearly understood checking the mathematical formulation in Sec. 5.2.

3.5– Proposed architecture

The proposed OFDM-PON architecture takes profit of all the aforementioned characteristics, consisting of ONUs using directly modulated lasers. Each of them has a distinct wavelength, located far enough as to avoid OBI. The electrical signals modulating these are formed by a set of subcarriers assigned to each ONU by the OLT through the MAC, and positioned in frequency respective to each carrier in order not to overlap with any other ONU electrical subcarriers. Prior to optical modulation, each ONU upconverts its baseband signal with a RF frequency suitable for correct assembly of the total OFDM band in the optical spectrum, as shown in Fig. 3.1.

Taking into account the massive number of ONUs of the PON technology that becomes standard that will be installed in the years to come, we can affirm that the optimum approach is to design them as cost-effective and simple as possible. Moreover, since one OLT serves several (maybe more than a hundred) ONUs, most complex parts can be incorporated
in them, making the amortization more feasible. Therefore, the OFDM-PON upstream architecture proposed in this work is fairly simple in terms of components, most of which are nowadays commercially available. But the techniques implied bring about several impairments that are not present in more complex approaches. It is then our objective to analyze these in order to characterize them and find the system parameters related to them that lead to an optimum performance. Some of these impairments are studied in Chapter 5 for point-to-point links, and their incidence in multipoint-to-point cases is similar. There are new phenomena as well, and they are treated in Chapter 6, before implementing a system comprising all of the studied effects.
Chapter 4

Simulations

Until less than 20 years ago, photonic systems were point-to-point and involved little more than a light source, fiber and a photodiode. Later on, data rates increased, and photonic circuits for switching, data regeneration and demultiplexing started to be used. Some years after, WDM, wavelength routing in packet-switched networks, tunable lasers and different amplifiers came along, increasing the complexity of photonic circuits even more. Although (even until nowadays) these usually included less components than their electronic counterparts, by that time photonic circuits design was already quite a daunting task because of the interactions between different components, which required a careful study. It was among these circumstances that photonic device and optical circuit and system simulation tools appeared [60].

The one with which the design in the present work was performed and simulated is VPItransmissionMakerTM 1 [20]. Its working principles and issues related to its use are not to be discussed here for sake of space and because of it being beyond the scope of this work. Vast information is available for the interested reader in the software’s manuals, and in references [19, 61, 62]. Issues concerning the design parameters of the simulated systems and how they are modeled in the simulations are to be indeed commented in this chapter.

4.1– Single-user scenario

The starting point for the creation of scenarios with which to initially test an OFDM system has been a demo scenario already available in VPI libraries, called OFDM for Long-Haul Transmission, which is based on [31]. It can be seen in Fig. 4.1. It features a 10 Gbps point-to-point IM/DD OFDM system with up/downconversion using a MZM followed by optical filtering to produce a single sideband signal. Such a system can be used for transmitting over several hundreds of kilometers.

This demo design uses the VPI feature of declaring global parameters that can be used for referencing those that belong to the specific modules or galaxies. This is a worthy characteristic, in the sense that it allows for simplifying the configuration and making the scenario somewhat more fool-proof. Although in the custom-made scenario we have improved the profit that can be taken from such a feature, in the demo, some of the parameters commented in Sec. 2.3 can be modified thanks to this. Namely, cyclic prefix may be added, as a variable portion of the OFDM symbol, zero padding can be employed, the number of subcarriers can be changed, as well as the mapping scheme (constant for all subcarriers, through the bits per symbol (BPS) parameter), though it can only be some form of QAM. Equalization is possible as well, if the correction coefficients for each subcarrier are input by the user. A more detailed description of this scenario can be obtained from [61, 62].

A crucial asset of the scenario is that the receiver provides a calculation of two important magnitudes: the error vector magnitude (EVM) and the symbol error rate (SER).  

1 From this point on, the simulation software will just be referred to as 'VPI' for simplicity and neatness.
Figure 4.1: Demo CO-OFDM system available in VPI.

The former is defined as follows:

\[
EVM = \frac{\sum_{i=1}^{N} |IQ_{tx,i} - IQ_{rx,i}|^2}{\sum_{i=1}^{N} |IQ_{tx,i}|^2} \tag{4.1}
\]

where \(IQ_{tx,i}\) and \(IQ_{rx,i}\) are the sent and received, respectively, QAM symbols in the complex plane, for each symbol \(i\) of the \(N\) transmitted. Note that they are taken before symbol decision process. SER, on the other hand, is defined as the total count of symbols incorrectly decoded. If for instance 8QAM is used, thus accounting for 3 bits/symbol, if one of the three bits is decoded faultily, the symbol is regarded as incorrect for the SER calculation. Another remarkable possibility is the constellation plotting after reception, which becomes fundamental when qualitatively assessing system performance or for recognizing patterns in the distortions.

All of these features makes of the demo an interesting didactic tool for getting insight on O-OFDM, and even for trying some sophisticated techniques like electronic dispersion compensation (EDC) through the per-subcarrier equalization. However, it results extremely constrained when it comes to altering other parameters that are not accessible from the universe or galaxies\(^2\) options. For instance, the receiver digital signal processing part cannot be changed, and the OFDM signal generation is somewhat obscure, since the code that delivers it is not accessible. This is why fully controllable scenarios have had to be created.

As research in the field of Optical OFDM has been going on at TSC (Departament de Teoria del Senyal i Comunicacions) in UPC for some years now, some scenarios are available at the moment of starting any work related to the field. However, it is a necessary task for everyone to get to know every aspect of the scenarios that are going to be used, since more flexibility in these cases means more necessary understanding and consciousness at the moment of putting them to the test.

For simulations of point-to-point O-OFDM links that are to be included in the present work, a custom scenario was designed. It is based on one built by members of the TSC. This, in turn, has a similar operation to the demo when it comes to discrete components, However, the main difference are the transmitter and receiver modules, which were built from scratch, allowing for much higher flexibility. The fundamental feature of VPI for doing this consists of its cosimulation interface with several programming languages. "Black-box" components can be included in the scenarios, and these interact with VPI through their inputs, outputs and magnitudes passed to the black-box as parameters. As expected,

\(^2\)In VPI, the file containing the overall system is called universe. This can as well contain smaller files or scenarios called galaxies, which are formed by several components on their own, and interact with the universe through their inputs and outputs.
4.1 Single-user scenario

This feature is far from straightforward to use, and in general several signal-conversion and numerical processing stages have to be performed in order to input and output data to and from these modules.

The built transmitter and receiver feature Matlab cosimulation modules, among the different possibilities. We shall call them coder and decoder, respectively, in order to distinguish them from the transmitter and receiver galaxies that contain them. The circuit for the whole system (universe) can be seen in Fig. 4.2, as well as the transmitter and receiver galaxies. For the sake of space, not all of the blocks can be explained and discussed here. Some of them need patient observation and study in order to fully understand their performance, and some require many simulations in order to get to the proper tuning. Nevertheless, having explained an O-OFDM system as it has been done in Chap. 2, the principle of operation of all of them becomes clear by analogy with a real system. The blocks in dark gray are used for measurement, display or calculation, and are not important for the system performance.

In order to replace the transmitter employed in the demo scenario, all the processes between the input sequence of bits and the output signal in the time domain (as explained in Sec. 2.3) need to be programmed in the coder (which is written in Matlab). Similarly for the receiver and the decoder, where a detected electrical signal is input and a demodulated bit sequence output. This, among others, implies calculating all the necessary parameters, mapping, performing IFFT calculation, adding cyclic prefix, training sequences, removing them, performing FFT, demapping, calculating channel estimation, equalizing and calculating the performance estimates.

VPI (in the mode used for our work) works based on samples of the electrical signals it models. Therefore, the signals it interchanges with cosimulation modules are samples of signals in the time domain. The coder utilizes therefore an upsampling function, while the decoder employs downsampling, which are not to be dwelt upon here. However, and although the immediate results may not be reflected in this work, study and discussion took place centered in these processes. This is due to the nature of the simulation software, which employs DTFT calculations, hence bearing a parallelism with the DTFT calculations used for modulation/demodulation of OFDM. This entails some consequences that might need to be considered for some particular uses. To give an example, some filtering processes had to be developed in frequency domain with Matlab in order to get the expected behavior.

This single-user scenario has all of the features available in the long-haul demo, plus some others. At a level of system parameters, zero padding introduction is possible, choosing which particular subcarriers should be turned off. This offers a wide range of possibilities that include, as an example, eliminating the RF up/downconversion stages at the transmitter and receiver, and in change consider a baseband signal whose low-frequency subcarriers have been left in 0. This is also important for multi-user networks, as later explained.

Besides EVM, which is still possible to obtain, bit error rate (BER) calculation has been incorporated, which is somewhat closer to the ultimately experienced system performance than SER. BER is the proportion of the total bits transmitted that were incorrectly decoded.

An improvement in user-friendliness has been achieved with this custom scenario as compared to the demo. For a user that fairly understands the system basics, many of the design parameters are easier to change. The difference may be said to be in the "intelligence" level of the scenario, because by changing some parameters, the ones depending on them automatically change. As compared to this, the demo is less foolproof in the sense that in order to change a system characteristic, several parameters may need to be changed, even in different modules. Task that, if not done correctly, can lead to faulty performance and
Figure 4.2: Pont-to-point O-OFMD system, transmitter and receiver. In the universe, the blocks with blue "UPC" and "EPSC" drawings are the transmitter and receiver, respectively.
incorrect conclusions. An easy example may be given with the number of subcarriers or the cyclic prefix length, which need to be a global parameter in point-to-point links, and may lead to errors if set differently in transmitter and receiver.

Regarding strictly system parameters, unless otherwise is stated in each particular case, the settings used for the single-user system simulations are as follows. The **carrier number** $N_{\text{FFT}}$ has been set to 64, which is typical throughout publications; **QPSK** modulation format has been used, thus yielding 2 bits/symbol; 4 **training symbols** have sent for channel estimation and equalization in many cases, as well as a cyclic prefix of 10% of the symbol duration. The **laser linewidth** has been set to 1 kHz when its effects are not particularly the subject of study, and to 1 MHz, which corresponds to the spectral width of commercially available lasers, when assessing the performance of the proposed architecture. This is not a completely unrealistic value, but it is certainly much smaller than the linewidths of lasers that may be used for cost-effective applications like ours.

In many sections, bitrates of 1 Gbps and 10 Gbps have been put to test. With QPSK modulation, they correspond to RF bandwidths of 500 MHz and 5 GHz, respectively.

![Figure 4.3: Point-to-point O-OFDM system featuring a MZM.](image)

The chosen VPI module for the directly modulated **distributed feedback laser** (DFB) is **LaserAnalog DSM** (Laser Analog, Data Sheet Model), whose response is defined by datasheet parameters, what makes it suitable for emulating lab experiments as is done in further sections of this work. The main parameters to define are linewidth, threshold current, bias current, slope efficiency, **driver transconductance**, linewidth enhancement factor and adiabatic chirp factor. The first has already been discussed. The last two parameters are related to chirp, which is discussed in Sec. 5.5. The Linewidth Enhancement Factor value has been set to 3, which is a common alpha factor, and the Adiabatic Chirp Factor to 1 GHz/W. Regarding the rest of the parameters, a set of realistic values has been taken, and they result as well numerically convenient, as seen in Sec. 5.6. This values are: threshold current $I_T = 20$ mA, bias current $I_B = 60$ mA, slope efficiency $\eta = 0.2$ W/A, driver transconductance $G = 40$ mA/V.

One change at a component level has been necessary for some sections. Although Fig. 4.2 displays a directly modulated laser as IM modulator, in some cases it has been chosen to avoid an effect related to it that is **chirp**. This phenomenon is to be studied in Sec. 5.5. As with MZMs chirp is much lower, it was often this the used modulator. A scenario featuring a MZM can be seen in Fig. 4.3. One component of the circuit that needs to be discussed is the optical filter located after modulation. Throughout the presentation so far we have advocated against the use of such components in passive optical networks, hence not including any in our proposal and working with a double sideband optical signal. However, in Sec. 5.2 it becomes convenient to have a OSSB signal, for which end we use...
the filter for. For eliminating one of the bands, the filter is a rectangular band-stop of the following parameters:

\[
\begin{align*}
\text{CenterFrequency} &= f_o - CF \\
\text{Bandwidth} &= BW \\
\text{BandstopAttenuation} &= 20\text{dB}
\end{align*}
\]

where \(f_o\) is the optical carrier frequency (193.1 THz), \(BW\) is the OFDM signal bandwidth, and \(CF\) is the electrical carrier frequency.

The number of transmitted symbols responds to the time window parameter in VPI, which defines the temporal duration of the simulation. It is usually set to be an entire number of OFDM symbols. As computational time is always a scarce resource, when the simulations have had a qualitative end, although employing a symbol number as high as possible, no strict requirements have been put. There are some studies, like those in Secs. 5.2 and 5.6, that required sweeps over parameters, which means repeating the simulation for each different value of the magnitude under study. Sometimes, nested sweeps of various parameters have been performed, which exponentially increases the needed time.

However, when assertions regarding system performance have been made, more computational time has been invested. Assessing the performance of such systems imply studying the EVM and BER. **Forward error correction** (FEC) algorithms demand a maximum BER of \(10^{-3}\). Lower values than this would theoretically permit a successful communication over the channel. In order to distinguish up to BERs of the order of \(10^{-3}\), it is a recommended practice to simulate a sufficient number of bits for getting around 100 errors, which in this case means generating \(2^{17}\) bits for each ONU.\(^3\)

### 4.2– Multi-user scenario

For testing the phenomena appearing in a multipoint-to-point link, and ultimately the proposed architecture, a VPI multi-user scenario is clearly needed. Based on an already existing one, a configurable and user-friendly multi-user VPI universe was created. The original one, that was still in development phase, worked properly for a simplest-case uplink, i.e. two ONUs with equal bandwidths, mapping scheme and even **pseudo-random bit sequences** (PRBS). As some intrinsically PON-environment features like **dynamic bandwidth assignment** (DBA) were to be put in practice, these functionalities were not enough for our objectives.

The created scenario features a two-user uplink O-OFDM system of similar characteristics to the single-user system regarding each of the transmitters, and the receiver. It can be observed in Fig. 4.4, where some important parameters and measured values can be seen as well.

This scenario’s parameters have been set whenever possible to those of the proposal. Therefore, the OLT has a bandwidth of 5 GHz, which with QPSK modulation format yields a total bitrate of 10 Gbps. 128 subcarriers are available, out of which 64 are assigned to each of the 2 ONUs in the beginning. The spectral location of the ONU signals is done by choosing the corresponding RF carrier frequencies for each of them. For this first setup in which both ONUs manage 64 subcarriers. As the so-called Nyquist subcarrier (number \(N_{\text{FFT}}/2+1\)) is located at both sides of the OFDM band for any subcarrier number (see

\(^3\)Strictly, \(100 \times \frac{1}{10^{-3}} = 10^5\) bits need to be generated for each ONU, but due to the nature of VPI simulation environment, it is much more convenient to work with powers of 2. Therefore, a slightly larger number of \(2^{17}\) is conservatively taken.
4.2 Multi-user scenario

Sec. 2.3), it has to be deactivated for every ONU, because those of ONUs adjacent in frequency overlap. This is done using zero padding.

There still are other design parameters to comment. These are the electrical carrier frequencies, optical carriers separation and lasers drive amplitude. Let us only mention them at this point, in order to conclude an overall view of the whole system, since they are subject of deeper analysis in the forthcoming sections. The optical spectra at each of the lasers’ output and after being coupled can be seen in Fig. 4.5, as well as the detected electrical spectrum. The received constellations for both ONUs are displayed in Fig. 4.6.

The system features two OFDM coders, which modulate a suitable electrical signal, based on the knowledge of the number of subcarriers assigned to them, and so the corresponding bandwidth, as well as the carrier frequency for RF upconversion. These signals modulate 1MHz-linewidth DMLs emitting 2 mW of optical power at different wavelengths. The optical signals are joined in a 50/50 coupler, after which the optical signal is transmitted through 25 km of SMF. This length has been used because it corresponds to a reach called urban scenario, meaning that it is the maximum fiber distance between OLT and ONU in a city or town.

At the output of the fiber an attenuator is placed, decreasing 12 dB. Optical splitters are the distribution method used in PONs. In their simplest form, they ideally divide their input power perfectly into two outputs. As no active devices are placed in PONs, this is the way in which several ONUs share the same physical medium. Cascaded optical splitters can be used for serving several ONUs, creating what is called 1xN optical splitters, where N is the number of output ports. Each power division by a factor of 2 represents a loss of 3 dB in optical power. It can be seen that if A is the power loss in dB, \( N = 2^{A/3} \). Therefore, 12 dB of attenuation represents the loss given by a 1x16 optical splitter. These are bidirectional, so the loss is suffered both in uplink and downlink.

Next, a preamplifier is included to supply the PIN photodiode with -9 dBm of optical power. Its noise figure has been set to a typical value of 6 dB. For the detector, typical values have been used as well: responsivity of 1 A/W, thermal noise of 21e-12 A/Hz^2 [42].
Simulations

Figure 4.5: Spectra at several points of the OOFDMA-PON scenario.

(a) Optical spectrum from ONU1.
(b) Optical spectrum from ONU2.
(c) Total optical spectrum input to the fiber.
(d) Detected electrical spectrum.

Figure 4.6: Received constellations of the described simulated system.

(a) ONU1
(b) ONU2
Chapter 5

Point-to-point O-OFDM

In Chapter 2 we have discussed the principles of OFDM, and a fairly detailed description of the most important processes that take place in a transmitter and a receiver has been given as well. Particularly in Sec. 2.4 the advantages of O-OFDMA in order to contest for becoming the modulation and access technology for next-generation passive optical networks were highlighted. Based on all of this, and taking into account the requirements of such an access network, an architecture proposal has been presented, focusing on cost-effectiveness, and analyzing mainly the upstream link. Specifically, intensity modulation and direct detection of a double sideband optical OFDM signal is advocated. While Chapter 3 anticipates transmitter and receiver side operation, there are specific phenomena that need to be studied in order to fully understand their nature and to design methods to alleviate their effects, when detrimental for the system. This is done in the present chapter.

5.1– Chromatic dispersion and equalization

Chromatic dispersion the phenomenon by which the group velocity of a wave depends on its frequency in certain medium (in this case, the optical fiber). This dependency causes the fiber to have a transfer function of the type:

\[ H(\omega) = \exp\left(\frac{1}{2}j\beta_2 L (\omega - \omega_{ref})^2 \right) \]  \hspace{1cm} (5.1)

where \( L \) is the fiber length, \( \beta_2 = \frac{d^2\beta}{d\omega^2} \) and \( \beta \) is the propagation constant in direction parallel to the fiber axis. \( \omega_{ref} \) is a reference frequency, with respect to which time and phase shifts are considered. In our work, whenever not stated otherwise, \( \omega \) has been always set as the optical carrier frequency. \( \beta_2 \) is a characteristic parameter of the fiber, but the one commonly used is the dispersion coefficient \( D \), which is related to it by:

\[ \beta_2 = -\frac{c}{2\pi f_{ref}^2} D \]  \hspace{1cm} (5.2)

where \( c \) is the speed of light. So, with respect to \( D \), and expressing using linear frequencies:

\[ H(\omega) = \exp\left(-j\frac{cL}{4\pi f_{ref}^2} D (2\pi)^2 (f - f_{ref})^2 \right) = \exp\left(-j\pi cLD \left(\frac{f - f_{ref}}{f_{ref}}\right)^2 \right) \]  \hspace{1cm} (5.3)

One of the consequences of chromatic dispersion is a phase shift between spectral components of a signal, as would be subcarriers in our case. So the phase shift between a subcarrier at frequency \( f_i \) and the frequency reference is:

\[ \Delta \phi^i = \pi cLD \left(\frac{f_i - f_{ref}}{f_{ref}}\right)^2 \]  \hspace{1cm} (5.4)

From Eq. 5.4 it can be noticed that the phase shift depends linearly on the fiber length and quadratically on the frequency difference with the reference. This is used in following sections as well, as when explaining amplitude fading in Sec. 5.3.
We have already discussed the effects of these phase shifts in Sec. 2.28. They manifest themselves as deviations in the inphase-quadrature plane. If their magnitude is high enough, a symbol may be incorrectly decoded after the decision step in the receiver when no equalization is performed. Fig. 5.1 shows the received constellations with and without chromatic dispersion correction.

5.2 – Direct detection and frequency guard band

The current resulting from the detection process in the photodetector has a square-module relation with the incident optical field.

\[ i_d(t) = |E(t)|^2 = E(t) \times E^*(t) \]  

(5.5)

where the * superscript denotes complex conjugation. This corresponds to a convolution in the frequency domain:

\[ i_d(\omega) = E(f) \ast E^*(-\omega) \]  

(5.6)

where * means convolution product. \( \mathfrak{F} \{E^*(t)\} = E^*(-\omega) \) is obtained from Fourier Transform properties [64]. This quadratic detection produces three important spectral components as seen in Fig. 5.2. One is the product of the optical carrier, which appears as a DC component in the electrical detected spectrum. The other is the product of the sideband multiplied by the optical carrier, which is the useful data signal. And the product of the sideband multiplied by itself gives unwanted mixing products.

This last term is sometimes called signal-to-signal mixing interference (SSMI). When IM/DD is employed, it appears only if dispersion has taken place in the channel. On the contrary, if AM/DD is used, SSMI appears with or without chromatic dispersion. However, as intensity modulation is to be utilized in this work, we do not consider this last case in henceforth. If these products overlap with the detected sideband, they act as interference. There exists one interesting point to notice, which is manifested in Fig. 5.2 as well. Generally, because of the convolution, the amplitude of the \( S_{OFMD} \ast S_{OFMD} \) products diminishes towards the higher frequencies. This fact proves useful when assessing the possibility of shortening the frequency guard band, ahead in this section.

Fig. 5.2 already shows a way around this problem, which is to offset the sideband in frequency. This is called frequency guard band. If a void frequency span is left between the optical carrier and the sideband, of width equal to that of the band, after detection, the unwanted mixing products do not overlap with the OFDM band, as seen in Fig. 5.3.
5.2 Direct detection and frequency guard band

This technique is has been named offset SSB-OFDM, when performed with optical single sideband [4].

Technically, this solution can be implemented in two different ways. One of them consists in upconverting the OFDM band before optical modulation with an intermediate RF carrier. It is not hard to appreciate from Fig. 5.3 that \( f_{RF} = 1.5B \). This is generally called offset DDO-OFDM [4, 31]. The other possibility is to leave the low frequency half of the subcarriers unmodulated, by utilizing the Zero Padding feature explained in 2.3. Either way, leaving such a frequency guard band implies reducing spectral efficiency by half. Moreover, performing so with the second method, it also means underusing the transmitter IFFT because of not modulating half of the subcarriers.

By now it is clear that the need for a frequency guard band is well founded when dealing with direct detection after a dispersive channel, as in our case. However, as sketched in Fig. 5.2, the amplitude of the mixing products lower towards higher frequencies. Also, the overall amplitude depends on the magnitude of accumulated dispersion. So, in links where accumulated dispersion remains under certain levels, it may be possible to reduce the guard band. Frequency is a scarce resource in most systems, and specially in ours, because demanding higher bandwidths of components such as modulators (either MZM, EAM, or DML) means critically increasing systems’ costs. Making more efficient use of frequency also means serving more ONUs per OLT. Therefore it is worth analyzing whether the guard band can be shortened to some extent.

As the amplitude of mixing products tends to decrease towards higher frequencies, bringing the OFDM band closer to the carrier may not be completely catastrophic for the detection process. In order to probe this idea, a point-to-point scenario has been used for simulation in VPI, as explained in Sec. 4.1. A MZM has been employed, and also a filter for optical single sideband generation.

Bit Error Rate and Error Vector Magnitude have been obtained for a series of carrier frequencies and fiber lengths. Because of the nature of our system, the minimum carrier frequency permitted is \( \frac{1}{2} \times B \), which corresponds to the OFDM band being right next to the optical carrier in the spectrum. The usual 'safe' value of carrier frequency is \( \frac{3}{2} \times B \), which is

---

1A way to realize of this is to think that when the bands are convoluted with each other, the resulting amplitude is higher when larger parts of them overlap, i.e. for lower frequencies.
used in almost all our other simulations. This corresponds to the situation depicted in Fig. 5.3. The maximum carrier frequency has been set higher than this value for simulations, but no considerable advantage is got, So the range below $\frac{3}{2} \times B$ is of much more interest for us at this point.

Simulations have been performed for a bitrate of 1 Gbps, which is coherent with the aim of the present work. The results are shown in Fig. 5.4. Starting from the minimum possible value of 250 MHz, both BER and EVM decrease as the guard band becomes larger. For all fiber lengths, BER remains at levels lower than $10^{-3}$ for carrier frequencies of 300 MHz and above. This happens due to the narrower signal bandwidth, which results in lesser chromatic dispersion over the band, and thus less significant mixing products.

![Figure 5.4: BER and EVM vs carrier frequency for varying fiber lengths. Bitrate: 1 Gbps. Those points in the BER curve corresponding to 0 have been plotted as crosses on the horizontal axis.](image)

These results mean that a significant part of the usually applied guard band in the offset DDO-OFDM approach (equal to the $B$) may not be so necessary in links whose lengths make chromatic dispersion not to be of great incidence. Therefore, for cases in which the spectral efficiency is crucial (like passive optical networks), the guard band could be reduced without being extremely detrimental for the system. Moreover, spectral efficiency is not the only reason for being interested in locating the OFDM band close to the optical carrier. This has other advantages like reducing the necessary modulators’ bandwidths. Also, this reduces the frequency-selective amplitude fading effect, as can be seen in Sec. 5.3.

### 5.3 – Double sideband and amplitude fading

Chromatic dispersion affects the system not only by introducing phase shifts and delays among the subcarriers, but also by causing an effect known as frequency-selective amplitude fading. We henceforth call it simply ‘amplitude fading.’ It takes place when a double sideband optical signal is directly detected at the receiver. As it is discussed in Sec. 2.3.1, directly modulating a semiconductor laser (or a MZM) with a driving electrical signal produces sidebands at both sides of the optical carrier, resulting in what is called optical double sideband (ODSB) signal. Let us consider that a single tone is be used as modulating signal. The optical modulated signal is then similar to what is observed in Fig. 5.5.
After it suffers dispersion in the optical fiber, the detected signal is:

\[ i_d(t) = |E(t)|^2 = A \left| e^{j\omega_0 t} + m e^{j(\omega_0 t - \omega_{RF} t - \phi^-)} + m e^{j(\omega_0 t + \omega_{RF} t + \phi^+)} \right|^2 \]

\[ = A^2 \left[ \cos(\omega_0 t) + m \cos(\omega_0 t - \omega_{RF} t - \phi^-) + m \cos(\omega_0 t + \omega_{RF} t + \phi^+) \right]^2 \]

\[ + \left( \sin(\omega_0 t) + m \sin(\omega_0 t - \omega_{RF} t - \phi^-) + m \sin(\omega_0 t + \omega_{RF} t + \phi^+) \right)^2 \]  

(5.7)

(5.8)

here, \( \phi^- \) and \( \phi^+ \) are the subcarrier phase given by the modulation plus the phase shift imposed by the chromatic dispersion to each of the tones. With the sign convention used, in a non-dispersive medium, they should be equal in magnitude and sign. \( m \) represents the optical modulation index. This leads to a considerable number of terms. We only care for the frequencies corresponding to \( \pm \omega_{RF} \), leaving the others aside, which are filtered out after detection. The remaining terms are proportional to:

\[ \cos(\omega_{RF} t + \phi^+) + \cos(\omega_{RF} t + \phi^-) = 2 \cos \left( \omega_{RF} t + \frac{\phi^+ + \phi^-}{2} \right) \cos \left( \frac{\phi^+ - \phi^-}{2} \right) \]  

(5.9)

This means that the detected electrical signal presents a component at frequency \( \omega_{RF} \) with a phase shift given as the average of those of each of the sidebands (tones, in this case). The last term is what produces Amplitude Fading. In a non-dispersive medium, \( \phi^+ = \phi^- \) and there would be no such effect. But in a general case, if \( \phi^+ - \phi^- \approx (2n + 1) \pi \), the detected component at \( \omega_{RF} \) will have low amplitude. Fig. 5.6 shows the amplitude fading factor \( \cos \left( \frac{\phi^+ - \phi^-}{2} \right) \) for a single tone being modulated in double sideband, vs its frequency.

The Optical OFDM signal can be regarded as formed by several subcarriers identical to the one used in the previous study, with varying frequencies \( \omega_{RF} \). Therefore, the phase shifts \( \phi^- \) and \( \phi^+ \) are different for each subcarrier, which leads to the amplitude fading effect to vary among them. Fig. 5.7 clearly depicts amplitude fading affecting a signal. Observe that in the optical domain, before detection, the OFDM band is flat, as opposed to the output of the photodiode, which is deeply affected by fading close to its center. For assessing the importance of the Double Sideband problem in OOFDM systems, a simulation in VPI is issued. As the aim is to analyze the consequences of transmitting double sideband in an optical fiber, a Mach-Zehnder Modulator is used, in order to diminish as much as possible all chirp-related effects (this will be addressed in Sec. 5.5). The bitrate is set to 10Gbps, which with QPSK modulation yields a signal bandwidth of 5GHz. The Double Sideband optical spectrum can be seen in Fig. 5.8. Different lengths of fiber are examined, spanning from 0 km to 100 km, in steps of 5 km. The EVM and BER graphs are shown in Figs. 5.9 and 5.10 respectively. Standard SMF characteristics are used. For lengths over 25 km, values of BER greater than \( 10^{-3} \) are obtained, thus showing an intrinsic limitation of the technique. For better insight into the phenomenon, a per-carrier study can be done.

---

**Figure 5.5:** Optical spectrum of a signal modulated by a single tone.
As explained before, this sort of impairment stems from the subcarriers adding to each other in pairs located exactly opposite to the optical carrier, with phases that add up to odd multiples of $\pi$. As explained in Sec. 5.1, the phase shift of a subcarrier $i$ can be
5.3 Double sideband and amplitude fading

\[ \Delta \phi_i = \frac{\pi cL D}{f_{\text{ref}}^2} \Delta f_i^2 \]  

where \( c \) is the speed of light, \( L \) the link length, \( D \) the fiber dispersion coefficient, \( f_{\text{ref}} \) the reference frequency in the fiber model (taken as the optical carrier frequency in our simulations), and \( \Delta f_i \) being the frequency difference between the subcarrier \( i \) and the optical carrier. Doing so, and summing the contributions of opposite pairs, a graph as shown in Fig. 5.11 is obtained. Conclusions similar to those that resulting from the simulations can be drawn, since these have showed that for lengths larger than approximately 30 km BER went up abruptly. Correspondingly, Fig. 5.11 shows that these are the cases for which some subcarriers accumulate a phase shift close to \( \pi \), which leads to an almost complete loss of amplitude.

Though the analysis so far has served to understand the principles and dangers of double sideband modulation in OOFDM systems, the chosen parameters were not realistic. For the purpose of this work, a bitrate of 1 Gbps is sufficient (considering that the upstream part of the system is the subject of study). A similar simulation as the one shown above is presented next, with the differences being the reduction in bitrate and that now we include
Figure 5.11: Accumulated phases of different subcarriers. Note that when the phase approaches $\pi$, BER leaps up.

Figure 5.12: DSB spectrum for 1Gbps.

an attenuation of 0.2 dB/km (as standard for SMF). The transmitted spectrum can be seen in Fig. 5.12.

Fig. 5.13 shows the evolution of the EVM for increasing fiber lengths, when a bitrate of 1Gbps, and hence a bandwidth of 500 MHz, has been set. BER results are not shown graphically, because for the simulation time window no errors appeared at any length, giving so values of BER of 0. This is straightforwardly understood by observing in Fig. 5.14 that no subcarrier presents added phase (due to its two sidebands components) any close to $\pi$. Therefore, if this was the class of link to design and evaluate, it is clear that this phenomenon would not be a great limitation for the system.

We have so far analyzed how the fact of having both sidebands affects the detection of a signal of an ONU with a particular location in frequency, i.e. equal to the OFDM
5.3 Double sideband and amplitude fading

bandwidth. This is to say, the closest to the optical carrier, after modulation. But, as it is explained in Sec. 6, when many ONUs are intended to share the same physical medium, this is not always the case. On the contrary, ONUs may have to position their signals over a large bandwidth.

For the next simulation, the bandwidth available for upstream is considered to be 16 GHz and, for reasons to be explained in Chapter 6, it spans from 16 GHz to 32 GHz. Due to its dependence on frequency and length, depending on the physical distance between ONU and OLT the placement in frequency where the signal is most affected by amplitude fading varies. This is shown in Fig. 5.15, where BER vs $f_{RF}$ is displayed for different fiber lengths. This graph results, for each length, complementary to Fig. 5.6, because low received amplitude results in high BER. So, when $f_{RF}$ is such that an amplitude minimum lies within the ONU signal band, a growth in BER is experienced.

From contemplating Fig. 5.15 arises the realization that different positions in the spectrum (different carrier frequencies) are convenient for different fiber lengths. Thus, we could think of a strategy that assigns ONU carrier frequencies depending on the fiber length of
each of them. For certain length, the carrier frequency to be used would be one of those that presents the smallest amplitude fading, and therefore lowest BER. Doing so, optimized performance as well as full spectrum utilization can be achieved.

5.4 – Phase Noise

While an ideal oscillator can be considered to emit only one frequency or wavelength, real oscillators as lasers present random fluctuations in the phase of their waveforms. These are caused by time-domain instabilities, but are most commonly expressed in frequency-domain as a broadening of the emission spectrum. Whereas an ideal oscillator would be represented by a delta function in frequency, lasers are characterized by a finite linewidth, caused by the random phase variation. The linewidth is inversely related to the coherence time, i.e. greater linewidths lasers have lower coherence times and lengths [1].

One of the claimed advantages of DDO-OFDM is its robustness against phase noise, because the sidebands are detected by their convolution with their own optical carrier, which means that any phase noise term will cancel upon detection. However, when combined with optical fiber dispersion, phase noise provokes walk-off between the optical carrier and the subcarriers. For links above a certain length, phase coherence between them may be completely lost, causing detrimental effects on transmission due to phase noise.

These effects take place in the form of power degradation, phase rotation term (PRT) and inter-carrier interference (ICI), [50]. An useful way of assessing the role of each of these in the signal degradation is to analyze its received constellation. Altogether, these deteriorate the system performance for wider laser linewidths, which clearly is a result of the constellation degradation that can be detected in Fig. 5.16. The consequences of this constellation widening can be recognized in Fig. 5.17, which evaluates system performance.
5.4 Phase Noise

(a) Linewidth: 100 kHz  
(b) Linewidth: 1 MHz  
(c) Linewidth: 5 MHz  
(d) Linewidth: 10 MHz

Figure 5.16: Directly detected constellations for a bitrate of 1 Gbps, constant input optical power of 3 mW, after 10 km standard SMF. Single Sideband modulation.

for a bitrate of 1 Gbps vs fiber length, for several laser linewidths. Performance degradation is observed, as expected, when increasing both laser linewidth and fiber length. Observe how for narrow linewidths, as seen in Fig. 5.16a, despite there being some phase noise, the detected symbols fall not too far from the ideal locations ($[\pm1,\pm1]$ in I-Q coordinates), giving thus no bit errors. But for growing linewidths, errors begin to arise in greater number, since symbols are wrongly decoded more and more often.

Figure 5.17: BER and EVM vs fiber length for several values of laser linewidth.

Attention should be given to the difference between the distortions in the constellations caused by chromatic dispersion alone, and by phase noise combined with it. We have explained in Sec 2.3.1 that chromatic dispersion may bring about inter-symbol interference (ISI). But this is avoided by adding a cyclic prefix to the transmission. Then, the 1-tap equalizer can correct for subcarrier phase shifts, and any linear impairment in general, provided enough guard band is left, in IM/DD. Therefore, OOFDM systems are intrinsically capable of coping with the effects of chromatic dispersion. However, phase noise along with CD mainly results in a different type of distortion that cannot be corrected as it has just been explained. Phase noise bandwidth can be even larger than 100 MHz, which inevitably means inter-carrier interference (ICI), making PN compensators very complex to design [50]. In fact, PN compensation in DD systems has not been explored much yet, and there are very few publications on the topic. Due to this, in the present work we deal with phase
noise as a sort of impairment that limits the fiber length for a given linewidth.

5.5 – Frequency chirp

Chirp is the phenomenon by which the frequency of a transmitted signal varies with time. This inherently causes spectral broadening of the signal, thus diminishing its tolerance to chromatic dispersion [67]. Fundamentally, chirp is takes place because the phase and amplitude of the optical signal coming out of a DML depend on each other, and its behavior is governed by [68]:

$$
\Delta \nu(t) = \frac{\alpha}{4\pi} \left[ \frac{1}{P(t)} \frac{dP(t)}{dt} + \kappa P(t) \right]
$$

(5.11)

where $\nu$ is the instantaneous signal frequency, $\alpha$ is the linewidth enhancement factor, $\kappa$ is the adiabatic frequency chirp coefficient and $P(t)$ is the DML optical output power [69]. The first term on the righthand side of Eq. 5.11 denotes the DML transient frequency chirp, which is proportional to the time variation of the optical signal waveform. The second term denotes the DML adiabatic frequency chirp, which depends on the instantaneous optical signal waveform. The latter can be appreciated in Fig. 5.18b, where the optical carrier is slightly shifted from the reference frequency, where it would be if there were no adiabatic frequency chirp.

![Optical spectrum. Mach-Zehnder Modulator.](image1)

(a) Optical spectrum. Mach-Zehnder Modulator.

![Optical spectrum. DML. $\alpha$: 3; $\kappa$: 10 GHz/W.](image2)

(b) Optical spectrum. DML. $\alpha$: 3; $\kappa$: 10 GHz/W.

![Received constellation. MZM](image3)

(c) Rx constellation. MZM

![Received constellation. DML. $\alpha$: 3; $\kappa$: 10 GHz/W.](image4)

(d) Rx constellation. DML. $\alpha$: 3; $\kappa$: 10 GHz/W.

Figure 5.18: Received constellations and spectra of OOFDM signals modulated using two different Intensity Modulation techniques.

The received constellations of identical systems using Mach-Zehnder Modulator and Directly Modulated Laser are compared in Fig. 5.18c and 5.18d. MZMs have a much lower $\alpha$ factor than DMLs, of around 0.2 for dual-drive MZM with equally fed arms, or even much
closer to 0 \[71\]. Compared to the usual range of DML α of 2-6, they are considerably small values. Therefore, the chirp effect is much more remarkable when dealing with directly modulated lasers.

Fig. 5.19 depicts the consequences on performance of including chirp effects in the modulator model. Fig 5.19a shows the variation of BER due to the chirp-generated spectral broadening together with chromatic dispersion. Fig. 5.19b displays the same effects on maximum signal line rate with respect to the transmission distance. As compared to the ideal intensity modulator, chirp in DMLs cause a reduction in transmission capacity of up to 25% \[69\]. This means that IM/DD OOFDM systems employing DML are chirp-limited up to link lengths of 80 km. As in PONs, which are the object of study throughout this work, distances are shorter than that, this enables us to give more attention to the effects of chirp than those of attenuation, which becomes more important for longer distances.

![Figure 5.19: Effects of including chirp in the modulator model. \[70\]](image)

We have so far been using MZM in the simulated systems, which with the correct parameter set performs much more similar to an ideal intensity modulator than a DML. Henceforth, a new impairment is to be taken into account, stemming from frequency chirp. This is a complex phenomenon to study and discuss, and an important part of the ongoing work related to IM/DD OOFDM is devoted to do so. Very recently, a DFB frequency chirp compensation technique for OOFDM IM/DD PON systems has been published \[69\]. However, these techniques are coming about while the present work is being performed, and are consequently beyond its scope. Therefore, chirp has been treated here as an limiting effect of utilizing DMLs, which does not prevent us from successfully investigating other phenomena and features of OOFDM(A) systems. The effect of chirp that has just been studied affects OOFDM systems regardless of them featuring single or double sideband. However, when directly modulated lasers are used to modulate an optical double sideband signal which is transmitted over a dispersive optical fiber, an effect that has not yet been mentioned appears. In Sec. 5.3, the amplitude fading phenomenon has been explained. It arises from the fact of having subcarriers symmetrically located at both sides of the optical carrier, which after direct detection and due to the accumulated phase shift generated by chromatic dispersion, cancel each other. Hence the low amplitudes at frequencies suffering from this, and the impossibility to successfully use them for conveying data. The calculation of the phase shift caused by chromatic dispersion can be done with Eq. 5.10 in Sec. 5.3. However, due to the chirp, and especially to the adiabatic chirp, there takes place a new phase shift factor.

As explained above, the chirp manifests itself as a frequency shift, which therefore causes
a phase shift, given by:

$$\Delta \phi(t) = 2\pi \int_0^t \Delta \nu(\tau) d\tau$$  \hspace{1cm} (5.12)$$

This new phase shift can be recognized in Fig. 5.20, where the optical signals and their instantaneous phase and chirp are compared in time, for a MZM and a Directly Modulated Laser. This is in consonance with the spectra observed in Fig. 5.18. Some characteristics of the graphs should be noticed. First of all, that both the phase and chirp are two orders of magnitude greater in the case of the DML compared to that of the MZM. Also, that the DML phase is continuously increasing. This is due to the adiabatic chirp term, because the intensity is always positive (see Eq. 5.11).

![Waveform graphs](image)

(a) Directly Modulated Laser.  \hspace{1cm} (b) Mach Zehnder Modulator.

Figure 5.20: Instantaneous intensity, phase and chirp of two differently modulated optical signals with equal power input to the fiber.

The effect of this phase shift is changing the location of the frequency components that suffer the total (chromatic-dispersion- plus chirp-generated) amplitude fading the most, with respect of the case of only considering chromatic dispersion. It is not the aim of this work to provide an analytical description of this particular feature, but it can be treated phenomenologically. In order to do that, Fig. 5.21 shows the effects of amplitude fading on the detected signal for three different carrier frequency values. It is clear from comparing these situations that an optimum value for which the amplitude fading affects the signal the least can be found. This is what is done in Sec. 6.1 for our own system design. As it happens with the chromatic-dispersion-generated amplitude fading, the spectral location of the affected subcarriers depend on the fiber length, which enables one to think that full spectrum utilization can be achieved in a Passive Optical Network if there are ONUs located at several different distances from the OLT, as suggested already when considering only the first studied type of amplitude fading.

5.6– Drive amplitude

The drive amplitude makes reference to the magnitude of the signal with which the laser is modulated. It is a value of considerable importance, because it is related, as already commented before, with the detected signal amplitude, and with nonlinear effects in the fiber, mainly through the carrier optical power. In order to study it, first let us analyze the direct modulation of a laser. A DML transfer function can be seen in Fig. 5.22. The driving voltage $v(t)$ is got from the OFDM electrical voltage signal passing through a laser driver module, that multiplies its input signal $s(t)$ by a gain value $m$. So the output of
5.6 Drive amplitude

(a) Normalized Carrier Frequency = 0.8.  
(b) Normalized Carrier Frequency = 1.75.  
(c) Normalized Carrier Frequency = 2.

Figure 5.21: Detected signals severely affected by amplitude fading. The carrier frequencies are divided by $\frac{3}{4}$ Bandwidth in order to get the Normalized Carrier Frequency.

Figure 5.22: Laser diode response.

the driver is $v(t) = m s(t)$, which is transformed into a driving current $i(t)$ in the laser, by the driver transconductance $G$ (see Fig. 5.23). So, if $i(t) = G v(t)$, the output optical

Figure 5.23: OFDM coder, Laser Driver and Laser.
power of a directly modulated laser in its linear regime is given by:

\[ P(t) = \eta (I_B - I_T + i(t)) \]  

(5.13)

where \( I_T \) is the threshold current, \( I_B \) the bias current and \( \eta \) the slope efficiency, as seen in Fig. 5.22. Then, replacing \( i(t) \) in Eq. 5.13:

\[ = \eta (I_B - I_T) + \eta G v(t) \]  

(5.14)

\[ = P_o \left( 1 + \frac{1}{I_B - I_T} G v(t) \right) \]  

(5.15)

Taking the following values results in:

\[ I_T = 0.02 \text{ A} \quad I_B = 0.06 \text{ A} \quad \eta = 0.2 \text{ W/A} \quad G = 0.04 \text{ A/V} \Rightarrow \frac{G}{I_B - I_T} = \frac{1}{N} \]  

(5.17)

\[ P_o = \eta (I_B - I_T) = 8 \text{ mW} \]  

(5.18)

So we get:

\[ P(t) = P_o \left( 1 + \frac{v(t)}{V} \right) \]  

(5.19)

where \( V \) is the [Volt] symbol.

In the way it is generated by the programmed transmitter in VPI, its outcome signal \( s(t) \) is normalized so that the maximum possible temporal sample amplitude is 1 \( \text{V} \). This means that \( s(t) \) oscillates in the range [-1;1]. Let us see, on the other hand, which is the maximum range in which \( v(t) \) should oscillate in order not to incur in nonlinear distortion during the modulation. For this, we can assume (as it is very possible to happen) that the closest source of distortion is when it causes \( i(t) \) to escape the linear region shown in Fig. 5.22 by being smaller than \( I_B - I_T \). So for the minimum wanted \( i(t) \):

\[ I_B + i_{\text{min}} \geq I_T \]  

(5.20)

and reminding that:

\[ i_{\text{min}} = G v_{\text{min}} \quad \text{and} \quad v_{\text{min}} = m s_{\text{min}} = m (-1) \]  

(5.21)

\[ \Rightarrow I_B - m G \geq I_T \Rightarrow m \leq \frac{G}{I_B - I_T} = 1 \]  

(5.22)

So we see that the maximum conceivable value of \( m \) for avoiding nonlinear distortion is 1. If the maximum value of 1 is taken, then in the case when the OFDM signal \( s(t) \) has amplitude of 1, the laser will be driven to the extreme of its linear regime.

However, OFDM signals consist of the sum of many independently modulated subcarriers. So when the number of subcarriers \( N \) is large (greater than 64 is usually set as enough), central limit theorem applies and the real and imaginary parts of the samples at the output of the transmitter IFFT have Gaussian distribution. The probability density function and cumulative distribution of each sample power \( |x_m|^2 \) can be observed in Fig. 5.24. Although OFDM has indeed high signal peaks, these occur relatively rarely. As depicted in Fig. 5.24b, less than one out of 1000 values is 8 dB above the mean [7].

As a result of this characteristic of OFDM signals, when these are normalized against their maximum possible value, as is usually done before subsequent transmitter stages, signals with dynamic range notably smaller than the maximum possible of [-1;1] are obtained. Fig 5.25 shows an electrical OFDM signal in time. As a result, larger values of \( m \) are suitable to be taken, in order to take advantage of a wider portion of the linear region of the laser.

\footnote{This normalization involves the \( \frac{1}{N} \) factor due to the IFFT calculation, and one that depends on the mapping scheme. For each QAM type, a different factor is used, which enforces that the symbols located the furthest away from the origin in the complex plane are at a distance of 1.}
5.6 Drive amplitude

(a) Probability density function.  
(b) Cumulative distribution.

Figure 5.24: Distribution power of baseband OFDM signal. [7]

The parameter \( m \) is called **electrical modulation index** (EMI). As it is introduced in the simulation environment by the laser driver, we call it as well **drive amplitude**, as the module parameter. We can freely choose this parameter’s value, thus producing a wider or narrower sweep in the laser response curve. However, care should be taken, because as soon as \( m > 1 \) is imposed, for some bit sequences transmitted (albeit few when \( N \) is large), the laser will escape the linear regime. Doing so results in **clipping** (a nonlinear effect we want to avoid). This phenomenon has been studied in order to avoid its effects hence lowering the optical power [65], [66]. However, the approach proposed by us consists in letting \( m \) be greater than 1 upto a point in which clipping worsens the performance to a greater degree than the benefits brought by a higher modulation index.

The optimum drive amplitude (or electrical modulation index) \( m \) for a point-to-point link has been found out by simulations, yielding an optimal EMI of 1.25 for a bandwidth of 2.5 GHz (5 Gbps with QPSK modulation), emulating an ONU of the multi-user architecture proposed (128 subcarriers). Back-to-back configuration has been set, as the DML behavior is to be assessed as isolated as possible. Fig. 5.26 shows the incidence of varying the drive amplitude on the error vector magnitude of the received signal. 5 different pseudo-random sequences have been used, as the extreme values of the baseband signal strongly depend on these. 256 symbols were simulated for each run.

Having explained the electrical modulation index \( m \), now its relation with the **optical modulation index** (OMI) is discussed. The OMI is defined as the amplitude ratio between
the first sideband and the optical carrier. The electric field amplitude after direct laser modulation is given by:

\[ E(t) = \sqrt{P(t)} = \sqrt{P_0} \sqrt{1 + m s(t)} \approx \text{Taylor series exp.} \ E_o \left( 1 + \frac{m}{2} s(t) - \frac{m^2}{8} s(t)^2 + \ldots \right) \] (5.22)

If \( m s(t) \ll 1 \), which happens in a great majority of cases, the small-signal approximation can be taken, and the higher order terms discarded, resulting in:

\[ E(t) \approx E_o \left( 1 + \frac{m}{2} s(t) \right) \] (5.23)

As discussed in Sec. 5.3, intensity modulation produces a double sideband optical signal, i.e. sidebands at both sides of the optical carrier, as can be noted in Eq. 5.24 for the case of a modulation with a single tone:

\[ E(t) = E_o \left( 1 + \frac{m}{2} \cos(\omega t) \right) \]

\[ \Rightarrow E(f) = E_o \left[ \delta(f) + \frac{m}{4} \left( \delta \left( f - \frac{\omega}{2\pi} \right) + \delta \left( f + \frac{\omega}{2\pi} \right) \right) \right] \] (5.24)

So as the OMI is measured as the ratio between one of the sidebands and the optical carrier, the relation between OMI and EMI is given by (remember that \( m = \text{EMI} \)):

\[ \text{OMI}_{\text{max}} = \frac{\text{EMI}}{4} \] (5.25)

The fact of the OMI estimated in Eq. 5.25 being the maximum is because, as discussed above, it depends on whether the modulating signal has extreme values of -1 and +1, as does the cosine used in Eq. 5.24. In dB, this relation corresponds to:

\[ \text{OMI}_{\text{max}}|_{\text{dB}} = \text{EMI}|_{\text{dB}} - 6 \text{ dB} \] (5.26)

One way to measure the OMI is to modulate the laser with a signal that presents maximum values of 1. With OFDM signals, this can be done by inputting a sequence of consecutive 1s to the coder. This, after QPSK modulation and IFFT, gives a coded temporal output of a maximum value of 1 followed by 0s, as seen in Fig. 5.27. Measuring the maximum
value of the optical signal ($P_o + m$) and the average one ($P_o$), $m$ can be straightforwardly calculated as the difference in dB between them. As in this case having a maximum value of $s(t)$ of 1 is ensured, the OMI is simply 6 dB below $m$. This means that the optimal OMI$_{max}$ found is 0.3125.

Figure 5.27: Directly modulated laser output seen in time. A method for measuring the EMI, OMI and PAPR is depicted.

Fig. 5.27 can be used as well to introduce another important concept related to the two already mentioned; one usually referred to in relation to OFDM. It is the peak-to-average power ratio (PAPR), which is calculated as $\text{PAPR} = \frac{P_o + m}{P_o}$.

5.7– Nonlinear effects

Nonlinear effects were precisely one of the main problems O-OFDM was accused of suffering irreparably in its first years. However, it has been readily shown that the numerous intermodulation products that it is supposed to yield do not constitute a fatal problem [6]. There are some insightful publications that study this aspect, and for a formal treatment the reader should consult, for instance, [16].

As a mathematical treatment of nonlinear effects shall take long time and effort, in this work, they have been taken into account by setting the corresponding parameters in VPI, following influential work in the field ([75], [42]). However, in order to show that the nonlinear effects do not have a remarkable influence in a system working at the bitrates and optical power we deal with, Fig. 5.28 shows the received constellations of an OFDM system with our default characteristics (see Sec. 4.1), with and without nonlinear effects taken into account, with a launching power of 4 mW.
Figure 5.28: Received constellations after equalization. BER = 0 for both cases for the number of transmitted symbols, and EVMs differ in less than 2%.
Chapter 6

Multipoint-to-point O-OFDMA

The objective of this chapter is mainly to test the next-generation passive optical network architecture that we have proposed. Its detailed explanation can be found in Chapter 3. However, for a correct dimensioning, several aspects of the underlying phenomena have to be understood. In the last chapter, the ones affecting a point-to-point DML IM/DD O-OFDM system were discussed. In the first part of this chapter, the ones related to multipoint-to-point architecture are considered. With those results in mind, we later elaborate on system dimensioning.

The results of the second part of this chapter have been obtained from simulations used for emulating a system built in the lab. The results of that work can be found in [72]. This system is in almost every sense identical to the scenario used for the simulations presented here. The slight existing differences, which are due to the wider possibilities of a simulation environment, are mentioned when there are any. The similarity in the results of both works is an indicator of the importance of a tool like photonic systems simulation, which enables to predict the performance of practical systems, while offering wider tunability in diversity of parameters, not suffering from the many impairments related to laboratory work.

6.1– Electrical carrier frequencies

The location of the OFDM band in the optical spectrum is an important design feature in such a system. As studied in Secs. 5.3, 5.5 and 5.4, if the sideband is located far from the optical carrier in frequency, as double sideband modulation followed by direct detection is used, amplitude fading takes place (Sec. 5.3), depleting the received amplitude of part of the subcarriers. On the other hand, the bands cannot be located too close to the optical carrier because certain frequency guardband is needed (Sec. 5.2). Therefore a compromise should be reached regarding the sideband spectral position and thus the electrical carrier frequencies. The optimum position of the sideband has been found by sweeping the electrical RF frequencies coherently (note however that they do not coincide in value) along the possible range, and finding the minimum EVM/BER figures. This is shown in Fig. 6.1, where the x-axis corresponds to:

\[
\text{Normalized Carrier Frequency} = \frac{\text{Carrier Frequency}_{\text{OLT}}}{\frac{3}{2} \text{BW}_{\text{OLT}}}
\]

and the relation between each of the ONUs’ carrier frequencies and the OLT one is given by:

\[
\begin{align*}
\text{Carrier Frequency}_{\text{ONU1}} &= \text{Carrier Frequency}_{\text{OLT}} - \frac{\text{BW}_{\text{OLT}}}{2} + \frac{\text{BW}_{\text{ONU1}}}{2} \\
\text{Carrier Frequency}_{\text{ONU2}} &= \text{Carrier Frequency}_{\text{OLT}} + \frac{\text{BW}_{\text{OLT}}}{2} - \frac{\text{BW}_{\text{ONU2}}}{2}
\end{align*}
\]

where \(\text{BW}_{\text{OLT}} = 5 \text{ GHz}\) and \(\text{BW}_{\text{ONU1}}\) vary.

The Normalized Carrier Frequency that yields the best EVM and BER results is around 1.75. This means that \(\text{Carrier Frequency}_{\text{OLT}} = 1.75 \times \frac{3}{2} \text{BW}_{\text{OLT}} = 13.125 \text{ GHz}\), then Carrier
Frequency\textsubscript{ONU1} = 11.875 and Carrier Frequency\textsubscript{ONU2} = 14.375 GHz when both ONUs are assigned 64 subcarriers each. For different number of subcarriers assigned to them, Carrier Frequency\textsubscript{OLT} remains the same, but the other two change correspondingly such that the sideband is always well formed.

6.2– Optical beat interference and optical carriers separation

We have seen how the optical upstream signal received by the OLT is constituted by several optical carriers at different wavelengths, each modulated by a RF-upconverted electrical signal. Upon detection at the OLT of \( M \) user signals, as the detected photocurrent can be expressed as:

\[
i(t) = \frac{R}{2} \left| \sum_{k=1}^{M} E_k(t) \right|^2 = R \left( \sum_{k=1}^{M} P_k(t) + \sum_{n=1}^{M} \sum_{k \neq n} \sqrt{P_k(t) P_n(t)} \cos(2\pi(v_k - v_n)t + \phi_k(t) - \phi_n(t)) \right) \cos\theta_{kn}
\]

(6.1)

where \( R \) is the photodiode responsivity and \( \theta_{kn} \) is the angle between the two polarization directions.

\[
E_k(t) = \sqrt{2P_k(t)} e^{j(2\pi v_k t + \phi_k(t))}
\]

(6.2)

is the \( k^{th} \) optical carrier amplitude, with frequency \( \nu_k \), optical phase \( \phi_k \) and an optical power of \( P_k = \frac{E_k^2}{2} (1 + m_k v_k(t)) \). Here, \( m_k \) the electrical modulation index and \( v_k(t) \) the electrical data signal. For an detailed mathematical formulation of this, the reader is referred to [12].

The first term in Eq. 6.1 shows a DC component, plus a band that is proportional to the electrical data signal, i.e. the coded information. The second summation, on the other hand, expresses a phenomenon called \textbf{optical beat interference} (OBI). These are products located around the difference frequencies \( \nu_k - \nu_n \). Their electrical power spectrum
shape results from the convolution of the beating signals optical spectra. The cosine term is related to the angle between the polarizations of the two laser sources, and it can be taken as 1 in a worst case scenario. An example of two laser sources being modulated by electrical signals producing OBI at detection can be seen in Fig. 6.2.

![Figure 6.2: OBI produced by two optical signals at different wavelengths detected together.](image)

It is useful in order to understand the complex phenomenon of OBI to analyze the case of two uncorrelated optical sources on the same wavelength. In such case Eq. 6.1 would give the received signal for each user as:

\[
i_1(t) = R \left( P_1(t) + \sum P_1(t) P_2(t) e^{j(\phi_1(t) - \phi_2(t))} \right) \quad (6.3)
\]

\[
i_2(t) = R \left( P_2(t) + \sum P_2(t) P_1(t) e^{j(\phi_2(t) - \phi_1(t))} \right) \quad (6.4)
\]

where the frequency difference term has vanished, and \( \cos \theta_{kn} \) has been taken as 1. Each source corresponds to an ONU modulating all of their subcarriers with QPSK modulation format. The received constellations can be observed in Fig. 6.3. One way to understand what happens here is to think of the photodiode as performing two detections: one with a correlated optical carrier, and one with an uncorrelated one. From this point of view, the constellations can be seen as the sum of that of a properly received QPSK transmission (correlated carrier), plus another one with random phase. It should be reminded that QPSK modulation yields constant amplitude for every symbol, and thus the constellation of symbols suffering random phase shifts would look as a circle around the origin. Detection of an OFDM band with an uncorrelated optical carrier is equivalent to suffering random phase shifts; hence the circles that appear around each of the positions where the properly detected (with the correlated carrier) symbols would be placed.\(^1\)

The situation is even worsened when more ONUs and therefore laser sources share the same physical medium. This suggests that successful performance of such a system without a strategy for alleviating OBI is practically impossible. The aim of such a technique should be somehow to avoid the aforementioned multiple detection situation.

Several OBI-mitigation techniques have been proposed in the literature, depending on the relation among the wavelengths used by the different ONUs ([73], [74]). If the approach is to assign every ONU the same nominal wavelength, one possibility is to filter out the optical carriers at the receiver and to add one prior to photodetection. The effect of this

---

\(^1\)The reason for the circles not appearing complete is the limited time window assigned to the simulation. Had the simulated time been longer, the circles would appear complete.
is having only one carrier for photodetection, with which multiple detection is avoided. A phase shift term due to the phase difference between the transmitters and receiver lasers would still appear, but it can be corrected.

Another possibility could be to include a second optical carrier at the other side of the OFDM band, as shown in Fig. 6.4. By carefully choosing the bands’ electrical frequencies so that detection products due to the two wavelengths do not overlap, the detected sidebands do not present multiple detection effects, even when the transmitting optical carrier is not eliminated. A phase noise term is present here as well, but it can be corrected. A great advantage of this technique is that the modulation becomes single sideband, as there are no mirrored copies of the OFDM bands with respect to the detection optical carrier.

As it has been explained, the solution to the OBI problem proposed in our approach is to let the free-running laser sources emit at wavelengths as far away from each other as to avoid that the unwanted mixing products fall inside the OFDM sideband after detection. So, analyzing Fig. 6.5, one can see that:

$$\Delta f_1 = \Delta f_\lambda - 2 \left( BW_{OLT} + GB \right)$$  \hspace{1cm} (6.5)
where $\Delta f_1$ can be understood from Fig. 6.5, $\Delta f_\lambda$ is the frequency separation between optical carriers, $BW_{OLT}$ is the width of the total sideband that the OLT receives, and $GB$ is the guard band. In order for the mixing products of the two adjacent bands from different users to fall at a frequency that does not overlap with the OFDM bands after detection:

$$\Delta f_1 > BW_{OLT} + GB$$  \hspace{1cm} (6.6)

Putting together Eqs. 6.5 and 6.6:

$$\Delta f_\lambda > 3 (BW_{OLT} + GB)$$  \hspace{1cm} (6.7)

which gives the minimum separation between the optical carriers needed.

The guard band topic needs further discussion, as we are dealing now with signals of variable spectral width located at variable distances from their corresponding optical carriers. In this wavelength-distributed approach, the total OFDM band is only contiguous after detection, but not in the optical domain. This means that the maximum frequency of the SSMI is only equal to the maximum signal bandwidth that can be assigned to a single user. Then, we do not need to leave a frequency guard band as wide as the entire OFDM band recover at the OLT, as it would be necessary in a point-to-point system that employed such bandwidth. This allows us to reduce the required guard band. Fig. 6.6 may be of worth for understanding this situation. Note that the guard band set is as wide as the maximum user bandwidth.

Let us now put some figures in order to get an idea of the carrier separation needed in the scenario to simulate. Regarding the guard band $GB$, as a main objective of this chapter is to extensively test the Dynamic Bandwidth Assignment technique, we will conservatively consider that the maximum electrical bandwidth given to an ONU is equal to the whole OLT OFDM band. This fact, in a real case, would cancel some of the advantages of the OOFMDA-PON principles, like the ONUs needing slower components than the OLT. Using the figures mentioned in Sec. 4.2, $BW_{OLT} = 5$ GHz and $GB = 5$ GHz as well. So, a carrier separation of $3 \times (5 \text{ GHz} + 5 \text{ GHz}) = 30$ GHz needs to be taken.

Although the analysis done in the last paragraphs based on Fig. 6.5 represents an useful first approach towards the OBI phenomenon and how to avoid it by setting the optical carriers far enough in wavelength, it may be insufficient. This is due to the nature
of Intensity Modulation, which because of its square-root nature, produces sidebands other than just the first one, as happens in Amplitude Modulation. Given certain conditions, as explained in Sec. 5.6, IM can be considered as AM and thus superior sidebands can be ignored. However, those conditions are not fulfilled in our setup, mainly because of the large Electrical Modulation Index used. As a consequence, higher-order sidebands need to be taken into account, and the spectrum sketched in Fig. 6.5 turns out to be not completely accurate.

As sidebands other than the first should be considered in the spectrum, larger wavelength separations may need to be set. Formal mathematical analysis in this regard can result intricate, because it implies relating lasers spectral shape and width, EMI, OMI, time-domain signals of each ONU, DBA configuration and other variables. On the other hand, an optimum $\Delta \lambda$ can be experimentally obtained as the one for which the BER values of both ONUs remain at a desirable range, but the wavelengths are close enough as to make good use of the photodetector optical bandwidth. As one of the objectives of the present work is to emulate the results obtained in [72], the chosen $\Delta \lambda$ is the one used there, which was empirically obtained in the laboratory, being the value for which the system yielded the best performance. This value is $\Delta \lambda = 0.5252 \text{ nm} = 65.2277 \text{ GHz}$.

### 6.3 – Drive amplitude

In order to obtain the optimum Drive Amplitudes for the two DML to use, a simulation is performed, where 64 subcarriers have been assigned to each of the ONUs, from a total spectrum of 5 GHz. Back-to-back configuration is implemented, for analyzing only the effects of varying the drive amplitude. Error vector magnitude turns out to be a better estimator of the performance than the bit error rate once again, since the latter is 0 for several regions of the simulated domain. Fig. 6.7 presents the simulations’ outcomes, that show that EVM stabilizes around a minimum value for drive amplitudes greater than 10 approximately.

Henceforth, a drive amplitude of 20 is used, in order to work inside the best performance regime, as seen in Fig. 6.7. Greater drive amplitudes are not wanted, for avoiding too big signal amplitudes, having in mind that, though seldom, OFDM symbols having large extreme values do appear, and that departing from the linear regime is detrimental for the system. Fig. 6.8 shows that in the 64 simulated symbols, no clipping appears, and in fact the extreme values of the temporal signal are still far away from the zero output power level.
6.4 Dynamic bandwidth assignment and variable received power

A key concept of optical orthogonal frequency division multiple access is the possibility of assigning different number of subcarriers to optical network units. This feature is called dynamic bandwidth assignment (DBA). The general scope of this feature has been discussed in Sec. 2.4. In the following, we study the effect of having a variable number of subcarriers per ONU.

The ONUs having assigned different number of subcarriers would be unimportant to the OLT, if the path lengths were quite similar for all ONUs. In that case, the OLT would have to be aware of which subcarriers correspond to which ONU, but the situation would not result in extra impairments when it comes to demodulation. On the other hand, if path
Multipoint-to-point O-OFDMA

lengths are considerably different, as it would happen in real cases since users are located at variable distances from the optical splitter, this results in disparate attenuation, dispersion, time delay and phase shift between the user upstream signals. Moreover, this happens even when physical distances are not too different, but distinct types of components (optical splitters, for instance) compose each one’s last mile\(^2\).

In the case of phase shift, the equalization step at the receiver, as it acts on subcarriers independently, corrects it as if it were a point-to-point signal. A similar situation happens with chromatic dispersion. Regarding time delay, we assume synchronization is performed by a mechanism that is not going to be explained here, but that has been performed in the laboratory experiment, as can be read in [72]. Therefore, we concentrate our efforts in analyzing the effects of the ONUs’ signals experiencing different attenuation, together with them having diverse number of subcarriers.

When the number of subcarriers assigned to a user changes, its carrier frequency need to vary too, in order to properly locate the ONU signal inside the OFDM band. In this 2-users system, as all subcarriers are constantly assigned to a certain user, both CFs change when the subcarrier assignment is varied, as explained in 6.1.

Different subcarrier assignment configurations have been taken, namely 4/124, 8/120, 16/112, 32/96, 64/64, 96/32, 112/16, 120/8 and 124/4 subcarriers for ONU1/ONU2 respectively. For each of them, the optical power that ONU2 inputs to the coupler is varied, while that of ONU1 remains constant. \( \Delta P_{\text{dB}} = P_{\text{ONU1dBm}} - P_{\text{ONU2dBm}} \) has been used for plotting the obtained BERs and EVMs as a function of the varying power ratio. The total optical spectrum is displayed in Fig. 6.9, and the curves resulting from the simulations in Figs. 6.10 to 6.16. In the case of the BER curves, as the vertical axis is in logarithmic scale, the values of BER of 0 were assigned the value \( 10^{-7} \) for plotting.

An important feature of OFDMA-PONs to test with an experiment of this kind is the power control mechanism. In order to maintain an optimal performance in the network, the OLT needs to control the power emitted by each of the ONUs, so avoiding extreme interference between them, due to some lasers emitting too high powers. Nonlinear effects in the fiber are avoided by a strategy like this as well. Although we do not discuss such a mechanism here, we assume that it is to be carried out by the MAC protocol, as with the DBA. Nevertheless, the results obtained here are of fundamental importance for it, since they foresee the precision of the needed power control.

From the analysis of Fig. 6.13, interesting results already arise. When both ONUs have equal bandwidths (same number of subcarriers) ONU1 shows a better performance. This can be noticed in the fact that the shape of the BER curves is almost identical for ONU1 and ONU2, but when having equal powers (\( \Delta P = 0 \text{ dB} \)), ONU1 has a much lower BER. The useful range in which both ONUs yield a BER of less than \( 10^{-3} \) occurs for a range of \( \Delta P \) of over 5 dB for the 64/64 subcarriers case. This means that the difference of received powers between the ONUs at the OLT can vary in more than 5 dB and still result in successful transmission by both. This sets how efficient and fast the power control mechanisms implemented by the OLT to avoid interference between users needs to be.

Generally speaking, BER curves slope increases for lower numbers of subcarriers assigned to the corresponding user. In the cases when an ONU is assigned a high number of subcarriers, its BER curve turns out quite flat. This means that when some ONUs are assigned a narrower portion of spectrum to transmit, power control techniques become more important, since small variations in \( \Delta P \) lead to important variations in the number

\(^2\)The term ‘last mile’ is used to designate the part of the access networks that is closest to the ONU. Here, it means the fiber after the optical splitter. That is to say, the one that is uniquely used by a particular ONU.
6.4 Dynamic bandwidth assignment and variable received power

Figure 6.9: Optical power spectrum after joining the signals coming from the 2 ONUs.

(a) $\Delta P = 0$ dB

(b) $\Delta P = 3$ dB

(c) $\Delta P = -3$ dB

Figure 6.10: ONU1: 4, ONU2: 124 subcarriers.

(a) Bit Error Rate vs $\Delta P$

(b) Error Vector Magnitude vs $\Delta P$

Figure 6.11: ONU1: 16, ONU2: 112 subcarriers.

(a) Bit Error Rate vs $\Delta P$

(b) Error Vector Magnitude vs $\Delta P$

Figure 6.12: ONU1: 32, ONU2: 96 subcarriers.
Figure 6.13: ONU1: 64, ONU2: 64 subcarriers.

Figure 6.14: ONU1: 96, ONU2: 32 subcarriers.

Figure 6.15: ONU1: 112, ONU2: 16 subcarriers.

Figure 6.16: ONU1: 124, ONU2: 4 subcarriers.
of errors at reception. However, as a general trend, ONU2 BER curves are less steep than those of ONU1. This can be attributed to the OFDM band of former being located at higher frequencies. Therefore, it suffers the effects of chirp in a stronger way.

As a way to analyzing the obtained curves, some parameters have been measured on the curves, in order to compare them and enable further insight. \( \Delta(P) \) has been defined as the range of \( \Delta P \) for which both users yield BER of less than 10\(^{-3}\). \( X_{EVM} \) is the value of \( \Delta P \) at which the two EVM curves intersect. The results are shown in Table 6.1.

The figures in Table 6.1 show that the different DBA setups do not result in immensely different performances. They show that the useful range is located at fairly the same values of \( \Delta P \). However, it should be noticed that it is the total power of ONU2 the one that is being varied. This results in remarkably different power per subcarrier among diverse DBA configurations. For example, for a constant value of \( \Delta P \), the same optical power can be divided among the carrier and 4 subcarriers, or among the carrier and 124 subcarriers.

One may think that in order to get similar performances, the power per subcarrier should be similar. However, comparable performances are seen at similar values of \( \Delta P \). What different DBA configurations with the same \( \Delta P \) value do have in common, as opposed to the per-subcarrier power, is the total power, as well as the optical carrier power, which depends only on the laser parameters (see \( P_o \) in Eq. 5.17, Sec. 5.6). So from this it is seen that at constant drive amplitudes, it is rather the optical carrier power than the power per subcarrier what defines the system performance. A consequence of this is that ONUs may not need to change their laser output power when having their assigned bandwidth modified. This is due to there being a range of around 2 dB of \( \Delta P \) for which any DBA configuration works properly, and to the fact that emitted power does not change significantly when a different number of subcarriers is modulated.

As a result of these simulations, we have proved that a DBA mechanism is suitable to be put in practice in an OFDMA-PON like the proposed. Such a feature would work for ONUs whose signals arrive to the OLT with quite different optical powers. The range in which the power difference can vary and still an acceptable performance be obtained is shown in Table 6.1 for different bandwidth allocations. In general, they oscillate in the order of 4 to 5 dB. Therefore, control power mechanisms will have considerable margin for operation, without needing to be neither extremely fast nor precise. This constitutes another strength for OFDMA-PONs.

<table>
<thead>
<tr>
<th>SCs: ONU1_ONU2</th>
<th>4_124</th>
<th>16_112</th>
<th>32_96</th>
<th>64_64</th>
<th>96_32</th>
<th>112_16</th>
<th>124_4</th>
</tr>
</thead>
<tbody>
<tr>
<td>( \Delta(\Delta P) )(_{10^{-3}} ) [dB]</td>
<td>(-1.75 : 2)</td>
<td>(-3.1 : 2.4)</td>
<td>(-3 : 2.4)</td>
<td>(-3.5 : 1.8)</td>
<td>(-3.5 : 0.75)</td>
<td>(-2.5 : 2.5)</td>
<td>(-2.5 : 0.4)</td>
</tr>
<tr>
<td>( X_{EVM} ) [dB]</td>
<td>-0.9</td>
<td>-0.9</td>
<td>-0.8</td>
<td>-0.9</td>
<td>-1.5</td>
<td>-1.2</td>
<td>-1.1</td>
</tr>
</tbody>
</table>

Table 6.1: Measurements on curves of Figs. 6.10 to 6.16.
Chapter 7

Conclusions

As a result of the study, experimentation, analysis and debate described throughout the present work, some meaningful conclusions can be drawn:

- We have described the context in optical communications and how OFDM has already proved to be a very competent technology for high-capacity long-haul systems. Due to this success and based on further advantages of OFDM compared to other technologies, we have shown why OFDMA appears as a promising multiple access technique for next-generation passive optical networks, highlighting its key point for addressing their particular needs.

- We have studied the entire process of OFDM modulation, transmission, reception and demodulation. The different possibilities in many of the functional parts of the system were compared, taking into account state-of-the-art techniques. After this comparison, we have proposed a NG-PON architecture, detailing both fundamental and more technical aspects of remarkable importance. The driving principle when choosing among diverse options has been cost-effectiveness and compatibility with already installed fiber deployments. As a result, a directly modulated laser intensity modulation and direct detection OFDMA-PON has been proposed.

- In order to study the underlying phenomena in such a network, the photonic systems simulation software VPItransmissionMaker\textsuperscript{TM} has been used. As existing tools were not flexible and scalable, and could not control important features of a system like the design, new tools have been created. These include new VPI scenarios for emulating the network, transmitter and receiver, as well as Matlab functions for high-tunability coders and decoders. With these, a new level of freedom for O-OFDMA simulation has been reached. These new tools can make further research in the field easier, and should constitute a stepping stone from where to continue improving O-OFDMA simulation at UPC.

- We have studied the phenomena affecting a single-user and multi-user OFDM system through mathematical formulation and software simulations. We have seen that due to the nature of the proposed network architecture, some of them affect our case in a clearly distinct way, compared to other systems reported. Chromatic dispersion has proved to affect our system by inducing different phase shifts on the subcarriers. On the other hand, it also brings about the requirement of a frequency guard band. Also, especially when combined with laser chirp and phase noise, CD imposes some important constraints on the system performance, in the form of frequency selective amplitude fading due to double sideband modulation. Also, when multiple users share the same fiber, the wavelength at which they transmit imposes a problem, causing OBI if they are close. If the users remodulate a unique carrier sent from the OLT, other problems arise, and phase noise still takes place. All this suggests drawbacks of our proposal, despite its advantages. However, the restrictions imposed by its particular characteristics can be alleviated with techniques that can be formulated with proper understanding of the underlying phenomena.
The strategies to overcome the proposal’s limitations have been explained. The phase shifts induced by CD can be corrected with 1-tap equalization. Due to the urban scenario under study, the necessary frequency guard band can be reduced considerably. Amplitude fading cannot be completely eliminated, but the spectral location of the OFDM band can be chosen in order to avoid the most affected frequencies. Moreover, as different users are located at variable distances from the OLT, the entire spectrum can be utilized, because prohibitive frequency ranges differ with fiber length. Also, when the different users utilize wavelengths that are separated by a sufficient spectral distance, OBI disappears. Therefore, based on realistic NG-PON necessities and characteristics, and putting into practice the mentioned strategies, an operating cost-effective and legacy-network-compatible system has been proved.

As users sharing the optical distribution network may be at considerably different distances from the optical line terminal, and may be affected by distinct attenuations, the OLT needs to be able to cope with receiving different signal levels from the ONUs. This has been tested over a wide range of power differentials between the users. The results and their implications have been discussed.

Moreover, a fundamental feature of NG-PON like dynamic bandwidth assignment, which is intrinsic of OFDMA and so constitutes an important advantage, has been successfully implemented. ONUs have been assigned different numbers of subcarriers, emulating different service requirements, and thus bandwidth needs. This has been combined with different attenuations suffered by the users’ signals, in order to yield an estimation of the performance of a NG-PON featuring dynamic bandwidth assignment with high-granularity. The results obtained have been commented with regard to their meaning for the evaluation of OFDMA as a competent technology for NG-PON. They show that regardless of the specific bandwidth allocation for each user, the system presents a considerably wide range of attenuation differences that the diverse signals can suffer, still ensuring correct transmission.

The performed simulations show that for the target BER that would ensure effective communications, a system working at 10 Gbps for a distance of 25 km is feasible. It has been designed employing directly modulated lasers intensity modulation and direct detection, and no optical filters are used, which corresponds to a very cost-effective proposal, which is as well compatible with current IM/DD installed systems. Moreover, the system can support fundamental PON features, which have been proved to work over a considerable range of received optical powers.

Based on the study and obtained results, we conceive this work as a prove for the suitability of directly modulated laser intensity modulation and direct detection OFDMA as a competent technology for next-generation passive optical networks.
Chapter 8

Future lines of work

While doing the preliminary studies, the development of the simulation scenarios and related programming code, the simulations themselves and the present report, several possible lines of work have sprung. Some of them, albeit not having been considered in the beginning, have found their way into the present work. Others, due to lack of time and impossibility of covering such distinct topics, phenomena, scenarios and techniques, have been left aside. In the following, some of them are commented.

- Completely flexible multi-user scenario. Although for the matters of this work the developed VPI scenarios and Matlab codes were sufficient, many improvements can still be done to them. One major task to tackle, although it may require months or at least several weeks of work, is the development of a VPI universe that support a flexible number of users. While this may seem a simple extrapolation of what has been done, it is not. As far as VPI is concerned, parameters have to be treated in vectorial form, with indexing and relating to TCL code in a not quite straightforward fashion. Concerning Matlab, it means employing 3-dimensional arrays, and requires functions intrinsically built for that purpose.

- Simulations with more than 2 users. Simulations done in Sec. 6.4 should be performed for a larger number of users, ideally utilizing a multi-user scenario like the one just described.

- Higher capacity tests. The experimental parts of the present work have been focused in the analysis of the phenomena affecting systems like our proposal, and features of PONs that can be addressed with them. Therefore, capacities have not been pushed to the limit, which can be done in order to investigate the maximum capacity that such a system could offer.

- Detection carrier insertion. While studying OBI mitigation techniques, the possibility of a detection-carrier insertion came about (Fig. 6.4). This carrier should be placed on the other side of one of the optical sidebands, i.e. opposed to the transmission-carrier. Direct detection in this case is that of an optical single sideband signal, since there is no other sideband with respect to the detection-carrier. This would alleviate the drawbacks related to double sideband detection, and deserves further study.

- Adaptive Modulation. A necessary technique to implement when testing high capacities, so much so that it is increasingly uncommon to see this not implemented in publications that force very high bitrates. This should be implemented in the scenarios we have developed. Coders and decoders with this feature have been developed at UPC for point-to-point scenarios. Merge is needed in order to get improved-capacities multi-user scenarios capable of testing the limits of OFDMA regarding bitrate.

- Chirp alleviation. A chirp compensation technique has been very recently proposed for directly modulated laser and direct detection OFDM [69]. This possibility should be taken into account and, if not implemented, at least it should alter the view on chirp regarding its pernicious effects.
• Phase noise alleviation. There have been proposals of phase noise alleviation techniques [48], which should be analyzed, discussed and incorporated in the model and simulation scenario.
Bibliography


[63] IEEE 802.11.a standardization (ISO/IEC 8802-11:1999/Amd 1:2000(E)).


