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Arbitrary waveform generation based on
Microwave Photonics Technology for Ultrawideband
applications

TESIS DOCTORAL

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*This dissertation is dedicated to my parents: Silvana
and José Luis, to my siblings Andrea and Luis to my partner
Emilio and to my beloved grandma: María Alejandrina*

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*“Nothing in life is to be feared, it is only
to be understood. Now is the time to understand
more, so that we may fear less”*
Marie Skłodowska-Curie

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Abstract

The herein presented Ph.D. dissertation finds its application niche in pulse generation for optical communication schemes, specifically for Ultrawideband (UWB) purposes. In this sense, as the requirements in terms of capacity and bandwidth per user in the field of broadband communication services continuously increase, different technological techniques such as hybrid wireless-optical approaches including UWB systems and close competitors like the Worldwide Interoperability for Microwave Access (WIMAX) standard, Long Term Evolution (LTE) and the newcomer Light Fidelity (LI-FI) continue to be investigated as viable solutions to provide data rates exceeding Gigabit per second per user. Currently, Hybrid wireless-optical networks are the key providers for simpler backhaul and they are destined to play an important role in next-generation access networks which will demand flexible deployment, high capacity, upgradeability, scalability and feasibility.

On the other hand, Arbitrary signal creation is a fundamental affair since it is widely employed for a myriad of practical approaches such as imaging layouts, radars, instrumentation, biomedical design and communications purposes. During the last years, all the efforts have been focused on proposals for direct generation within the optical domain because of the limitations presented by traditional electrical solutions when operating at high frequencies with large bandwidth. An extensive review and analysis of the previously proposed set ups has allowed us to establish the more relevant generation techniques at the same time that has provided the basis for conceptualizing designs of our own.

The main objective of this doctoral thesis lies on the aforementioned mentioned analysis of the existent state of the art of photonics based pulse generation solutions as well as the proposal and experimental validation of new techniques that are proven to be successful in producing UWB high order pulses, which represent the best option when searching for compliance with the settle standard by the Federal Communications Commission (FCC). Specifically, the two main approaches that are handled in this research are:

- 1) The employment of non-linear effects to generate pulses with UWB features: In this segment of experiments I successfully took advantage of the nonlinear

effects inherent to the Mach-Zehnder Interferometer (MZI) employed to produce the aspired waveforms. First, we test the designs abilities to generate low-order derivatives (monocycles and doublets) and finally the flexibility of this technique also allowed obtaining high-order pulses that provided a proper FCC spectral fitting. Other valuables characteristics identified in these set ups are the plausibility of excellent tuning and a wide reconfiguration range.

- 2) The use of PM-IM conversion: Although the initial set of proposals succeeded at creating the desired waveforms, working in the linear regime increases the systems stability. Hence, for the second set of experiments it was decided to work with layouts based on the phase modulation-to-intensity modulation conversion process. One key structure for these experiments was a customization stage located after the before mentioned conversion process.

In both cases, a very important aspect that was always taken into consideration was the capacity of the designed systems to adapt themselves to integrated structures. Therefore, all the contributions originated from this work can be considered of special interest not only as pulse generation designs but, also as valid photonic processing structures.

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Resumen

El trabajo de doctorado aquí presentado tiene su nicho de aplicación dentro de los esquemas de generación de pulsos para comunicaciones ópticas, específicamente para aplicaciones de Ultra banda Ancha (Ultrawideband, UWB). En este sentido, a medida que la capacidad y ancho de banda requerido por los usuarios para los servicios de comunicaciones de banda ancha aumenta constantemente, diferentes tecnologías como híbridos entre esquemas ópticos-no cableados, sistemas UWB en conjunto con competidores como la norma de interoperabilidad mundial para acceso por microondas (Worldwide Interoperability for Microwave Access, WIMAX), el estándar de Evolución a largo plazo (Long Term Evolution, LTE) y el reciente Light Fidelity (LI-FI) continúan siendo investigadas como soluciones viables para proporcionar tasas de velocidades de datos que superan los Gigabits por segundo por usuario. Actualmente, las redes híbridas ópticas no cableadas son los proveedores claves para las redes de retorno y definitivamente jugarán un papel estratégico en la siguiente generación de redes de acceso, las cuales requieren una estrategia de despliegue flexible, alta capacidad, capacidad de actualización, escalabilidad y factibilidad.

Por otro lado, la creación de señales arbitrarias es un tema fundamental dada su amplia utilización para una miríada de aplicaciones prácticas como: ensambles para sistemas de imagen, radares, instrumentación, diseño de estructuras para biomedicina y soluciones de comunicaciones. Durante los últimos años, todos los esfuerzos han estado dirigidos hacia propuestas enfocadas en la generación directamente dentro del dominio óptico por las limitaciones presentes en las soluciones tradicionales eléctricas cuando operan a altas frecuencias y requieren largos anchos de banda. Una extensa revisión y un meticuloso análisis de las propuestas vigentes me ha permitido establecer las más relevantes técnicas de generación de señales, al mismo tiempo que ha proporcionado las bases para la conceptualización de diseños propios.

El objetivo principal de esta tesis doctoral recae en el ya mencionado análisis del estado del arte de las soluciones fotónicas para generación de pulsos ya mencionadas, en conjunto con la propuesta y validación experimental de nuevas técnicas que sean exitosas en producir pulsos UWB de alto orden, los cuales representan la mejor alternativa para satisfacer el estándar establecido por la FCC. Específicamente, los dos enfoques estudiados en esta investigación son:

- 1) La utilización de los efectos no lineales para generar pulsos con características UWB: En este segmento de experimentos, exitosamente se tomó ventaja de los efectos no lineales inherentes al Interferómetro Mach- Zehnder (MZI) empleado para producir las formas de ondas deseadas. Primero, se probaron las habilidades del diseño en generar derivadas gaussianas de bajo orden (conocidas en el vocabulario de la tecnología UWB como “monociclos” y “dobletes”) y finalmente la flexibilidad de esta técnica para obtener aquellas derivadas de alto orden, que son las que proporcionan una concordancia apropiada en términos de la normativa espectral establecida por la FCC. Otros rasgos resaltantes identificados en estos esquemas son la factibilidad para su reproducción t excelente capacidad de sintonización y reconfiguración.
- 2) El uso de conversión PM-IM: A pesar de que las propuestas iniciales mostraron resultados óptimos en la búsqueda de las señales requeridas, operar dentro del régimen lineal incrementa la estabilidad del sistema, algo que definitivamente no era un factor favorable en los diseños iniciales. Por ende, para el segundo conjunto de experimentos llevados a cabo en esta investigación, se decidió trabajar con diseños basados en el proceso de conversión de modulación de fase a modulación de intensidad. Una de las estructuras claves para estos experimentos fue una unidad de reconfiguración de pulsos ubicada después del comentado proceso de conversión

En ambos casos, un aspecto muy importante que ha sido tomado en cuenta en ambos casos fue la capacidad del sistema diseñado de adaptarse al esquema de estructuras ópticas integradas. Por lo tanto, todas las contribuciones originadas de este trabajo pueden ser estudiadas y consideradas de interés especial no solo como esquemas de generación de pulsos pero también como estructuras de procesamiento óptico válidas.

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Acronyms List

APT:	Asia Pacific Region
AWG:	Arbitrary Waveform Generation
AWGs:	Array Waveguide Gratings
BER:	Bit Error Probability
BI-SMF:	Bend-Insensitive Single-Mode- Fiber
BPAM:	Binary Pulse Amplitude Modulation
BPM:	Bi-Phase Modulation
BW:	Bandwidth
CEPT:	European Conference of Postal and Telecommunications
DAA:	Detect and Avoid
DARPA:	Defense Advanced Research Project Agency
EIRP:	Effective Isotropic Radiated Power
ETSI:	European Telecommunications Standards Institute
FBG:	Fiber Bragg Grating
FCC:	Federal Communications Commission
FFT:	Fast Fourier Transform
FWM:	Four Wave Mixing
GVD:	Group Velocity Dispersion
IEEE:	Institute of Electrical and Electronics Engineers
IFFT:	Inverse Fast Fourier Transform
IMT:	International mobile telecommunications
IR:	Impulse Radio
IR-UWB:	Impulse Radio- Ultrawideband
ITU:	International Telecommunications Union
LI-FI:	Light Fidelity
LTE:	Long Term Evolution
MB-OFDM:	Multiband- Orthogonal Frequency Division Multiplexing
MMF:	Multi-mode Fiber

MWP:	Microwave Photonics
MZI:	Mach- Zehnder Interferometer
OFDM	Orthogonal Frequency Division Multiplexing
OOK:	On-Off Keying
OPU:	Optical Processing Unit
PAM:	Pulse Amplitude Modulation
PM:	Phase Modulation
PM-IM:	Phase to Intensity Modulation
POF:	Polymer optical fibers
PoIM:	Polarization Modulator
PPM:	Pulse Position Modulation
PRF:	Pulse Repetition Frequency
PSD:	Power Spectral Density
PSM:	Pulse Shape Modulation
QoS:	Quality of Service
R&O:	Report & Order
RF:	Radiofrequency
RF-AWG:	Radiofrequency
RoF:	Radio over Fiber
SIR:	Signal-to-Interference Ratio
SMF:	Single-Mode- Fiber
SOA:	Semiconductor Optical Amplifiers
UFZ:	Ultrawideband Friendly Zone
UMTS:	Universal Mobile Telecommunications System
UWB:	Ultrawideband
UWBoF:	Ultrawideband over Fiber
UWBWG:	Ultrawideband Working Group
VDL:	Variable Delay Line
WDM:	Wavelength-Division Multiplexing
WIMAX	Worldwide Interoperability for Microwave Access

WLANs: Wireless Local Area Networks
WPANs: Wireless Personal Area Networks
XGM: Cross- Gain Modulation
XPM: Cross-Phase Modulation

CHAPTER 1

Introduction

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1.1 UWB as a key enabling technology.

This introduction, represents a complete set of empirical arguments that will not only justify the necessity of developing full optical generation schemes for UWB applications following different approaches, but also will present the key goals to be fulfilled by this research work. The structure of this chapter will be, first, to present an overview of the current trends for optical-wireless networks. Second, to explain the relevance they play in the access networks of today and finally, the research objectives of this dissertation will be exposed, along with the employed methodology.

Nowadays, when the mandatory speed and bandwidth levels never seem to be covered, and the demanded capacity per user for broadband communication services continuous to increase; state of the art technologies such as WiMAX [IEEE04], Long Term Evolution (LTE) [Aldm13], LI-FI [Scop15] and Ultra-wideband (UWB) radio, as well as many other mainstream indoor and outdoor optical wireless solutions such as infrared light and visible light are being studied and investigated as reliable options to guarantee data rates that exceed the desired bit/rate per user value.

Apart from high-data rates, the future wireless personal-area networks demand a low complexity, low cost, low power consumptions and scalable approaches within the personal and corporative operating environments. However, these networks are restricted by the reality of radio system engineering, where very limited free spectrum resources are available in the industrialized and developed regions of the world. Under these premises, UWB radio schemes rise up as an optimal alternative in terms of spectra sharing capacity without requiring new available spectral bands, since they can adapt to operate within already accessible ones [Ghav04, Oppe04].

UWB solutions come with numerous valuable advantages that originally had been appreciated for some time in the radar and communications communities [Vick91, Full91, and Wong98]. They offer the possibility of operating within a very broad frequency spectrum and providing large bandwidth capabilities. In this sense, the main feature of an UWB design is the ability to spread its signal across a significantly wide bandwidth while guaranteeing a low power spectral density (PSD) with negligible interferences on neighboring wireless systems [Ghau12, Giul04, Isol07]. Moreover, with bandwidth resources becoming increasingly scarce, UWB radio was a “latent necessity” waiting to be fulfilled.

The situation regarding this technology suffered a drastic milestone in the spring of 2002, when the U.S Federal Communications Commission (FCC) authorized the released of an official spectral mask allowing commercial operation of UWB radios at the noise floor, but over an enormous bandwidth [FCC 02]. This large “new bandwidth” opened the door for an

unprecedented number of new applications [Verd00, He14, Sun08, and Belt09] and stimulated the rapidly increasing efforts undertaken by several research institutions, industry, and government agencies to assess and exploit the potential of UWB radios in various areas. In the specific case of wireless communications, the FCC regulated power levels are very low (below -41.3 dBm), this permits UWB technology to overlay already available services such as the global positioning system (GPS) and the IEEE 802.11 wireless local area networks (WLANs) that coexist in the 3.6–10.1 GHz band. This remarkable spectrum cohabitation capacity is what puts it one step ahead over its closest competitors.

Figure 1.1: Comparative chart of UWB with neighboring wireless technologies.

Figure 1.1 depicts a comparative plot of the position held by UWB when compared to other close in application technologies, from where we can extract that even with the significant power restrictions; UWB holds enormous potential for wireless ad-hoc and peer-to-peer networks [Rao08]. One of the major potential advantages in impulse-radio-based systems is the ability to trade data rate for link distance by simply using more or less concatenated pulses to define a bit. Without dramatically changing the air interface, the data rate can be changed by orders of magnitude depending on the system requirements. The narrow time domain pulse presence also implies that UWB offers the possibility for very high positioning accuracy [Bell08].

Despite all the valuable remarked features, due to the established low PSD of a standard UWB signal and although UWB signals could theoretically propagate through great distances at high power levels, current FCC regulations only enable high-rate (above 110 MB/s) data transmissions over a short range (10–15 m) at very low power. Evidentially, we are facing a clear tradeoff between the achievable data rate and the communication distance. Such a short-range network would operate mostly in stand-alone mode, with nonexistent integration into the fixed wired and wireless wide-area infrastructures which will represent an underuse of this

valuable technology. So, to increase the covered area and at the same time offer the availability of uninterrupted service across different platforms, the technique based on the combination of UWB and microwave photonics (MWP) for signal distribution and generation purposes has emerged as a valid and efficient solution with demonstrated applications [Yao07,Kasz08, Wang10, Pere10].

1.2 Generation of UWB pulses directly in the optical domain.

Originally the fusion between UWB and MWP was merely adopted to solve area coverage issues and was commonly referred as “UWB over Fiber” (UWBoF). In these systems, optical UWB signals were usually generated in a central office and distributed to a single (or various) access points via optical fiber links [Yao07]. One of the fundamental benefits provided by an UWBoF architecture is the lack of necessity for extra electrical to optical domain conversions since the system’s input pulses are directly generated in the optical domain, hence no optical power is squandered. In addition, the generation of UWB signals in the optical domain offers remarkable characteristics that are intrinsic to the optical domain such as light weight, small size, large tunability and immunity to electromagnetic interference.

The versatility of UWB along with its interoperability with other wireless technologies and optical distribution methods represented a “golden opportunity” to put forward the so aspired fully-converged optical schemes that our current needs demand. Such designs will provide a complete access over the networks, optimize resources and will maximize the efficiency of the services. Figure 2 illustrates a recent proposal [Pere14] focused on an optical wireless hybrid architecture approach.

Figure 1.2: Theoretical application scenario for an optically hybrid wireless techniques and MWP based design.

After having a general idea of the contemporary panorama in the UWB generation field and the potential solutions that can be developed in order to optimize the current systems, it's clear that there is still a long path to follow in order to accomplish full optically enhanced and integrated designs for UWB applications. Therefore, by means of this thesis several methods for UWB pulses generation will be proposed and experimentally demonstrated. Here in, employing both active and passive based proposals we will generate both low and high order UWB waveforms.

1.3 Objectives and Thesis Outline.

The main objective pursued by this thesis consists of proposing fully optical UWB generation schemes and demonstrating their validity for low and high order waveforms creation. The submitted systems do not only hold generation capabilities, but in most cases modulation ones as well. In these order of ideas and looking forward to accomplish this main goal, the thesis covers a set of specific objectives that can be listed as follows:

- To know and understand in detail the background behind the rise of UWB fundamentals as a key enabling technology.
- To analyze the current UWB regulation and its inherent limitations in terms of certain operational features.
- To scrutinize UWB's original main competitor and perform a comparison analysis between both technologies.
- To summarize the most relevant optical UWB generation techniques, acknowledging their strengths and potentiality and comparing their advantages and disadvantages.
- To propose schemes for low and high order pulses creation based on both active and passive components. These methods will be review not only for their legitimacy as generators but also in features as important as: flexibility, scalability and integration.
- To demonstrate experimentally the validity of the proposed schemes, to verify not only their generation capacity but also the possibility of modulation performance.

In this way, in order to accomplish the before mentioned goals, the herein presented thesis holds the following outline:

In chapter 2, the current bibliography related to the UWB technology field, the origins of the fusion between UWB and MWP and the established optical approaches focused on UWB pulses generation are deeply studied so it can be possible to determine the key advantages of the most significant proposals and the existing trade-off between FCC Mask compliance, scalability and wavelength efficiency.

In chapter 3, once the current available techniques for optical waveform generation are analyzed, a final selection of two methods as base technology for the proposed designs is

performed. In this sense, Semiconductor optical amplifiers (SOAs) and phase-to-intensity (PM-IM) conversion based schemes are the chosen approaches and a theoretical analysis of them both two is presented. Such examination will help set the required premises for a better understanding of the proposals to be exposed in the next chapter, at the same time that expands the application basis of these methods.

In Chapter 4, several structures for arbitrary waveform generation based on Microwave photonics technology for UWB applications are proposed. Such schemes rely on the employment of the inherent non-linear effects present in a SOA-MZI device or in a PM-IM conversion process. In the case of the last such procedure, it is accomplished by means of the combination of a phase modulator (PM) with a dispersive element (a fiber link) or the PM with a customized Fiber-Bragg Grating (FBG) pulse shaping structure. Theoretical and experimental measurements of the FCC compliant generated pulses are exposed, in both time and frequency domain.

At last, in chapter 5 summarizes the author's general conclusions and it indicates the possible directions for future research.

Finally, It is mandatory to remark that the work carried out in this thesis has been done within the scoped of work of the following Spanish national projects: TEC2010-21303-C04-02, TEC2011-26642 and TEC2014-58048-C2-2-P, which were funded by the Spanish Ministry of Economy and Competitiveness ("Ministerio de Economía y Competitividad") along with the projects UPVOV08-3E-008 and UPVOV10-3E-492 funded by the European Regional Development Fund (ERDF).

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CHAPTER 2

UWB migration from the electrical to the optical domain: techniques and applications.

In this chapter, we discuss and analyze the state of the art related to the UWB technology field, the origins of the fusion between UWB and MWP and the established optical approaches focused on UWB pulses generation. The high interest and expectations about developing novel and valid schemes within this area rely on its promising adaptability into a wide range of applications. First, we will expose a brief introduction corresponding to UWB fundamentals, the meaningful benefits behind this technology, its limits, its evolution in terms of regulations and practical solutions through the years. Then, a detailed exposition of well-known optical methods to generate lower and higher order UWB pulses is presented, from which we would be able to determine the advantages of some features of these proposals and the existing trade-off between FCC Mask compliance, scalability and wavelength efficiency. Finally, we would present a comparison among a vast majority of available approaches which would allow us to establish our own feasible solutions for UWB signals creation.

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2.1 Highlights of Ultrawideband (UWB) Technology

Ultra-wideband (UWB) technology has changed intensely through recent history. For over 20 years, UWB schemes had been used for radar, sensing, military communications, and niche applications. A substantial change occurred in February 2002 when the Federal Communications Commission (FCC) [FCC02] issued a ruling that UWB could be used for data communications as well as for positioning and safety applications. In this context, the scope of influence of UWB technology has grown larger and a deep interest towards its exploitation has been addressed from different sectors.

There are several advantages that make UWB an attractive solution for communications applications. In particular, UWB systems hold potentially low complexity, low cost, a resistance to severe multipath and jamming and a very good time domain resolution employed for location and tracking purposes, among many other benefits. The low complexity and low cost present in UWB systems arise from the baseband nature of the signal transmission [Chon06]. Unlike conventional radio schemes, UWB transmitters produce a very short time domain pulse, which is able to propagate without the need of any additional radio frequency (RF) mixing signal to a frequency which has desirable propagation characteristics. The wideband nature of UWB signals means these pulses propagate through frequencies commonly employed as carrier frequencies. Therefore, such signals will propagate well without the need of extra up-conversion and amplification, a trait that makes them suitable candidates for migration to optical schemes. The reverse process of down conversion is not required in the UWB receivers.

One of the key issues when discussing UWB communication systems is frequency allocation. Standard radio systems have traditionally used dedicated frequency spectrum for sending and receiving radio signals in order to mitigate interference effects. Contrary to this approach, Ultra wideband (UWB) technology as its name claims, operates by spreading its signal across a very wide frequency band (typical bandwidth values are larger than 500 MHz).

UWB transmissions shall usually be low-power and will operate in frequency segments that are already active to other radio and wireless technologies (including satellite services, microwave links, fixed wireless techniques, to name a few) that

might be currently licensed to commercial entities for their exploitation [Vald10]. As a result of this fundamental difference, UWB has affected the way frequency spectrum is managed and its standardization has turned out to be an absolute challenge through the last decades.

2.2 UWB Regulatory bodies and current legislation

Under these circumstances, the regions have proposed different approaches. Some companies in the USA are working towards removing the restrictions from the FCC's regulations for applications regarding UWB technology. These companies have established an Ultrawideband Working Group (UWBWG) to negotiate with the FCC. At the moment, there are no dedicated frequency bands for UWB applications for Europe in the ETSI (European Telecommunications Standards Institute) or ITU (International Telecommunications Union) recommendations. A more detailed depiction of this panorama will be discussed in the next paragraphs.

2.2.1 UWB Regulation in USA

Initially, in view of the unclear impacts that might be caused by UWB devices, the FCC adopted a cautious approach in devising its standards in the First Report & Order (R&O) on UWB technology. FCC had strong beliefs that the combination of technical standards and operational restrictions would help to ensure coexistence of UWB devices and provide adequate protection to other operational radio communication services.

Before the FCC submitted their first report and order on 2002, there was a significant effort by several industrial parties to convince the FCC to allow UWB technology to operate under the FCC Part 15 regulation status, and hence to permit license-free employment of UWB products. In this First Report and Order [FCC02], the FCC concluded that low-power UWB devices indeed could operate within the existing spectrum without causing significant interference, this threshold being represented by a 1 dB noise floor. The Commission also assumed cellular services to have a higher level of received signal than what it is often the case in practice. Hence, the FCC amended Part 15 of its rules to permit UWB operation without a license or the need for frequency coordination.

However, for regulatory purposes, the Commission established various technical

standards and operating restrictions for 3 types of UWB devices: communications systems, imaging systems (including ground penetrating radars) and vehicular radar systems. This classification was based on their potential to causing interference to the existing licensed users in the 3.1-10.6 GHz frequency band. Accordingly, emission limits were designed for each type of UWB devices to further mitigate interference

Finally, in December 2004, the FCC released its Second Report and Order and a Second Memorandum Opinion & Order [FCC04] on UWB to provide greater flexibility for the introduction of wide-bandwidth devices and systems which do not fall into the classification of UWB technology.

Under the new rules, the peak power limits for wide-bandwidth device emission was raised to the same level as UWB devices in the three frequency bands that had already been made available for unlicensed operation, such frequencies were: 5925-7250 MHz, 16.2-17.2 GHz and 23.12-29 GHz. In addition, the Commission amended its measurement procedures to permit frequency hopping, swept frequency and gated systems operating in these bands to be measured in their normal operating mode. For the purposes of this work, we will use the parameters settled for the working band allocated between 3.1 and 10.6 GHz. A summarized overview of the FCC established radiation levels for indoor and outdoor units can be appreciated in table 2.1.

	Indoor	Outdoor
Frequency in GHz	EIRP in dBm	
From 0.96 to	-75.3	-75.3
From 1.61 to	-53.3	-63.3
From 1.99 to 3.1	-51.3	-61.3
From 3.1 to 10.6	-41.3	-41.3
Above 10.6	-51.3	-61.3

Table 2.1 FCC radiation limits for indoor and outdoor communication applications in terms of effective isotropic radiated power (EIRP).

2.2.2 UWB Regulation in Europe

The European approach is somewhat more cautious than the one faced by the USA and its scope of work was based on the concept of “the importance of designing

suitable UWB regulations” as European standards require for every new technology to demonstrate its lack of side effects or harm to previously existing ones. Still, it is fair to say that European organizations have been heavily influenced by the FCC’s decision. Currently in Europe, the recommendations for short-range devices belong to the CEPT (European Conference of Postal and Telecommunications) working group CEPT/ ERC/ REC 70- 03 (Ultra-Wideband Working Group, 1999) [Bere 09].

In a comparative note between the European Commission (ECC) and the FCC rules, the ECC recommendations appear to be more stringent and its decision on the emission limit in the 2.7-3.4 GHz band remains divided. Unlike the FCC which has 3 types of UWB devices, ECC only considers, at this stage, 2 main types of UWB systems that operate below 10.6 GHz: UWB communications systems and UWB imaging schemes. The latest regulation was announced in February 2007 [FCC07], and it permits the employment of spectrum by UWB equipment using and to harmonize the conditions throughout Europe.

2.2.3 UWB Regulation in the Asia Pacific Region (APT)

Regulatory operations inside the Asia Pacific region have been quite active as well; some APT member countries like Japan and Korea have implemented a regulatory framework that allows unlicensed use of UWB devices in indoor environment while others like Singapore and Hong Kong have established UWB trials within specific localized zones. In the case of Singapore, a more experimental approach is taken into consideration and a UWB Friendly Zone (UFZ) was created to carry out innovative designs. On the other hand, Hong Kong established a well-defined UWB frequency working range in the 3.1-10.6 GHz band for UWB indoor trials and the 4.2-10.6 GHz band for outdoor trials. One of the most notorious stakeholders in this dynamic is Japan. A preliminary mask for UWB was announced in September 2005. Subsequent compatibility studies revealed that UWB devices operating in the 3.40-4.20 GHz band should incorporate interference mitigation techniques. At last, in Korea the spectrum allocation for UWB ranges in the frequency band from 3.1 to 10.2 GHz.

Considering the importance of harmful interference avoidance to International mobile telecommunications (IMT) system and broadcasting relay system, interference mitigation technology such as Detect and Avoid (DAA) technology

must be used for the UWB devices operating in the 3.10 to 4.20 GHz band. Figure 2.1 permits to visualize the current standardized UWB operational frequency limitations in terms of Power Spectral Density (PSD) values.

Figure 2.1 FCC radiation limits according to the current regulation agents for UWB technologies [FCC 07].

Administration	License Regime	Max. E.I.R.P	Frequency bands/mask	Mitigation techniques	Operational conditions
US	Unlicensed	-41.3 dBm/MHz	3.1 to 10.6 GHz with mask	<ul style="list-style-type: none"> Not specified 	<ul style="list-style-type: none"> Indoor only. If outdoor, not attached to a fixed antenna. Only transmit when sending information to associated receiver. Handheld outdoor devices do not employ a fixed structure. 10 second handshake rule only for handheld devices. Prohibited operation onboard ships, aircraft or satellite.
Europe	Unlicensed	-41.3 dBm/MHz	3.1 to 10.6 GHz with mask	<ul style="list-style-type: none"> Low duty cycle restrictions in the 3.4 -4.8 GHz band. Expects products without mitigation techniques to operate exclusively above 6 GHz in the longer term. 	<ul style="list-style-type: none"> Indoor only. If outdoor, not attached to a fixed location. Transmit power control for devices installed in road and rail vehicles in the 4.2-4.8 GHz and 6-8.5 GHz bands. Time limited deployment for devices operation in the 4.2-4.8 GHz. Resolution not applicable to aircrafts and other aviation applications.
Japan	Unlicensed	-41.3 dBm/MHz	3.4 to 10.25 GHz with mask	<ul style="list-style-type: none"> Mitigation techniques applicable between 3.4 - 4.8 GHz (DAA) 	<ul style="list-style-type: none"> Indoor use only Time limited operation for devices without mitigation techniques in 4.2 -4.8 GHz band.
Korea	Unlicensed	-41.3 dBm/MHz	3.1 to 10.2 GHz with mask	<ul style="list-style-type: none"> DAA technology between 3.1-4.23 GHz from 2007 and 4.2-4.8 GHz from 2010. 	<ul style="list-style-type: none"> Indoor only Only transmit when sending information to associated receiver Devices must include 10 second handshake acknowledgment.

Table 2.2 Regulations Comparison divided in the most relevant active UWB working regions.

2.3 Basic Ultrawideband (UWB) Concepts

Figure 2.2 displays a comparison between UWB spectral features versus other conventional working technologies.

Figure 2.2 Comparative between UWB spectral features versus other conventional working technologies.

As it can be appreciated when comparing conventional radio transmission systems and UWB signals the previous reveal a small fractional bandwidth component. As an example we can comment about Universal Mobile Telecommunications System (UMTS) which operates around 2 GHz with a bandwidth of 5 MHz. This system is often referred as “wideband”, but following the premises of equation 2.2, the fractional bandwidth metric of this standard is 2.5%, which is significantly smaller than the regulated one. One key aspect in UWB signals is the power spectral density (PSD), which is defined as the power (P) transmitted in watts (W) within a range of bandwidth (BW) of the signal in hertz (Hz), and the unit of PSD is watts/hertz (W/Hz). For UWB systems the power spectral density tends to be considered extremely low, especially for communication applications.

$$PSD = \frac{P}{BW} \quad (2.1)$$

If we consider BW as the signal bandwidth and f_c as its center frequency that can be obtained by means of f_H and f_L which are the effective highest and lowest frequency components of the signal, we can calculate the fractional bandwidth (B_f) parameter of a pulse since, it is defined as the ratio of signal bandwidth and the center frequency [Gezi08]:

$$B_f = \frac{BW}{f_c} = \frac{(f_H - f_L)}{(f_H + f)} 100\% \quad (2.2)$$

2.3.1 Current Definition of UWB

According to the current FCC UWB issued rule, for a pulse to be considered a UWB signal it must possess a fractional bandwidth factor (B_f) of at least 20% or to hold a signal bandwidth of 500MHz or more. This translates to:

$$\begin{aligned} B_f &\geq 20\% \\ BW &> 500\text{MHz} \end{aligned} \quad (2.3)$$

In this context, the energy used to transmit a wireless signal is not infinite and in general, should be as low as possible, especially for modern electronic devices. In the case of a fixed amount of energy, we can either transmit a great deal of energy density over a small bandwidth or a very small amount of energy density over a large bandwidth. For UWB systems, the energy is spread out over a very large bandwidth (hence the name Ultrawideband) and, in general, is of a very low power spectral density. The major exception to this general rule of thumb is UWB radar systems which transmit at high power over a large bandwidth. However, here we will restrict ourselves to the communications area. Table 2.3 depicts a comparison between UWB and other common wireless broadcast and communication Systems in terms of power spectral density:

System	Transmission power (W)	Bandwidth (Hz)	Power spectral density (W/MHz)	Classification
Radio	50 kW	75 kHz	666,600	Narrowband
Television	100 kW	6 MHz	16,700	Narrowband
2G Cellular	10 mW	8.33 kHz	12	Narrowband
802.11a (WLAN systems)	1 W	20 MHz	0.05	Wideband
UWB	1 mW	7.5 GHz	0.013	Ultrawideband

Table 2.3 UWB and fellow narrowband technologies comparison in terms of PSD.

The outstanding ability depicted by UWB technologies to provide very high data rates for short ranges has made it for a long time an excellent candidate for the physical layer of wireless personal area networks (WPANs) [IEEE15a]. Since,

Ultrawide Band communication using impulse radio (IR-UWB) has many interesting benefits for wireless applications; it has attracted much research and industrial attention [Bene03, Scho93]. UWB signals using impulse radio can be divided into two groups: single band (pulsed) systems and multi-band systems using a single carrier in each sub-band.

In this sense, two opposing groups of UWB developers have appeared over the years. The single-band technique supporters claim that impulse radio is the original approach to UWB by using narrow pulses that occupy a large portion of the spectrum which translates into a large bandwidth. On the other hand, the multiband approach upholders praise their proposal's ability to maximizing the available UWB frequency spectrum (3.1 GHz to 10.6 GHz) by dividing it into multiple smaller and non-overlapping bands with bandwidths greater than 500 MHz to obey the FCC's definition of UWB signals. To date, numerous proposals from both groups have been submitted and analyzed by the scientific community and the decision of which one is better is yet to be made because both technologies are impressive and have technical credibility. The following subsections discuss briefly these two relevant and promising techniques: impulse radio and multiband system.

Traditional Impulse radio (IR) is a baseband digital communication technique whereby information symbols are conveyed to a receiver by an information bearing signal without a sinusoidal carrier, the employed baseband signal is usually composed of sub-nanosecond pulses [Shim15]. Carriers are typically employed in the majority of modern communication, radar and imaging systems, in the specific case of ultra-wideband (UWB) IR techniques, since such methods employ extremely short bursts of baseband pulses we found that large bandwidths are required. The relationship between a measurable function in time and its frequency representation has been well established by the Fourier Transform [Carl09]. Following these principles, It is obvious to mention that signals that have an extremely short time duration occupy a very large bandwidth; hence the definition, ultra-wideband (UWB) impulse radio. The potential simplicity of UWB IR has driven attention towards this approach for quite some time. Actually, It can be stated that UWB technology has a story dating back to one hundred years ago when G. Marconi sent the first ever wireless transmission from the Isle of Wight Cornwall on the British mainland using spark-gap transmitters [Sark06]

As for implementation procedures, a classical UWB transmitter ideally consists of a few basic components such as a voltage source, a fast switch, and a UWB antenna that can efficiently radiate the UWB pulse with high fidelity (i.e., minimum distortion). On the other hand, the receiver only requires a properly designed UWB antenna [Pire08], a simple matched filter and a decision device. Unfortunately, design and manufacture of UWB systems has found some difficulties through the years because of the extremely large bandwidth required.

In the case of the Multiband approach, a typical multi-band UWB system consists in dividing the bandwidth into several sub-bands in a fixed manner or using a frequency hopping technique to address each sub-band [Batr04]. The frequency hopping depends on pseudo-random code sequences.

Another multi-band system based on orthogonal frequency division multiplexing (OFDM) techniques on each sub-band has been proposed (MB-OFDM). During its standardization attempt by IEEE 802.15.3a group [Ieee15b], its good properties for capturing efficiently multi-path energy, mitigating interference and achieving large throughput [Sabe03, Batr03] have been demonstrated. In the MB-OFDM scheme, the signals consist of 128 sub-carriers spaced 4.125 MHz apart and utilize the 3.1 to 10.6 GHz range in 14 sub-bands, the bandwidth of each sub-band being set to 528 MHz. But the drawback of this type of system is that the transmitter is slightly more complex because it requires an inverse FFT (IFFT) and the peak-to-average ratio may be slightly higher than that of the other pulsed multi-band approach.

All multi-band systems are attractive radio techniques that coexist with other narrowband signals and have interesting properties. For example, the information can be processed over a much smaller bandwidth, thus reducing the complexity of the design, reducing the power consumption, lowering the cost, and improving spectral flexibility and worldwide compliance. However, the disadvantage of a pulsed multi-band system is the difficulty in collecting significant multi-path energy using a single RF chain. Moreover, the MB-OFDM system requires relatively large computational power because of the required fast Fourier transform (FFT) processors.

2.3.2 FCC Mask Efficiency

As it has been clearly established, although UWB holds abundant possibilities for spectra maximization, this technology is also very constrained by its corresponding regulations. When evaluating a technology and its scope of advantages we require metrics that permit us to measure quantifiable metrics. For the scope of work of this research, the key parameter to analyze and evaluate UWB is represented by the spectral or FCC mask efficiency.

In this way, the FCC imposed mask (that it is possible to appreciate in Fig 2.1) stands as $M(f)$ and it bounds the power spectral density $S_{EIRP}(f)$ of the equivalent isotropic radiated power (EIRP) of a UWB pulse [FCC 02]. According to the specific application and adopted regulation, the detailed shape of $M(f)$ will vary. For the vast majority of modulation formats, the power spectral density (PSD) of a modulated train of pulses will be defined by the single function of the modulation process and the Fourier transform $S(f)$ of a UWB pulse shape $s(t)$. The transformation from the modulated pulse train on which the EIRP mask is imposed, is a linear transformation which we represent by the equivalent system function that can be named $H_{eq}(f)$. Therefore, the power spectral density $S_{EIRP}(f)$ is represented by:

$$S_{EIRP}(f) = S_{mod}(f) |S(f)|^2 |H_{eq}(f)|^2 \quad (2.4)$$

Where $S_{mod}(f)$ stands as the modulating signal in the frequency domain. Here, we assume that for further modulation process in terms of the PSD, this parameter is close to a constant for all frequencies (f) of interest by means of a a_c constant, without loss of generality and based on we assume then:

$$S_{EIRP}(f) \approx a_c |S(f)|^2 |H_{eq}(f)|^2 \quad (2.5)$$

This a_c constant includes factors that hold conversion properties and contributes to scaling the constant spectral density of the modulation signal. Hence, the FCC requires that for all frequencies (f) in the settled range (F_{min}, F_{max}) the following requirement is accomplished:

$$a_c |S(f)|^2 |H_{eq}(f)|^2 \leq M(f) \quad (2.6)$$

In this order of ideas, since we have established the principles around the FCC Mask constraint, it is logical to affirm that there is an intrinsic mask-filling efficiency factor (η) present in any UWB design that can be defined as the ratio of the power contained in the resulting FCC-compliant PSD $S_{EIRP}(f)$ to the total power that could be contained under the Mask $M(f)$ levels, both confined between the effective UWB frequency range [3.1 GHz (F_{min}), 10.6 GHz (F_{max})]. So, η gets defined as:

$$\eta \triangleq \frac{\int_{F_{min}}^{F_{max}} |S(f)|^2 |H_{eq}(f)|^2 df}{\int_{F_{min}}^{F_{max}} M(f) df} \quad (2.7)$$

Here, the mask-filling efficiency measured for a standard pulse waveform embodies the effects of the transmitter, antenna and propagation on signals. Being this mask-filling efficiency factor along with the fractional bandwidth the main characteristics we will take into consideration for analysis of UWB pulses, it becomes essential to establish the adequate characteristics and limitations for η . Ideally speaking, if there were no constraints on the pulse shape, then it could be possible to determine a waveform that fulfills a 100% mask-filling factor by solving:

$$|S(f)|^2 = \frac{M(f)}{a_c |H_{eq}(f)|^2} \quad (2.8)$$

Here, the ratio on the right side of equation 2.8 represents the squared magnitude of a finite energy waveform. Factorization of this type to recover the time domain waveform can be carried out since equation 2.42 satisfies Paley Wiener conditions [Arfk12]. Following the approach of [Shen03] the following conditions will be considered:

- a) $H_{eq}(f)$ is not an ideal all-pass filter. For any application of the proposed method, they must assume $H_{eq}(f)=1$.
- b) Since it involves a work focused on FCC regulated UWB pulses, it is constrained by the PSD Mask and within the frequency range limits associated to it. In the band from 3.1 GHz to 10.6 GHz, UWB will apply the FCC regulations with a peak value of -41.3 dBm/MHz. Outside of this band, the PSD must decrease. From 0.96 GHz to

1.61 GHz, the reduction in admissible transmitted power is necessary to protect GPS transmissions.

Although desirable, we will realize that in practical implementations percentages close to 100% are never accomplished and the plausible range of adequate performance is allocated between 50% and 60% for higher order pulses [Li13, Li08, and Gao09].

2.4 UWB pulse forming and modulation

As known, different UWB designs can operate as notorious pulse waveforms such as: Hermite Pulses, Legendre and Prolate Spheroidal Functions. In this thesis, Gaussian based waveforms will represent the standard UWB pulses.

2.4.1 Gaussian based low order UWB pulses

As it was previously expressed, the most commonly employed method to generate UWB pulses is by means of Gaussian pulses and their derivatives, since they can be easily produced by modern pulse generators (when compared to rectangular pulses with very short rise and fall time) [Goya14]. The typical Gaussian pulse considered for UWB purposes is given by:

$$s(t) = \frac{A}{\sqrt{2\pi}\sigma} e^{-\frac{1}{2}\left(\frac{t}{\sigma}\right)^2} \quad (2.9)$$

Where A and σ denote the amplitude and width of the Gaussian pulse respectively and t allocates itself within the range $-\infty < t < \infty$. Figure 2.3 plots the typical pulse shape and Power spectral density (PSD).

Figure 2.3: (a) Time domain waveform and (b) PSD of a standard Gaussian pulse.

Regularly, when dealing with Gaussian derivatives, the higher the derivatives order the better candidates they are for UWB transmission purposes. This is mainly due to the DC value of the Gaussian pulse. Therefore, the best option is to use the Gaussian derivatives with smaller DC components since they will guarantee a more appropriate fit in terms of regulatory rules. In the following section, we will analyze UWB signals with a focus on the FCC established regulations, in terms of the spectral efficiency parameter [Lewi04] and the fractional bandwidth [Mill03]. The n^{th} derivative of a Gaussian pulse can be obtained recursively by means of the following expression:

$$s_n(t) = -\frac{n-1}{\sigma^2} s^{(n-2)}(t) - \frac{t}{\sigma^2} s^{(n-1)}(t) \quad (2.10)$$

As it was mentioned before, the efficiency of these pulses are related to their spectral representation. Therefore, it's imperative to hold a general adaptable Fourier transform of 2.10, such requirement is expressed in:

$$S_n(f) = A(j2\pi f)^n e^{-\left[\frac{(2\pi f\sigma)^2}{2}\right]} \quad (2.11)$$

Where, if we consider the real component of the spectrum we deal with the amplitude magnitude of the n th derivative:

$$|S_n(f)| = A(2\pi f)^n e^{-\left[\frac{(2\pi f\sigma)^2}{2}\right]} \quad (2.12)$$

The frequency where the maximum value of the n^{th} derivative is reached corresponds to the peak emission frequency which in this case we will tag as f_M and can be found by differentiating equation 2.12 and setting it to zero. Hence, the peak emission frequency must satisfy $2\pi f_M \sigma = \sqrt{n}$, and the maximum value of the amplitude spectrum is obtained through the expression:

$$|S_n(f_M)| = A \left(\frac{\sqrt{n}}{\sigma} \right)^n e^{\left(-\frac{n}{2} \right)} \quad (2.13)$$

With Equation 2.13, it is easier to obtain the normalized PSD within the frequency domain by considering equations 2.12 and 2.13. This expression has a peak value of 1 (0 dB) and It's given by:

$$|PSD_n(f)| \triangleq \frac{|S_n(f)|^2}{|S_n(f_M)|^2} = \frac{(2\pi f \sigma)^{2n} e^{[-(2\pi f \sigma)^2]}}{n^n e^{(-n)}} \quad (2.14)$$

In this approach, the n^{th} derivative of the Gaussian corresponds to the UWB transmitted pulse waveform. The PSD of the transmitted signal can be expressed in terms of the peak PSD permitted by the FCC (A_{max}) by means of the following expression:

$$|PSD_{A_{max}}(f)| = A_{max} |PSD_n(f)| = A_{max} \frac{(2\pi f \sigma)^{2n} e^{[-(2\pi f \sigma)^2]}}{n^n e^{(-n)}} \quad (2.15)$$

In equation 2.15, the n and σ parameters can be modified in order to comply with the FCC settled mask requirements. Considering the specific FCC spectral emission mask conditions, we can identify the presence of some “corner” points that must be complied by the pulses in question and therefore hold values that either pass through or are located below such points. This is the case for frequency points located at 0.96 GHz, 1.61 GHz, 1.99 GHz, 3.1 GHz and 10.6 GHz. Through the years, different spectrum fitting techniques for UWB systems have come up [Well02], all oriented to fulfill in the most accurate way possible the FCC mask standards. In the next section, we will present a detailed analysis of both Gaussian based lower and higher order pulses. To start, among the lower derivatives, we will analyze both monocycles and doublets. We understand them as the corresponding

first and second derivative of a Gaussian pulse. The time representation can be identified as $s_I(t)$ and defined in the following way:

$$s_I(t) = -\frac{At}{\sqrt{2\pi}\sigma^3} e^{\left\{-\frac{t^2}{2\sigma^2}\right\}} \quad (2.16)$$

While its amplitude spectrum is expressed in:

$$|S_I(f)| = A(2\pi f)^2 e^{\left\{-\frac{(2\pi f\sigma)^2}{2}\right\}} \quad (2.17)$$

Figure 2.4 plots the time and frequency domain of a monocycle.

Figure 2.4 (a) Time domain depiction of a positive and a negative monocycle waveform and (b) Frequency domain representation of the positive monocycle

Rigorously speaking, the duration of the Gaussian pulse and all of its derivatives is infinite. In this case, we can define a pulse width parameter of T_p , as the interval in which 99.99% of the energy of the pulse is contained. Using this definition, current literature states a monocycle pulse's spectrum does not contain low-frequency components including DC. This attribute is useful for some applications, as it facilitates the design of other system components including antennas, amplifiers, and sampling downconverters. Better signal characteristics, such as more precisely defined pulse shape and reduced distortion for the transmitting and receiving signals, can thus be easier to obtain but It is clear that PSD of this pulse does not fulfill FCC mask standards no matter what value of pulse width is employed.

In the case of the doublet or the second order derivative, its time domain representation is defined as $s_2(t)$ in the following way:

$$s_2(t) = A \left(\frac{t^2}{\sqrt{2\pi}\sigma^5} - \frac{1}{\sqrt{2\pi}\sigma^3} \right) e^{-\frac{t^2}{2\sigma^2}} \quad (2.18)$$

While its amplitude spectrum is expressed in:

$$|S_2(f)| = A (2\pi f)^2 e^{-\frac{(2\pi f\sigma)^2}{2}} \quad (2.19)$$

The corresponding time and frequency domain representation is depicted by Figure 2.5.

Figure 2.5 (a) Time domain depiction of a positive and a doublet and (b) Frequency domain representation of such doublet.

2.4.2 Gaussian based Higher order UWB Pulses

In this context, within the time domain, the higher-order derivatives of the Gaussian pulse resemble sin waves modulated by a Gaussian pulse-shaped envelope. As the order of the derivative increases, the number of zero crossings in time also increases; more zero crossings in the same pulse width correspond to a “higher” carrier frequency sin wave modulated by an equivalent Gaussian waveform. As for FCC mask compliance, in the case of indoor applications a fifth-order derivative is sufficient while, for the second at least a seventh or higher derivative is required. For practical reasons we will consider the indoor applications scheme. In this assumed

system, following the general equation expressed in 2.10 the transmitted pulse with FCC mask compliance within the time domain gets expressed as:

$$s_5(t) = A \left(-\frac{t^5}{\sqrt{2\pi}\sigma^{11}} + \frac{10t^3}{\sqrt{2\pi}\sigma^9} - \frac{15t}{\sqrt{2\pi}\sigma^7} \right) e\left(-\frac{t^2}{2\sigma^2}\right) \quad (2.20)$$

And the amplitude spectrum is:

$$|S_5(f)| = A(2\pi f)^5 e\left\{-\frac{(2\pi f\sigma)^2}{2}\right\} \quad (2.21)$$

Typical UWB high order pulses can be visualized in figure 2.6, in the shape of a third, fourth and fifth derivative of a Gaussian Pulse.

Figure 2.6: Time Domain Representation of higher order Gaussian derivatives.

In order to support earlier statements in this work that declared how higher-order derivatives of the Gaussian pulses represent the best candidates for UWB transmission, figure 2.10 depicts the normalized PSD for various Gaussian derivatives (starting in $n=1$). The normalization factor is the peak value permitted by the FCC, -41.3 dBm/MHz. It is obvious that the PSD corresponding to the first derivative does not meet FCC requirements, regardless of modification of other parameters (i.e. pulse width). On the quest for a suitable waveform set, interesting options as frequency shifting and bandwidth adjustment could be acknowledge, but such technique would require modulation of the low order pulse with external waveforms.

Figure 2.7: PSD of the higher-order derivatives of the Gaussian pulse for UWB for indoor systems.[Shen03]

However, since UWB is a carrier-less system so, modulation will increase the cost and complexity. Alternative solutions are necessary to obtain pulses that guarantee FCC mask compliance. In this context, higher order derivatives ($n > 3$) and their corresponding PSD masks hold a more accurate fit.

2.4.3. Modulation of UWB pulses:

For practical and functional UWBoF systems, the information should be encoded, which is done through different pulse modulation schemes. Traditionally, most modulation schemes had been demonstrated exclusively within the electrical domain. In this order of ideas, standard modulation formats such as on-off keying (OOK), pulse-position modulation (PPM), bi-phase modulation (BPM), pulse-amplitude modulation (PAM) and pulse shape modulation (PSM) have been proposed and demonstrated over UWB architectures. In this section, we will approach a brief explanation of the technical fundamentals of each as well as several experimental implementations.

Pulse Amplitude Modulation (PAM)

In the case of the classical Binary Pulse Amplitude Modulation (BPAM) is implemented using two antipodal Gaussian pulses. The transmitted binary pulse amplitude modulated signal $S_{BPAM}(t)$ can be represented as:

$$s_{3BPAM}(t) = d_k s_3(t) \quad (2.22)$$

Where $s_3(t)$ is the UWB pulse waveform and in this case it will be represented by the third derivative of the Gaussian pulse. “ k ” represents the transmitted bits (either “0” or “1”) and d_k is employed for the antipodal representation of the transmitted bit k . For this modulation format d_k is defined as:

$$d_k = \begin{cases} -1 & \text{if } k = 0 \\ 1 & \text{if } k = 1 \end{cases} \quad (2.23)$$

Figure 2.8 illustrates the time domain representation of a BPAM modulation scheme.

Figure 2.8: Binary PAM waveforms for “1” and “0” bits in a triplet pulse.

On-Off Keying (OOK):

This modulation format is depicted in Figure 2.9. When compare with PAM, the difference relies on the fact that in On-Off Keying (OOK), no signal is transmitted in the case of bit “0”. The pulse sequence employed for this scheme (d_k) is presented as:

$$d_k = \begin{cases} 0 & \text{if } k = 0 \\ 1 & \text{if } k = 1 \end{cases} \quad (2.24)$$

Figure 2.9: Set of OOK pulses employed to portrait “1” and “0” bits

The major difference between OOK and PAM is that nothing is transmitted in OOK when bit ‘0’ is chosen.

Pulse Position Modulation (PPM)

For Pulse Position Modulation (PPM) the information of the data bit to be transmitted is encoded by the position of the transmitted UWB pulse with respect to a nominal position. In this context, while bit “0” is depicted as a pulse originating at the time instant 0, bit “1” gets shifted in time by the range between c and 0. Analytically, the PPM signal can be represented as:

$$s_{IPPM}(t) = s_I(t - \delta k_d) \quad (2.25)$$

Here $s_I(t)$ denotes the UWB signal (a monocycle in this example), δ indicates the time between two states of the PPM modulation and once again d_k is employed as the modulating bit sequence. The value of δ could be chosen according to the autocorrelation characteristics of the pulse. For the purpose of this example, d_k assumes the following values, depending on the bit chosen to be transmitted:

$$d_k = \begin{cases} 0, & k = 0 \\ 1, & k = 1 \end{cases} \quad (2.26)$$

Figure 2.10 plots a binary PPM where data bit “1” is delayed by a fractional time interval δ whereas a data bit “0” is sent at the nominal time.

Figure 2.10: PPM waveforms for “1” and “0” bits

For PPM, the optimal modulation index is independent of pulse width because of the definition of relative fraction of pulse width. As the order of the derivative increases, the minimum bit error rate is reached for a lower value of δ and better bit error probability (BER) performance is achieved.

Pulse Shape Modulation (PSM):

Pulse Shape Modulation (PSM) represents an alternative to PAM or PPM modulation formats. In PSM, information data is encoded by different pulse shapes. This requires a suitable set of pulses for higher order modulations. Modified Hermite polynomials functions [Alle03] and Prolate Spheroidal Wave functions [Zhua03] have been implemented in current literature as pulses sets for PSM systems. Figure 2.11 depicts a PSM modulation scheme.

Figure 2.11: Depiction of monocycle and doublet waveforms employed for PSM modulation.

2.5 UWB and wireless coexistence.

Given the fact that UWB systems operate in a very large bandwidth, they need to share the spectrum with other users and other communication technologies and hence interferences are likely to occur. Besides from this interference coming from users, the UWB propagation channel itself might cause disturbances. Both type of interferences are important and should be considered in the design, evaluation and implementation of UWB systems [Ghas03]. In Fig. 2.12 the spectrum of UWB system and other wireless systems are shown. As seen from this figure, several other services exist in or in the neighborhood of the UWB band. For example, the 2.4 GHz band WLANs and the GPS system (at 1.5 GHz), mobile cellular system (at 800 MHz and 1.80 GHz band) represent latent main sources of interference to UWB systems. Mutually, UWB systems may affect existing wireless or navigation systems and cause interference to systems such as 802.11a, wireless systems in ISM band, Mobile cellular, and GPS.

Figure 2.12 Frequency allocation of wireless communication standards up to 10.6 GHz, showing narrowband technologies which could be potentially interfered by UWB. [Niko 09]

Due to the intrinsic nature of UWB systems in cohabiting in a large size of spectrum, it is important to study the effect of the UWB interference on spectral neighboring systems. For many years, UWB's closest contender has been WIMAX, so the next fragment will focus on a general comparison of UWB with narrowband technologies in terms of interference and specifically with WIMAX.

2.5.1 UWB vs. Narrowband Systems.

The performance of narrowband fellow receivers in the presence of UWB interference depends on a vast number of parameters such as the pulse repetition frequency (PRF) of the UWB interferer with respect to the bandwidth of narrowband signals, the ratio of narrowband signal power to the UWB signal power.

For the purpose of this brief analysis is important to mention that UWB waveforms tend to be modelled occupying a much larger bandwidth than the narrowband receivers, hence, in the frequency domain, it could be viewed as being relatively constant within the bandwidth of the narrowband receiver.

Figure 2.13: UWB pulse model considered for analysis.

Following the approach of [Foer01], it is assumed that the UWB waveform has a theoretical bandpass shape (such response can be appreciated in figure 2.13) where the PRF $\approx Bp$ and $Bp=1/Tp$ (being Tp the pulse repetition period).

This model assumes that $\omega T = 2\pi m$ for some integer “ m ” and ignores the double frequency term results, since this term will be negligible after integration. In this context, it is also assumed that there is no inter-symbol interference at the output of the matched filter at the sampling instance.

As for the noise, components are independent and identically distributed random variables with a Gaussian distribution and a zero mean. In this sense, the bit error probability (BER) can be derived by conditioning on the random variables that make up the UWB interference terms. The analysis comes to the realization that since the UWB signal occupies a much larger bandwidth than the narrowband receiver. Therefore, we can consider its spectral representation as relatively constant in the bandwidth range of the narrowband system. Furthermore, the signal-to-interference ratio (SIR) within the bandwidth of the narrowband receiver can be obtained in function of the power spectral density of UWB standard pulses (PSD) and the bandwidth of the narrowband receiver.

Finally, the analysis concludes that the performance of the narrowband neighboring system of a UWB interferer depends on several parameters such as the PRF of UWB interferer with respect to the bandwidth of narrowband signal as well as the ratio of narrowband signal power to the UWB signal power within the bandwidth of the narrowband receiver.

2.5.2 UWB vs. WiMAX.

A notorious example of the effect of UWB interference with close competitors can be seen in WiMax receivers. Such analysis is quite significant since Wimax and UWB not only share spectral resources and a common applications' target but they struggle to reach a very similar clients' niche. WiMax systems are based on the IEEE 802.16 standard, work in the frequency range of 2–11 GHz and provide high data rates (in the order of 75 Mbps) to a broad geographical area [Ghos05-Ieee04]. In this way, Wimax systems operate within the 3.1–10.6 GHz range and therefore share this band with UWB.

When studying the coexistence of these two systems, a proper assumption would be to consider a UWB system that works in the vicinity of the victim WiMax system. In this specific case, the UWB signal power at the location of WiMax system is expressed as:

$$P = \frac{\int_{F_{min}}^{F_{max}} PSD(f)df}{PL(r)} \quad (2.27)$$

Where F_{min} and F_{max} are respectively the lowest and highest frequencies of the UWB band, and $PL(r)$ is the path loss in terms of the distance “ r ” between UWB and WiMax systems. As for this path loss, we would consider it as a free space formula, so $PL(r)$ can be written as:

$$PL(r) = PL(r_0) \left(\frac{r}{r_0} \right)^2 \quad (2.28)$$

Here $PL(r_0)$ is the path loss at reference distance r_0 . In the case we are working with a set of UWB systems working in a hot spot and distributed ratio allocated between “ r_{min} and r_{max} ” in the vicinity of a WiMax receiver the total interference power at the victim receiver will reveal a direct dependence on the power spectral density of these UWB interferers. Hence, knowing the number of UWB interferers, and the min and max distances as well as standard distance and its path loss, the received power can be calculated and compared with the maximum tolerable interference for WiMax. In [Sarf05] it is concluded that for different values of the

WiMax receiver bandwidths and up to 50 UWB interferers, the total interference maintains the threshold levels shown in their Table 2.4.

Maximum tolerable interference for WiMax				
BW(MHz)	1.25	2.5	10	20
OFDM WiMax (dBW)	-134.5	-131.5	-127	-123.9

Table 2.4 Maximum tolerable interferences between UWB and WiMax.

2.6 MWP technology as a promising solution for UWB generation.

2.6.1 MWP overcoming UWB limitations.

As it has been stated, the key feature of UWB systems is their capacity to spread a signal over a significant wide bandwidth to guarantee a low power spectral density (PSD) with negligible interference from current wireless system, therefore ensuring different wireless technologies cohabitation. In this context, aligned to the modern FCC approved regulation transmitted levels PSD must hold a value less than -41.3 dBm/MHz for indoor communications. Due to this low PSD in transmission regime, there is a trade-off between the achievable data rate and the feasible coverage area. Though undesirable, within these schemes higher data rate results in shorter communication distances, while a lower rate translates into a relatively longer distances. So, a UWB typical functional range tends to work within a limited distance of a few meters or tens of meters tops. Such short-range wireless networks can operate mainly in an indoor environment in stand-alone mode with virtually no integration into wireless wide-area infrastructures or external long haul fixed wired networks. In order to overcome this coverage limitation and offer availability of uninterrupted service across different technologies, a technique focused on the combination of microwave photonics (MWP) and standard UWB schemes to distribute UWB signals over fiber (UWBoF) has emerged.

The concept of Microwave photonics emerged as the combination of photonic devices operating at microwave frequencies and their application into optical schemes [Seed03-Nova07]. One of the main applications of MWP consists on the transport and distribution of radio or wireless signals over optical media [Nova07], a profitable and very active research field for over more than 25 years. The junction between fiber-optics technologies and wireless networks is commonly known as

hybrid-fiber-radio (HFR) schemes and it represents a baseband technology for the provision of untethered access to broadband wireless communications. The inherent advantages of transmission over optical fiber make it an ideal solution for signal transmission from a central office to remotely located or difficult access external sites, experimental approaches have already benefit from these features [Yang10]. There are several possible approaches when it comes to transporting radio signals over optical fiber each one of them with certain trade-offs. Among the numerous proposal, the intermediate-frequency-over-fiber and the baseband-over-fiber techniques [Qian05, Chia03, Pers06] are the most suitable to make use of mature and reliable RF and digital hardware for signal processing when dealing with central office and remote antenna site architecture. Even though, complications arise in the form of the required complex frequency conversion in the receiver module, MWP in general has succeeded in developing and providing a vast range of devices and solutions.

In the specific case of UWB signal transport, Microwave Photonics does not only solve the geographic distribution constraint, but also supports effective procedures to perform optical modulation formats and signal generation. Originally, UWB was approach as a RF technology and UWB pulses were generated electrically by means of different techniques. For instance, using an electronic circuit, such as a microwave ring filter, a microwave resonator-based bandpass filter, or a microwave differentiator but to distribute UWB pulses over optical fiber it is recommendable that such UWB signals are directly generated in the optical domain. One of the main advantages of UWBoF schemes, with optically generated input UWB signals is the lack of necessity for extra electrical to optical conversion. Furthermore, the generation of UWB pulses directly in the optical domain provides other valuable advantages such as light weight, small size, large tenability and immunity to electromagnetic interference.

2.6.2. Techniques for optical pulses generation.

Firstly, techniques to generate arbitrary waveforms within the optical domain shall be discussed. Moreover, As for the specific case of UWB pulses, the currently available techniques can be classified into three main categories: 1) UWB pulse generation based on Phase-modulation-to-intensity modulation (PM-IM) conversion

through its two principal approaches: employing either a dispersive device or an optical frequency discriminator) , 2) UWB waveform generation using photonic filtering with negative coefficients, where an intrinsic relationship with SOAs nonlinearities is established and 3) UWB pulse generation based on optical spectral shaping and dispersion-induced frequency –to-time mapping. A broad analysis dissection of these procedures will be presented to not only obtain a complete vision of them, but to properly select which ones stand as the better ground base for our own designs.

Spectral line by line Pulse Shaping

Mode-locked lasers generate uniform output optical pulse trains with periodicity defined by the inverse of the laser repetition frequency. Manipulation of continuous pulse trains has attracted interests for many applications including the always expanding niche of high speed optical fiber communications, ultrafast optical signal processing and photonic assisted generation microwave waveforms. The simplest examples for continuous pulse train operation are repetition rate multiplication (RRM) [Azañ01], which are usually implemented by spectral intensity or spectral phase filtering. In addition to simple RRM, continuous pulse trains with controllable pattern and envelope are also proposed and demonstrated [McKi03]. Among the almost arbitrary waveforms enabled by pulse shaping, a subset of particular interest is optical pulse train generation and manipulation.

In the frequency domain, mode-locked laser pulses can be characterized by a series of discrete spectral lines (optical frequency comb) with the frequency interval equal to the longitudinal mode spacing of the laser, or equivalently to the pulse repetition rate [Cund02]. From the pulse shaping perspective, if it is possible to independently manipulate both the amplitude and phase of individual spectral lines which translates into the capacity of achieving full pulse shape control. Specifically, waveforms produced from adjacent pulses can overlap with each other. However, in typical pulse shapers [Wein00] spectral lines are manipulated in groups rather than individually. This is primarily due to the practical difficulty of building a pulse shaper capable of resolving each spectral line for typical mode-locked lasers with repetition rates below 1 GHz. Group-of-lines pulse shaping is illustrated in Fig. 2.14 (a) , where f_{rep} is the repetition rate. Assuming that the pulse shaping occurs M lines

at a time, the shaped pulses have maximum duration $\sim 1/(Mf_{rep})$ and repeat with period $T = 1/f_{rep}$. Accordingly, the pulses are isolated in time. In contrast, for line-by-line pulse shaping ($M \approx 1$) as shown in Fig. 2.14 (b), the shaped pulses can overlap with each other, something that leads to interference between contributions from different input pulses in the overlapped region.

Figure 2.14: Illustration of pulse shaping (a) Manipulating groups of lines.

(a) Manipulating individual lines. [Wein00]

The capability of line-by-line pulse shaping immediately leads to efforts in pursuing true optical arbitrary waveform generation [Wein02], here we demonstrate optical arbitrary pulse train generation. Other approaches include [Kuo10], where a qualitative agreement with a theoretical analysis of spectral power enhancement by optical short pulses comprised of equi-amplitude frequency lines over sinusoidal excitation is accomplished in the way of high-modulation-depth photonic millimeter-wave (MMW) waveforms generator by applying line-by-line pulse shaping on a phase-modulated continuous-wave frequency comb.

Coherent frequency to-time mapping

The Coherent frequency to-time mapping technique has been popular over the last decades. The so-called frequency-to-time mapping phenomenon relies on the fact that, after linear propagation in a dispersive medium, the temporal intensity profile of an ultrashort optical pulse becomes a scaled version of its optical spectrum [Tong97]. The ability to measure or manipulate the spectral information in the temporal domain has enabled new metrology and signal processing schemes with a superior performance in terms of compactness and operation speed with respect to more conventional approaches. This solution has been employed in a myriad of interesting application fields such as spectroscopy [Soll08], imaging [Goda09], optical

communications [Mcki10], pulse shaping [Thom10], and arbitrary radio-frequency waveform generation (RF-AWG) [Khan10].

Applications of this solution are widely spread in current bibliography, in [Abta09] we can appreciate an outstanding optical approach in which a simple and practical UWB waveforms generator with high precision and arbitrary characteristics is achieved. In this scheme, waveforms are pre-calculated and uploaded to the pulse generator as required. The spectral pulse shaping is performed using a FBG with temperature controlled apodization, along with a series of heating elements that are used to change locally the temperature of a uniform linearly chirped FBG. The required spectral resolution of the pulse shaping device can be achieved by using the appropriate size heating elements.

Incoherent frequency to-time mapping

With the before mentioned popularity of optical arbitrary waveform generation and its associated approaches, many interesting breakthroughs have come to light along with novel and promising techniques, each category has its own advantages and in the case of wavelength-to-time mapping, this solution proves to be very flexible and robust since the spectral shape is easily changed in the frequency domain, including both coherent techniques and incoherent techniques. Coherent techniques offer high-fidelity output waveforms but at the expense of expensive coherent sources. The incoherent wavelength-to-time mapping concept was proposed and demonstrated first by [Dorr04] offering a low-cost solution but the output waveform had a low signal-to-noise ratio. In this sense, the incoherent frequency to time mapping phenomenon establishes that after temporal external amplitude modulation and transmission through a dispersive media of a spectrally incoherent optical source, the averaged intensity profile at the so-called temporal far-zone regime coalesces with a magnified replica of the spectral density function of the source.

A very interesting application of the generalization of the frequency-to-time mapping operation to the incoherent applied to the field of arbitrary RF waveform generation is depicted in [Comp08]. Here, a spectrally incoherent stationary light source is manipulated in amplitude with a reconfigurable filter. The modified optical source is actively modulated with an electro-optic modulator providing optical pulses

with a temporal duration larger than the inverse of the spectral resolution of the shaper in order to satisfy the condition of quasi-homogeneous light pulse. After this, the modulated light signal is launched into a GDD medium, hence satisfying the mandatory condition of [Comp07]. Finally, the synthesized spectral density is transferred to the averaged temporal profile and this pulse is finally mapped into the electrical domain by using a high-speed photodiode.

2.6.3 Experimental solutions focused on UWB signals generation:

Phase Modulation-to-Intensity Modulation:

A simple solution to achieve first or second order differentiation is to implement optical phase modulation and phase modulation to intensity-modulation (PM-IM) [Yao10b]. Figure 2.15 shows these two main approaches' fundamental schemes and Chapter 3 will delve into the mathematical analysis of these proposals and their contribution to our experimental work. In these layouts, in order to recover the information being carried by the optical phase and avoid a cancelation effect as a result of the optical-electrical stage after a phase modulation process, current literature [Zeng04, Zeng05, Wang05]. is based on two main solutions to convert a phase-modulated signal into an intensity modulated one. The first method [Zeng05, Wang05] involves a to use a dispersive device to change the phase relationship among the two first-order sidebands from out of phase to partially or fully in phase, in the second method [Zeng06] an optical filter is used to act as an optical filter to perform PM-IM conversion.

Figure 2.15: PM-IM conversion based generation techniques. (a) Chromatic dispersion based PM-IM conversion layout (b) UWB pulses generation based on PM-IM conversion through an optical frequency discriminator. [Yao10b]

Microwave Photonic filtering

A Gaussian monocycle or doublet can be produced by performing the first or second-order derivative of a Gaussian pulse respectively [Chen02]. Mathematically, the first-or-second-order derivative can be approximated by the first- or second-order difference, which can be implemented using a two-or three tap photonic microwave delay-line with coefficients $[1, -1]$ in the case of the monocycle and $[0.5, -1, 0.5]$ for a doublet. To successfully implement this filtering solution, a negative coefficient must be generated. A variety of approaches have been proposed for the generation of negative coefficients. Traditionally, a photonic microwave delay-line filter with incoherent detection can usually have positive coefficients only [Seed02, Mina06, Capm06]. Among the techniques applied within this field we can mention [Wang06] based on cross gain modulation (XGM) in a SOA and [Pan10] which employs a PolM (polarization modulator). Fig 2.16 depicts both schemes and just as its predecessor, further mathematical analysis will be performed in the coming chapter.

Figure 2.16: PM-IM conversion based generation techniques. (a) Chromatic dispersion based PM-IM conversion layout (b) UWB pulses generation based on PM-IM conversion through an optical frequency discriminator. [Yao 07]

2.6.4 Solutions for high order pulses

In this context, as stated in section 2.4.1, doublet and monocycle pulses are considered standard pulses for UWB applications but they fail in satisfying fully FCC regulations, so higher order pulses are found more profitable pulses. Among the most relevant techniques related to obtain high-order pulses, we can find traditional arbitrary waveform generators based on microwave photonic filtering [Bole09] and nonlinearities exploitation by means of Semiconductor Optical Amplifiers (SOAs) [Feng14]. Although effective, these approaches require the insertion of one coefficient into the system for each Gaussian based pulse handled or their generation structure relies on complex non-linear effects that disable any reconfiguration capabilities. Therefore, these designs cannot stand as fully optimized proposals because they comprise a constrained or costly solution. In this way, a very valuable alternative technique has been recently proposed for obtaining UWB high order signals consisting on the merge of various low order UWB pulses, such as monocycles and doublets to create triplets or quadruplets [Zhou10, Abra11a] and be visualized in Figure 2.21.

Figure 2.17: Graphic depiction of higher order pulses forming through merge of lower order Gaussian derivatives.

Once again, though these approaches can be considered competent compared to the abovementioned schemes which only employ Gaussian based pulses, they present some deficiencies such as lack of flexibility, low configuration capacity related to the

reconfiguration and scalability and, finally, limited integration feature. This research is strictly focused on the presentation and experimental validation of highly flexible photonic architectures, which will be presented on chapter 4.

Finally, many reports have focused on the optical generation of impulse-radio UWB signals in recent years. Not only generation but also the performance of the entire UWBoF design has also been taken into consideration, but also Radio-over-fiber distribution of impulse-radio and OFDM UWB signals in FTTH networks has been proposed and compared [Llor08a, Llor08b]. Generation procedures vary according to the aspired UWB pulse and the complexity of the system. To make an easier general visualization of these concepts, tables 2.5 and 2.6 summarize the state-of-the-art generation techniques and their advantages and constraints in terms of fractional bandwidth and spectral efficiency for lower and higher order derivatives.

Ref.	Generated waveforms	Technique	Scalability	Wavelength Efficiency	FCC Mask fit	Considerations
[Chen07]	Monocycles Doublets	Monocycle: polarity inverse Gaussian like pulses by means of birefringence time delay. Doublet: gain saturation and recovery effect in SOA	NO	YES	Lower order pulses do not fully fulfill FCC's settled mask.	<ul style="list-style-type: none"> • Monocycle's fractional bandwidth $\approx 140\%$. • Doublet's fractional bandwidth $\approx 160\%$. • Feasibility for PPM format performance.
[Zeng06]	Monocycles Doublets	PM-IM conversion by employment of FBG as frequency discriminator. By locating the optical carrier at the linear or the quadrature slope of the FBG reflection spectrum, several monocycles or doublets can be generated.	NO	YES.	Even though working with low order derivatives, customization of these signals could provide	<ul style="list-style-type: none"> • Monocycle's FWHM= 52 ps. • Doublet's FWHM= 42 ps. • Stability of the output pulses relies on the wavelength stability of the laser source and the FBG employed.
[Wang06]	Doublets	Pulse shaping through tuning the bias voltage of an EOM near minimum and maximum transmission points. By switching the BIAS points the polarity of the obtained doublet can be reversed.	NO	YES	Poorly and not holding pulse customization capabilities.	<ul style="list-style-type: none"> • Doublet's width ≈ 270 ps. • Capacity for BPM format performance.
[Lin05]	Monocycles Doublets	Open-loop reflection-mode dispersive Fourier transform based pulse shaping.	NO	YES	Poorly and not holding pulse customization capabilities.	<ul style="list-style-type: none"> • Monocycles's width ≈ 200 ps. • Doublet's width ≈ 450 ps. • Possibility for PPM format performance.
[Dong08]	Monocycles Doublets	Pulses generation based on the nonlinear effects (XGM, SGS and FWM) intrinsic to a SOA device	NO	YES	Poorly and not holding pulse customization capabilities	<ul style="list-style-type: none"> • Doublet's FWHM=53 ps and fractional bandwidth $\approx 125\%$. • Presence of three wavelength components requires compensation modules in transmission schemes

Table 2.5: Comparative Matrix corresponding to available schemes for low order UWB pulses generation.

Ref.	Generated waveforms	Technique	Scalability	Wavelength Efficiency	FCC Mask fit	Considerations
[Bole09]	Monocycles Doublets Triplets	Microwave photonic filtering and adaptable phase inversion	YES	NO	Yes, in the case of the higher order pulses	<ul style="list-style-type: none"> • Scheme easily adapted to produce modulations formats such as: PAM, OOK, OPM, PPM and BPM.
[Abta08]	Monocycles Doublets Triplets	Optical pulse shaping	YES	YES	Yes, in the case of the triplet signal.	<ul style="list-style-type: none"> • Monocycle's width ≈ 62.4 ps • Doublet's width ≈ 66.3 ps • System is quite stable and insensitive to environmental changes.
[Li10]	Monocycles Doublets Triplets Quintuplets	Summation of multiple monocycle pulses with inverted polarities	YES	YES	Yes, for the higher order derivatives.	<ul style="list-style-type: none"> • Monocycle's fractional bandwidth $\approx 177\%$. • Doublets's fractional bandwidth $\approx 166\%$ • Triplet's fractional bandwidth $\approx 98\%$. • Quintuplet's fractional bandwidth $\approx 113\%$. • OOK, BPSK, PPM and PSM can be implemented
[Zhou10]	Monocycles Doublets Triplets	Summation of asymmetric complementary pulses obtained through multiple PM-IM conversions processes.	YES	NO	Yes, in the case of the accomplished triplet.	<ul style="list-style-type: none"> • Triplet's spectral efficiency $\approx 50.59\%$. • High speed transmission without pulse overlapping and distortion feasibility.
[Feng14]	Monocycles Doublets Triplets Quadruplets	Exploitation of non-linear effects intrinsic on SOAs for low order waveforms and overlapping of these pulses with orthogonal polarization for higher order ones.	YES	YES	No, not even in the higher orders signals.	<ul style="list-style-type: none"> • Monocycle's fractional bandwidth $\approx 176\%$. • Doublets's fractional bandwidth $\approx 140\%$ • Triplet's fractional bandwidth $\approx 63\%$. • Quintuplet's fractional bandwidth $\approx 69\%$. • High order UWB signals are tolerance to fiber dispersion, transmission through long length fiber links

Table 2.6: Comparative Matrix corresponding to available schemes for high order UWB pulses generation

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CHAPTER 3

Fundamental concepts of SOA nonlinearities and PM-IM conversion for optical pulse generation

In this chapter, we will analyze the selected techniques to generate UWB signals directly within the optical domain. Their theoretical analysis will set the required premises for a better understanding of the proposals to be exposed in the next chapter, at the same time that expands the already established basis of these systems. In the first segment, nonlinear effects present in an SOA are employed to expose a model that permits to describe both lower and high order UWB pulses. The second part of this section will reveal a detailed explanation of PM-IM based solutions that not only permit high order UWB pulses creation but also provide wavelength employment efficiency.

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On the last chapter, the state of the art techniques related to UWB Communications technology was presented. Tables 2.5 and 2.6 revealed which techniques offer a vast majority of advantages and the expected performance for each one of them. A clear preference towards fully optical pulses generation procedures was found.

In this first section, we will present the current solutions based on the employment of nonlinearities intrinsic to SOAs and then an analysis of the Cross-phase-modulation (XPM) effect inherent to the designed proposed.

3.1 Theoretical principles for SOAs non-linear effects.

Since non-linear effects will be vastly discussed through this chapter, it is essential to have a clear picture of the fundamentals related to Semiconductor Optical Amplifiers (SOAs). They are optoelectronic devices that under suitable operational parameters can amplify an input light signal. In a basic configuration, they are very similar to lasers except they have no reflecting facets [Sait88, Maho88]. A typical amplifier chip value is ranged between 0.6 to 2 mm long. It has a p-cladding layer, an n-cladding layer and a gain region all of which are grown under epitaxial behavior over a binary substrate. Figure 3.1 depicts a schematic plot of a basic SOA. The amplifier chips have cleaved facets with anti-reflection coating (and possibly other modifications) which reduce its reflectivity to nearly zero. Just like the laser, the amplifier has a p-n junction which is forward biased the external injected current (BIAS) stands as the energy source that enables gain to take place. However, the optical confinement is weak and the output signal is accompanied by noise. This additive noise is generated as a direct result of an amplification process and cannot be completely avoided.

Figure 3.1: Diagram of a Semiconductor Optical Amplifier (SOA) [Conn04]

The principle of amplification in an SOA is based on the injection of SOA carriers (or electrons) from an external current source into the active region. These carriers occupy

energy states in the conduction band (CB) of the active region material, leaving holes in the valence band (VB). Three well differentiated radiative stages are present in the semiconductor structure [Dutt06]. The first one is a loss process called stimulated absorption, where an incident light photon of sufficient energy can stimulate a carrier from the VB to the CB, resulting this into the photo extinction. The second one, it is known as stimulated emission. In this case, a state of optical gain takes place by means of a stimulated recombination process of a CB carrier with a VB hole from a photon of light with suitable energy levels. Finally, we can observe, spontaneous emission processes, where there exists a non-zero probability per unit time of carrier and hole recombination which translates to the emission of photons with random phase and direction, which essentially represent the inherent noise present in SOAs. Therefore to recapitulate, stimulated processes are proportional to the intensity of the inducing radiation while the spontaneous emission process is independent of it.

As for their applications, SOAs were originally conceptualized as amplifiers in fiber communication systems or other types of optical amplifier applications. However, they can also fit into functions that are particularly useful in optical transparent networks. This versatility is sustained by their capacities of high gain and gain bandwidth, negligible facet reflectivity, low polarization sensitivity, high saturation output power, additive noise near the theoretical limit, insensitivity to the input signal modulation characteristics and multichannel amplification with no crosstalk capacity [Zhan10]. Besides, the development within the integrated photonics research area has opened the window of opportunity for less bulky and hybrid SOAs integrated structures [Andr13- Grub08] which has also contributed to a fast pace growth for this technology.

In this context, SOAs can be employed for optical signal processing by means of intensity or phase modulation of a continuous wave (CW) signal. The refractive index of an SOA active region is not constant but is dependent on the carrier density and hence the material.

This translates to the phase and gain of a propagating optical wave to be coupled through gain saturation. The magnitude of this coupling is related to the linewidth enhancement factor (α_l), also known as Henry factor or Alpha factor. Its name derives from its direct influence on the spectral linewidth of semiconductor lasers. It is expressed as:

$$\alpha_l = -\frac{\pi}{\lambda_0} \frac{dn_e / dn}{dg_m / dn} \quad (3.1)$$

Here, λ_0 represents the free-space wavelength, n_e the amplifier waveguide effective index, g_m the material gain coefficient and n the carrier density [Osin87]. Intensity modulation can be accomplished by simply modulating the amplifier current with a suitable data signal as illustrated in figure 3.2. The amplifier gain and the optical input are modulated by the current modulation. The modulation bandwidth is limited by the amplifier carrier lifetime; hence the output signal finite rise and fall times depend on the non-zero carrier lifetime. The linearity requirements of analog optical transmission systems generally preclude the possibility of introducing SOAs to their schemes as external modulators due to carrier density induced second order non-linear distortion. Plausibility of an SOA as external modulator in a multichannel subcarrier multiplexed digital video link is demonstrated in [Ali97].

Figure 3.2: SOA application as an external intensity modulator.

The phase of a propagating signal can also be modulated via carrier density dependence of the amplifier effective index. This phase modulation response is directly proportional to α_L . The linewidth enhancement factor increases proportionally with the photon energy in the proximity of the bandgap energy. Therefore, the optimum phase response of the amplifier might be obtained when selecting an operating wavelength on the long-wavelength side of the amplifier gain spectrum. Phase modulators are highly useful in communications links [Zhan11, Mads06, and Herc13].

Wavelength conversion is another appealing functionality SOAs hold. All optical wavelength converters play a key role in current optical networks since they increase flexibility and capacity features [Durh96]. SOA based wavelength converters benefit from the intrinsic SOAs' nonlinearities [Dosh03, Cao02]. Hence, it is fair to say that among the great dossier of interesting feasible applications for SOAs, exploitation of these devices nonlinearities represents a valuable and strategic asset for optical communications. Nonlinearities in SOAs are principally a consequence of carrier density changes induced by the input signal. The four principal non-linear effects that might take place in semiconductor optical amplifiers are Cross-gain modulation (XGM), Cross-phase modulation (XPM), Self-

phase modulation (SPM) and Four-wave mixing (FWM). As for the proposals to be presented, these designs rely on the premises of an interferometric configuration to convert phase changes into intensity ones using constructive or destructive interference therefore, a detailed understanding of XPM basis is required.

3.2 Cross-phase modulation (XPM).

As it was established in equation 3.1, the linewidth enhancement factor increases rapidly as the photon energy nears the bandgap energy due to the sensitivity of the material gain coefficient to carrier density in the region. This α factor also holds a direct relationship with the carrier density, when a signal pulse is injected into an SOA it causes carrier density changes as well as a change in its propagation coefficient. Given the case that more than one signal is injected into an SOA, there will be cross-phase modulation (XPM) between the signals. XPM can be used to create wavelength converters and other functional devices. However, as abovementioned this nonlinearity demands to be placed in some interferometric structure to be capable of performing PM-IM conversion. The traditional interferometric configurations are the Mach Zehnder, the Michelson and the Sagnac [Sale 91]. Although all of the previous have been applied in the UWB pulses generation field [Bahr12, Li08, Ober97], it is the MZI based solution the selected one in our first approach.

All-fiber Mach-Zehnder interferometers are constructed by connecting two fiber couplers in series. The first coupler splits the input signal into two parts, which acquire different phase shifts if arm lengths are different before they interfere at the second coupler. So, basically, an optical input beam is divided into two beams that travel across different paths before they are combined at the output stage. Moreover, an MZI can be unbalanced by simply using different lengths for its two arms since the two optical fields inside it take physically separated paths. However, the same feature also makes the interferometer susceptible to environmental fluctuations.

For a standard SOA-MZI configuration, two inputs are required. The probe signal (P_{wc}) at a specific wavelength of λ_{cw} and the pump signal (P_s) at its corresponding wavelength λ_s . Depending of the propagating direction of both signals, two basic schemes employed in SOA-MZI configuration are defined: the co-propagating and the counter-propagating layout. As it can be visualized in figure 3.3, the pump and probe signals can be launched together or in opposite directions into the amplifier.

The pump signal is in charge of modifying the carrier density of the upper branch SOA (SOA_1) according to its modulation and hence taking into consideration both gain and phase factors. This alteration might affect the probe signal in terms of amplitude (as a result

of XGM) and in phase (as a direct effect of XPM) in SOA₁. The couplet output recombines the probe signal where they can interfere constructively or destructively according to the selected branch for signal entrance and output for such probe. This process defines the information transfer from the pump to the probe.

It has been demonstrated that under certain conditions the co-propagation scheme reveals a superior noise performance [Ober 97]. However, the main advantage of the counter-propagation scheme is that no optical filter is required to divide the modulated probe signal from the pump signal, in both configurations the converted probe signal is inverted compared to the input signal.

Figure 3.3: Layout for (a) co-propagation and (b) counter-propagation approach.

The main component for our first designed approach [More13] is constituted a symmetric all optical monolithically integrated Mach Zehnder interferometer with two integrated SOAs of 1 mm length. The device holds five ports tagged as P1 or P3 for the input optical data signal, P2 for the wavelength converted optical data signal output in the case of a counter-directional approach and P4 or P5 for the co-propagation configuration. In the aforementioned solution, we proposed and experimentally demonstrated a novel scheme to generate UWB doublets based on exploiting the nonlinearities present in an integrated interferometric structure with a counter propagation configuration. The conversion process depends directly of the currents applied in each branch. A pump signal is introduced to the setup to induce the XPM effect in the SOA-MZI and as a result of the non-linear

characteristics of the transfer function we can achieve the generation of UWB doublets at a specific operation point that is dictated by the saturation of the SOAs. The theoretical fundamentals of this approach rely on an accurate knowledge on the equations that rule the performance of SOA-MZI structures.

The presented scheme holds a counter propagation layout and though either P1 or P3 can be employed as the input port, the pump signal (P_s) will be introduced through P1. Moreover, ports P4 and P5 will be employed as input ports for the probe signals (P_{WC}) while P2 will represent the system's output. In figure 3.3 (b), the factors a, b, c and d will represent the coupling parameters. Following the analysis presented in [Manz13] the general expression for the electromagnetic field $E_{OUT}^{P2}(t)$ in output P2, when the probe $E_{CW}^{P4}(t)$ goes through P4:

$$E_{OUT}^{P2}(t)|_{P4} = \sqrt{I-d} \cdot \sqrt{G_1(t)} \cdot \left[j\sqrt{b} \cdot E_s^{P1}(t) + \sqrt{I-b} \sqrt{I-a} \cdot E_{CW}^{P4}(t) \right] \cdot e^{j\phi_1(t)} + j\sqrt{d} \cdot \sqrt{G_2(t)} \cdot \left[j\sqrt{I-c} \sqrt{a} \cdot E_{CW}^{P4}(t) \right] \cdot e^{j\phi_2(t)} \quad (3.2)$$

To simplify this model, an ideal case of all the coupling factors holding the same value of 0.5 has been assumed, hence $a=b=c=d=0.5$. The corresponding converted pump signal of:

$$E_{OUT}^{P2}(t)|_{P4} = \frac{I}{2^{3/2}} \cdot \left[\sqrt{G_1(t)} \cdot e^{j\phi_1(t)} + \sqrt{G_2(t)} \cdot e^{j\phi_2} \right] \cdot E_{CW}^{P4}(t) \quad (3.3)$$

Another important consideration consists on assuming that the variations of the gain factor G_I , are negligible when compared to the phase variations ϕ_I , since for this model a very small modulation index feature is considered. Hence, it can be postulated that $G_I(t) \approx G_I$.

And finally the total output power based on the electromagnetic field for input through the ports P4 and P5 respectively can be concluded as:

$$P_{OUT}^{P2}(t)|_{P4} = \frac{P_{CW}}{8} \cdot \left[G_I + G_2 - 2\sqrt{G_I G_2} \cdot \cos(\phi_{RF}(t) - \Delta\phi) \right] \quad (3.4)$$

$$P_{OUT}^{P2}(t)|_{P5} = \frac{P_{CW}}{8} \cdot \left[G_I + G_2 + 2\sqrt{G_I G_2} \cdot \cos(\phi_{RF}(t) - \Delta\phi) \right] \quad (3.5)$$

For a co-propagation scheme the analysis approach follows the same premises, the main difference relies on how to obtain the complementarity among the output ports. As for the

counter-propagation layout such complementarity is obtained by means of the input ports of the SOA-MZI (P4 and P5 in the case of this configuration).

3.3 Phase-to-intensity modulation conversion.

As previously stated, various designs have been proposed in this thesis to accomplish full optical pulses generation. Phase modulation-to-intensity modulation (PM-IM) conversion stands as the second selected solution, because of their scalability and flexibility. In this segment, the main two conceptual approaches for such technique are described at the same time. In this order of ideas, the first depicted method is plotted in Figure 3.4. Here, a basic layout of PM-IM conversion based on the combination of a phase modulation stage with a dispersive element is presented. The principle of this technique can be explained through the magnitude relationships among the sidebands and the carrier are changed, leading to PM-IM conversion.

Figure 3.4: Generic chromatic-dispersion based PM-IM conversion.

In this case, if this phase modulated signal was directly detected by a photodetector (PD) no signal except a DC will be generated. Hence to recover the information that is carried by the optical phase, a dispersive device is employed to change the phase relationship between the two first-order sidebands from out of phase to partially or fully in phase. The following analysis will explain in detailed this phenomenon. Figure 3.5 portrays a basic phase modulation scheme with a photo detection stage.

Figure 3.5: Basic Phase-Modulation (PM) with photodetector scheme.

From figure 3.5, the electrical field of a phase modulated signal will be $e_{PM}(t)$ and can be expressed as:

$$e_{PM}(t) = e_0 \cos[\omega_0 t + Y_{PM} \cdot f(t)] \quad (3.6)$$

Where e_0 and ω_0 are the amplitude and the angular frequency of the optical carrier, Y_{PM} is the phase modulation index defined as the phase change of the optical carrier when a unit voltage is applied (radians per volt) and $f(t)$ is the electrical modulating signal. Such modulating pulse is a single-frequency sinusoidal signal with a zero initial phase that can be expressed in terms of its angular frequency ($\omega_m = 2\pi f_m$) and its amplitude (A_e) as:

$$f(t) = A_e \cos \omega_m t \quad (3.7)$$

The electric field expression can be stated in terms of Bessel functions of the first kind from where we can extract that the phase modulation process generates a series of sidebands with amplitude coefficients that are determined by these Bessel functions. In the case of small-signal modulation, only the first-order upper and lower sidebands need to be considered while the higher order sidebands can be ignored. Under these considerations, the electrical phase modulated field is represented by:

$$e_{PM}(t) \cong e_0 \left\{ \begin{aligned} &J_0(Y_{PM} A_e) \cos \omega_0 t + J_1(Y_{PM} A_e) \cos \left[(\omega_0 + \omega_m) t + \frac{\pi}{2} \right] \\ &+ J_{-1}(Y_{PM} A_e) \cos \left[(\omega_0 + \omega_m) t - \frac{\pi}{2} \right] \end{aligned} \right\} \quad (3.8)$$

Applying the Fourier transform on equation 3.8, we obtain the spectra of this phase modulated signal in which the lower and upper sidebands are exactly out of phase. Finally, equation 3.9 illustrates the frequency domain expression:

$$\begin{aligned} E_{PM}(\omega) &\approx \pi e_0 J_0(Y_{PM} A_e) [\delta(\omega - \omega_0) + \delta(\omega + \omega_0)] \\ &- j\pi e_0 J_1(Y_{PM} A_e) [\delta(\omega + \omega_0 + \omega_m) - \delta(\omega - \omega_0 - \omega_m)] \\ &- j\pi e_0 J_{-1}(Y_{PM} A_e) [\delta(\omega + \omega_0 - \omega_m) - \delta(\omega - \omega_0 + \omega_m)] \end{aligned} \quad (3.9)$$

Figure 3.9 plots the spectral depiction of the modulating signal and the obtained phase modulated one corresponding to equation (3.9):

Figure 3.6: Schematic spectra representation of (a) the single-frequency modulating signal and (b) the obtained phase modulated optical pulse.

3.3.1 Chromatic-Dispersion based PM-IM conversion.

Taking again the figure 3.4 as reference, in this scheme the phase modulated signal propagates through a dispersive device and then fed to a PD. The dispersive media is assumed to hold a magnitude response with low loss and a quadratic phase response. For example, the dispersive media might be represented by a single mode fiber (SMF) link.

Under these conditions, the optical signal at the output of the SMF link can be presented in terms of the phase shifts experienced by the optical carrier which can be tagged as θ_0 , θ_{-1} and θ_{+1} . In the particular case of these phase shifts, they hold a direct relationship with the fiber length since $\theta = \beta z$, where β represents the propagation constant and z the length of the traveled distance. Mathematically the effects of fiber dispersion are accounted for by expanding the mode propagation constant in Taylor series about the frequency ω_0 at which the pulse spectrum is centered [Agra01]. Therefore, we have that:

$$\theta = z\beta_0(\omega_0) + z\beta_1(\omega_0)(\omega - \omega_0) + \frac{1}{2}z\beta_2(\omega_0)(\omega - \omega_0)^2 + \dots \quad (3.10)$$

Where β_1 and β_2 are the first- and second- order derivatives of β with respect to the optical angular frequency. The parameter β_2 represents the dispersion of the group velocity known as the GVD parameter. The group delay $\tau(\omega)$ is established by differentiating, with respect to optical frequency, the phase response. Neglecting third and higher order derivatives of β , we can define the group delay as:

$$\tau(\omega) = \frac{d\theta}{d\omega} = z\beta_1(\omega_0) + z\beta_2(\omega_0)(\omega - \omega_0) \quad (3.11)$$

Here, the first term $\tau_0 = z\beta_1(\omega_0)$ is the experienced group delay τ_0 by the optical carrier at frequency ω_0 and the second term describes the group delay variation as a linear function with respect to angular frequency ω . As for the first-order chromatic dispersion, this feature is defined as the first-order derivative of the group delay with respect to angular frequency ω , which is constant and represented by $D_\omega = z\beta_2(\omega_0)$. Evaluating θ at the optical carrier of ω_0 and the sidebands of $\omega_{0\pm} \omega_m$ we can find that:

$$\begin{aligned}
 \theta_0 &= z\beta_0(\omega_0) \\
 \theta_{-1} &= z\beta_0(\omega_0) - \tau_0\omega_m + \frac{1}{2}D_\omega\omega_m^2 \\
 \theta_1 &= z\beta_0(\omega_0) + \tau_0\omega_m + \frac{1}{2}D_\omega\omega_m^2
 \end{aligned} \tag{3.12}$$

Once the light wave passes through the dispersive element stage, a photocurrent is generated at the PD. If we only assume the RF signal that is centered at the modulating frequency ω_m as valid and neglect the DC current along with the higher order harmonics, we obtain:

$$i_{PD} \propto \sin\left(\frac{\theta_{+1} + \theta_{-1}}{2} - \theta_0\right) \cos\left(\omega_m t + \frac{\theta_{+1} - \theta_{-1}}{2}\right) = \sin\left(\frac{1}{2}D_\omega\omega_m^2\right) \cos(\omega_m t - \tau_0\omega_m) \tag{3.13}$$

Equation 3.13 depicts in its first term the frequency response of the chromatic-dispersion-based PM-IM conversion with direct detection. The frequency response can be visualized in figure 3.7 from which we can affirm that we are dealing with a function with a notch at DC and a sinusoidal dependence with the square of the RF frequency. It is also possible to extract that the first peak and the second notch by setting up $D_\omega\omega_m^2/2 = \pi/2$ and π , respectively.

Figure 3.7: Frequency response of the chromatic-dispersion-based PM-IM conversion.

The frequency response between these first two notches creates a passband filter which can be employed to shape the spectrum of the modulating signal. The effective transfer function of this phase-modulation-based microwave filter can be expressed as the multiplication of the dispersion effects induced frequency response $H_1(\omega)$ and the response from the summation of all the recovered RF signals $H_2(\omega)$ by means of the expression:

$$H(\omega) = H_1(\omega)H_2(\omega) \tag{3.14}$$

In the case of $H_1(\omega)$ when considering the entire array of optical inputs, each laser source can be considered as an independent optical carrier. If we take the assumption that the wavelength spacing between any adjacent laser diode is very small and all the laser sources hold an identical input power at the phase modulator, then the aforementioned response for each carrier can be considered identical. Figure 3.8 plots the experimental frequency response of $H_1(\omega)$ according to [Wang05].

Figure 3.8: Frequency response corresponding to $H_1(\omega)$. [Wang05]

But the phase of the n^{th} RF signal recovered from the n^{th} modulated optical signal is not identical because different carriers travel at different velocity along the SMF link. In this order of ideas, the sum of all the recovered RF signals will determine the $H_2(\omega)$ response through the expression:

$$H_2(\omega) \propto \sum_{n=1}^N e(j\theta_n) = \sum_{n=1}^N e[j\omega_m(n-1)T] \quad (3.15)$$

Where N is the number of optical inputs employed, φ_n is the phase of the n^{th} recovered RF signal, ω_m is the modulating angle frequency and T stands for the time interval of any two adjacent taps. Figure 3.9 shows the corresponding frequency response of $H_2(\omega)$ within a phase range allocated between 0° and 45° according to the reference approach [Wang 05].

Figure 3.9: Frequency response corresponding to $H_2(\omega)$. [Wang 05]

Based on the features present in $H_1(\omega)$ and $H_2(\omega)$ it is possible to conclude that the baseband resonance of the low pass filtering function due to the conventional intensity-modulation direct detection (IM-DD) scheme can be eliminated, and this all-optical equivalent band pass filter is accomplished. For this structure, if the input modulating signal is a Gaussian pulse, the spectrum at lower frequency will be filtered out by the frequency response of the PM-IM conversion and finally the spectra corresponding to a Gaussian monocycle or doublet can be generated.

3.3.2 Frequency discriminator employment for PM-IM conversion.

Optical PM-IM conversion can also be implemented by means of an optical frequency discriminator. As it was discussed in the previous segment, a phase modulated signal holds two out of phase sidebands, whose relationship can be changed to partially or totally in phase through the PM-IM conversion process. In this approach, if the mentioned phase modulated pulse is sent to a properly design frequency discriminator, depending on the location of the optical carrier at the linear or quadrature slope, a UWB monocycle or doublet could be generated. In addition to the PM-IM functionality, the location of the optical carrier at the positive or negative slope would lead to the generation of UWB pulses with opposite polarity. The frequency response of an ideal optical filter can be appreciated in figure 3.10.

Figure 3.10: Frequency response of an ideally designed optical filter with two linear slopes and a flat top [Yao07.]

Mathematically, the frequency response of this filter can be expressed:

$$|H_d(j\omega)| = \begin{cases} \text{Left_slope:} & \text{Center:} & \text{Right_slope:} & \text{Otherwise} \\ K\omega - K(\omega_1 - \Delta\omega) & K\Delta\omega & K(\omega_2 + \Delta\omega) - K\omega & 0 \\ \omega_1 - \Delta\omega \leq \omega \leq \omega_1 & \omega_1 \leq \omega \leq \omega_2 & \omega_2 \leq \omega \leq \omega_2 + \Delta\omega & \end{cases} \quad (3.16)$$

Herein, K stands for the slope of the filter ($K > 0$) and ω is the optical frequency. From equation 3.16, it is possible to determine that this spectral response is fundamentally constituted by three linear sections: the left slope, the center and the right slope. Based on the premises assumed by [Yao07], we will assume that since the filter to employ is FBG, in each and every one of these segments the phase response is linear. Furthermore, if the phase modulated signal has a narrow bandwidth and the optical carrier is properly selected to make it located at the center of each spectral section of the frequency response of the optical filter, the impulse of the optical filter $h_d(t)$ can be expressed as:

$$h_d(t) \approx \begin{cases} -K(\omega_1 - \Delta\omega)\delta(t) - jK\delta_1(t) & \text{Left_slope} \\ K\Delta\omega\delta(t) & \text{Center} \\ K(\omega_2 + \Delta\omega)\delta(t) + jK\delta_1(t) & \text{Right_slope} \end{cases} \quad (3.17)$$

Where the derived group delay from the linear phase response is neglected $\delta(t)$ is the unit impulse and $\delta_1(t)$ is the first order derivative of the unit impulse. The resulting optical field after passing through the optical filter will be tagged as $e_d(t)$ and is the result of the convolution of $e_{PM}(t)$ and $h_d(t)$.

$$e_d(t) \approx \left\{ \begin{array}{ll} e_{PM}(t) [K(\omega_0 - \omega_l + \Delta\omega) + KY_{PM}f_l(t)] & \text{Left_slope} \\ e_{PM}(t) K\Delta\omega & \text{Center} \\ e_{PM}(t) [K(\omega_2 + \Delta\omega - \omega_0) - KY_{PM}f_l(t)] & \text{Right_slope} \end{array} \right\} \quad (3.18)$$

Where $f'(t)$ is the first-order derivative of the modulating signal labeled as $f(t)$, hence in the case of a Gaussian modulating pulse $f_l(t)$ corresponds to an UWB monocycle waveform. Under these parameters, the photocurrent at the output of the PD is provided by:

$$i_{PD}(t) \sim \left\{ \begin{array}{l} PK^2 \left\{ (\omega_0 - \omega_l + \Delta\omega)^2 + [Y_{PM}f'(t)]^2 \right\}, \\ PK^2 \Delta\omega^2, \\ PK^2 \left\{ (\omega_2 + \Delta\omega - \omega_0)^2 + [Y_{PM}f'(t)]^2 \right\}, \end{array} \right\} \quad (3.19)$$

In 3.19, P is the optical power at the input of the PD. When the optical carrier ω_0 is located at the left or right slope with an assumption of small-signal modulation, the second term is significantly smaller than the third term and can be neglected. In the end, the obtained recovered RF signal results in:

$$recovered_{RF}(t) \sim \left\{ \begin{array}{ll} 2PK^2(\omega_0 - \omega_l + \Delta\omega)Y_{PM}f'(t), & \text{Left_slope} \\ 0, & \text{Center} \\ -2PK^2(\omega_2 + \Delta\omega - \omega_0)Y_{PM}f'(t), & \text{Right_slope} \end{array} \right\} \quad (3.20)$$

From this analysis and by means of equation 3.20, we can state that no signal can be recovered if the optical carrier is located at the center passband of the optical filter, the recovered signal is proportional to the first-order derivative of a Gaussian modulating pulse when the optical carrier is positioned at either slope of the optical filter and finally that the detected pulses hold different signs when the mentioned carriers are located at the opposite slopes, a feature that allows for valuable implementation of modulation possibilities.

3.4 Employment of PM-IM techniques and SOAs in current UWB generation schemes.

To conclude this chapter, a review of some of the mainstream established PM-IM based techniques for UWB systems will be presented. First, we will introduce a fully optical approach [Yao07] exhibited in detail in figure 3.11, here the optical Gaussian “seed” pulse is generated using a femtosecond pulse laser (FSPL), and the phase modulation is implemented based on XPM in a length of nonlinear fiber. To control the pulse width, a tunable bandpass filter is incorporated after the FSPL. The generated optical pulse is then injected with a

continuous wave (CW) probe into a length of dispersion-shifted fiber (DSF) to achieve optical cross-phase modulation (XPM). The phase-modulated signal that is carried by the probe is then converted to an intensity modulated signal at an FBG (serving as a frequency discriminator). The FBG also serves as an optical bandpass filter to remove the residual pump and the amplified spontaneous emission (ASE) noise from the erbium-doped fiber amplifier (EDFA)

Figure 3.11: All-optical experimental UWB pulse generator. In this scheme, The Gaussian pulse is generated by the FSPL with proper spectrum slicing to control the pulse width.

Furthermore, other interesting approach is presented by [Feng14b] and depicted in figure 3.12. This solution stands as a valid UWB signal generation scheme with reconfigurable differential orders based on cascaded reflective semiconductor optical amplifiers (RSOAs). In this scheme, gain saturation effect is controlled by adjusting bias currents of two discrete RSOAs. Thus this cascaded RSOAs structure is converted to a quasi-high-pass RF filter with tunable cut-off frequency. Here, in the electrical domain, Gaussian like pulse train is filtered by a high pass RF filter with cut-off frequency f_c to suppress low frequency part and pass high frequency part. Through changing the cut-off frequency f_c of this filter, various orders Gaussian like pulses can be generated.

Figure 3.12: Experimental setup of reconfigurable orders UWB signals generation Scheme based on cascaded RSOAs.

In this solution, the lowest cut-off frequency (which can be assigned as “ f_1 ”) corresponds to the first order derivative (UWB monocycle). After increasing cut-off frequency to a new value tagged as “ f_2 ” ($f_1 < f_2$) a doublet UWB signal is obtained by second order differential operation. To further increasing cut-off frequency (and upgrading the pulse generation technique towards high order pulses generation) a new “ f_3 ” is added ($f_1 < f_2 < f_3$) and hence a third order derivative (triplet UWB signal) is produced.

As it was analyzed in the previous chapter there is an extensive spectrum of possibilities within the signal generation field. In this section, the feasibility and popularity of the employment of both SOAs and PM-IM conversion schemes was exposed. Another main aspect is a deep comprehension of the theoretical fundamentals behind these two approaches so a maximum advantage of their inherent benefits can be exploited in research solutions. The next chapter is dedicated exclusively to the presentation and explanation of the designed schemes for this thesis.

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CHAPTER 4

Experimental results on UWB generation schemes

In this chapter, several structures for arbitrary waveform generation based on Microwave photonics technology for UWB applications are proposed and experimentally tested. Such schemes rely on the employment of non-linear effects and phase – to- intensity (PM-IM) conversion. In this context, in the case of PM-IM conversion, two techniques were analyzed. In the first one, various low order UWB pulses are produced through a dispersive media and then, these obtained pulses were combined in an optical processing unit (OPU) which allowed reconfiguration of the generated signals. Finally, in the last PM-IM scheme, the conversion process was achieved by means of a customized fiber Bragg grating (FBG) array that operated as a frequency discriminator.

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4.1 Arbitrary waveform generation of UWB pulses based on non-linear effects.

4.1.1 UWB Doublet generation using cross-phase modulation in a semiconductor optical amplifier Mach-Zehnder Interferometer.

Results corresponding to the conceptualized experimental approaches of this thesis will be presented in this section. The first proposal consists on a generation scheme where an UWB doublet is accomplished, by exploiting the nonlinearities inherent to an integrated SOA interferometric structure. Even though, under its initial configuration this set up does not accomplish UWB high order pulses, it succeeds at maximizing the benefits of one wavelength operation. Since, only one optical laser source is employed and a second-order Gaussian derivative is obtained.

In this implementation, the key component is an integrated semiconductor optical amplifier (SOA) Mach-Zehnder interferometer (MZI) which is comprised by two 1-mm InGaAsP-InP SOAs with low polarization sensitivity and it is configured under a counter-propagation regime. The design stands as a new method to generate UWB doublet pulses based on the cross phase modulation (XPM) effect present in the SOA-MZI and its validity is experimentally demonstrated. A modulated optical pulse is employed as a pump signal to convert the input optical signal into a doublet through an integrated SOA-MZI device. The experimental analysis of the transfer function permits to identify the optimum operation point, which leads to the capacity of generating the desired doublet by means of the electrical currents applied to each one of the SOAs that compose the interferometric structure. Figure 4.1 shows a detailed depiction of the referenced interferometric assemblage.

Figure 4.1: Semiconductor optical amplifier Mach-Zehnder interferometer device with counter-propagation configuration.

The aforementioned solution holds a similar approach to previous proposals that are based on the nonlinear response of the transfer function of an Electro-Optical Modulator [Wang06]. Such operational stage can be achieved for currents far from maximum performance zones. When comparing, this system achieves higher extinction ratio and better modulation efficiency, therefore improving the performance of the aspired UWB pulse

generation process. In this technique, generation of UWB doublets is controlled by tuning different values for the current applied to the SOA located in the second lower branch (I_2) and maintaining a fixed value for the other Bulk SOA current (I_1). Working operation currents for this nonlinear device ranges up to 300 mA. Once the relation between the nonlinear response and the current values is found, it is possible to control the polarity of the obtained pulse.

Performance of the system was analyzed through an experimental representation of the transfer function. This experimental characterization was obtained by means of a radio frequency (RF) tone in low modulation regime. The pump signal was modulated with a radio frequency tone of 10 GHz, which at the same time had been generated and analyzed by a lightwave component analyzer (LCA). The representation involved the conversion efficiency factor and the relative phase between the output signal and the pump signal. Both features hold a direct relationship with the amplitude and polarity of the outcome produced by the device.

Figure 4.2: Experimental depiction of the transfer function of the SOA-MZI: Conversion amplitude at the output P_2 (Black Solid line) and the corresponding relative phase between input port P_1 and output port P_2 (red dashed line). Representative operation points are labeled with letters a, b and c.

As it can be observed in figure 4.2, there are several distinguishable operation points. Operation points (a) and (c) correspond to a maximum conversion position that holds a quasi-linear response between the output and input RF modulation. In the case of (a), such operational stage corresponds to the highest working limits within the current range. On the contrary, operation point (b) reveals a conversion minimum to the SOA-MZI transfer function with a quadratic response. In addition, this approach permits to control the polarity of the produced pulses since the experimental transfer function reveals different regions where the relative phase between output and input changes from 0° to 180° . These specific working stages can be perceived in the area constrained between points (a) and (c).

Considering the electromagnetic field analysis performed in [Manz13] and based on the consideration that $P(t) = \langle E(t).E^*(t) \rangle$ permits calculation of the optical power for counter propagation configurations, we obtain an expression to quantify this outcome power, corresponding to ports P_4 and P_5 of the SOA-MZI device.

$$P_{out(P4)}(t) = \frac{P_{probeP4}}{8} \cdot \left[G_1 + G_2 - 2\sqrt{G_1 G_2} \cdot \cos(\phi_{RF}(t) - \Delta\phi) \right] \quad (4.1)$$

$$P_{out(P5)}(t) = \frac{P_{probeP5}}{8} \cdot \left[G_1 + G_2 + 2\sqrt{G_1 G_2} \cdot \cos(\phi_{RF}(t) - \Delta\phi) \right] \quad (4.2)$$

Where G_1 , G_2 stand for the gains and ϕ_1, ϕ_2 stand for phase variation of the semiconductor optical amplifiers that comprised the SOA-MZI. Firstly, we can observe how both outputs are complementary among them as a result of the cross-phase modulation effect. Secondly, we can also perceive that the extinction ratio is determined by gain values G_1 and G_2 , hence this ratio will reach its peak value when both gains are the same.

As it can be appreciated, efficiency of the mentioned conversion relies on pump and probe parameters. Therefore, it is essential a detailed analysis of such pump signal. In principle, both cross-gain modulation (XGM) and cross-phase modulation (XPM) processes could be present in the SOAs. However, the linewidth enhancement factor of the SOAs is large enough to neglect the XGM effect. For this application, the pump signal of the system is obtained by external electro-optical modulation of a continuous wave (CW) optical carrier which will perform as the probe with an optical wavelength that represents the pump. Such pump signal is introduced to the design to induce the XPM in the SOA-MZI. This pump laser source is externally modulated with a quasi-Gaussian pulse train originated at a BERT with repetition rate of 12.5 Gbit/s and amplitude of 0.250 Vpp. The electrical signal that feeds the EOM has a fixed pattern ‘1000 0000 0000 0000’ summing up a total of 64 bits, in this sense the equivalent modulating signal holds a time domain period of 5.12 ns and a full width half maximum (FWHM) of around 80 ps.

The XPM process depends on the average optical power of the pump and probe signals, as well as of the electrical currents applied to each SOA branch and the bias voltage V_{BIAS} injected into the EOM. Consequently, we achieve the generation of an UWB doublet within a specific operation point which is determined by the saturation of the SOAs. The pulse employed as pump is biased in a negative region in order to obtain a negative pulse with

high power levels and therefore guaranteeing saturation of the upper SOA branch. Figure 4.3 plots the electrical pulsed signal that feeds the EOM and the modulated pump pulse.

Figure 4.3: (a) Electrically generated modulating signal and (b) optically modulated pump pulse.

In this context, the initial value for I_2 is set at 300 mA. Such current value corresponds to an operation region in the transfer function with negative slope at the maximum current value. As expected, the output pulse turns out inverted when compared to the pump signal leading to the generation of a quasi- Gaussian pulse. The obtained pulse and its corresponding spectrum can be appreciated in figure 4.4.

Figure 4.4: Cross phase modulated output pulse obtained at P_2 with $I_2=300$ mA in (a) time domain and (b) spectral domain.

Figure 4.5 displays the experimental layout for the UWB doublets generator. In this design, the pump signal is centered at 1535.00 nm (λ_{PUMP}) and introduced by port P_1 , the probe signal is centered at 1550.00 nm (λ_{PROBE}) and goes in by port P_4 . This implies the conversion process is carried out in counter-propagation configuration and hence, optical filtering is not required to separate pump from probe. The average optical input power of the system is directly related to both pump and probe signals. Consequently, optical power value of these parameters and the biasing voltage V_{BIAS} injected into the EOM must be controlled in order to optimize the nonlinear XPM process.

In the case of the electro-optical modulation stage, this task is carried out by an AVANEX EOM device. This equipment introduces insertion losses up to 3.2 dB, offers an extinction ratio ≥ 20 dB and holds a $V\pi$ voltage of 5 V. Optical power values were established at 13 dBm and 5.4 dBm for the pump and probe signals, respectively. In the case of the bias voltage, a fixed value of 1.5 volts was established, all in order to guarantee high input power levels and hence operation within the saturation regime. Finally, at the output port P_2 the optical processing is performed and the optically produced signal is measured by means of a photodetector (PD) in an oscilloscope or and electrical spectrum analyzer (ESA).

Figure 4.5: Experimental UWB generator based on a SOA-MZI.

In the previous scenario, control of the SOAs working currents is a fundamental affair. Operation current I_1 is set into a fixed value of 300 mA while I_2 fluctuates between a range of 0 and 300 mA. If current I_2 is continuously tuned, it is possible to appreciate different created waveforms through the complete working cycle of the SOA-MZI. In order to validate this statement, the value of I_2 is gradually increased in order to observe and measure the resulting waveform and spectrum after the PD through a digital communication analyzer (DCA) and an electrical spectrum analyzer (ESA).

Demonstration of this principle was performed by means of set of measurements plotted in figure 4.6. The most representative stages for this operational scheme appears when current I_2 are located at specific values as shown in figures 4.6 and 4.7

Figure 4.6: Main working stages for different I_2 values, a) 300 mA, b) 285 mA, (c) 270 mA, (d) 255 mA,

Figure 4.7: Main working stages for a different set of I_2 values (a) 225 mA, (b) 206 mA, (c) 176 mA, (d) 150 mA.

After analysis of the obtained results, we can identify two clear stages that best suit the conditions for an UWB doublet pulse. The mentioned operational points can be distinguished for I_2 current bias of 270 mA and 176 mA. When working at 270 mA, the conversion cycle is located within the saturation range of the SOA-MZI which is aligned to the maximum conversion efficiency area, therefore generating the desired doublet waveform. Polarity switch of the created waveforms can be accomplished at 176 mA.

Figure 4.8: (a) Waveform of time domain representation (solid red line) and (b) electrical spectrum of the UWB doublet pulse (solid black line) compare to the FCC settled mask for an electrical current applied to SOA2 of 270 mA. (c) Waveform of time domain representation (solid red line) and (d) electrical spectrum of the UWB doublet pulse for an electrical current applied to SOA2 of 176 mA.

When examining the obtained UWB doublets in terms of the temporal representations plotted in figure 4.8, it is noticeable that there exists a relevant difference in the amplitude depicted by the produced optical signals. Behind this dissimilarity lies the fact that conversion efficiency is lower for a 176 mA than for 270 mA, as it was evidenced in figure 4.2. In terms of spectral representation, it can be perceived that the obtained RF power can be reduced in acceptable levels as for FCC requirements. In these results, we will not come across with a fully FCC compliant pulse, but this is standard for UWB doublets. This first experiment represents an efficient and novel procedure to generate UWB doublet pulses, based on a widely studied nonlinearity as cross-phase modulation. The XPM effect is present in the interferometric structure employed and validity of the design was demonstrated through experimental measurements.

When compared to previous works [Wang10], we witness for this application a better performance in terms of noise features as the MZI structure improves the extinction ratio of the conversion process. Other implementations include employment of an optical carrier phase-shifting method by cascading polarization modulators (PoIMs) [Li13], PM-IM conversion based approaches [Marp11] and birefringence tailored time delays using SOAs

[Chen07]. These approaches produce low derivative UWB waveforms which do not fully comply with the FCC mask requirements and at the same time contemplate complex reconfiguration scenarios, such as manipulation of polarization states.

The scope of this thesis is focused on the optimized generation of high order pulses. Encouraged for the promising results provided by the introduced interferometric structure, the next step in our scope of work was trying to take advantage on the feasible combination of the nonlinearities present in SOAs and standard microwave photonics filtering techniques. In the case of photonics filters some implementations are restricted to positive taps because the coefficients of the optical samples are proportional to the optical power. Nevertheless, the implementation of negative coefficients is beneficial since it provides flexibility in the design of the transfer function of a filter, and therefore offers customization features or into a communications set up. Such flexibility feature was considered in the coming proposal to create a UWB pulses generator.

4.1.2 High order UWB pulses photonic generator based on the cross-phase modulation effect of a SOA-MZI.

Our next design exemplifies a high-order generation scheme. This approach succeeds as a generator that permits to obtain high performance with easily scalable features. We cannot ignore the fact that engaging in the utilization of electro-optical modulators (EOMs) and semiconductor optical amplifiers (SOAs) might increase the complexity of this experimental assembly. However, in current bibliography there are references to photonic integration possibilities with EOMs, SOAs and AWGs on a single chip, revealing enhanced improvement of performance, robustness as well as reduction of size, weight, cost and power consumption. Therefore, this next set up represented a key milestone in the path towards an efficient high-order UWB generator based on the benefits of nonlinearities.

In this experimental set up, a simple and novel design for generation of UWB monocycles with data modulation capabilities is proposed and demonstrated. This application exploits the nonlinearities present in the previously employed SOA-MZI device at the same time that introduces reconfiguration and customization capacities by means of an optical processing unit (OPU). This UWB pulses generator is based on the combination of Gaussian profile positive and negative pulses. The creation of these waveforms is accomplished by means of the sum of various modulated continuous wave probe signals that hold polarity switch among them. Figure 4.8 displays a detailed illustration of the generic integrable high order generator proposed.

Figure 4.9: Experimental layout of a scalable optical generator for high order pulses based on a non-linear interferometric structure and an optical processor unit.

In order to verify the operational principle, a first stage focused on a low-order derivative pulse generation with two laser inputs was developed. Parameters corresponding to the pump signal shall not suffer any modification in comparison to the design of section 4.1. Therefore, this pump pulse is centered at an optical wavelength (λ_{PUMP}) of 1535.04 nm, electro-optically modulated by a Gaussian like pulse train of 80 ps width approximately and holds an optical power of 13 dBm and we are dealing with two continuous probe pulses (labeled as λ_1 and λ_2) centered at 1550.12 and 1552.52 nm with an optical power values of 5.29 dBm and 4.44 dBm, respectively.

Once again, the pump pulse enters the system through port P_1 while λ_1 and λ_2 are introduced into the assemblage through ports P_4 and P_5 . In this order of ideas, the architecture employs again a counter-propagation configuration, thus filtering stage to separate probes signals from the pump is not required. As it was considered in the prior application, current for the upper SOA branch was fixed at a high constant current of 300 mA which determines the conversion speed of the continuous input signal and performance under saturation regime. Conversion efficiency of the gain depends on the current injected to SOA2 with a threshold current around 40 mA. A dynamic range allocated between 150 and 300 mA was employed for the current I_2 .

The main breakthrough achieved in this proposal is its flexibility and reconfiguration possibility by inserting an assembly identified as the “optical processing unit” (OPU). Functionality of this OPU consists on the control of the amplitude and delay parameters of the generated optical taps. The OPU segment is composed of the concatenation of two demultiplexing/multiplexing stages, in this case represented by an array waveguide grating (AWG). For this implementation, the employed AWGs holds 1x40 channels equally spaced by 100 GHz (0.8 nm) according to ITU regulations and with an optical bandwidth of 0.4 nm. All generated signals coming out from the port P_2 of the SOA-MZI are introduced into another AWG stage, which once again plays the role of a demultiplexing device. Hence, every signal pulse carried at a different optical wavelength is treated independently. The

amplitude and delay features are customized by means of a set of tunable attenuators and variable delay lines (VDLs) that introduce a differential delay of $\Delta\tau$. Finally, the output signal is obtained by combining all processed optical pulses through a final multiplexing stage.

Once it is distinctly manifested the capacity to produce both positive and negative optical taps, we proceeded to measure the time-domain base obtained waveforms. Such signals, which are measured before the reach the OPU reconfiguration stage, can be observed in Figure 4.9. They correspond to two optical taps of temporal Gaussian like profile with inverse polarities among them and a time separation around 440 ps. In this case scenario, the first maximum of the spectrum is close to the GPS band (0.9-1.61 GHz) since it depicts a free spectral range (FSR) of approximately 2.2 GHz ($\text{FSR} \approx 1/\tau$).

Figure 4.10: (a) Generated pulses in their original position, before any kind of delay configuration, (b) corresponding spectral representation

The relevance and necessity of the OPU relies on the fact that enables the feasibility to reshape the output pulse and hence provide as the desired low-order UWB pulse (a monocycle for the employment of two laser sources). The separation between the positive and negative pulse is reduced to an optimized value close to 60 ps as shown in Figure 4.11(a). When analyzing the corresponding spectrum of the generated pulse, which is depicted in Figure 4.11 (b), we perceive that this waveform reveals a better adjustment in terms of fitting the FCC spectral mask parameters. In this case, the FSR is 16.7 GHz approximately which is compatible with the UWB mask. Even though monocycles do not fully satisfy successfully the FCC spectral mask restriction as predicted theoretically, they provide remarkable advantages in terms of achievability, simplicity and integration.

Figure 4.11: (a) Obtained monocycle waveform after final delay configuration and (b) spectrum for the corresponding UWB monocycle pulse.

Versatility of the proposal extends to the implementation of data transmission under certain modulation formats. In our first test, we transmitted data with Pulse Amplitude Modulation (PAM) format by modifying the bit sequence of the BERT generated component of the pump signal. Figure 4.12 plots the transmission of three different data strings as well as their corresponding spectra. Firstly, a data scheme composed by a bit sequence of all ones “1” was transmitted as plotted in figure 4.12(a). A second data series comprised by both alternated ones “1” and zeros “0” bits was also sent as shown in figure 4.12(b). The third data sequence to be transmitted was constituted with “0” and “1” distributed randomly as plotted and observed in figure 4.12(c). Moreover, figures 4.12(b), (d), and (f) depict the corresponding spectra.

Alternatively, PAM can be implemented by means of variable optical attenuators that modify the amplitude of each generated optical pulse. Pulse Position Modulation (PPM) can be also achieved by controlling the relative position within the time domain of the processed optical signals through the variable delay lines that compose the OPU. Finally, Bi-Phase Modulation (BPM) can be also performed by adding an optical switch prior to the conversion stage (SOA-MZI). We could modify the polarity of the obtained signals and therefore achieve two complementary monocycle pulses.

On the other hand, this solution is not only efficient and flexible but also scalable, offering the opportunity to generate high order pulses by simply introducing additional optical taps. In this context, data modulation could also be plausible by means of precise control of the delay and amplitude parameters through the OPU. In order to demonstrate such concept, our next experimental approach embraced the activation of four CWs lasers to perform the before explained optical conversion.

Figure 4.12 (a) First data scheme transmitted with a bit sequence of all ones “1” and (b) corresponding spectral representation. (c) Second data series composed of both ones “1” and zeros “0” bits and (d) respective spectrum. (e) Third data sequence transmitted and (f) its electrical spectrum.

The sources were centered at four different wavelengths λ_1 , λ_2 , λ_3 and λ_4 at 1550.12 nm, 1550.92 nm, 1551.72 nm and 1552.52 nm, respectively. Wavelengths λ_1 and λ_2 are launched into Port 4 of the interferometric device and wavelengths λ_3 and λ_4 into port P5. Optical power delivered by the probe continuous wave sources was set at 5 dBm for λ_1 and λ_2 and 8 dBm for λ_3 and λ_4 . As it has been established the principle of the proposed system is based on the use of XPM in the SOA-MZI structure, which permits to obtain pulses with positive and negative polarity according to the input port to which the optical source is applied.

Indeed, the sign of the optical tap is set by selecting the input Ports 4 or 5, resulting complementary outputs correspond to port P2.

Since the operation principle remains unaltered, then just as in the predecessor's implementations, the process of cross-phase modulation (XPM) in the interferometric structure depends on the average optical power of the pump and probe signals and on the electrical currents applied to each SOA. Therefore, the optical power of the pump and probe signals must be controlled in order to optimize the conversion process. Another parameter to take into consideration is the electrical current applied to each SOA since it determines the XPM efficiency. The rest of standard working parameters will not be modified, so SOA1 and SOA2 in the upper and lower branch respectively have a high constant current of 300 mA which determines the conversion speed of the continuous probe signal and current. In this scheme, the limiting factor could be related to the XPM process which permits conversion up to 10 Gb/s. The electro-optical modulator and photodiode have a response of 40 Gb/s and 50 GHz, respectively. Consequently, our system is fast enough for UWB applications which feature bit rates ranging from hundreds of Mb/s to a several Gb/s.

To demonstrate the flexible operation as an UWB generator of the design, different signals were produced and analyzed. The main objective in each one of them was to determine the proper matching between their spectrum and the FCC settled specifications. The proposed set up comprised four laser sources, whose activation is directly related to the number of required of taps in the generated signal. Power level at the output port P2 is approximately 4 dBm and the optical loss in the optical processor is close to 9 dB. Since a monocycle has already been implemented by means of two laser sources. Measurements were focused on producing higher order waveforms. Firstly, a doublet was produced using three optical lasers (λ_1 , λ_2 and λ_3). As in the monocycle generation configuration case, λ_1 was launched into port P4, λ_3 into port P5 and the incorporated source λ_2 was introduced through port P4. By adjusting the amplitude and delay of the introduced sources by means of the optical processing unit, it was possible to obtain a doublet with the following coefficients [-0.5, 1, -0.5]. Temporal and frequency responses of the doublet pulse are displayed in Figures 4.13(a) and 4.13 (b).

Figure 4.13: (a) Generated doublet (solid black line), simulated time domain representation is plotted with a red line and (b) Corresponding spectrum (black line). FCC mask illustrated with a red line. Theoretical spectral representation is plotted with a blue line.

As predicted for UWB doublets, the experimental spectrum measurements do not fit properly the FCC mask established requirements. But, when compared to the previously obtained monocycle a clear improvement can be observed when passing from 2 to 3 coefficients. This proves that addition of suitably designed coefficients can lead to a better fit within the FCC mask.

The system is flexible enough to allow for the addition of more coefficients. For instance, by activating an additional laser source the system can generate a third-order UWB pulse. In this scenario, the previously employed inputs (λ_1 , λ_2 and λ_3) maintained the already specified connections and λ_4 was introduced by port P5. By means of the optical processing unit, we adjusted the amplitude and delay of the additional channel to obtain a combined waveform with coefficients $[-0.3, 1, -1, 0.3]$. For this UWB triplet, the odd optical pulses have negative polarity while the even ones have a positive one. Figure 4.14 (a) plots the corresponding UWB pulse in the time domain and Fig. 4.14 (b) exhibits the spectral representation, which, in this case matched the FCC mask requirements more properly than previously generated signals. In our case, the coefficients are determined quantitatively by means of a process of synthesis. Also, the system offers more flexibility than plotted since the optical processor could permit the control of 40 coefficients (1x40 WDM) but we demonstrated the system with 4 taps. In this case, the limiting factor is related to the laboratory facilities.

Figure 4.14: (a) Generated triplet (solid black line), simulated time domain representation is plotted with a red line and (b) Corresponding spectrum (black line). FCC mask illustrated with a red line. Theoretical spectral representation is plotted with a blue line.

These results reveal that our proposed scheme features a high degree of flexibility and reconfigurability. Implementation of different modulation formats employed in UWB communication systems such as PPM and BPM is feasible. In the case of BPM, it can be implemented by changing the input port of the each optical tap for a given UWB pulse implementation.

On the other hand, PPM can be achieved by customizing the signals amplitudes and delays according to our requirements through the optical processing unit. Figure 4.15 displays the measured results for generated UWB-PPM and UWB-BPM modulation signals using the monocyte, doublet and 4-tap pulse configurations.

Figure 4.15: PPM and BPM modulation for the generated UWB pulses. For PPM, original pulse (black line) and delayed pulses (red line) for (a) monocycle, (b) doublet and (c) four coefficients pulse. For BPM, original pulses (black line) and inverted pulses (red line) for (d) monocycle, (e) doublet and (f) four-coefficients pulses.

4.2 Arbitrary waveform generation of UWB pulses based on phase-to-intensity (PM-IM) conversion.

In the previous section various solutions based on SOA non-linearities were proposed and demonstrated. These approaches even though novel and effective, require for each Gaussian base pulse created one coefficient inserted to the system. In order to provide system optimization, we expose the necessity of working with flexible proposals that permit efficient exploitation of the wavelengths involved. This principle represented the starting point for our next segment of experiments, which were focused on merging various low order UWB pulses, such as monocycles and doublets to create high-order ones. To fully satisfy the FCC requirements complex and sometimes costly structures are required, these set of submitted solutions represent simple, scalable and flexible approaches that can definitely establish a difference within the UWB signals generation field.

4.2.1 UWB Scalable UWB photonic generator based on the combination of doublet pulses:

The primal premise of this experiment consists of the possibility of increasing a Gaussian derivative order when summing up two UWB doublets with inverted polarities. Therefore, the fitting features are upgrading in terms of FCC standards, which results in more versatile and specific applications in the UWB scope of work. Figure 4.16 exposes the validity of such statement, by depicting the lineal sum of two low order UWB waveforms such as doublets, at the same time that permits to realize simplicity of the concept

Figure 4.16: (a) Theoretical simulation of controlled generated UWB doublets and (b) the customized sum of these waveforms to obtain a quadruplet.

One key aspect in this approach is the employed method to generate to mentioned base pulses. In relation to this, within standard optical UWB pulses generation techniques, PM-IM conversion stands out for its effectiveness, versatility and simplicity. In [Zeng06], PM-

IM conversion is accomplished by means of the junction of an electro-optic phase modulator (EOPM), a phase modulated optical carrier and a length of single mode fiber (SMF). Thanks to the SMF-induced chromatic dispersion, which is usually considered a negative effect in traditional intensity-modulation and direct-detection (IM-DD) systems, the combination of the EOPM and the SMF link forms an all-optical microwave band-pass filter that can be designed to shape the input Gaussian pulses into UWB pulses.

Selection of a suitable dispersive element is not a trivial issue, since the quality of the spectrum corresponding to the UWB doublet generated determines directly the high-order pulse to be produced. For instance, Figure 4.17 depicts experimentally generated and simulated UWB doublets with their corresponding theoretical obtained by employing various SMF links as a dispersive element. The dispersion into consideration ranged between 170 ps/nm (10km SMF fiber link) and 425 ps/nm (25 km SMF fiber link). Apart from the band pass filtering effect of the phase-to-intensity conversion as predicted, it is shown that a notch filter is achieved at a frequency that depends on the amount of dispersion for a given Gaussian pulse driving the PM.

Figure 4.17: (a) Simulated doublets with various lengths dispersive elements and (b) their respective theoretical spectral representations. Green lines correspond to a dispersion equivalent to 10 km SMF link, blue lines correspond to 20 km and black lines correspond to 25 km. FCC mask is plotted in red line.

For further validation of the performance of the proposed configuration, we first need to determine this optimal value for dispersion. After analyzing the theoretical and experimental measurements previously exposed, the most suitable scenario when comparing with the FCC mask was found at an accumulated dispersion around 340 ps/nm. Henceforth, this experimental implementation considers a 20 km SMF fiber length as a dispersive element taking into account that the SMF-28 fiber has a chromatic dispersion of approximately 17 ps/nm.km at the operating window (1550 nm). The mask-filling efficiency η of any particular signal can be defined as the ratio of the power contained in the resulting FCC-compliant power spectral density $S_{\text{EIRP}}(f)$ to the total power that could be contained under

the FCC imposed mask $M(f)$, both being evaluated in the range confined between 3.1 and 10.6 GHz. Hence it is calculated by:

$$\eta \triangleq \frac{a_c \int_{F_{min}}^{F_{max}} |S(f)|^2 df}{\int_{F_{min}}^{F_{max}} M(f) df} \quad (4.3)$$

On the other hand, based on the 2002 FCC definition, another key feature is represented by the fractional bandwidth, in this sense it is remarked that any signal having -10 dB larger than 20% is considered as UWB. Such parameter is determined by the lower and the higher -10 dB cutoff frequencies, therefore it is established by the ratio among them: $B = 2(f_H - f_L / f_H + f_L)$.

Figure 4.18 plots the theoretical and experimental representations in both time and spectral domain, for the selected doublet pulse. The pulse holds a full-width at half-maximum (FWHM) of about 45 ps. First of all, it is remarkable that the doublet pulse has a notch around the GPS band that fully complies with the FCC spectral mask regulation. This generated doublet achieves a fractional bandwidth of about 133 % and a spectral efficiency of about 23 %.

Figure 4.18: (a) Time domain representation of obtained doublet with a 20km SMF length and (b) corresponds to spectral representation. Black lines represent experimental results and blue lines correspond to theoretical prediction. In addition, FCC mask is plotted in red line.

So, this work comprises a novel proposal for a UWB high order generator which operation principle consists on the combination of different doublets generated by means of phase-modulation and a dispersive element. An optical processing unit (OPU) is added just like in previous schemes, in order to provide the reconfigurability feature. Such unit is composed by a set of AWGs, attenuators and variable delay lines (VDLs) that permit to increase the order of the generated pulse. The waveform shaping is achieved by means of the independent

control of the amplitude and time delay for each doublet. In this solution, in order to modify the polarity of the outcome pulse, a differential detection stage was added. Figure 4.19 plots in detail the implemented scheme to generate high-order UWB pulses.

Figure 4.19: Scalable scheme for generating high-order UWB pulses.

In the first segment of the proposed designed, light from a set of N optical sources is launched to a phase modulator (PM) by means an array waveguide grating (AWG). The PM is modulated by Gaussian pulses coming from an electrical pulse generator providing an OOK data sequence with a fixed pattern of one “1” and sixty three “0” (total 64 bits) to be transmitted at a 12.5 Gb/s bit rate. The PM has an insertion loss of 3.5 dB and a 3dB bandwidth of 20 GHz. Then, the phase-modulated optical signal is applied to a dispersive element that provides the required phase-to-intensity conversion. Indeed, the combination of PM and dispersive element provides a doublet pulse by means of its equivalent microwave photonic bandpass filter, for each single-wavelength laser.

After the phase-to-intensity conversion process, the set of optical taps are launched into the optical processor unit (OPU) to reconfigure the high order pulses. At the OPU, each UWB doublet pulse coming from one single-wavelength source is treated independently and is distinguishable from other doublets by means of their corresponding optical wavelength using a second AWG. The OPU allows to customize the amplitude and delay parameters for each doublet separately. Therefore, by means of a demultiplexing structure combined with a set of attenuators and variable delay lines (VDLs), the signal outputs are multiplexed into two branches that feed a balanced detector that determines the positive or negative polarity of the taps. Note that an Erbium Doped Fiber Amplifier (EDFA) is placed before OPU to compensate optical losses .

The use of the balanced photodetector has a significant importance in the process, beyond being a simple optical-electrical convertor since, to gain a derivative order, the low order pulses to be reconfigured and summed up need to be previously inverted in polarity. In this way, our proposal consists on an equivalent N-tap microwave photonic filter with positive and negative coefficients with reconfigurable amplitude and optical delays at the OPU. Each optical tap is obtained by the combination of phase modulation and a dispersive element. Scalability of the system depends on the availability of multiple WDM channels. In our case, the number of optical sources is reduced as compared to those schemes that implement each optical tap with an independent Gaussian pulse [Bole09, Feng14]. Since, the main objective in this experimental setup is to determine the proper matching between the generated spectrum and the specifications settled by the FCC, a FCC mask compliant UWB triplet is generated.

The operation principle of the design relies on the combination of two second-order derivatives of Gaussian pulses separated in time around 130 ps and generated by means of two laser sources, labeled with the emission wavelength λ_1 and λ_2 . Optical sources λ_1 and λ_2 are centered at 1550.12 nm and 1550.92 nm, respectively, with an optical power of 10.50 dBm each. As it was already established, the base low-order doublets are accomplished through PM-IM conversion and the generated signals are introduced into the reconfiguration stage (OPU) by means of an AWG. In this sense, each doublet is customized in terms of amplitude and delay through a set of attenuators and variable delay lines (VDLs), in order to optimize the output signal.

For this implementation, one of the VDLs introduces a delay of 142 ps between both optical taps in order to compensate the chromatic dispersion introduced by the fiber link (272 ps) due to their wavelength separation. Finally, because of the differential detection provided by the balanced photodetector, the pulse introduced through the upper arm is combined with its inverted version produced in the lower arm, leading to the expected higher order UWB waveform.

Figure 4.20(a) exhibits the obtained UWB triplet waveform, where the blue line corresponds to the theoretical prediction and Figure 4.20(b) plots its corresponding spectral representation. This UWB triplet depicts a fractional bandwidth of 116 % along with a spectral efficiency with a value as high as 60 %. As expected, a perceptible increase of the spectral efficiency occurs when comparing the previously obtained doublet in Figure 4.18, which shows a spectral efficiency around 23 %.

Figure 4.20: (a) Generated UWB triplet with its time domain representation and (b) corresponding spectrum. FCC mask is plotted in red line.

Since scalability is a key feature within this proposal, demonstration of this principle was mandatory within the experimental trials performed for this set up. For generation of an UWB quadruplet pulse, the same operational principle corresponding to the triplet was employed. In this case, the combination of three doublet pulses separated in time enabled the creation of the aspired waveform. These low-order derivative base pulses were generated by means of three optical sources, labeled as λ_1 , λ_2 and λ_3 . Optical sources λ_1 and λ_2 are centered at the same wavelengths as before and λ_3 at 1551.72 nm, respectively. Figure 4.21(a) exhibits the obtained UWB quadruplet pulse, where the blue line corresponds to the theoretical prediction and Figure 4.21(b) plots its corresponding spectral representation.

Figure 4.21: (a) Generated UWB quadruplet with its time domain representation and (b) corresponding spectrum. FCC mask is plotted in red line.

Although this previous approach is competent compared to aforementioned schemes which only employ Gaussian based pulses, it can be also noted that presents a bulky structure by the presence of the 20 km SMF link and, hence, limited integration feature. Since, accomplishing an integrated and compact design is one of the main concerns behind all these experimental schemes, the search for an alternative structure within this technique was essential. In [Zeng06], an optical carrier is phase modulated by a Gaussian pulse via an EOPM and the desired PM-IM conversion is achieved by use of a fiber Bragg grating (FBG)

that serves as a frequency discriminator. In this set up, by locating the optical carrier at the linear or the quadrature slope of the FBG reflection spectrum, monocycle or doublet pulses can be obtained at the output of a photodetector (PD). In addition, UWB pulses with opposite polarities can be generated by locating the optical carrier in the right or left side of the FBG reflection spectrum. This property is highly significant, because it is essential to generate high order pulses by means of the sum of low base polarity shifted ones. Besides, the intrinsic capacity of the system for this polarity inversion, allows to remove the balanced detection stage, which simplifies the structure.

4.2.2 UWB Scalable UWB photonic generator based on the combination of doublet pulses.

In this work, a highly flexible photonic architecture has been proposed and experimentally demonstrated to generate UWB pulses based on a Fiber Bragg Gratings (FBGs) superstructure. The principle of operation relies on the combination of a phase modulator with a customized FBG array for a simultaneous process of PM-IM conversion and pulse multiplexing. Figure 4.22 illustrates a scheme of the proposed high-order pulses generator with modulation capabilities by employing a FBG-based superstructure.

In the experimental setup, the optical signal from a set of 4 optical sources is launched to a phase modulator (PM) by means of a 4x1 optical coupler. The PM is modulated by a Gaussian pulse sequence with a fixed pattern of one “1” and sixty three “0” (total 64 bits) at a 12.5 Gb/s bit rate, just like previous applications. The PM device has an insertion loss of 3.5 dB and a 3dB-bandwidth of 20 GHz. Once modulated, the obtained pulse is amplified with an Erbium Doped Fiber Amplifier (EDFA) and introduced to the FBG superstructure through an optical circulator which will perform the desired PM-IM conversion in function of the optical wavelength selected for each optical source. Inset (a) of Figure 4.22 plots a diagram of the designed FBG superstructure.

The FBG structure consists on an array constituted by 4 FBGs designed with an apodized profile considering a precise separation d and length L in order to maximize the PM-IM conversion with an exact optical delay between adjacent optical taps. For this implementation, the PM-IM conversion process relies on the employment of a fiber Bragg grating (FBG) which operates as a frequency discriminator.

Figure 4.22: Experimental setup for the generation of high-order pulses generation FBGs based on a FBG photonic superstructure.

The design of a suitable FBG array is a significant matter since the quality of the reflected spectrum and the distance between each one will determine the quality of the high-order pulse to be produced. Therefore, the key improvement revealed by this work is represented by the FBGs array employed in comparison to the previous proposal. As it is pointed out earlier, accomplishing an integrated and compact design is one of the main concerns behind this proposal. The key improvement revealed by this work is represented by the FBGs array superstructure employed. As it is well known, FBGs are optical fiber devices that selectively reflect only those light signals with the wavelength matched to the grating period of refractive index modulation. As it was pointed out earlier, accomplishing an integrated and compact design was one of the main concerns behind this scheme. Therefore, using a customized FBGs structure plays a fundamental role in the proposal of size limited architectures.

Design of a suitable array is a significant matter since the quality of the reflected spectrum and the distance between each one of the generated pulses (monocycles or doublets according to the operation point) will determine the quality of the high-order pulse to be generated. As shown in figure 4.23, the photonic structure consists on an array constituted by 4 FBGs designed with a specific apodized profile considering a specific separation d and length L in order to maximize the PM-IM conversion with a specific optical delay between adjacent optical taps.

Figure 4.23: Principle of operation for a Fiber Bragg Grating Array.

First of all, when designing the array, we had to consider a specific apodized profile to reduce the secondary lobes of the FBG spectrum in order to minimize the crosstalk between channels. The FBG inscription has been performed employing an Argon ion frequency-doubled laser at a wavelength of 244 nm. The inscription procedure is based on the use of a uniform phase mask combined with a precise relative movement between the phase mask and the optical fiber [Cole95]. Such technique permits the apodization of the refractive index modulation along the optical fiber in order to tune the FBGs optical spectra. The use of a boron co-doped optical fiber increases the photosensitivity and a Kaiser apodization is employed in these FBGs in order to obtain the desired spectrum. The reflectivity of the FBGs obtained through this process varies between 82 % and 90 %.

Secondly, the separation d is also an important factor to consider when fabricating the superstructure, this parameter holds a direct relation with the optical delay between taps ($\Delta\tau = 2n_0d/c$). Where n_0 is the effective index, c the light velocity at vacuum and d the separation factor we are referring to. In order to be working with a FSR/2 around 6.45 GHz, in agreement with the central frequency of the UWB spectral band we require a FSR ($1/\tau$) of approximately 12.9 GHz and an optical delay ($\Delta\tau$) of 77ps. In this context, d must hold a value approximate to 8mm. Finally, we evaluate the FBG shaper in function of the length. For instance, figure 4.24 portraits each one of the power reflected spectrums of various FBGs sets that were manufactured in our lab facilities specifically for this experiment. FBGs of 6, 7 and 8 mm length were fabricated and tested. Apart from verifying the phase-to-intensity conversion stated as represented in Figure 4.25, we corroborate the accuracy of the spectral distance and shape between the generated waveforms as well as the absence of undesired pulse broadening. In this context, the length of the FBG is a flexible parameter, not being this the case for the separation between the FBGS, which as we will corroborate later represents a determining factor for the suitable generation of high-order UWB pulses.

Figure 4.24: Comparison for variations in the length parameter in the FBG array design. Reflectivity profile for 6, 7 and 8 mm gratings.

Figure 4.24: (b) produced UWB monocycle for the afore mentioned lengths and (c) theoretical spectral representation for the generated pulses.

To demonstrate and validate the capabilities for the generation and low-order pulses modulation capacities of the proposed scheme, different signals were produced and analysed. Firstly, we examined the lower-order waveforms by means of the different FBGs that compose the array with the aim of obtaining the exact wavelength values in which they are generated.

Figure 4.25: (Upper) Power reflection spectrum of the selected array and (Lower) experimental measurements (black lines) and theoretical simulations (red lines) corresponding to the four obtained waveforms (a) monocycle, (b) doublet, (c) inverted doublet and (d) inverted monocycle.

One of the main properties featured by the superstructure is its capacity of generating UWB high order pulses by combining several low-order ones. In spite of the fact that monocycle and doublet pulses are widely used due to their simplicity, high-order derivatives of Gaussian pulses are more suitable candidates for UWB systems in terms of spectral efficiency and FCC mask compliance. To achieve this measurement, the proposed set up employed two lasers with an optical power value of 5.5 dBm for each one. Activation of such optical sources is proportional to the number of lower order pulses to be reconfigured in the outcome signal. Wavelengths employed for these optical devices were 1549.852 nm and 1550.702 nm. These values are directly related with the base UWB waveforms to be employed and hence the PM-IM conversion working point. For instance, in this case a triplet was accomplished by means of the lineal sum of two inverted doublet pulses, corresponding to the “C” and “B” operation points. This UWB triplet depicts a fractional bandwidth of 99.8 % with a spectral efficiency of 65 % in the 3.1-10.6 GHz band. The temporal and frequency response of the obtained triplet pulse are displayed in Figures 4.26 (a) and 4.26 (b), respectively.

Figure 4.26: (a) Experimental (black) and theoretical (red) waveforms for the accomplished triplet pulse and (b) Corresponding spectrum for the generated triplet (black line). FCC mask illustrated with a red line.

The second main capability of the proposed system is its modulation features. First, BPM was achieved by the juxtaposition of two inverted pulses. This was accomplished by means of one FBG and two optical sources. The FBG1 used as the frequency discriminator for this measurement had a central wavelength of 1549.808 nm. Figure 4.27 (a) shows the two low-order UWB pulses obtained for BPM by switching the optical wavelength between $\lambda_{1A} = 1549.721$ nm (red line) and $\lambda_{1D} = 1549.871$ (black line) nm. The first one (red line) corresponds to the “A” working point, which produces a positive monocycle and the inverted pulse is related to the “D” working point and thus to a negative monocycle (black line). In order to appreciate the scalability and versatility of the superstructure, the rest of the FBGs of the array were also taken into consideration. In this case, PPM modulation can be achieved by employing the entire FBGs array and selecting the proper wavelength for each FBG. The wavelengths can be tagged as λ_{1A} , λ_{2A} , λ_{3A} and λ_{4A} with corresponding values of 1549.721, 1550.631, 1551.568 and 1552.561 nm. Figure 4.27 (b) depicts the generated pulses for PPM modulation, in which there is a separation between the original pulse and the delayed ones of approximately 77 ps, 154 ps and 231 ps.

Figure 4.27: BPM and PPM modulation for the generated UWB pulses. (a) For BPM, original pulse (red line) and inverted pulse (black line) and (b) for PPM with a separation of 77 ps.

In this context, multilevel modulation was also verified by establishing the positive monocycle as the state “0” and the negative monocycle as the state “1”. In this implementation, two optical sources and two FBGs of the array were required. Figure 4.28 depicts a multilevel modulation format comprised by four states that could be associated to levels 00, 01, 10 and 11.

Figure 4.28: Multilevel modulation for the generated UWB pulses associated to (a) level 00 obtained with two positive monocycles, levels (b) 01 and (c) 10 obtained with two pulses with inverted polarity and (d) level 11 with two negative monocycles.

In order to maximize the functionalities of the system, along with high-order waveforms generation and binary format modulation, other modulation schemes were demonstrated through the presented design. Since the sum of different types of lower order pulses to obtain higher order ones can be exploited, a four level pulse shape modulation (PSM) scheme can take place. In this way, depending on the selected conversion point of the PM-IM cycle we can control the obtained pulse in the time domain. Therefore, the opportunity of generating high order pulses permitted us to accomplish four different waveforms for an UWB PSM format. Performance of this concept was carried out experimentally and its corresponding results are depicted in Figure 4.29. In this sense, Figure 4.29(a) and Figure 4.29(b) reveal lower-order signals (monocycle and doublet, respectively) whereas Figure 4.29(c) and Figure 4.29(d) show the maximum system capability to generate higher-order pulse such as a triplet and a quadruplet, respectively.

Figure 4.29: Four levels of pulse shape modulation: (a) monocycle, (b) doublet, (c) triplet and (d) a quadruplet. The pulses can be related to levels 00, 01, 01 and 11.

Based on the exhibited results is accurate to state that by means of this architecture it is possible to obtain UWB triplet and quadruplet pulses compatible with FCC standard purposes by using various single-wavelength lasers and a PM, which do not require any control as previous reports [Abra11]. Finally, it is interesting to remark the potentiality of the system to be integrated on a single chip by means of the available technology [Marp13]. Specifically, the building blocks which can be implemented with current technology are the PM-IM convertor [Zhua13] and the OPU including the differential detection stage [Sale10, Nish12]. Reference [Zhua13] proposes and demonstrates an approach for a PM-IM convertor with performance improvement of phase-modulated microwave links based on ring resonator. On the other hand, for the implementation of delay lines of the OPU one could choose the compact approach based on dispersive photonic crystal waveguide reported in [Sale10]. Moreover, a novel and interesting solution is proposed by [Nish14] which work involves the conceptualization of an AWG made of Si-rich silica (SiOx) waveguides.

Finally, it can be concluded that we have proposed and demonstrated several efficient flexible high-order UWB pulse generation system by means of nonlinear effects and PM-IM conversion. Along all the exploited techniques. The key point was to generate high-order pulses by taking advantage of the benefits provided by active and passive approaches and at the same time simplifying size structures. In terms of FCC mask compliance; there is a noticeable upgrade when comparing a low order derivative pulse with a higher order one,

spectral efficiencies experience significant increase meanwhile fractional bandwidths appeared to be reduced. These entire set of solutions portrait an interesting flexibility for adaptation to optical modulation formats such as BPM, PPM and PSM.

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CHAPTER 5

Conclusions and Future works

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5.1 Conclusions

This research establishes from early on not only the importance and advantages of Ultra wideband (UWB) schemes but also the challenges this technology has faced from a long time. On one side, it possesses the wonderful capacity of employing a significative large amount of spectrum that enables it to coexist with other wireless oriented technologies, but the standard low power levels settled by the FCC limits its coverage area range. To solve this predicament, Radio-over-fiber (RoF) techniques emerged and have been subject of intense research for some time and they hold applications in a myriad of areas such as: in optical signal processing (photonic analogue-to-digital conversion, photonic, microwave filters, arbitrary waveform generation), antenna array beamforming, imaging and spectroscopy, among many others. As it can be deduced, the particular drawback of lack of a vast coverage represented the key factor that led to the MWP-UWB working alliance.

Traditionally speaking, signal distribution by means of fiber has been considered the most successful application of RoF technology. These schemes are known as “Hybrid fiber-radio networks” have been extensively deployed in the last decade due to the increasing demand for higher data rates in communication services in the field of optical access networks. Among the present limitations within fully optical RoF systems, one can mention: the available modulation techniques, the optical sources costs and the corresponding input pulses employed for such layouts.

To comprehend the current importance of Microwave Photonics (MWP) and its role in the current photonics scenario, It is vital to understand its past and origins. As it has been already affirmed generation within the electrical domain has played a key role in the last decades with MWP techniques setting a noticeable improvement by providing several key breakthroughs in terms of configurability and tunability. When merging these two technological niches, we are talking about electrical signal generation by means of optical set ups. In this work, an extensive research of the state of the art has permitted to understand the basis of photonic schemes focused on signal generation. Parallel to this, acknowledging its evolution along the years and identifying the most promising approaches was what lead us to set up the fundamentals for our own designs.

The main objective of this thesis has been the experimental implementation of arbitrary waveform generation techniques within ultra-wideband (UWB) schemes. In this way, several methods and structures haven been approached and all of them have proven to be successful in producing pulses that are compliant with the standards settled by the Federal Communications Commission (FCC), such accomplishments also hold significant upgrades when compared to previous proposals since besides fulfilling the before mentioned regulations, the proposed solutions reveal valuable features such as flexibility, scalability and integration capabilities.

Throughout this dissertation a detailed description of the theoretical basis of the proposed systems has been presented and analyzed. In this way, the inherent non-linear effects of an SOA-MZI and Phase modulation-to-Intensity modulation (PM-IM) stand as the base for the carried out experiments. This thesis has been specifically focused on the generation schemes for higher order pulses, since as it was mentioned before such waveforms are the best option when searching FCC standards compliance. In this sense, the settled objectives were accomplished through the following premises:

- To know and understand in detail the background behind the rise of UWB fundamentals as a key enabling technology: It is not secret that although UWB's potential is as valid as it ever was, nowadays there is a trend that considers this technology a "has been niche" within the wireless field. It would be very ambiguous to determine a single theory or specific insight to back up this statement, but instead it can be considered the result of the combination (in hard-to-figure proportions) of several factors such as:
 - In many cases, it does not seem to be clear what actual market need UWB solutions should be serving. Regarding this fact, many of the generated enthusiasm for UWB technology was originally based on commercial self-interests rather than meeting genuine end-user requirements.
 - As it seems to happen with many state of the art developments, UWB technology was overhyped by its proponents and yet under-delivered. This could be explained by the fact that the design of a fully solid and reliable protocol for UWB with all the operational conditions that it implies is accompanied by significant costs that might be higher than those the market could accept.

In these order of ideas, although the challenges that imply to push forward this niche are clear, it is also evident that most of them fit into a more commercial category, hence technologically speaking the legitimacy of UWB can be considered solid and well-grounded. In this case, as it has occurred with previous breakthroughs, it might be possible that the technology was ahead of its time.

- To analyze the current UWB regulation and its inherent limitations in terms of certain operational features: Ultra wideband (UWB) standardization has come a long way, and although it has succeeded in developing different regulations according to geographical locations, it has failed in unifying these regulatory bodies. One of the identified factors responsible for this "pending task" was the intense involvement displayed by commercial entities from the beginning to promote and facilitate more "convenient standards" for their particular purposes. Regarding this matter, an appropriate way to tackle this still current problem would be to re-start the UWB forums scheme but with a majority input coming from independent research and development entities.
- To scrutinize UWB's original main competitor and perform a comparison analysis between both technologies: In chapter two, a detailed report of coexistence between UWB and one of his strongest competitors (WiMAX) is provided. Such analysis is quite significant since WiMAX and UWB not only share spectral resources and a common applications' target but they struggle to reach a very similar client

target. It is appealing to notice how their coexistence is feasible and balanced, opening a window of opportunity for more diverse indoor hybrid architectures that could be connected through optical media to central offices.

- To summarize the most relevant optical UWB generation techniques, acknowledging their strengths and potentiality and comparing their advantages and disadvantages: UWB low order pulse generation can be accomplished by solutions based on first- or second-order differentiation of a Gaussian pulse. As it was explained in chapter two, this can be achieved by means of a myriad of techniques. One very powerful and exploited technique involves carrying out phase-modulation-to-intensity-modulation (PM-IM) conversion employing an optical frequency discriminator. A method that can be carried out using a fixed-long-length dispersive fiber or a frequency discriminator that acts as an optical filter, such as a fiber Bragg grating.

Combining delayed optical Gaussian pulses with opposite polarities is also a concept approached within this dissertation. This method implicates to merge complementary low order UWB signals to create higher order ones. Chapter three details the fundamentals of this method and chapter four depicts our experimental results when implementing this method.

Exploitation of the nonlinear effects also stands as an engaging approach for UWB pulse generation. In this way, Optical phase modulation can be implemented employing electro-optic phase modulation or cross-phase modulation (XPM) in nonlinear fiber or Semiconductor optical amplifiers (SOAs) for all-optical generation. This was other of the employed techniques taken as base to our experiments in chapter four.

Another possible implementation is to employ a microwave photonic filter. Gaussian monocycle or doublet pulses can be generated employing two-coefficient or three-coefficient filters, respectively. The polarity inversion can be performed based on cross-polarization modulation and employing two MZM biased at opposite slopes. One significant difficulty to overcome by microwave filters is the capacity managing the control of more than two coefficients and the amount of resources (specifically input optical sources) that they might demand to generate additional coefficients. Other remarkable challenges for these layouts are flexibility and tunability.

Finally, we must mention Optical spectrum shaping and frequency-to-time conversion. In this method, usually the spectrum of an ultra-short pulse is shaped by a spatial light modulator (SLM) or optical filtering. The shaped spectrum is converted to the time domain employing a dispersive element with a minimum required total dispersion, such as a dispersive fiber, which at the same time distributes the UWB pulse to a remote site, or a linearly-chirped FBG to simplify the system. These designs reveal more flexibility in producing UWB signals with arbitrary waveforms. However, most of the times, they are based on complex implementations requiring costly mode-locked laser sources, bulky free-space optics components or custom environment-sensitive devices

- To propose schemes for low and high order pulses creation based on both active and passive components. These methods will be reviewed not only for their legitimacy as generators but also in features as important as: flexibility, scalability and integration: One of the key aspects of this research has been to establish the relation among the spectrum exploited displayed by lower order Gaussian derivatives and higher order ones. This relationship is explained by means of the spectrum efficiency measurement, which employs the FCC settled regulations as the reference to fulfill.

Throughout this work the following main techniques were employed: PM-IM conversion, merging of complementary lower order Gaussian derivatives and exploitation of non-linear effects intrinsic to a Mach-Zehnder Interferometer (MZI) structure.

Chapter four is fully dedicated to explain in details the proposed solutions. The first experiment is an UWB doublet pulses based on the XPM effect present in a SOA-MZI. In this design, a modulated optical pulse is employed as a pump signal to convert the input optical signal into a doublet through an integrated SOA-MZI. The experimental analysis of the transfer function permits to identify the optimum operation point, which leads to the capacity of generating the desired doublet by means of the electrical currents applied to each SOA that compose the interferometric structure. For this design, we obtain an outstanding performance in terms of noise characteristics as the MZI structure improves the extinction ratio of the conversion process. Even though this scheme does not provide high order pulses, it stands as a valid solution for lower order pulses generation that can be merged for the creation of higher order ones.

Our next approach also took advantage of the non-linear effects inherent to SOA devices and presents an UWB monocycle generator based on the combination of Gaussian positive and negative pulses. The polarity of the pulses is carried out by exploiting the cross-phase modulation (XPM) effect present in an interferometric structure. Moreover, the reconfiguration of the pulses is achieved by means of an optical processor unit (OPU) which permits to control the amplitude and time delay of each optical pulse. The scalability of the system is represented by its capacity of obtaining higher order pulses by introducing additional optical probe signals and adaptation of the OPU. In this sense, by activating additional optical input sources, the system permits the addition of more coefficients and therefore generation of higher-order UWB pulses that satisfy fully the spectrum requirements of the FCC mask.

As for the schemes based on PM-IM techniques, we first proposed and experimentally demonstrated a novel UWB high-order generator with an operation principle that consists on the combination of different doublets generated by means of phase modulation and a dispersive element (a standard SMF link). In this design we employed again the concept of an OPU, composed by a set of attenuators and variable delay lines (VDLs) permits to increase the order of the generated pulse. The waveform reconfigurability is once again achieved by the independent control of the amplitude and time delay for each doublet. Another novelty present in this structure was the addition of a differential detection stage to accomplish polarity switch of the produced waveforms. At last, an upgrade of this scheme was submitted with the replacement of single mode fiber link for a customized FBG array as the dispersive element responsible for the PM-IM

conversion. In this process, the design method of the FBG structure implied the proper selection of key parameters such as the separation, length and a specific apodized profile, factors that would guarantee the production of the aspired pulses. Finally, besides the obvious reduction of size stipulated by the elimination of the fiber link, this option enable the elimination of the differential detection stage and yet maintained the capacity of complementary waveforms creation. One key merit that can be attributed to these both solutions is the production of higher order pulses with spectral efficiency values that range between 60% and 65%.

Another important observation to notice is that as it has been proven throughout the performed experiments, one of the main elements that determines the proper operation of the submitted optical systems is the input signal. In previous submitted works experimental set ups employed mainly ASE sources, which constitutes a cheap and valid alternative but such sources introduce a significant amount of noise into the systems. In the case of all the previously presented layouts, independent laser sources are managed to enter into the systems, hence they reveal better noise levels when compared to ASE based structures and the amplitude of the generated coefficients is directly determined by the optical power of each laser, aligned to this in the case of lasers that hold a capacity of wavelength tuning, there is a possibility of delay control among the samples. Such characteristic remarks again the important tunability feature present in these designs and the noticeable optimization they offer when compared to previous solutions.

- To demonstrate experimentally the validity of the proposed schemes, to verify not only their generation capacity but also the possibility of modulation performance: the clear flexibility displayed for the developed solutions permits their adaptation to the main modulations formats available in standard data transmission schemes, such as PAM, PSM, BPM and PPM.

5.2. Future Research:

Finally to conclude this thesis, different productive research lines can be identified:

- For the PM-IM approaches, the dispersive elements employed cover standard single-mode fiber (SMF) links and Fiber Bragg Gratings (FBGs), within this scope employing alternative components such as customized photonic crystals. The employment of customized dispersive devices in this PM-IM conversion schemes would provide the opportunity of full tunability into future designs, at the same time that size gest reduced.
- Aligned with the previous remark, using integrated photonic circuits should be a solution to improve performance and to reduce sizes and costs, hence a very attractive line of work is definitely focused on migration of the different stages in the submitted systems into an integrated approach. However, although the size is reduced, current silicon photonic technology does not allow full integration of the entire filter system, especially the light source, which cannot be produced using silicon since silicon is an indirect bandgap material and thus will produce no light emission. To produce effective light emission, direct bandgap materials must be used. One possible solution is to employ hybrid integration, to produce light

emission and light amplification using a direct bandgap material and using silicon for other functions (modulation, time delay, and photodetection). With the introduction of hybrid approaches, fully integrated microwave photonic filters would perform better and be ready for practical applications.

- For the conceptualized and designed layouts proposed in this thesis, only single-band oriented pulses were considered since we consider it to be the technique that automatically translates into larger bandwidth exploitation, but it would be interesting to also propose and test multiband approaches that could provide an appealing capacity to maximize the available frequency spectrum.
- Throughout this work, all the different presented schemes designed for signal generation employed Single mode fibers (SMFs) as a dispersive and transmission element. An alternative open line in this field could be represented by a detailed study of the employment of other optical fibers like silica multimode fibers (MMFs), Polymer optical fibers (POFs), bend-insensitive fiber single-mode (BI-SMF) and/or multi-mode (BI-MMF). This concept is not entirely new since for short-distance communications networks used in-vehicle, the “last mile”, and in-building home/office/factory networks, different optical media have been considered as plausible alternatives to standard SMFs.
- If discussing the advantages of alternative optical media, in the case of multi-mode fibers their larger core diameter (typically 50 μm or 62.5 μm) offers easier installation and simpler splicing when comparing them with SMFs. MMFs are also more efficient to couple to transceivers, which translate into reduced costs. For polymer optical fibers, the high elasticity and ductility of polymers allows for larger sizes compared with silica SMFs and MMFs. Further advantages of POF fibers over silica fibers are light weight, robustness, tight bend radius and better tolerance to tensile load and stress. As for the recently developed bend-insensitive single-mode fibers (BI-SMF) they can maintain the transmission and interconnection properties of SMF with much lower bending loss at lower bend radius and tighter dimensional specifications posing an interesting opportunity for a wide variety of feasible applications.

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APPENDIX A

ORIGINAL CONTRIBUTIONS

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Original Contributions in peer reviewed journals:

[More13a] **V. Moreno**, M. Rius, J. Mora, M.A. Muriel and J. Capmany, “Integrable high order pulse photonic generator based on cross-phase modulation in a SOA-MZI,” *Optics Express*, vol. 21, no. 19, pp. 22911- 22917, (2013).

[More13b] **V. Moreno**, M. Rius, J. Mora, M.A. Muriel and J. Capmany, “UWB doublet generation employing cross-phase modulation in a semiconductor optical amplifier Mach-Zehnder Interferometer,” *IEEE Photonics Journal*, vol. 05, no. 06, (2013).

[More14a] **V. Moreno**, M. Rius, J. Mora, M.A. Muriel and J. Capmany, “UWB monocycle generator based on the non-linear effects of an SOA-Integrated structure,” *IEEE Photonics Technology Letters*, vol. 26, no. 07, pp. 690-693, (2014).

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[More15a] **V. Moreno**, M. Rius, J. Mora, M.A. Muriel and J. Capmany, “Scalable high-order UWB pulse generation employing an FBG based photonic superstructure,” *IEEE Photonics Technology Letters*, vol. 27, no. 20, pp. 2146-2149, (2015).

Original Contributions in national and international conferences:

[More13c] M. Rius, **V. Moreno**, J. Mora, M.A. Muriel and J. Capmany, “UWB doublet generation using cross-phase modulation in a Semiconductor Optical Amplifier Mach-Zehnder interferometer” in *International Conference on Transparent Optical Networks (ICTON)*, Cartagena (Spain), pp. 1-4, (2013).

[More13d] **V. Moreno**, M. Rius, J. Mora, J. Capmany and M.A. Muriel, “UWB doublet generation employing non-linear effects of a Semiconductor Optical Amplifier Mach-Zehnder interferometer,” in *Reunión Española de Optoelectrónica (OPTOEL)*, Alcalá de Henares (Spain), pp. 208-212, (2013).

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[More15b] **V. Moreno**, M. Rius, J. Mora, J. Capmany and M.A. Muriel “Customized FBG based photonic superstructure for UWB signal generation applications”, in Reunión Española de Optoelectrónica (OPTOEL), Salamanca (Spain), (2015).

[More15c] **V. Moreno**, M. Rius, J. Mora, J. Capmany and M.A. Muriel “High Order UWB Pulses Generation based on a Scalable Phase-to-intensity Technique,” in European Conference and Exhibition on Optical Communication (ECOC), Valencia (Spain), paper We.1.6.5, (2015).

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