MB-OFDM metropolitan networks with concatenation of optical add-drop multiplexers

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Electrical and Computer Engineering

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Abstract

Multi-band (MB) orthogonal frequency division multiplexing (OFDM) signals have recently been proposed to be used in optical fibre telecommunications systems. The major reason for this is the increased bandwidth allocation flexibility, high spectral efficiency, higher capacity provisioning granularity and high tolerance to linear fibre distortion effects. In transparent metropolitan (metro) networks, the filtering concatenation effect due to the use of optical add-drop multiplexers (OADM) in consecutive nodes has been identified as the main limiting factor of the number of transparent nodes that can be traversed with acceptable performance.

The objective of this dissertation is to study and evaluate numerically (using a numerical simulator developed in MATLAB®) the transmission performance of 42.8 Gb/s virtual carrier-assisted direct detection MB-OFDM signals along a set of concatenated metro network nodes.

In this dissertation, the challenges presented by the transmission of MB-OFDM signals employing virtual carriers are identified. Various possible reconfigurable OADM (ROADM) architectures that can be employed in the MB-OFDM metro network are studied. An extension of a physical model of the wavelength-selective switch (WSS), which is the enabling technology of the ROADM, is proposed to take into account the group delay introduced by the WSS. The bit error ratio (BER) along a set of concatenated MB-OFDM metro network nodes is evaluated. It is shown that, for a 3-band MB-OFDM system, up to 26 ROADMs can be traversed and, for a 4-band MB-OFDM system, up to 18 ROADMs can be traversed, with BER<10^{-3}.

**Keywords:** Metropolitan networks, direct detection, orthogonal frequency-division multiplexing, multi-band, reconfigurable optical add/drop multiplexers, wavelength-selective switches.
Resumo

Sinais multi-banda (MB) com multiplexagem por divisão ortogonal na frequência (OFDM) foram propostos recentemente em sistemas de telecomunicações por fibra óptica. A principal razão para tal é devido à maior flexibilidade de alocação de banda, elevada eficiência espectral, superior capacidade de provisionamento de granularidades e facilidade em mitigar os efeitos de distorção linear na fibra. Em redes metropolitanas (metro) transparentes, o efeito de concatenação de filtros devido ao uso de multiplexadores ópticos de inserção/extracção (OADM) em nós consecutivos foi identificado como sendo o principal factor limitativo do número de nós transparentes que podem ser atravessados com desempenho aceitável.

O objectivo desta dissertação consiste em estudar e avaliar (recorrendo a um simulador numérico desenvolvido em MATLAB®) o desempenho de sistemas MB-OFDM a 42.8 Gb/s, usando portadoras virtuais para auxiliar a detecção directa, ao longo de uma série de nós concatenados numa rede metro.

Nesta dissertação, são identificados os desafios associados à transmissão de sinais MB-OFDM usando portadoras virtuais. Várias arquitecturas de OADMs reconfiguráveis (ROADM) são estudadas no contexto da rede metro MB-OFDM. É apresentada uma proposta de extensão de um modelo físico do comutador selectivo de comprimento de onda (WSS), a tecnologia de suporte da ROADM, para ter em conta o atraso de grupo introduzido pelo WSS. O rácio de erro de bit (BER) é avaliado ao longo de uma série de nós concatenados na rede metro MB-OFDM. Os resultados mostram que, para um sistema MB-OFDM com 3 bandas, podem ser atravessados até 26 ROADMs e, para um sistema MB-OFDM com 4 bandas, podem ser atravessados até 18 ROADMs, com BER<10\(^{-3}\).

**Palavras-chave:** Redes metropolitanas, detecção directa, multiplexagem por divisão ortogonal na frequência, multi-banda, multiplexadores ópticos de inserção/extracção, comutadores selectivos de comprimento de onda.
# Table of Contents

Acknowledgements v  
Abstract vii  
Resumo ix  
Table of Contents x  
List of Figures xii  
List of Tables xiv  
List of Acronyms xvii  
List of Symbols xxi  

1 Introduction 1  
1.1 Scope of the work . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . 1  
1.1.1 OFDM for optical communications . . . . . . . . . . . . . . . . . . . . . . 1  
1.1.2 Metropolitan optical networks . . . . . . . . . . . . . . . . . . . . . . . . . 3  
1.1.3 Reconfigurable optical add-drop multiplexers . . . . . . . . . . . . . . . . 5  
1.1.4 Multi-band OFDM approach . . . . . . . . . . . . . . . . . . . . . . . . . . 8  
1.2 Motivation . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . 9  
1.3 Objectives and structure of the dissertation . . . . . . . . . . . . . . . . . . . . . 10  
1.4 Main contributions . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . 11  

2 Description of the MB-OFDM system 13  
2.1 Fundamentals of OFDM . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . 13  
2.1.1 Mathematical description of an OFDM signal . . . . . . . . . . . . . . . . . 13  
2.1.2 Guard interval . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . 15  
2.1.3 Spectral efficiency . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . 16  
2.1.4 Peak-to-average power ratio . . . . . . . . . . . . . . . . . . . . . . . . . . 18  
2.1.5 OFDM transmitter and receiver . . . . . . . . . . . . . . . . . . . . . . . . . 19  
2.2 Optical transmission and direct detection . . . . . . . . . . . . . . . . . . . . . . 20  
2.2.1 Optical transmitter . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . 21  
2.2.2 Optical receiver . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . 24  
2.3 Main features of the MB-OFDM system . . . . . . . . . . . . . . . . . . . . . . . 26  
2.3.1 Virtual carrier-assisted transmission . . . . . . . . . . . . . . . . . . . . . . 27  
2.3.2 Band-selector . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . 29  
2.3.3 Virtual carrier-to-band gap . . . . . . . . . . . . . . . . . . . . . . . . . . . 29  
2.3.4 Band gap . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . 30  
2.3.5 Virtual carrier-to-band power ratio . . . . . . . . . . . . . . . . . . . . . . . 31  
2.4 Conclusions . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . 31
3 MB-OFDM metro network: operation principles and modelling
  3.1 Structure of the MB-OFDM metro network ........................................... 33
  3.2 Structure of the MB-OFDM metro network nodes .................................... 35
  3.3 Extraction and insertion of OFDM bands .............................................. 58
  3.4 Analytical model of the WSS ............................................................... 39
     3.4.1 Derivation of the frequency response of the WSS ......................... 40
     3.4.2 Group delay originated by the WSS .............................................. 42
     3.4.3 Matching the predicted group delay to experimental data ............... 45
  3.5 Dependence of the predicted WSS amplitude response and group delay on the model parameters ......................................................... 47
  3.6 Conclusions ........................................................................................... 52

4 Performance degradation along a concatenation of MB-OFDM metro network nodes
  4.1 MB-OFDM system parameters ................................................................. 53
  4.2 Results without fibre dispersion ......................................................... 54
  4.3 Results with fibre dispersion ............................................................... 56
  4.4 Analysis of the impact of the electrical noise on the MB-OFDM system performance ................................................................. 59
  4.5 Analysis of the impact of the group delay of the WSS on the MB-OFDM system performance ......................................................... 61
  4.6 Conclusions ........................................................................................... 63

5 Conclusions and future work
  5.1 Final conclusions .................................................................................... 65
  5.2 Future work ........................................................................................... 66

References ..................................................................................................... 69

A MB-OFDM system details
  A.1 Electrical noise ...................................................................................... 75
  A.2 Optical noise .......................................................................................... 75
List of Figures

1.1 Metro network in ring topology with five nodes. ........................................ 4
1.2 Architecture of a WB-based ROADM, extracted from [21]. ...................... 5
1.3 Functional diagram of a WSS, adapted from [22]. ............................... 6
1.4 Simplified diagram of a four-degree node, adapted from [22]. .................. 7
1.5 Illustration of sub-wavelength switching at a network node, adapted from [32]. 8
2.1 Two consecutive OFDM symbols with cyclic prefix. .............................. 16
2.2 Illustration of the power spectrum of an OFDM signal. ......................... 17
2.3 Block diagram of the electrical OFDM transmitter. .............................. 19
2.4 Block diagram of the electrical OFDM receiver. .................................. 20
2.5 Simplified diagram of a SSB DD-OFDM optical transmission system. .... 21
2.6 Structure of the DP-MZM. ................................................................. 22
2.7 Amplitude and phase response of the 90° hybrid coupler. ..................... 24
2.8 Spectrum of an OFDM signal after photodetection. ............................. 25
2.9 SSB MB-OFDM signal with four OFDM bands. .................................. 26
2.10 Use of a dual band optical filter to select the carrier and the OFDM band.   27
2.11 MB-OFDM signal with virtual carriers to assist the direct detection. ..... 28
2.12 Simplified diagram of the virtual carrier-assisted MB-OFDM system. ..... 29
2.13 Amplitude response, in dB, of the Gaussian and 2nd order super-Gaussian BS. 29
2.14 Illustration of the VBG. ................................................................. 30
2.15 Illustration of the BG. ........................................................................ 30
3.1 Simplified diagram of the MB-OFDM metro network. ........................... 34
3.2 Schematic diagram of a two-degree MB-OFDM node employing two WSSs. 35
3.3 Schematic diagram of a two-degree MB-OFDM node employing a splitter and a WSS. .................................................................................................................. 36
3.4 Schematic diagram of a MB-OFDM node employing an additional mux to create extra add ports. ................................................................. 37
3.5 Schematic diagram of the MEB. ............................................................. 38
3.6 Schematic diagram of a generic WSS, adapted from [48]. ....... 39
3.7 Schematic diagram of a LCOS-based WSS, showing the spatially dispersed light focused onto the aperture plane, adapted from [52]. ...................... 41
3.8 Amplitude response of a WSS (continuous line) and a Gaussian filter (dashed line), with -3 dB bandwidth of 45 GHz. ......................................... 42
3.9 Group delay of a WSS. (a) $\xi = 0$, $\zeta = 1.61 \times 10^{-21}$ s$^2$; (b) $\xi = 2.5 \times 10^{-31}$ s$^3$, $\zeta = 0$. ......................................................... 45
3.10 Group delay predicted by the extension of the WSS model, used to match the measured group delay. ................................................................. 46
3.11 Matching of the parabolic group delay obtained by the extension of the WSS model and the measured group delay [22]. ................................. 46
3.12 WSS selectivity dependence on $B/B_{OTF}$ (a), WSS selectivity dependence on $B_{OTF}$, with $B = 25$ GHz (b). ..................................................... 47
3.13 Amplitude response of the WSS for different $B/B_{\text{OTF}}$ ratios, with $B = 25$ GHz. 48
3.14 Group delay (peak-to-peak) dependence on $\zeta$, with $B = 25$ GHz and $B_{\text{OTF}} = 5.2$ GHz. 49
3.15 Group delay (peak) dependence on $\xi$, with $B = 25$ GHz and $B_{\text{OTF}} = 5.2$ GHz. 50
3.16 $m_\zeta$ dependence on $B/B_{\text{OTF}}$, for $B = 25$ GHz (a), $m_\xi$ dependence on $B/B_{\text{OTF}}$, for $B = 25$ GHz (b). 51
4.1 Signal spectra of the 42.8 Gb/s a) 3-band and b) 4-band SSB MB-OFDM signal at the output of the electro-optical modulator. 51
4.2 Schematic diagram of the MB-OFDM network model. 55
4.3 Amplitude response of a cascade of ROADMs with 1, 10 and 20 WSSs. $\xi = 0$, $\zeta = 0$. 56
4.4 BER after a cascade of nodes for the 3-band MB-OFDM signal employing a) a Gaussian BS and b) a 2nd order SG-BS. Fibre dispersion was not considered. 56
4.5 Signal spectrum of the 3-band MB-OFDM signal after a cascade of ROADMs. 57
4.6 BER after a cascade of nodes for the 4-band MB-OFDM signal employing a 2nd order SG-BS. Fibre dispersion was not considered. 58
4.7 BER after a cascade of nodes for the 3-band MB-OFDM signal employing a) a Gaussian BS and b) a 2nd order SG-BS. 58
4.8 BER after a cascade of nodes for the 4-band MB-OFDM signal employing a 2nd order SG-BS. 59
4.9 BER after a cascade of nodes for the 3-band MB-OFDM signal employing a) a Gaussian BS and b) a 2nd order SG-BS. Electrical noise of the receiver was not considered. 59
4.10 BER after a cascade of nodes for the 4-band MB-OFDM signal employing a 2nd order SG-BS. Electrical noise of the receiver was not considered. 60
4.11 Group delay introduced by a cascade of ROADMs with 1, 10 and 20 WSSs. a) $\xi = 0$, $\zeta = 1.61 \times 10^{-21}$ s$^2$; b) $\xi = 2.5 \times 10^{-31}$ s$^3$, $\zeta = 0$. 61
4.12 BER as a function of the number of ROADMs for a 4-band MB-OFDM system employing WSSs with null delay (continuous lines) and WSSs with linear delay in the passband (dashed lines). Band 1 (circles), band 2 (squares), band 3 (triangles), band 4 (stars). 61
4.13 BER as a function of the number of ROADMs for a 4-band MB-OFDM system employing WSSs with null delay (continuous lines) and WSSs with parabolic delay in the passband (dashed lines). Band 1 (circles), band 2 (squares), band 3 (triangles), band 4 (stars). 62
4.14 BER as a function of the number of ROADMs for a 3-band MB-OFDM system employing WSSs with null delay (continuous lines) and WSSs with parabolic delay in the passband (dashed lines). Band 1 (circles), band 2 (squares), band 3 (triangles). 63
List of Tables

3.1 Selectivity and bandwidth (BW) dependence on $B_{OTF}$, with $B = 25$ GHz. . . . . 48
3.2 Group delay (peak-to-peak), selectivity and bandwidth dependence on $\zeta$, with $B = 25$ GHz and $B_{OTF} = 5.2$ GHz. . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . 49
3.3 Group delay (peak), selectivity and bandwidth dependence on $\xi$, with $B = 25$ GHz and $B_{OTF} = 5.2$ GHz. . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . 50
4.1 Parameters of the MB-OFDM system. . . . . . . . . . . . . . . . . . . . . . . . . . . 55
# List of Acronyms

<table>
<thead>
<tr>
<th>Acronym</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>ADC</td>
<td>Analog-to-Digital Converter</td>
</tr>
<tr>
<td>ADM</td>
<td>Add/drop Multiplexer</td>
</tr>
<tr>
<td>ASE</td>
<td>Amplified Spontaneous Emission</td>
</tr>
<tr>
<td>ATM</td>
<td>Asynchronous Transfer Mode</td>
</tr>
<tr>
<td>AWG</td>
<td>Arrayed Waveguide Grating</td>
</tr>
<tr>
<td>BB</td>
<td>Band-blocker</td>
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<tr>
<td>BER</td>
<td>Bit Error Ratio</td>
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<tr>
<td>BG</td>
<td>Band Gap</td>
</tr>
<tr>
<td>BS</td>
<td>Band-selector</td>
</tr>
<tr>
<td>CDC</td>
<td>Colorless, Directionless, Contentionless</td>
</tr>
<tr>
<td>CDIPF</td>
<td>Chromatic Dispersion-induced Power Fading</td>
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<td>CMOS</td>
<td>Complementary Metal–oxide–semiconductor</td>
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<td>CO-OFDM</td>
<td>Coherent OFDM</td>
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<td>CP</td>
<td>Cyclic Prefix</td>
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<td>CW</td>
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<td>DAC</td>
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<td>DD</td>
<td>Direct Detection</td>
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<td>Direct Detection OFDM</td>
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<tr>
<td>DFT</td>
<td>Discrete Fourier Transform</td>
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<tr>
<td>DP-MZM</td>
<td>Dual Parallel Mach Zehnder Modulator</td>
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<td>DSB</td>
<td>Double Sideband</td>
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<td>DWDM</td>
<td>Dense Wavelength-division Multiplexing</td>
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<td>DXC</td>
<td>Digital Cross Connect</td>
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<td>EAM</td>
<td>Electro-absorption Modulator</td>
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<td>EDFA</td>
<td>Erbium-doped Fibre Amplifier</td>
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<td>EGA</td>
<td>Exhaustive Gaussian Approach</td>
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<td>EO</td>
<td>Electrical-optical</td>
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<td>FFT</td>
<td>Fast Fourier Transform</td>
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<tr>
<td>GD</td>
<td>Group Delay</td>
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<td>GDR</td>
<td>Group Delay Ripple</td>
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<td>HC</td>
<td>Hybrid Coupler</td>
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<td>Hilbert Transform</td>
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<td>Acronym</td>
<td>Description</td>
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<td>--------------------------------------------------</td>
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<tr>
<td>I/Q</td>
<td>In-phase/In-quadrature</td>
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<tr>
<td>ICI</td>
<td>Inter-carrier Interference</td>
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<tr>
<td>IDFT</td>
<td>Inverse Discrete Fourier Transform</td>
</tr>
<tr>
<td>IP</td>
<td>Internet Protocol</td>
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<tr>
<td>ISI</td>
<td>Inter-symbol Interference</td>
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<tr>
<td>LC</td>
<td>Liquid Crystal</td>
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<td>LCOS</td>
<td>Liquid Crystal on Silicon</td>
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<td>LTE</td>
<td>Long-term Evolution</td>
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<td>MB</td>
<td>Multi-band</td>
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<td>MB-OFDM</td>
<td>Multi-band Orthogonal Frequency-division Multiplexing</td>
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<td>MCM</td>
<td>Multi-carrier Modulation</td>
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<tr>
<td>MEB</td>
<td>MORFEUS Extraction Block</td>
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<td>MEMS</td>
<td>Micro-electro-mechanical Systems</td>
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<td>MORFEUS Insertion Block</td>
</tr>
<tr>
<td>MMF</td>
<td>Multi-mode Fibre</td>
</tr>
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<td>Metro Networks Based on Multi-band Orthogonal Frequency-division Multiplexing Signals</td>
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<td>MSR</td>
<td>Multiple Signal Representation</td>
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<td>MZM</td>
<td>Mach-Zehnder Modulator</td>
</tr>
<tr>
<td>OA</td>
<td>Optical Amplifier</td>
</tr>
<tr>
<td>OE</td>
<td>Optical-electrical</td>
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<td>OEO</td>
<td>Optical-electrical-optical</td>
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<td>OFDM</td>
<td>Orthogonal Frequency-division Multiplexing</td>
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<td>OSNR</td>
<td>Optical Signal-to-noise Ratio</td>
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<tr>
<td>OTF</td>
<td>Optical Transfer Function</td>
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<tr>
<td>P/S</td>
<td>Parallel/Serial</td>
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<td>PAPR</td>
<td>Peak-to-average Power Ratio</td>
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<tr>
<td>PIN</td>
<td>Positive-Intrinsic-Negative</td>
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<tr>
<td>PLC</td>
<td>Planar Lightwave Circuit</td>
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<td>PMD</td>
<td>Polarization Mode Dispersion</td>
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<td>PON</td>
<td>Passive Optical Network</td>
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<tr>
<td>PSD</td>
<td>Power Spectral Density</td>
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<td>PSK</td>
<td>Phase-shift Keying</td>
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<td>QAM</td>
<td>Quadrature Amplitude Modulation</td>
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<td>ROADM</td>
<td>Reconfigurable Optical Add/drop Multiplexer</td>
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<tr>
<td>S/P</td>
<td>Serial-to-parallel</td>
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<tr>
<td>SDH</td>
<td>Synchronous Digital Hierarchy</td>
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<td>SG-BS</td>
<td>Super-Gaussian Band-selector</td>
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<td>SLM</td>
<td>Spatial Light Modulator</td>
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<td>SMF</td>
<td>Single-mode Fibre</td>
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<td>SONET</td>
<td>Synchronous Optical Networking</td>
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<td>SSB</td>
<td>Single-sideband</td>
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<td>SSBI</td>
<td>Signal-signal Beat Interference</td>
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<tr>
<td>Acronym</td>
<td>Definition</td>
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<tr>
<td>VBG</td>
<td>Virtual carrier-to-band Gap</td>
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<td>VBPR</td>
<td>Virtual carrier-to-band Power Ratio</td>
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<tr>
<td>VLSI</td>
<td>Very-large-scale Integration</td>
</tr>
<tr>
<td>VOA</td>
<td>Variable Optical Attenuator</td>
</tr>
<tr>
<td>VoIP</td>
<td>Voice over Internet Protocol</td>
</tr>
<tr>
<td>WB</td>
<td>Wavelength-blocker</td>
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<tr>
<td>WDM</td>
<td>Wavelength-division Multiplexing</td>
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<td>WSS</td>
<td>Wavelength-selective Switch</td>
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## List of Symbols

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<tr>
<th>Symbol</th>
<th>Designation</th>
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<tr>
<td>$\Delta \lambda$</td>
<td>spectral width of the optical signal launched into the fibre</td>
</tr>
<tr>
<td>$\Delta f$</td>
<td>frequency spacing between adjacent subcarriers</td>
</tr>
<tr>
<td>$\zeta$</td>
<td>constant that controls the amount of group delay originated by the parabolic phase response of the switching element</td>
</tr>
<tr>
<td>$\eta$</td>
<td>spectral efficiency of the OFDM signal</td>
</tr>
<tr>
<td>$\xi$</td>
<td>constant that controls the amount of group delay originated by the cubic phase response of the switching element</td>
</tr>
<tr>
<td>$\sigma$</td>
<td>standard deviation of the Gaussian OTF</td>
</tr>
<tr>
<td>$\tau_{g.p}$</td>
<td>peak group delay</td>
</tr>
<tr>
<td>$\tau_{g.p-p}$</td>
<td>peak-to-peak group delay</td>
</tr>
<tr>
<td>$v_0$</td>
<td>optical frequency</td>
</tr>
<tr>
<td>$\phi(f)$</td>
<td>phase response of the switching element</td>
</tr>
<tr>
<td>$\phi_{\text{out}}(f)$</td>
<td>phase response of the WSS</td>
</tr>
<tr>
<td>$\parallel$</td>
<td>parallel polarization in the optical fibre</td>
</tr>
<tr>
<td>$\perp$</td>
<td>perpendicular polarization in the optical fibre</td>
</tr>
<tr>
<td>$B$</td>
<td>width of the rectangular aperture in frequency</td>
</tr>
<tr>
<td>$B_0$</td>
<td>optical reference bandwidth</td>
</tr>
<tr>
<td>$B_{\text{MB-OFDM}}$</td>
<td>MB-OFDM signal bandwidth</td>
</tr>
<tr>
<td>$B_{\text{OFDM}}$</td>
<td>OFDM signal bandwidth</td>
</tr>
<tr>
<td>$B_{\text{OTF}}$</td>
<td>-3 dB bandwidth of the Gaussian OTF</td>
</tr>
<tr>
<td>$B_{Gm}$</td>
<td>band gap between the $m$-th band and the $(m+1)$-th band</td>
</tr>
<tr>
<td>$D_{\lambda}$</td>
<td>chromatic dispersion parameter of the fibre</td>
</tr>
<tr>
<td>$\hat{e}_{\parallel}$</td>
<td>versor at the parallel polarization in the fibre</td>
</tr>
<tr>
<td>$\hat{e}_{\perp}$</td>
<td>versor at the perpendicular polarization in the fibre</td>
</tr>
<tr>
<td>$e_i(t)$</td>
<td>optical field at the PIN input</td>
</tr>
<tr>
<td>$e_{i,\text{EDFA}}(t)$</td>
<td>optical field at the EDFA input</td>
</tr>
<tr>
<td>$e_{n,\parallel}(t)$</td>
<td>optical noise field at the parallel polarization</td>
</tr>
<tr>
<td>$e_{n,\perp}(t)$</td>
<td>optical noise field at the perpendicular polarization</td>
</tr>
<tr>
<td>$e_{o,\text{EDFA}}(t)$</td>
<td>optical field at the EDFA output</td>
</tr>
<tr>
<td>$E_{\text{in}}$</td>
<td>input electrical field of the DP-MZM</td>
</tr>
<tr>
<td>$E_{o}(t)$</td>
<td>optical field at the DP-MZM output</td>
</tr>
<tr>
<td>$f_n$</td>
<td>frequency of the $n$-th subcarrier</td>
</tr>
<tr>
<td>$f_{n,e}$</td>
<td>noise figure of the electrical circuit of the receiver</td>
</tr>
<tr>
<td>$f_{n,o}$</td>
<td>noise figure of the optical amplifier</td>
</tr>
<tr>
<td>Symbol</td>
<td>Description</td>
</tr>
<tr>
<td>---------</td>
<td>-----------------------------------------------------------------------------</td>
</tr>
<tr>
<td>$g_{\text{EDFA}}$</td>
<td>gain of the EDFA</td>
</tr>
<tr>
<td>$h$</td>
<td>Planck constant</td>
</tr>
<tr>
<td>$H_{\text{HT}}$</td>
<td>ideal Hilbert transform transfer function</td>
</tr>
<tr>
<td>$i_{\text{pin}}$</td>
<td>photocurrent at the PIN output</td>
</tr>
<tr>
<td>$k_B$</td>
<td>Boltzmann constant</td>
</tr>
<tr>
<td>$K_{g,p}$</td>
<td>slope of the relation between $\xi$ and $\tau_{g,p}$</td>
</tr>
<tr>
<td>$K_{g,p-p}$</td>
<td>slope of the relation between $\zeta$ and $\tau_{g,p-p}$</td>
</tr>
<tr>
<td>$L(f)$</td>
<td>normalized Gaussian function</td>
</tr>
<tr>
<td>$L_f$</td>
<td>fibre length</td>
</tr>
<tr>
<td>$m$</td>
<td>modulation index of the MZM</td>
</tr>
<tr>
<td>$m_{\zeta}$</td>
<td>relation between $\tau_{g,p-p}$ and $\zeta$</td>
</tr>
<tr>
<td>$m_{\xi}$</td>
<td>relation between $\tau_{g,p}$ and $\xi$</td>
</tr>
<tr>
<td>$M$</td>
<td>modulation size alphabet</td>
</tr>
<tr>
<td>$n_b$</td>
<td>number of transmitted bits in an OFDM symbol</td>
</tr>
<tr>
<td>$n_{I,\parallel}(t)$</td>
<td>in-phase component of the optical field at the parallel polarization</td>
</tr>
<tr>
<td>$n_{I,\perp}(t)$</td>
<td>in-phase component of the optical field at the perpendicular polarization</td>
</tr>
<tr>
<td>$n_{Q,\parallel}(t)$</td>
<td>quadrature component of the optical field at the parallel polarization</td>
</tr>
<tr>
<td>$n_{Q,\perp}(t)$</td>
<td>quadrature component of the optical field at the perpendicular polarization</td>
</tr>
<tr>
<td>$N$</td>
<td>number of subcarriers</td>
</tr>
<tr>
<td>$N_{\text{bands}}$</td>
<td>number of bands composing the MB-OFDM signal</td>
</tr>
<tr>
<td>$p_{\text{band}}$</td>
<td>power of the OFDM band</td>
</tr>
<tr>
<td>$p_i(t)$</td>
<td>incident optical power on the PIN</td>
</tr>
<tr>
<td>$p_{\text{vc}}$</td>
<td>power of the virtual carrier</td>
</tr>
<tr>
<td>$p_{\text{ASE}}$</td>
<td>ASE noise power at the EDFA output</td>
</tr>
<tr>
<td>$r(t)$</td>
<td>OFDM-symbol shaping function</td>
</tr>
<tr>
<td>$R(f)$</td>
<td>aperture function of the WSS</td>
</tr>
<tr>
<td>$R_{\lambda}$</td>
<td>responsivity of the photodetector</td>
</tr>
<tr>
<td>$R_b$</td>
<td>OFDM bit rate</td>
</tr>
<tr>
<td>$R_{\text{bias}}$</td>
<td>bias resistance of the electrical part of the receiver</td>
</tr>
<tr>
<td>$R_{b,\text{MB-OFDM}}$</td>
<td>bit rate of the MB-OFDM signal</td>
</tr>
<tr>
<td>$R_{b,n}$</td>
<td>bit rate of the $n$-th OFDM band in the MB-OFDM signal</td>
</tr>
<tr>
<td>$R_s$</td>
<td>information symbol rate</td>
</tr>
<tr>
<td>$s(t)$</td>
<td>electrical MB-OFDM signal</td>
</tr>
<tr>
<td>$s_n(t)$</td>
<td>waveform of the $n$-th subcarrier</td>
</tr>
<tr>
<td>$s_H(t)$</td>
<td>Hilbert transform of $s(t)$</td>
</tr>
<tr>
<td>$s_{\text{SSB}}(t)$</td>
<td>SSB signal at the output of the DP-MZM</td>
</tr>
<tr>
<td>$S(f)$</td>
<td>frequency response of a bandpass filter created by the WSS</td>
</tr>
<tr>
<td>$S_c$</td>
<td>one-sided power spectrum density of the electrical circuit noise</td>
</tr>
<tr>
<td>$S_{\text{ASE}}$</td>
<td>power spectral density of the ASE noise</td>
</tr>
<tr>
<td>$t_d$</td>
<td>time spread related to the chromatic dispersion parameter of the fibre</td>
</tr>
<tr>
<td>$T_g$</td>
<td>guard interval duration</td>
</tr>
<tr>
<td>Symbol</td>
<td>Definition</td>
</tr>
<tr>
<td>----------</td>
<td>-----------------------------------------------------------------------------------------------</td>
</tr>
<tr>
<td>$T_r$</td>
<td>room temperature</td>
</tr>
<tr>
<td>$T_s$</td>
<td>duration of the OFDM symbol</td>
</tr>
<tr>
<td>$v_1(t)$</td>
<td>electrical signal at arm 1 of the DP-MZM</td>
</tr>
<tr>
<td>$v_2(t)$</td>
<td>electrical signal at arm 2 of the DP-MZM</td>
</tr>
<tr>
<td>$V_\pi$</td>
<td>switching voltage of the MZM</td>
</tr>
<tr>
<td>$V_{b,1}$</td>
<td>bias voltage of MZM 1 of the DP-MZM</td>
</tr>
<tr>
<td>$V_{b,2}$</td>
<td>bias voltage of MZM 2 of the DP-MZM</td>
</tr>
<tr>
<td>$V_{\text{rms}}$</td>
<td>root mean square (RMS) voltage of the electrical signals applied to the MZM arms</td>
</tr>
<tr>
<td>$x(t)$</td>
<td>transmitted OFDM signal represented in time domain</td>
</tr>
<tr>
<td>$X_{n,i}$</td>
<td>information symbol at the $n$-th subcarrier of the $i$-th OFDM symbol</td>
</tr>
</tbody>
</table>
Chapter 1

Introduction

In this chapter, an introduction to metropolitan (metro) optical networks using orthogonal frequency division multiplexing (OFDM) signals is provided. In section 1.1, the scope of the work is presented along with a description of optical OFDM implementations, as well as the technologies used by reconfigurable optical add-drop multiplexers (ROADMs), which are part of the metro network nodes. The motivation of this work is presented in section 1.2. The objectives and the organization of the dissertation are presented in section 1.3. The main contributions are identified in section 1.4.

1.1 Scope of the work

The scope of this work is to study the transmission of multi-band OFDM (MB-OFDM) signals along a set of concatenated nodes in metropolitan optical networks. Particularly, the impact of the filter concatenation effect on the transmission performance of these metro networks is addressed.

In this section, the main topics that constitute the framework of this dissertation are discussed. These include a brief history of OFDM and a description of the two main ways to implement optical OFDM. The state-of-the-art of metro optical networks is presented along with a description of the technologies used by ROADMs. The multi-band approach to OFDM transmission in metro networks is introduced.

1.1.1 OFDM for optical communications

OFDM is a modulation technique that encodes digital data on multiple carrier frequencies. It uses the principle of orthogonality to achieve high spectral efficiency and offers many advantages such as resilience to channel dispersion. In recent years, OFDM has been seen as a promising candidate for future optical networks [1]-[4]. The technical aspects of OFDM modulation are described in chapter 2.
**History of OFDM**

The concept of OFDM was formulated by Chang in 1966 [5]. It was first developed to be used in military applications. At that time, there was a lack of sufficiently powerful integrated electronic circuits to support the complex computation required by OFDM. In the 1990s, the maturing of very large scale integrated (VLSI) CMOS chips and the arrival of broadband digital applications prompted a surge of interest in OFDM. It is now used in a wide range of applications such as digital audio and video broadcasting, wireless local area networks and digital subscriber lines. OFDM has also been adopted in fourth-generation mobile communication systems based on long-term evolution (LTE) [1].

The application of OFDM to optical communications came late compared to its early adoption for radio frequency standards. The first paper on optical OFDM appeared in the literature in 1996 [6]. However, the robustness of OFDM against fibre dispersion was recognized as a fundamental advantage of OFDM for optical communications only in 2001. That work proposed the use of OFDM to mitigate the modal dispersion in multi-mode fibre (MMF) [7]. In recent years, research on optical OFDM has been focused on single-mode fibre (SMF), starting with proposals for long-haul applications [8][9]. Optical OFDM has also been seen as a likely candidate for future passive optical networks (PON) and metro networks, due to its ability to provide flexible bandwidth allocation [10][11].

**Optical OFDM implementations**

The two main implementations of optical OFDM are direct-detection optical OFDM (DD-OFDM) and coherent optical OFDM (CO-OFDM) [1]. In DD-OFDM, direct-detection is used at the receiver to convert the optical signal to electrical domain with a single photodiode. An optical carrier is sent along with the OFDM signal and, for this reason, a laser is not required at the receiver. This lessens the problem of sensitivity of OFDM to phase noise and frequency offset. DD-OFDM is less power efficient, because some of the available power has to be allocated to the optical carrier, which bears no information [2].

CO-OFDM offers the best performance in receiver sensitivity, spectral efficiency and is more robust against polarization mode dispersion (PMD) [11]. As its name suggests, the underlying principle is coherent detection, which means that the optical carrier does not need to be transmitted as a result of the received signal being mixed with a locally generated carrier for detection. This allows more power to be allocated to the signal and, due to its architecture, a frequency guard band to prevent intermodulation distortion is not needed. CO-OFDM requires lasers at both the transmitter and the receiver and is also more sensitive to frequency and phase noise,
leading to a complex and costly implementation, which can be an obstacle for future PON and metro systems [4]. Coherent detection is seen as a good candidate for long-haul optical transmission systems; however, due to its disadvantages, more cost effective solutions are preferred for metropolitan and optical access networks. DD-OFDM is a more suitable choice for these applications.

1.1.2 Metropolitan optical networks

The growing use of Internet applications such as cloud computing, video streaming and voice over IP (VoIP) results in increasing data rate requirements. In the past 5 years, annual global data traffic has increased more than four times, and will increase threefold over the next five years [12]. This growth has prompted research in flexible optical networks. Since the late 20th century, optical communications systems have been the main trend of study, as a result of electrical-based systems having reached a point of saturation in capacity and reach [1].

Metro networks are the part of the optical networks that usually lies within a large city or a region [13]. They aggregate the traffic coming from the access networks and provide the link to the long-haul core network. Metro networks can reach a few hundred kilometres, depending on the geographical area, but they typically extend to about 200 km [14]. Regional networks have a similar purpose, but they can have longer link lengths in order to serve sparse populated areas.

Long-haul networks usually represent strategic, long term investments. In contrast, metro networks need to have a cost effective architecture, because the network cost is divided among a smaller number of customers [14]. For this reason, there are space and power consumption constraints, and system complexity is kept to a minimum. In addition, metro networks need to cope with considerable fluctuations in traffic flow due to the aggregation of traffic from several users with different bitrate requirements. They also need to support different types of traffic coming from the access networks (IP, ATM, Ethernet, SDH, etc.). A study from Bell Labs predicts that metro traffic will grow about two times faster than long-haul traffic by 2017 [15]. As a result, metro networks must present high flexibility, dynamic reconfigurability and enabling scalability.

The preferred metro network physical architecture is the ring topology, as it is represented in figure 1.1. Rings are sparse but still provide an alternative path for traffic in case a link fails [13]. In the past, metro networks were designed to transport synchronous digital hierarchy (SDH) or synchronous optical networking (SONET) traffic. The nodes of the network consisted of add-drop multiplexers (ADM) and digital cross-connects (DXC) to interconnect adjacent rings. The traffic was carried by one wavelength per fibre (single-channel transmission) and the networks
INTRODUCTION

Figure 1.1: Metro network in ring topology with five nodes.

were opaque, i. e., optical-electrical-optical (OEO) conversion was required at each node [13]. The introduction of wavelength-division multiplexing (WDM) allowed the transmission of several wavelengths (WDM channels) per fibre providing increased transport capacity and network flexibility [16]. In addition, WDM systems are capable of wavelength routing, which prompted the development of optical add-drop multiplexers (OADMs) and optical cross-connects (OXCs) [14]. OADM offer traffic aggregation/extraction capabilities without the need of OEO conversion (transparent nodes).

The design of transparent WDM metro networks has to take into account several impairments [17][18]: (i) the noise introduced by the optical amplifiers and the distortion caused by the filter concatenation effect limit the distance between nodes and the number of nodes in the network, (ii) chromatic dispersion introduced by the optical fibre has to be compensated, (iii) linear crosstalk due to the finite selectivity of add and drop functions causes performance degradation. In dense wavelength-division multiplexing (DWDM) networks (channel spacing not exceeding 200 GHz), fibre nonlinearities can cause significant performance degradation if the optical power level is not appropriately selected [18].

In today’s metro networks, reconfigurable optical add-drop multiplexers (ROADMs) are able to automate the rearrangement of wavelengths on optical fibres leaving and entering network nodes [19]. Still, as channel capacity increases, the available levels of granularity in transparent WDM metro networks become too restrictive to allocate bandwidth to the users in a dynamic and effective manner [4]. OEO conversion of wavelengths is to be avoided, because it increases network cost and power consumption. It has been envisioned that metro and access networks could be integrated into a single hybrid optical network in order to meet users’ demands of higher data rate and to simplify network architecture [20].
1.1.3 Reconfigurable optical add-drop multiplexers

Over the past few years, there has been a trend to add wavelength reconfigurability to OADM nodes. As DWDM is deployed extensively in metro networks, there is a need to remotely reconfigure which wavelengths are added, dropped or passed-through the nodes [18]. In the past, the rearrangement was done by separating each wavelength at a node using an optical demultiplexer, manually rearranging wavelengths at an optical patch panel and combining them again using an optical multiplexer. This operation was time consuming and prone to human error. The use of a large number of wavelengths and unpredictable bandwidth demand have led to the development of ROADM s, which allow the rearrangement to be done under software control, simplifying manual operations [19].

ROADMs can be constructed using a variety of enabling component technologies which differ in functionality, performance and cost [18][21]. These technologies include wavelength blockers (WBs), planar lightwave circuits (PLCs) and wavelength-selective switches (WSSs). The first commercial available ROADM s were based on WBs and typically supported 80 channels with 50 GHz channel spacing for long-haul networks and 32 or 40 channels with 100 GHz channel spacing for metro networks [21]. Deployment of these ROADM s began in late 1990s [19].

Figure 1.2 shows the architecture of a ROADM that uses a demultiplexer (demux) and a variable multiplexer (vmux) to create a WB. A vmux integrates an array of variable optical attenuators (VOAs) and a multiplexer (mux). The WB is able to attenuate (block) or pass-through specific wavelengths, being referred as 1×1 switch (with one input port and one output port, no switching is actually performed) [18]. This type of ROADM s has a broadcast and select architecture, since the incoming optical signal is broadcast through a passive coupler. A part of the signal goes to the drop path and the other is sent to the WB (pass-through path) [13]. The
drop path leads to a demux or a passive splitter. The dropped wavelengths are subsequently
blocked from propagating further by the WB. At the output of the ROADM, wavelengths can be
added through a passive combiner. VOAs are usually employed to equalize the optical power of
added wavelengths [19]. The optical performance monitor (OPM) measures channel power and
provides feedback to the VOAs.

WBs are more commonly implemented with free space optics, using micro-electro-mechanical
systems (MEMS), liquid crystal (LC) or liquid crystal on silicon (LCOS) switching technolo-
gies [21]. These do not require VOAs, since wavelength power equalization is provided by these
devices [18]. A detailed description of the MEMS switching technology is given in chapter 3.

PLC-based ROADMks are slightly more complex than WB-based ROADMks. The functional
architecture differs from the one depicted in figure 1.2 in that switching is performed from either
the pass-through or add path to the output of the ROADM. PLCs integrate a demux to separate
the wavelengths from the incoming signal and a switch based on a Mach–Zehnder interfero-
meter [18]. PLCs technology is used mainly in 2×1 switches due to performance limitations.

The architectures based on WBs or PLCs generally have fixed wavelength-port mapping, i.e.,
each wavelength is associated with a specific add or drop port (fixed ROADM). The use of tunable
filter arrays at the drop ports and tunable lasers at the add ports allow colorless operation, but
at an increased cost [21]. Being colorless means that any wavelength can be routed to any drop
port and from any add port [19].

Wavelength-selective switch

The WSS has emerged as the dominant enabling technology in ROADMks [19] [23]. A functional
representation of a WSS is depicted in figure 1.3. A single WSS has equivalent functionality to a
module consisting of a large number of optical switches, VOAs and multiplexers. It is a colorless
device, so any individual or set of wavelengths at the input of a 1×N WSS can be switched to any

![Figure 1.3: Functional diagram of a WSS, adapted from 22.](image-url)
of the N output ports. The flexible nature of the WSS allows to increase the degree of the network nodes. In DWDM networks, the degree of a node corresponds to the number of fibres entering and leaving the node (considering traffic moving in one direction only) [19]. The architectures based on WBs and PLCs previously described are applicable in two-degree nodes [21]. On the other hand, WSS-based ROADM are able to provide extra expansions ports and therefore allow nodes to be mesh upgradeable [18] [23].

Four-degree nodes are particular relevant to metro networks, since they are used to interconnect two adjacent metro rings. In early DWDM networks, nodes with a degree higher than two (multi-degree nodes) employed multiple WBs, but this results in an architecture that is too bulky and costly for metro networks [18].

Figure 1.4 depicts a four-degree node, employing four 1:5 WSSs to interconnect two adjacent DWDM rings. Any wavelength entering the node can be switched to one of the three possible output directions. Wavelengths can also be added or dropped in the same way as two-degree nodes. Examples of multi-degree node architectures are given in [24].

WSS technology relies on free-space optics and a switching engine that directs the wavelengths to the desired ports. The switching engine performs this task by changing the polarization (using LC cells), angle (using tilting MEMS micromirrors) or phase (using a LCOS or MEMS-based spatial light modulator) of a wavelength-dispersed optical beam [25] [26]. The switching engine also enables the WSS to perform wavelength power equalization. Although the WSS has better pass-through performance than PLCs, the passband narrowing due to the concatenation of ROADM (filter concatenation effect) is still the major limiting factor in the number of trans-
parent nodes that can be traversed in metro networks [18][27]. The physical architecture of the WSS and the architectures of WSS-based ROADM are analysed in chapter [3].

Current WSSs support up to 96 controllable WDM channels with 50 GHz channel spacing and 1×20 port configuration. Flexible-grid WSSs allow to dynamically adjust channel bandwidth (and spacing) in 12.5 GHz increments [28][29].

Next generation ROADM architectures will focus on three key features: colorless, directionless and contentionless (CDC). Directionless means that any wavelength can be sent in any direction, e. g. any wavelength can be added or dropped at any port. Contentionless means that multiple copies of the same wavelength can be added or dropped at the same node [30].

1.1.4 Multi-band OFDM approach

Multi-band (MB) OFDM transmission provides access to finer capacity granularity levels, increasing network flexibility with no need for electrical switching. Instead of transmitting a single OFDM signal, multiple OFDM bands can be transmitted in the same wavelength (optical channel). Each band has its own carrier frequency and behaves as an independent OFDM signal. Thus, transparent routing and aggregation of bands with less capacity than the WDM channel are possible [31].

The rectangular spectral shape of the OFDM bands is particular suitable for a multi-band approach, although the inclusion of a guard band between adjacent OFDM bands reduces spectral efficiency [32]. As each individual OFDM band uses less bandwidth, the performance requirements of digital-to-analog converters (DACs) and analog-to-digital converters (ADCs) are greatly relaxed. At the transmitter and the receiver, the DACs and ADCs do not have to process the entire MB-OFDM signal, but only a selected OFDM band [31]. In addition, MB-OFDM maintains the same advantages of single-band OFDM, such as high tolerance to channel dispersion.

In order to insert and extract individual OFDM bands at a network node, additional functional blocks are required apart from the ROADM. These blocks provide sub-wavelength switch-
ing, but represent an increase in system complexity. Another challenge is the finite selectivity of the optical filters used in the process of insertion and extraction of OFDM bands, which leads to crosstalk between bands. The reduction of the passband width due to the filter concatenation effect has to be considered in the design of the MB-OFDM system.

Figure 1.5 shows an example of sub-wavelength aggregation of OFDM bands. Band 5 coming from input fibre 1 is extracted and dropped, leaving a temporary empty frequency band slot. This subsequently allows the other bands from the two input fibres to be aggregated and switched to the output fibre. The bands that are aggregated in the node must occupy the same wavelength position in the input fibres [32].

Recently, a direct detection MB-OFDM system was proposed (MORFEUS) [33] to address the future requirements of metro networks, such as high flexibility, enabling scalability, dynamic reconfigurability and transparency. A particularity of this system is the use of a virtual carrier close to each OFDM band to assist the direct detection.

The optical MB-OFDM system employed in this work is described in chapter 2. The operation principles of the MB-OFDM metro network, including the processes of insertion and extraction of OFDM bands are explained in detail in chapter 3.

1.2 Motivation

The expected growth in metro traffic over the next few years highlights the need for metro networks presenting high flexibility, enabling scalability, dynamic reconfigurability and transparency. DD-OFDM is a cost effective solution for metro networks, offering high spectral efficiency and resilience to channel dispersion. The transmission of MB-OFDM signals provides access to sub-wavelength granularity levels, increasing bandwidth allocation flexibility. The use of virtual carriers close to each OFDM band to assist the direct detection has been shown to be an efficient way of providing the features required by future metro networks [33].

In transparent metro networks, the passband narrowing due to the concatenation of ROADMs has been identified as the main limiting factor in the number of transparent nodes that can be traversed. Future metro network architectures are expected to support an optical reach of several hundred kilometres without regeneration and 16-24 pass-through nodes, in order to minimize costs [18]. Therefore, it is important to evaluate the transmission performance of MB-OFDM signals along a set of concatenated nodes in metro networks.

An accurate model of the ROADMs and their enabling technology, the WSS, needs to be obtained. This requires a careful study of the switching technologies used by these devices and the capabilities and limitations of the various ROADM architectures. The group delay ripple
originated by the WSS and its impact on the transmission performance of MB-OFDM signals also needs to be evaluated.

In this dissertation, the transmission performance of virtual carrier-assisted MB-OFDM signals in metro networks is evaluated after a set of concatenated ROADMs. Various MB-OFDM signal configurations are tested, in order to determine which has the best pass-through performance. The impact of other metro network impairments, such as fibre dispersion, optical noise and electrical noise of the receiver is also analysed and discussed.

1.3 Objectives and structure of the dissertation

The main objective of this dissertation is to evaluate the transmission performance of 42.8 Gb/s virtual carrier-assisted single-sideband (SSB) MB-OFDM signals along a set of concatenated metro network nodes. To this aim, a numerical simulator of the MB-OFDM system is developed using the software MATLAB®. Emphasis is put on the modelling of the ROADMs present in the network nodes and on the analysis of the impact of the filter concatenation effect on the transmission performance of metro networks. The maximum number of concatenated ROADMs allowed in the metro network while maintaining acceptable performance degradation is identified.

The dissertation is composed of five chapters and an appendix providing supplemental information about the MB-OFDM system.

In chapter 2 the MB-OFDM system is described. The fundamentals of OFDM modulation are presented, as well as the processes of optical SSB transmission and direct detection of optical OFDM signals. The optical transmitter, which uses a dual parallel Mach-Zehnder modulator (DP-MZM), is described. Also, the optical MB-OFDM signal and relevant optical parameters are introduced.

In chapter 3 the MB-OFDM metro network is characterized and its operation principles are described. A model for the metro network nodes is provided, including a thorough analytical description of the WSS, which is the enabling technology of the ROADMs. The dependence of the predicted WSS amplitude response and group delay on the model parameters is analysed in detail, and expressions are obtained for the relevant relations.

In chapter 4 the transmission performance of the MB-OFDM metro network is analysed. The impact of the passband narrowing and non-constant group delay of the cascade of ROADMs is evaluated for 3-band and 4-band MB-OFDM signals. The performance is evaluated in terms of bit error ratio (BER) after a cascade of nodes, with optical noise distributed along the network.

In chapter 5 the final conclusions of this dissertation are outlined and proposals for future work are provided.
1.4 Main contributions

In the author’s opinion, the main contributions of the work developed in this dissertation are:

- Implementation of a MB-OFDM metro network simulator allowing to evaluate the transmission performance along a set of concatenated nodes,

- Identification of the challenges presented by the transmission of MB-OFDM signals employing virtual carriers and proposal of solutions to address those challenges,

- Study of the various possible ROADM architectures that can be employed in the MB-OFDM metro network,

- Development of an extension of a physical model of the WSS to take into account the group delay introduced by MEMS-based WSS,

- Analysis of the dependence of the predicted WSS amplitude response and group delay on the model parameters,

- Evaluation of the BER along a set of concatenated MB-OFDM metro network nodes, taking into account the optical noise distributed along the network, fibre dispersion and electric noise introduced by the receiver,

- Assessment of the maximum number of concatenated MB-OFDM metro network nodes that leads to a BER $< 10^{-3}$, for 3-band and 4-band MB-OFDM signals.

The main results of this work were presented in the following papers:


- J. Rosário, T. Alves and A. Cartaxo, “Impact of the WSS delay distortion of the ROADM cascade performance in virtual carrier-assisted DD multi-band OFDM metro networks”, presented at International Telecommunications Symposium, São Paulo, Brazil, 17-20 August 2014. DOI: 10.1109/ITS.2014.6948007
Chapter 2

Description of the MB-OFDM system

In this chapter, the optical fibre communication system based on the transmission of MB-OFDM signals along metropolitan networks is described. In section 2.1, the fundamentals of OFDM signals are characterized and the most important concepts about OFDM transmission are presented, as well as the advantages and drawbacks of this type of modulation. In the same section, the architecture of the OFDM electrical transmitter and receiver is described. In section 2.2, the optical transmission and the direct detection process are explained, including a description of the optical transmitter and the optical receiver. The virtual carrier-assisted MB-OFDM optical signal is characterized in section 2.3.

2.1 Fundamentals of OFDM

OFDM encodes digital data on multiple carrier frequencies. It is a special case of multi-carrier modulation (MCM), where the stream of data is divided into parallel lower rate streams which are assigned to subcarriers. The particularity of OFDM is that the subcarriers are allowed to partially overlap because they are chosen to be orthogonal between themselves [1].

2.1.1 Mathematical description of an OFDM signal

An OFDM signal is a sequence of OFDM symbols, each composed of $N$ subcarriers. Input data is modulated using phase-shift keying (PSK) or M-ary quadrature amplitude modulation (M-QAM) such that an information symbol (in case of M-QAM, a complex number representing a specific QAM constellation point) is mapped to a single subcarrier. The transmitted OFDM signal is represented in time domain as [3]:

$$x(t) = \sum_{i=\infty}^{+\infty} \sum_{n=0}^{N-1} X_{n,i} s_n(t - iT_s) \quad (2.1)$$
where $X_{n,i}$ is the information symbol at the $n$-th subcarrier of the $i$-th OFDM symbol, $t$ is the time (in seconds) and $T_s$ is the duration of the OFDM symbol (in seconds). $s_n(t)$ is the waveform of the $n$-th subcarrier and is given by

$$s_n(t) = r(t) e^{j2\pi f_n t} \quad (2.2)$$

where $f_n$ is the frequency of the $n$-th subcarrier and $r(t)$ is the OFDM-symbol shaping function given by

$$r(t) = \begin{cases} 
1, & 0 < t \leq T_s \\
0, & t \leq 0, t > T_s 
\end{cases} \quad (2.3)$$

An important difference between OFDM and other MCM implementations is that the signals comprising each subcarrier overlap in frequency. However, since they are orthogonal among each other, they can be recovered without intercarrier interference (ICI). There is no need for a guard band between individual subcarriers. This property of OFDM signals leads to high spectral efficiency [2].

For two subcarriers $s_k(t)$ and $s_l(t)$ to be orthogonal, they must verify [3]

$$\langle s_k, s_l \rangle = \frac{1}{T_s} \int_0^{T_s} s_k(t)s_l^*(t) \, dt = 0 \quad (2.4)$$

This condition must hold true for all $\langle s_k, s_l \rangle$, $k = 0, 1, \ldots, N - 1$, $l = 0, 1, \ldots, N - 1$, as long as $k \neq l$. In other words, the orthogonality condition must be verified for any two subcarriers of the OFDM symbol. Hence, it can be shown that the frequency spacing $\Delta f$ between adjacent subcarriers must be a multiple of the inverse of the OFDM symbol duration [3]

$$\Delta f = \frac{m}{T_s}, \quad m \in \mathbb{N} \quad (2.5)$$

In order to maximize the spectral efficiency, the value of $m$ should be $m = 1$. The subcarrier frequencies can then be defined as

$$f_n = \frac{n}{T_s}, \quad n = 0, 1, \ldots, N - 1 \quad (2.6)$$

At the receiver, the subcarriers can be recovered with the use of matched filters and without ICI. There is a more elegant and efficient way of generating and recovering the OFDM signals using the discrete Fourier transform (DFT).
**DFT implementation**

In an OFDM system, a large number of subcarriers are needed so that the channel appears flat to each subcarrier [1]. To meet this requirement, several oscillators and filters are required at the transmitter and receiver, leading to a excessively complex architecture. It was demonstrated that OFDM modulation and multiplexing can instead be implemented digitally using the inverse discrete Fourier transform (IDFT) [34]. The demodulation and demultiplexing follow the inverse procedure, namely the DFT. In addition, the IDFT/DFT can be computed using a very efficient algorithm, the Fast Fourier Transform (FFT), which reduces the number of complex multiplications from $N^2$ when the algorithm is not used to $\frac{N}{2} \log_2 N$ when used. This means that the number of complex multiplications of FFT scales almost linearly with the number of subcarriers $N$ [3].

For simplicity, let us consider only one OFDM symbol. In this situation, equation 2.1 can be simplified to

$$x(t) = \sum_{n=0}^{N-1} X_n e^{j2\pi f_n t}$$

(2.7)

and combining equation 2.7 with the orthogonality condition in equation 2.6, we obtain

$$x(t) = \sum_{n=0}^{N-1} X_n e^{j2\pi nt/T_s}$$

(2.8)

If equation 2.8 is converted to the digital domain, by sampling at times $t = kT_s/N$, the $k$-th sample of $x(t)$ is given by

$$x[k] = \sum_{n=0}^{N-1} X_n e^{j2\pi nk/N}$$

(2.9)

$$x[k] = F^{-1}\{X_n\}$$

(2.10)

which is the IDFT of the transmitted signal. If a IDFT with $N$ points is taken from $N$ frequency domain information symbols, the time domain version of the corresponding OFDM signal with $N$ subcarriers is obtained.

### 2.1.2 Guard interval

When an OFDM signal is transmitted across a dispersive channel, the pulse spreads along time causing different subcarriers to arrive misaligned at the receiver. In this situation, two problems can occur: (i) intersymbol interference (ISI), when a subcarrier crosses the symbol boundary
of neighbouring OFDM symbol and (ii) ICI, when the waveform of a “slower” subcarrier is incomplete in the DFT window causing the orthogonality principle to be lost \cite{1}.

To avoid ISI, a guard interval is inserted between consecutive OFDM symbols. The guard interval duration \( T_g \), in which no signal is transmitted, needs to accommodate the time spread \( t_d \) due to the received OFDM symbol being longer than the original symbol before transmission. The condition for ISI-free transmission is then given by

\[
t_d < T_g
\]

where the time spread \( t_d \) is related to the chromatic dispersion parameter of the fibre \( D_\lambda \) (in ps/nm/km), the fibre length \( L_f \) (in km) and the spectral width of the optical signal launched into the fibre \( \Delta \lambda \) (in nm), and given by

\[
t_d = D_\lambda L_f \Delta \lambda \quad [\text{ps}]
\]

By filling the guard interval with a cyclic prefix (CP), the ICI penalty can also be avoided. CP insertion consists of prepending an end portion of the OFDM symbol to its beginning. As long as the CP length is longer than \( t_d \), the information symbols can be recovered by means of channel estimation, using a single-tap equalizer in the frequency domain \cite{1}. Figure 2.1 illustrates a sequence of two OFDM symbols with CP extension.

An OFDM signal consisting of a sequence of OFDM symbols with CP extension can be described by equation 2.1 if the pulse shaping function \( r(t) \) (equation 2.3) is extended to the guard interval \cite{1}, being given by

\[
r(t) = \begin{cases} 
1, & -T_g < t \leq T_s \\
0, & t \leq -T_g, t > T_s
\end{cases}
\]

\[ \text{(2.13)} \]

### 2.1.3 Spectral efficiency

Spectral efficiency refers to the information rate that can be transmitted over a certain bandwidth. OFDM has high spectral efficiency in comparison to other MCM implementations, because
the subcarriers are allowed to overlap in frequency. This aspect can be seen in figure 2.2 which shows the power spectrum of an OFDM signal. Due to the rectangular OFDM-symbol shaping function, each transmitted subcarrier has a sinc squared-shaped spectrum.

![Image of OFDM spectrum]

Figure 2.2: Illustration of the power spectrum of an OFDM signal.

In order to evaluate the spectral efficiency of OFDM signals, it is important to define some parameters. For every OFDM symbol, \( N \) subcarriers are transmitted in a period of \( T_s + T_g \) seconds, with each subcarrier carrying an information symbol. The information symbol rate can then be written as

\[
R_s = \frac{N}{T_s + T_g} \text{ [symbol/s]}
\]  

(2.14)

Each subcarrier contains a sequence of bits mapped as a M-PSK or M-QAM symbol. The number of transmitted bits in an OFDM symbol is given by

\[
n_b = N \log_2 M \text{ [bit]}
\]  

(2.15)

where \( M \) is the size of the modulation alphabet. The OFDM bit rate is then given by

\[
R_b = \frac{n_b}{T_s + T_g} = \frac{N \log_2 M}{T_s + T_g} \text{ [bit/s]}
\]  

(2.16)

The OFDM signal bandwidth is given by

\[
B_{OFDM} = N \Delta f = \frac{N}{T_s} \text{ [Hz]}
\]  

(2.17)

where \( \Delta f \) is the frequency spacing between the subcarriers, as defined by equation 2.5 with \( m = 1 \).
From the results of equations 2.16 and 2.17, the spectral efficiency $\eta$ of an OFDM signal can be defined as

$$\eta = R_b \frac{T_s}{B_{OFDM}} = \frac{T_s}{T_s + T_g} \log_2 M = \frac{1}{1 + \frac{T_g}{T_s}} \log_2 M \quad \text{[bit/s/Hz]} \quad (2.18)$$

2.1.4 Peak-to-average power ratio

OFDM signals possess a significantly high peak-to-average power ratio (PAPR), which constitutes a disadvantage of this type of modulation. High PAPR occurs due to the multi-carrier nature of OFDM. The independent phases of subcarriers often combine constructively, causing peaks [1].

In optical fibre communication systems, OFDM signals with high PAPR suffer from higher distortion due to the optical fibre nonlinearities, because the signal peaks enhance their effects [35]. In addition, the converters (ADCs and DACs) must have a wide dynamic range to support these peaks of power. A way to avoid this requirement is to set the signal power much lower than the components saturation power, but this leads to a poor signal-to-noise ratio (SNR). On the other hand, when components saturate, signal clipping occurs causing distortion.

The PAPR of an OFDM signal is defined as [1]

$$\text{PAPR} = \frac{\max\{|x(t)|^2\}}{E\{|x(t)|^2\}}, \quad t \in [0, T_s] \quad (2.19)$$

where $\max\{}$ stands for the maximum value function and $E\{}$ for the expected value operator. The theoretical maximum PAPR is $10 \log_{10}(N)$ [1]. If the number of subcarriers is $N = 128$, the maximum PAPR is 21 dB, which is excessively high [1]. The theoretical maximum only occurs very rarely, so a different metric is used instead. The complementary cumulative distribution (CCDF) of PAPR corresponds to the probability that PAPR exceeds a certain value [1].

There has been extensive research on PAPR reduction. Some techniques are based on coding, multiple signal representation (MSR) or signal clipping [2].

OFDM suffers from another disadvantage which is its sensitivity to frequency offset and phase noise. This problem is more severe when using CO-OFDM, particularly with higher order constellations. In this case, frequency estimation and compensation can be implemented to reduce the impact of frequency offset and adequate design of oscillators is essential [1].
2.1.5 OFDM transmitter and receiver

The architecture of the electrical OFDM transmitter is depicted in figure 2.3. It is composed of a digital transmitter and an analog front-end where the signal is up-converted. The operations performed by each block are explained as follows:

• In the serial-to-parallel (S/P) block, the binary data is split into \( N \) parallel streams (one for each subcarrier). The bits in each stream are mapped as M-PSK or M-QAM symbols.

• The mapped symbols are then subjected to zero padding, which consists in introducing additional subcarriers with zero amplitude. This is an oversampling procedure that relaxes the requirements of the reconstruction filter used after the DAC to prevent aliasing.

• In the IDFT block, the OFDM symbol is generated using the IFFT algorithm. A guard interval is appended to the beginning of each symbol and filled with a cyclic prefix. The symbols are then converted to serial format in the parallel-to-serial (P/S) block. The in-phase (I) and quadrature (Q) components are separated.

• The digital part of the transmitter ends here, as the I and Q signals are converted to an analog waveform at the digital-to-analog converter (DAC). The signals are then low-pass filtered to remove any aliasing products originated in the DAC conversion.

• The I and Q components are up-converted to a specific carrier frequency and combined to form a real-valued signal. A virtual carrier is transmitted together with the signal to assist the direct detection at the receiver. This process is explained in section 2.2.1.

The architecture of the electrical OFDM receiver is shown in figure 2.4. It is composed of an analog front-end where the received signal is down-converted and a digital receiver. The operations performed by each block are explained as follows:
The down-conversion process is performed using the same carrier frequency as in the transmitter. The I and Q components are retrieved and low-pass filtered to reduce out-of-band distortion components and to reduce noise power.

Discrete samples are obtained after analog-to-digital conversion at the ADC. Symbol synchronization is achieved by performing the autocorrelation of the received signal.

In the S/P block, the discrete sequence is transformed into a parallel stream of complex samples. The cyclic prefix is removed from each symbol. Each OFDM symbol is subsequently demodulated in the DFT block, using the FFT algorithm.

The complex values corresponding to each information symbol belonging to each subcarrier are then equalized by a single-tap equalizer (multiplication by a single coefficient). The equalizer is able to compensate for linear amplitude and phase distortion introduced by the channel. Training symbols are used to estimate the channel transfer function.

After equalization, the recovered symbols are demapped and converted to a serial stream of bits at the P/S block. These bits will be equal to the transmitted bits, if no transmission error occurs.

### 2.2 Optical transmission and direct detection

An optical transmitter is required to convert the electrical OFDM signal into optical form and launch the resulting optical signal into the optical fibre. Optical transmitters can use direct modulation, where the optical power at the output of the optical source is controlled by the driving current, or external modulation, where the emitted optical power is kept constant and the modulation is applied by an external device that actuates the optical beam. In this work, an optical transmitter using external modulation is employed. Directly modulated optical sources are associated with a high chirp, which combined with the dispersion of the optical fibre, severely limit the transmission distance at high bit rates.
In DD-OFDM, the transmitted signal is recovered at the receiver using direct detection. Since a carrier is sent along with the OFDM signal, only a photodiode is required at the receiver to convert the optical signal to the electrical domain. When transmitting double-sideband (DSB) signals through an optical fibre, chromatic dispersion induced power fading (CDIPF) occurs after photodetection, causing significant performance degradation. The accumulated dispersion of the link produces different phase shifts to the signal sidebands. CDIPF is subsequently caused by the beat between the two signal sidebands due to the square law characteristic of the photodetector [36]. There are two approaches to overcome this impairment: 1) to deploy optical dispersion compensators or 2) to employ SSB transmission. In this work, SSB transmission is employed because it simplifies network design, avoiding the deployment of optical dispersion compensators at each link. In addition, SSB transmission conserves bandwidth.

A block diagram representing a SSB DD-OFDM optical transmission system is shown in figure 2.5. An externally modulated continuous wave (CW) laser and a SSB filter are used to generate the optical SSB OFDM signal. The spectrum of the signal at the fibre input shows the SSB signal and the optical carrier that is used for detection at the optical receiver. After photodetection, the recovered OFDM signal then goes to the electrical OFDM receiver previously described in this chapter.

![Figure 2.5: Simplified diagram of a SSB DD-OFDM optical transmission system.](image)

### 2.2.1 Optical transmitter

The optical SSB signal can be generated by using a conventional optical modulator such as the Mach-Zehnder modulator (MZM) or the electro-absorption modulator (EAM) followed by an optical SSB filter to remove one sideband of the signal. This approach is illustrated in figure 2.6. Another option is to use a dual parallel Mach-Zehnder modulator (DP-MZM), which consists of four phase modulators in parallel [37]. The structure of the DP-MZM can also be described as a dual arm MZM. It is possible to generate a SSB signal by applying the electrically generated signal to one branch and the Hilbert transform (HT) of that signal to the other branch of the DP-MZM. This avoids the use of an optical SSB filter which presents a limitation for spectral efficiency and system performance due to its finite selectivity.
The structure of the DP-MZM is shown in figure 2.6. $v_1(t)$ and $v_2(t)$ are the electrical signals applied to the upper and lower MZM arms respectively. A third modulator in the outer structure (controlled by voltage $V_3$) introduces a $90^\circ$ phase difference between the optical fields $m_1(t)$ and $m_2(t)$ at the MZM arm outputs, which sets them in quadrature to each other.

In this study, the two inner MZMs of the DP-MZM are biased at the minimum transmission point, which generates a SSB signal with the optical carrier suppressed. As it will be shown later in this chapter, an electrically generated carrier is instead used to assist the direct detection. The optical field at the DP-MZM output is given by

$$E_o(t) = E_{in} \frac{1}{2} \left\{ \exp \left( -j \frac{\pi}{4} \right) \sin \left( \frac{\pi}{2V_\pi} v_1(t) \right) + \exp \left( j \frac{\pi}{4} \right) \sin \left[ \frac{\pi}{2V_\pi} v_2(t) \right] \right\}$$

(2.20)

where $E_{in}$ is the input electrical field of the DP-MZM (a CW signal), $V_\pi$ is the switching voltage of the MZMs and $V_{b,1}$, $V_{b,2}$ are the bias voltages of the inner MZMs. In order to bias the inner MZMs at the minimum transmission point, $V_{b,1} = V_{b,2} = V_\pi$. By setting $V_3 = V_\pi/2$, the outputs of the two inner MZMs are set in quadrature to each other. Substituting in equation 2.20 we obtain the optical field at the DP-MZM output with the inner MZMs biased at the minimum transmission point:

$$E_o(t) = E_{in} \frac{1}{2} \left\{ \exp \left( -j \frac{\pi}{4} \right) \sin \left( \frac{\pi}{2V_\pi} v_1(t) \right) + \exp \left( j \frac{\pi}{4} \right) \sin \left[ \frac{\pi}{2V_\pi} v_2(t) \right] \right\}$$

(2.21)
The electrical-optical (EO) conversion performed by the DP-MZM is a nonlinear process, as shown by equation 2.21. The EO conversion becomes more nonlinear with the increase of the voltage of the electrical signals applied to the MZM arms. The modulation index $m$ measures how the electrical signal is driven into the MZM arms and is defined as

$$m = \frac{V_{\text{rms}}}{V_{\pi}}$$

(2.22)

where $V_{\text{rms}}$ is the root mean square (RMS) voltage of the electrical signals applied to the MZM arms. In this study, $m = 5\%$ is used in most of the simulations. Optimization of the modulation index and other parameters related to the optical transmitter in MB-OFDM metro networks can be found in [38].

The SSB signal is obtained from the input DSB signal and the HT of that signal by

$$s_{\text{SSB}}(t) = s(t) \pm js_{\text{H}}(t)$$

(2.23)

where $s_{\text{H}}(t)$ is the HT of $s(t)$. The ideal HT transfer function is given by

$$H_{\text{HT}}(f) = js_{\text{gn}}(f)$$

(2.24)

where $s_{\text{gn}}(f)$ is the sign function defined as

$$s_{\text{gn}}(f) = \begin{cases} 
-1, & f < 0 \\
0, & f = 0 \\
1, & f > 0
\end{cases}$$

(2.25)

In this study, the HT is implemented using a 90° hybrid coupler (HC) with a -6 dB passband located between 1 GHz and 36 GHz. The transfer function of the HC was obtained from experimental measurements. Figure 2.7 shows the amplitude and phase response of the HC used in this work.
The HC generates a non-ideal HT of the signal, which causes an imperfect suppression of the sideband at the output of the DP-MZM. In this case, CDIPF may still be a source of performance degradation, albeit much less pronounced [38]. The passband ripple is also going to cause slight attenuation at some signal frequencies.

### 2.2.2 Optical receiver

The purpose of the optical receiver is to convert the optical signal into the electrical domain, so that the transmitted data can be recovered. The optical-electrical (OE) conversion is accomplished with the use of a photodetector.

Photodetectors are made of semiconductor materials and generate electron-hole pairs when incident photons are absorbed. When an external voltage is applied to the semiconductor these electron-hole pairs give rise to an electrical current known as photocurrent [13]. Typical photodetectors used in optical communications are the positive-intrinsic-negative (PIN) photodiode and the avalanche photodiode. In this work, only the PIN photodiode is used.

Photodetectors are characterized by their responsivity $R_\lambda$ which is defined as

$$R_\lambda = \frac{i_{\text{pin}}(t)}{p_i(t)}$$  

(2.26)
where $i_{\text{pin}}(t)$ is the photocurrent at the output of the PIN and $p_i(t)$ is the incident optical power. In this study, a responsivity of $R_\lambda = 1 \text{ A/W}$ is considered. Assuming that the optical field $e_i(t)$ at the PIN input is normalized so that its power is $|e_i(t)|^2$, the photocurrent is given by

$$i_{\text{pin}}(t) = p_i(t) = |e_i(t)|^2$$

(2.27)

From this expression comes the square-law of the photodetector mentioned before. Considering back-to-back operation (i.e. the DP-MZM is directly connected to the PIN input) and assuming the DP-MZM is a linear modulator, the optical field at the PIN input can be represented as

$$e_i(t) = A + Bs(t)$$

(2.28)

where $A$ and $B$ are constants and $s(t)$ is the OFDM signal. Substituting (2.28) in equation (2.27), the photocurrent generated after photodetection can be represented as

$$i_{\text{pin}}(t) = A^2 + 2AB \Re\{s(t)\} + B^2|s(t)|^2$$

(2.29)

where $A^2$ is a DC component corresponding to the carrier $\times$ carrier beating term, $2AB \Re\{s(t)\}$ is the carrier $\times$ signal beating term that allows to recover the transmitted OFDM signal and $B^2|s(t)|^2$ is an undesirable signal $\times$ signal beating term, also known as signal-signal beat interference (SSBI) \[39\]. The impact of the SSBI on the transmission performance can be mitigated using various techniques \[10\]-\[12\]: 1) employ a frequency gap between the carrier and the OFDM signal, 2) increase the power of the carrier in comparison to the power of the OFDM signal, 3) reconstruct and remove the SSBI from the photodetected signal at the receiver. In figure 2.8, the spectrum of the signal after photodetection is shown. A frequency gap was used so that the SSBI does not overlap with the OFDM signal.

![Figure 2.8: Spectrum of an OFDM signal after photodetection.](image-url)
In this work, the SSBI is removed after photodetection using an approach similar to the technique demonstrated in [42]. The analysis of the impact of the SSBI on the transmission performance of MB-OFDM metro networks can be found in [38].

2.3 Main features of the MB-OFDM system

The sharp rectangular shape of the OFDM spectrum allows each wavelength to carry several OFDM signals with a reduced guard band between adjacent signals. Therefore, MB-OFDM transmission therefore provides finer granularity for flexible bandwidth allocation while maintaining the same advantages of OFDM, such as high tolerance to channel dispersion. In addition, the performance requirements of DACs and ADCs are greatly relaxed when using MB-OFDM instead of just one band with the same bit rate.

The MB-OFDM concept was previously introduced in chapter [1]. Although OFDM systems have been widely studied, the implementation of MB-OFDM using direct detection is a recent concept that requires further investigation [33][43][44].

A MB-OFDM signal is composed by several individual OFDM signals located at different frequencies, hereafter designated as OFDM bands. Each OFDM band is generated by an OFDM transmitter with the architecture described in section [2.1] before EO conversion, the bands are added together electrically. Each OFDM band has a different central frequency within the optical channel, so that a guard band exists between adjacent bands to avoid interference. The same considerations expressed in section [2.1] for OFDM signals are valid for each individual OFDM band in the MB-OFDM signal.

\[ \lambda_N \]

Figure 2.9: SSB MB-OFDM signal with four OFDM bands.

An illustration of the spectrum of a MB-OFDM signal with four OFDM bands is shown in figure [2.9]. The signal is transmitted in SSB format to increase the spectral efficiency and to avoid CDIPF as explained in section [2.2]. In this study, each band composing the MB-OFDM signal occupies the same amount of bandwidth, as illustrated in figure [2.9]. Even so, the MB-OFDM system considered in this work allows to fine-tune the information rate of each OFDM band not only by choosing the bandwidth occupied by the band, but also by choosing the modulation...
order used in its different subcarriers. This capability provides additional flexibility and is useful to adjust the signal to the physical characteristics of the channel. The downside is the increase of system cost, which is a concern in metro networks. For this reason and also for the sake of simplicity in the analysis, this work will assume that each OFDM band carries the same amount of information. In chapter 3, the description of the architecture of the MB-OFDM metro network nodes provides a better understanding of these issues.

The bit rate of the MB-OFDM signal is given by

\[ R_{b,\text{MB-OFDM}} = \sum_{n=1}^{N_{\text{bands}}} R_{b,n} \]  

(2.30)

where \( R_{b,n} \) is the bit rate of the \( n \)-th OFDM band in the MB-OFDM signal, which is given by equation 2.16. If the bands were generated using the same OFDM signal parameters, equation 2.30 becomes

\[ R_{b,\text{MB-OFDM}} = N_{\text{bands}} R_b \]  

(2.31)

where \( R_b \) is the bit rate of one OFDM band. From equations 2.16, 2.30, 2.31 and figure 2.9, it can be seen that the target bit rate of the MB-OFDM signal can be attained by the interplay between the number of bands, the bandwidth occupied by each band and the modulation used in the subcarriers.

2.3.1 Virtual carrier-assisted transmission

After photodetection of the MB-OFDM signal with the spectrum illustrated in figure 2.9, several beating components will overlap the signal. For this reason, a frequency gap with width at least as large as the MB-OFDM signal bandwidth is necessary to avoid the SSBI \[39\]. This situation requires the photodetector to have a very large bandwidth (at least two times the bandwidth of the MB-OFDM signal). Furthermore, only the bands that are assigned to the node should be photodetected, as the others are assigned to different nodes. A better solution is to detect each OFDM band separately.

![Figure 2.10: Use of a dual band optical filter to select the carrier and the OFDM band.](image-url)
Figure 2.10 shows how a dual band optical filter can be employed to select the carrier and a specific OFDM band before photodetection (the dashed line represents the ideal filter response). This implementation leads to a significantly better spectral efficiency and reduces the required bandwidth of the photodetector. Although dual band optical filters are available, they are very expensive and it would be difficult to achieve the specifications (passband of a few GHz, high selectivity, frequency adjustable) needed for this end.

The solution adopted in this work is one based on virtual carriers [33]. As mentioned before, the optical carrier is suppressed in the DP-MZM and an electrically generated carrier (virtual carrier) is used instead to assist the direct detection. In order to achieve a low requirement for the photodetector bandwidth and to avoid the use of dual band filters, a virtual carrier is placed close to each band.

![Figure 2.11: MB-OFDM signal with virtual carriers to assist the direct detection.](image)

Figure 2.11 illustrates how an individual OFDM band is selected from a MB-OFDM signal employing virtual carriers to assist the direct detection. The optical filter used to select a band or a set of bands is hereafter designated as band-selector (BS). The BS is detailed later in this chapter. It should be noted, that in this solution, the frequency gap width that accommodates the SSBI corresponds to the width of the gap between the OFDM band and its virtual carrier. Since the virtual carrier is placed closed to the band to improve spectral efficiency, the SSBI will necessarily overlap the signal after photodetection. Therefore, it is necessary to employ a technique to mitigate the SSBI, as explained before in the description of the optical receiver.

In figure 2.12, the virtual carrier-assisted MB-OFDM optical transmission system is represented in block diagram form. The OFDM bands and virtual carriers are generated with different central frequencies within the wavelength and added electrically. At the output of the DP-MZM, a SSB MB-OFDM optical signal is obtained. At the receiver, the BS selects the band and the virtual carrier. After photodetection, a DC block is employed to remove the DC component of the signal. In the simplified diagram shown in figure 2.12, it is assumed that all bands composing the MB-OFDM signal are generated and inserted at the same node.
2.3.2 Band-selector

The BS is an optical filter that selects the OFDM band and its virtual carrier from the MB-OFDM signal at the receiver input. The bandwidth and the central frequency of the BS can be fixed or tunable. More insight into this aspect is given in chapter 3 when the architecture of the nodes of the MB-OFDM network is described.

Figure 2.13: Amplitude response, in dB, of the Gaussian and 2nd order super-Gaussian BS.

Figure 2.13 shows the amplitude response of the two BS filters used in this study: Gaussian and 2nd order super-Gaussian (SG-BS). Both have a -3dB bandwidth of 3.6 GHz in figure 2.13. The frequency scale is shifted so that, zero frequency in the plot corresponds to the central frequency of the BS. It can be seen that the 2nd order SG-BS has a higher selectivity than the Gaussian BS. The impact of the detuning of the central frequency of the BS relative to the central frequency of the selected OFDM band is analysed in [45].

2.3.3 Virtual carrier-to-band gap

The virtual carrier to-band-gap (VBG) concept is highlighted in figure 2.14. The VBG is defined as the width of the frequency gap between the OFDM band and its corresponding virtual carrier.
As mentioned before, if the VBG is smaller than the bandwidth occupied by the OFDM band, the SSBI spectrum will overlap the signal spectrum after photodetection. On the other hand, in order to maximize spectral efficiency, the VBG should be as small as possible. Since a technique to remove the SSBI is employed in this work, a VBG of only 20 MHz is used.

2.3.4 Band gap

The band gap (BG) is defined as the width of the frequency gap between two adjacent bands composing the MB-OFDM signal. The BG is highlighted in figure 2.15.

An appropriate BG is necessary due to the finite selectivity of the BS. It has been shown that if \( \text{BG} \geq B + 2 \text{VBG} \) the crosstalk between neighboring bands (inter-band crosstalk) does not overlap the signal after photodetection, using a Gaussian BS [38]. \( B \) is the bandwidth occupied by the OFDM band. Since the 2nd order super-Gaussian BS is more selective, a smaller BG can be used in that case to obtain similar transmission performance. The optimization of the BG is out of the scope of this dissertation, and the optimized values for the BG presented in [38] are used instead.

After establishing the VBG and BG definitions, the total bandwidth of the MB-OFDM signal is given by

\[
B_{\text{MB-OFDM}} = \sum_{n=1}^{N_{\text{bands}}} B_n + \sum_{m=1}^{N_{\text{bands}}-1} \text{BG}_m + \text{VBG} \tag{2.32}
\]

where \( B_n \) is the bandwidth occupied by the \( n \)-th band and \( \text{BG}_m \) is the BG between the \( m \)-th
band and the \((m + 1)\)-th band.

If all bands of the MB-OFDM signal occupy the same amount of bandwidth and the BG is kept the same between each two adjacent bands, equation 2.32 becomes

\[
B_{\text{MB-OFDM}} = N_{\text{bands}}B + (N_{\text{bands}} - 1)BG + VBG
\]  

(2.33)

where \(B\) is the bandwidth occupied by one OFDM band.

### 2.3.5 Virtual carrier-to-band power ratio

The virtual carrier-to-band power ratio (VBPR) is defined as the ratio between the power of the virtual carrier and the power of the corresponding OFDM band. The VBPR in dB is given by

\[
\text{VBPR} = 10 \log_{10} \left( \frac{p_{\text{vc}}}{p_{\text{band}}} \right)
\]

(2.34)

where \(p_{\text{band}}\) is the power of the band and \(p_{\text{vc}}\) is the power of its virtual carrier.

A high VBPR allows to reduce the performance degradation caused by the SSBI. If the power of the virtual carrier is increased in comparison to the power of the OFDM band, the impact of the SSBI on the transmission performance is diminished. From equation 2.29 if

\[
2AB \text{Re}\{s(t)\} \gg B^2|s(t)|^2
\]

(2.35)

the OFDM signal is recovered after photodetection without significant performance degradation caused by the SSBI. The downside of using a high VBPR is that, for the same average optical power, less optical power is allocated to the OFDM band, resulting in a degradation of the system sensitivity.

### 2.4 Conclusions

In this chapter, the virtual carrier-assisted MB-OFDM system was described. In the first section, the fundamentals of OFDM signals were discussed, including the mathematical description, DFT implementation and spectral efficiency. The use of a guard interval and a cyclic prefix were identified as being able to prevent the intersymbol interference caused by the fibre-induced chromatic dispersion. The architectures of the OFDM transmitter and receiver were presented and their main functionalities were highlighted.

Afterwards, the processes of optical transmission and direct detection were described. The necessity of employing SSB transmission due to the CDIPF was recognized. The model of the
DP-MZM, which is able to generate a SSB optical signal, was described. Since an electrical carrier is used to assist the direct detection, the inner MZMs of the DP-MZM are biased at the minimum transmission point, so that the optical carrier is suppressed. The beating components that appear after photodetection at the receiver were identified and the SSBI was recognized as a limiting factor of the transmission performance.

Following the description of the optical components, the main feature of the MB-OFDM optical system were introduced. The use of virtual carriers placed close to each OFDM band to assist the direct detection was identified as a distinctive feature of the system. Although the MB-OFDM system concept offers concrete advantages, it remains to be seen how it performs in metro networks where the concatenation of ROADMs will affect the virtual carriers and the bands of the MB-OFDM signal.
Chapter 3

MB-OFDM metro network: operation principles and modelling

In this chapter, the MB-OFDM metro network is characterized and the operation principles of MB-OFDM signals transmission along metro networks are described. To evaluate the transmission performance of the network, an appropriate model of the metro network nodes is described. These nodes include ROADMs based on WSSs, as identified in chapter 1.

In section 3.1, the structure of the MB-OFDM metro network is summarized. The nodes are described in section 3.2 as well as their internal structure. The processes of inserting and extracting an OFDM band are outlined in section 3.3. The analytical model of the WSS, which is the enabling switching technology of the ROADMs, is detailed in section 3.4. The dependence of the WSS frequency response on the model parameters is studied in section 3.5. Section 3.6 presents the main achievements and conclusions of the chapter.

3.1 Structure of the MB-OFDM metro network

The MB-OFDM metro network has a ring topology, as stated in chapter 1. In each node of the network, a ROADM is able to add and drop WDM channels and to equalize the optical power among the wavelengths. The ability of ROADMs to remotely change the add/drop traffic pattern across the network is particularly valuable in DWDM networks, where a large number of WDM channels exists and the bandwidth demand is unpredictable [23].

In figure 3.1, a simplified diagram of the MB-OFDM metro network is shown. In each node, besides the ROADM, two additional types of blocks are needed to insert and extract individual OFDM bands. These blocks are known as MORFEUS insertion block (MIB) and MORFEUS extraction block (MEB), and are unique to the MB-OFDM metro network.
The nodes of the MB-OFDM metro network are transparent, i.e., there is no OEO conversion at the network nodes. Thus, WDM channel switching and OFDM band switching are made entirely in the optical domain. The MB-OFDM network allows increasing the switching granularity: the minimum granularity is the capacity of the WDM channels and switching is accomplished at the ROADMs; the maximum granularity is the capacity of one OFDM band and switching is performed at the MIB and MEB. The internal structures of the ROADM and the MEB are described in the following section.

As it is shown in figure 3.1, optical amplifiers (OA) are employed before and after each node. These are erbium-doped fibre amplifiers (EDFA). The EDFA employed at the input of the node is used to compensate for link losses. Since the ROADM and other blocks in the node introduce substantial attenuation of the signal, a second EDFA is required at the output of the node to compensate for this loss. Alternatively, the network can operate with only one EDFA preceding each node to compensate for both the link losses and the node losses. The solution with two EDFAs is preferable, as it yields the best transmission performance, although it is has a higher cost [13]. The EDFAs introduce amplified spontaneous emission (ASE) noise, which has to be taken into account. The optical signal-to-noise-ratio (OSNR) will gradually degrade along the chain of nodes, causing the bit error ratio (BER) to deteriorate.
3.2 Structure of the MB-OFDM metro network nodes

In chapter 1, a distinction was made between two-degree and multi-degree nodes. Both are represented in the network shown in figure 3.1. Two-degree nodes are used solely to add and drop WDM channels and, in the case of the MB-OFDM metro network, insert and extract OFDM bands. On the other hand, multi-degree nodes can also be used to interconnect adjacent rings. In any case, when traversing a node, a MB-OFDM signal will generally go through two blocks in the pass-through path of the ROADM: a demultiplexer (demux) block, which allows the wavelength drop function, and a multiplexer (mux) block which allows the wavelength add function. The majority of deployed ROADMs are two degree nodes [19]. For this reason, the analysis on this section is focused on two-degree nodes.

Figure 3.2: Schematic diagram of a two-degree MB-OFDM node employing two WSSs.

In figure 3.2, the schematic diagram of a possible two-degree node to be used in the MB-OFDM network is shown. Two WSSs are used to perform the demux and mux functions. The WSS provides colorless switching of the WDM channels, i.e., any wavelength at the input of a WSS can be switched to any output port.

If 1-to-N WSSs are used, such as in the architecture represented in figure 3.2, the network supports at most N-1 WDM channel switching. Any wavelength coming from the input fibre can be switched to any output port of the demux WSS, which includes the drop ports and the forwarding port to the mux WSS.

The MIB and MEB are responsible for providing network granularity at the sub-wavelength level. The connection between the MEB and the MIB is necessary for one reason: to forward the remaining MB-OFDM signal after a band is extracted and an empty slot is created in that band location to allow the insertion of a band in the MIB. The processes of extracting and inserting an OFDM band will be explained in section 3.3.
Figure 3.3: Schematic diagram of a two-degree MB-OFDM node employing a splitter and a WSS.

Figure 3.3 shows the architecture of a two-degree MB-OFDM node employing only one WSS in the pass-through path. The demux function is performed by a passive splitter and a demultiplexer. The mux function is performed by a WSS. The pass-through path is taken by a MB-OFDM signal that is passing through the node, i.e., it does not undergo add/drop actions. In figure 3.2, the pass-through path is the horizontal line that intersects the ROADM. It should be clarified that each WDM channel carries one MB-OFDM signal.

As it is depicted in figure 3.3, the optical signal coming from the input fibre is split into two copies. One copy of the signal continues to the mux WSS and the other copy goes to the 1:N demux where all the WDM channels are available. The 1:N demux can either be an array waveguide grating (AWG) or a WSS [23]. The AWG has a fixed wavelength-port mapping; on the other hand, the WSS has colorless ports. The output ports of the 1:N demux are connected to the MEBs.

The ROADM architecture depicted in figure 3.3 is able to balance the flexibility of the WSS and the amount of spectral filtering performed at the node [19] [23]. In comparison, the ROADM architecture shown in figure 3.2 is most suitable for multi-degree nodes, since the loss of the splitter increases with the degree of the node [46]. On the contrary, the loss of the WSS has a fixed value [28] [29]. The disadvantage of this architecture is that it performs two filtering actions in the pass-through path due to the two WSSs. Therefore, to reduce the penalties arising from passband narrowing due to successive filtering actions, the architecture shown in figure 3.3 is preferable for nodes with degree less than four [23] [47].

There is another possible ROADM architecture, which replaces the passive splitter with a WSS, but additionally replaces the mux WSS with a passive combiner. This architecture is not a good choice for applications with a large number of channels due to the high insertion loss of the passive combiner [23] [47]. Furthermore, this architecture does not provide power level equalization.
of the add WDM channels, which leads to degradation of the transmission performance [23]. On the contrary, the use of a mux WSS allows the power of each WDM channel exiting the node to be individually attenuated. For these reasons, the architecture with a passive combiner is not adequate for the MB-OFDM metro network.

Commercially available WSS-based ROADMs support up to 96 controllable optical channels with 50 GHz spacing, but only have up to $1 \times 20$ port configuration [28][29]. Hence, if a 20:1 mux WSS is used in the architecture shown in figure 3.3 the node supports the simultaneous add of at most 19 WDM channels. It should be stressed that if a band is extracted from a MB-OFDM signal in a node, the WDM channel that carries that MB-OFDM signal needs to be added again in the same node. To address the limitation due to the small number of ports of the WSS, additional muxes are usually employed [29].

![Figure 3.4: Schematic diagram of a MB-OFDM node employing an additional mux to create extra add ports.](image)

In figure 3.4 a MB-OFDM node employing an additional mux is shown. In this configuration, the mux WSS has extra expansion ports that can be used to increase the degree of the node. Analogous to the 1:N demux used to drop WDM channels, the N:1 mux can be implemented with an AWG or a WSS.

The internal structure of the MEB is depicted in figure 3.5 [33]. It includes a splitter, the band-selector (BS) and a band-blocker (BB). The BS is an optical filter, previously described in chapter 2. The bandwidth and the central frequency of the BS can be fixed or tunable. The BB is used to remove the extracted OFDM band from the MB-OFDM signal that is being transmitted along the ring. The MIB includes a DP-MZM among other components not studied in this dissertation [33]. The number of required MEBs and MIBs in a given node depends on the
number of simultaneous insertions and extractions of OFDM bands that need to be performed at that node.

3.3 Extraction and insertion of OFDM bands

In this section, the operations for extraction and insertion of OFDM bands are described. These operations can be implemented with any one of the ROADM architectures described in section 3.2. The section that follows details the analytical model of the WSS, which is the device that introduces spectral filtering in the ROADM.

The operations for insertion and extraction of OFDM bands are summarized in the following steps:

1. The WDM channels entering the node, each carrying a MB-OFDM signal, are demultiplexed by the ROADM,

2. If at least one OFDM band carried by a WDM channel is assigned to be dropped, that WDM channel is switched to its respective MEB. The other WDM channels that do not have drop assignments follow the pass-through path of the ROADM,

3. In the MEB, the MB-OFDM signal is split into two branches. In one branch, the band-selector selects a band assigned to be dropped. That band is then photodetected and delivered to the client or routed to the access network,

4. In the other branch of the MEB, the band-blocker removes the extracted band by the BS from the MB-OFDM signal,
5. The MB-OFDM signal with the band removed is sent to the MIB. In the MIB, a new band can be inserted in an empty band slot of the MB-OFDM signal. This band is provided by the client in the electrical domain or is routed from the access network.

6. In the ROADM, the WDM channels coming from the MEBs and the ones that were passed-through are multiplexed and exit the node.

### 3.4 Analytical model of the WSS

The WSS was introduced in chapter 1. It is the most widely used optical switch in ROADMs, and it is also the ROADM component that introduces spectral filtering [23]. As such, it imposes a limitation on the number of nodes that can be traversed, before the MB-OFDM signal becomes excessively frequency-distorted.

![Schematic diagram of a generic WSS](image)

**Figure 3.6**: Schematic diagram of a generic WSS, adapted from [48].

In figure 3.6, the schematic diagram of a generic WSS is shown. In most current WSS designs, the linear separation between input/output fibres is mapped to an angular separation [25]. A diffraction grating provides spatial dispersion to separate the wavelengths. The diffracted light is then focused onto the switching engine, and the switching element redirects the input beam to the selected output port. The switching engine can be implemented using different technologies, the main ones being MEMS and LCOS, as mentioned in chapter 1. In MEMS-based WSSs, the switching elements are individual micromirrors and, in LCOS-based WSSs, the switching
elements are clusters of LCOS pixels [23]. Additional optical components, such as polarisation
diversity optics and anamorphic optics are used depending on the switching engine. Nevertheless,
the basic principle of operation is the same regardless the choice of switching engine.

When evaluating network performance, super-gaussian functions have been used to model the
WSS channel shape [27][49]. However, those functions are not representative of the underlying
physical process and do not predict an accurate passband shape. A recent model based on the
physical operation of the device has been proposed [50]. It offers an accurate match to measured
results and can be generalized for different switching technologies.

This dissertation aims at evaluating not only the impact of bandwidth narrowing due to
cascaded ROADMs, but also the penalty due to the group delay distortion originated by the
WSS. The model proposed in [50] only provides the amplitude response of the WSS. Therefore,
an extension of the model is developed to predict the phase response.

3.4.1 Derivation of the frequency response of the WSS

The frequency response of a WSS channel is modelled as a bandpass filter, created by an aperture
at the image plane of the spatially diffracted light [50]. The aperture function, \( R(f) \), is defined
in frequency space, as

\[
R(f) = \begin{cases}
\exp(j \cdot \phi(f)), & -B/2 \leq f \leq B/2 \\
0, & \text{otherwise}
\end{cases}
\]  

(3.1)

where \( B \) is the width of the rectangular aperture in frequency, i. e., the width of the switch-
ing element. \( B \) takes the value of the channel spacing (e. g. for a channel spacing of 50 GHz,
\( B = 50 \) GHz). The channel edges are usually aligned to the 50 GHz or 100 GHz ITU grids. In
comparison with [50], the phase response \( \phi(f) \) of the switching element was incorporated in \( R(f) \).
By adequately selecting this phase response, the group delay of the WSS can match experimental
data, as it will be shown later in subsection 3.4.3.

It should be stressed that \( f \) is a distance measured in the dispersion axis of the switching
element. It is related to the spatial dispersion of the grating by

\[
f = \frac{df}{dx} x
\]

(3.2)

where \( df/dx \) is the inverse of the spatial dispersion of the grating, in Hertz per meters. \( x \) is the
distance measured in the dispersion axis of the switching element, in meters. In figure 3.7 the
dispersion axis is shown next to the aperture plane. Each beam corresponds to a wavelength
and is focused onto its respective switching element. A theoretical description of the diffraction
Analytical model of the WSS

grating and other optics involved is provided in [51].

![Diagram of a LCOS-based WSS](image)

Figure 3.7: Schematic diagram of a LCOS-based WSS, showing the spatially dispersed light focused onto the aperture plane, adapted from [52].

The frequency response of a bandpass filter created by the WSS is predicted to be the convolution of the aperture function $R(f)$ with the optical transfer function (OTF) of the device $L(f)$ [50]. The inputs of the WSS are single-mode optical fibres with a Gaussian profile, hence the image on the switching element is assumed to be Gaussian [50]. In figure 3.7, the focused beams have elliptical Gaussian profiles, instead of circular ones, due to the use of anamorphic optics [51].

$L(f)$ is a normalized Gaussian function given by

$$L(f) = \exp\left(-\frac{f^2}{2\sigma^2}\right),$$

where $\sigma$ is the standard deviation of the Gaussian OTF and is related to the -3 dB bandwidth by

$$\sigma = \frac{B_{\text{OTF}}}{2\sqrt{2\log(2)}}$$

where $B_{\text{OTF}}$ is assumed constant over the frequency range of the device [50]. $B_{\text{OTF}}$ represents how the beam has been focused onto the aperture plane. A smaller $B_{\text{OTF}}$ indicates that the spot size in the switching element is smaller.
The lowpass equivalent of the frequency response of a bandpass filter created by the WSS is given by [50]

\[ S(f) = \int_{-B/2}^{B/2} R(f')L(f' - f)df'. \]  

(3.5)

When \( \phi = 0 \), the result is simply

\[ S(f) = \frac{1}{2} \sigma \sqrt{2\pi} \left[ \text{erf}\left(\frac{B/2 - f}{\sqrt{2}\sigma}\right) - \text{erf}\left(\frac{-B/2 - f}{\sqrt{2}\sigma}\right) \right], \]  

(3.6)

where \( \text{erf}(x) \) is the error function. Equation 3.6 is the expression given in [50] to predict the amplitude response of the device. The parameter \( \sigma \) can be adjusted to accurately match measured results [50].

![Figure 3.8: Amplitude response of a WSS (continuous line) and a Gaussian filter (dashed line), with -3 dB bandwidth of 45 GHz.](image)

In figure 3.8 a typical amplitude response of a WSS for a channel spacing of 50 GHz is shown. It can be observed that this WSS response has high selectivity and a flat passband, when compared with the response of a conventional Gaussian-shaped filter.

### 3.4.2 Group delay originated by the WSS

The group delay originated by the WSS depends on the switching technology. When evaluating the ROADM cascade penalty, the phase response of the device is usually neglected [53]. The explanation for this assumption is that the group delay is assumed almost constant inside the passband. Even considering small group delay ripples (GDR), the ripple variance is reduced as the number of cascaded devices increases. Despite this assumption, MEMS-based devices are known to be susceptible to amplitude and phase effects due to the curvature of the micro-mirrors [54].
The group delay of a bandpass filter created by the WSS is defined as

\[ \tau_g(f) = -\frac{1}{2\pi} \frac{d\phi_{\text{out}}(f)}{df} \text{[ps]}, \]  

where \( \phi_{\text{out}}(f) \) is the phase response of the device. It should be stressed that \( \phi_{\text{out}}(f) \) is not the same as \( \phi(f) \) in equation 3.1, which is the phase response introduced by the switching element at the aperture plane.

A parabolic and cubic phase responses of the switching element are considered in this analysis. A parabolic phase response generates a group delay at the WSS with constant non-zero slope in the passband. This occurs when the micromirror has an uniform radius of curvature. This effect has been observed in several experiments with MEMS-based devices [55]-[57]. A phase response with a higher-order dependence on \( f \) may occur if the micromirror has a non-uniform radius of curvature. A cubic phase response means that the radius of curvature has a linear dependence on frequency. Since the width of the micromirror is defined in the frequency space (equation 3.1), as the beam moves away from the centre of the micromirror, it experiences a different radius of curvature. A group delay, corresponding to a phase response of the switching element which can be approximated by a cubic phase response, can be observed in a measurement shown in a JDSU white paper [22].

In general, the phase response \( \phi(f) \) is then given by

\[ \phi(f) = -(\xi f^3 + \zeta f^2), \]  

where \( \xi \text{[s]}^3 \) and \( \zeta \text{[s]}^2 \) are constants that control the amount of group delay originated by the cubic and parabolic phase responses of the switching element, respectively.

Calculating the integral from eq. 3.5 including only a parabolic phase response (\( \xi = 0, \zeta \neq 0 \)), we obtain the following response:

\[ S(f) = -\frac{\sqrt{j} \exp \left( \frac{f^2\zeta}{j - 2\sigma^2\zeta} \right)}{\sqrt{-2j + 4\sigma^2\zeta}} \sqrt{\pi \sigma} \left[ \text{erfi} \left( \frac{\frac{1}{4} + \frac{j}{4}}{\frac{B + 2f + 2jB\sigma^2\zeta}{\sigma \sqrt{-j + 2\sigma^2\zeta}}} \right) \right] \]

\[ - \text{erfi} \left( \frac{\sqrt{j} \left( f + B \left( -\frac{1}{2} - j\sigma^2\zeta \right) \right)}{\sigma \sqrt{-2j + 4\sigma^2\zeta}} \right), \quad (3.9) \]
where \( \text{erfi} \) is the imaginary error function, defined as

\[
\text{erfi}(z) = -j \frac{1}{\sqrt{\pi}} e^{-z^2}
\]

(3.10)

If we set \( \zeta = 0 \) in eq. 3.9, the following expression is obtained:

\[
S(f) = \left(\frac{1}{4} + \frac{j}{4}\right) \sqrt{j} \sqrt{\pi} \sigma \left[ \text{erfi} \left( \frac{\left(\frac{1}{4} + \frac{j}{4}\right) \sqrt{j}}{\sigma} (-2B + 4f) \right) \right. \\
- \left. \text{erfi} \left( \frac{\left(\frac{1}{4} + \frac{j}{4}\right) \sqrt{j}}{\sigma} (2B + 4f) \right) \right]
\]

(3.11)

and using the definition of the imaginary error function (eq. 3.10) we get:

\[
S(f) = -j \left(\frac{1}{4} + \frac{j}{4}\right) \sqrt{j} \sqrt{\pi} \sigma \left[ \text{erf} \left( \frac{j \left(\frac{1}{4} + \frac{j}{4}\right) \sqrt{j}}{\sigma} (-2B + 4f) \right) \right. \\
- \left. \text{erf} \left( \frac{j \left(\frac{1}{4} + \frac{j}{4}\right) \sqrt{j}}{\sigma} (2B + 4f) \right) \right]
\]

(3.12)

Using de Moivre’s formula \( \sqrt{j} = \frac{1}{\sqrt{2}} - \frac{j}{\sqrt{2}} \), and so \( \left(\frac{1}{4} + \frac{j}{4}\right) \sqrt{j} = \frac{1}{2\sqrt{2}} \). Substituting the equality in eq. 3.12 yields

\[
S(f) = \frac{1}{2\sigma \sqrt{2\pi}} \left[ \text{erf} \left( \frac{B/2 - f}{\sqrt{2\sigma}} \right) - \text{erf} \left( \frac{-B/2 - f}{\sqrt{2\sigma}} \right) \right],
\]

(3.13)

which is the same result given in equation 3.6. Therefore, equation 3.9 is a general case of equation 3.6. Calculating the integral including only a cubic phase response (\( \xi \neq 0, \zeta = 0 \)), a closed-form expression is not easy to obtain, but the integral can be evaluated numerically.
Figure 3.9 shows the group delay introduced by a WSS, obtained by calculating the integral in equation 3.5 for (a) a parabolic phase response in $R(f)$ and (b) a cubic phase response in $R(f)$. A shift in frequency was performed so that the central frequency of the bandpass filter created by the WSS coincides with the central frequency of a MB-OFDM signal in a 25 GHz channel. Inside the passband, a group delay with (a) a linear shape and (b) a parabolic shape is observed. $\tau_{g,p-p}$ is defined as the peak-to-peak group delay, when a parabolic phase response is introduced in $R(f)$. Likewise, $\tau_{g,p}$ is defined as the peak group delay, when a cubic phase response is introduced in $R(f)$.

### 3.4.3 Matching the predicted group delay to experimental data

Figure 3.10 shows the group delay obtained by calculating the integral in equation 3.5 for a parabolic phase response in $R(f)$ and performing a shift in frequency in order for the central frequency of the bandpass filter created by the WSS to coincide with the central frequency of a channel measured in [22]. The $\xi$ parameter was adjusted so that a 7 ps peak group delay is obtained.
Figure 3.10: Group delay predicted by the extension of the WSS model, used to match the measured group delay.

Figure 3.11: Matching of the parabolic group delay obtained by the extension of the WSS model and the measured group delay [22].

Figure 3.11 shows the measurement of the insertion loss and group delay for an odd/even channel pattern [22]. It can be seen that the group delay predicted by the model matches the measured group delay.
3.5 Dependence of the predicted WSS amplitude response and group delay on the model parameters

The selectivity of the WSS is dependent not only on the $B_{\text{OTF}}$ alone, but also on the relation between $B$ and $B_{\text{OTF}}$. This statement is valid because the band-shape reflects on how large the diameter of the Gaussian beam is compared to the width of the switching element. In equation 3.13, it can be noticed that the error function imposing band limitation depends on $B/\sigma$ and consequently on $B/B_{\text{OTF}}$.

![Graph](a)

![Graph](b)

Figure 3.12: WSS selectivity dependence on $B/B_{\text{OTF}}$ (a), WSS selectivity dependence on $B_{\text{OTF}}$, with $B = 25$ GHz (b).

Figure 3.12 shows the selectivity dependence of the amplitude response of the WSS on $B/B_{\text{OTF}}$. As the ratio $B/B_{\text{OTF}}$ grows, the WSS amplitude response becomes more selective, showing that a larger portion of the Gaussian beam is confined in the switching element.

The definition of selectivity of a filter used in this dissertation is given by

$$S = \frac{B_{-3\text{dB}}}{B_{-20\text{dB}}}$$

(3.14)

where $B_{-3\text{dB}}$ is the -3 dB bandwidth and $B_{-20\text{dB}}$ is the -20 dB bandwidth of the filter.
Figure 3.13: Amplitude response of the WSS for different $B/B_{OTF}$ ratios, with $B = 25$ GHz.

Table 3.1: Selectivity and bandwidth (BW) dependence on $B_{OTF}$, with $B = 25$ GHz.

<table>
<thead>
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<th>$B_{OTF}$ [GHz]</th>
<th>Selectivity</th>
<th>-0.5dB BW [GHz]</th>
<th>-3dB BW [GHz]</th>
<th>-20dB BW [GHz]</th>
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</thead>
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<tr>
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<td>0.81</td>
<td>20.3</td>
<td>23.4</td>
<td>28.8</td>
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<td>0.77</td>
<td>18.9</td>
<td>22.9</td>
<td>29.9</td>
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<tr>
<td>5.5</td>
<td>0.72</td>
<td>17.6</td>
<td>22.4</td>
<td>31.0</td>
</tr>
<tr>
<td>6.5</td>
<td>0.69</td>
<td>16.2</td>
<td>22.0</td>
<td>32.1</td>
</tr>
<tr>
<td>7.5</td>
<td>0.65</td>
<td>14.9</td>
<td>21.5</td>
<td>33.2</td>
</tr>
<tr>
<td>8.5</td>
<td>0.61</td>
<td>13.6</td>
<td>21.1</td>
<td>34.3</td>
</tr>
</tbody>
</table>

Figure 3.13 shows the amplitude response of the WSS for different $B/B_{OTF}$ ratios. Table 3.1 shows the selectivity and -0.5 dB, -3 dB and -20 dB bandwidth dependence on $B_{OTF}$, with $B = 25$ GHz. It is evident that a smaller $B_{OTF}$ is desired to obtain a higher selectivity. Unfortunately, the spot size is usually limited either by the switching technology or by the optical components of the WSS [58].

In [50], two devices were reported, one with $B_{OTF} = 10.4$ GHz and another with $B_{OTF} = 11.1$ GHz. This corresponds to $B/B_{OTF} \approx 5$. In this analysis, $B_{OTF} = 10.4$ GHz will be considered for 50 GHz channel spacing. To maintain the same selectivity, $B_{OTF} = 5.2$ GHz will be used for 25 GHz channel spacing. The impact of bandwidth narrowing is predicted to be severe after a few cascaded ROADMs, so the smaller $B_{OTF}$ of the two was chosen.

In figure 3.12 (a), for $B/B_{OTF} < 2$, the dependence of the selectivity on $B/B_{OTF}$ exhibits a steeper slope variation. This occurs when the -3 dB bandwidth stops increasing as $B/B_{OTF}$ decreases. In this region, the only contributing factor to the decrease of the selectivity is the increase of the -20 dB bandwidth.
The following analysis will determine the range of values attributed to $\xi$ and $\zeta$, that introduces the predicted group delay.

Table 3.2: Group delay (peak-to-peak), selectivity and bandwidth dependence on $\zeta$, with $B = 25$ GHz and $B_{\text{OTF}} = 5.2$ GHz.

<table>
<thead>
<tr>
<th>$\zeta$ [$s^{-2}$]</th>
<th>GD (p-p) [ps]</th>
<th>Selectivity</th>
<th>-0.5dB BW [GHz]</th>
<th>-3dB [GHz]</th>
<th>-20dB BW [GHz]</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>0</td>
<td>0.74</td>
<td>18.0</td>
<td>22.6</td>
<td>30.7</td>
</tr>
<tr>
<td>$3.0 \times 10^{-22}$</td>
<td>1.31</td>
<td>0.74</td>
<td>18.0</td>
<td>22.6</td>
<td>30.7</td>
</tr>
<tr>
<td>$6.0 \times 10^{-22}$</td>
<td>2.62</td>
<td>0.74</td>
<td>18.0</td>
<td>22.6</td>
<td>30.7</td>
</tr>
<tr>
<td>$9.0 \times 10^{-22}$</td>
<td>3.93</td>
<td>0.74</td>
<td>18.0</td>
<td>22.6</td>
<td>30.7</td>
</tr>
<tr>
<td>$1.20 \times 10^{-21}$</td>
<td>5.24</td>
<td>0.74</td>
<td>18.0</td>
<td>22.6</td>
<td>30.7</td>
</tr>
<tr>
<td>$1.50 \times 10^{-21}$</td>
<td>6.55</td>
<td>0.74</td>
<td>17.9</td>
<td>22.6</td>
<td>30.7</td>
</tr>
<tr>
<td>$1.80 \times 10^{-21}$</td>
<td>7.86</td>
<td>0.74</td>
<td>17.9</td>
<td>22.6</td>
<td>30.7</td>
</tr>
<tr>
<td>$2.10 \times 10^{-21}$</td>
<td>9.16</td>
<td>0.74</td>
<td>17.9</td>
<td>22.6</td>
<td>30.7</td>
</tr>
<tr>
<td>$2.40 \times 10^{-21}$</td>
<td>10.48</td>
<td>0.74</td>
<td>17.9</td>
<td>22.6</td>
<td>30.7</td>
</tr>
<tr>
<td>$2.70 \times 10^{-21}$</td>
<td>11.80</td>
<td>0.74</td>
<td>17.8</td>
<td>22.6</td>
<td>30.7</td>
</tr>
</tbody>
</table>

![Figure 3.14: Group delay (peak-to-peak) dependence on $\zeta$, with $B = 25$ GHz and $B_{\text{OTF}} = 5.2$ GHz.](image)

Table 3.2 and figure 3.14 show that the relation between the output peak-to-peak (p-p) group delay (GD) induced by the second-order phase distortion, and the parameter $\zeta$, is approximately linear when the p-p group delay reaches a few picoseconds. This relation can be written as

$$\zeta \approx 2.292 \times 10^{-10} \cdot \tau_{\text{g,p-p}} \ [s^2], \quad (3.15)$$

where $\tau_{\text{g,p-p}}$ is the p-p group delay. Equation 3.15 is valid only for a channel spacing of 25 GHz, but relations can be found for other values of $B$.

In table 3.2 it is shown that the passband is slightly reduced as the group delay increases.
This effect is in qualitative agreement with [57] and can be attributed to the curvature of the micro-mirror.

This analysis was made for up to a few picoseconds because this is the typical peak group delay measured in commercial devices [22].

Table 3.3: Group delay (peak), selectivity and bandwidth dependence on $\xi$, with $B = 25$ GHz and $B_{OTF} = 5.2$ GHz.

<table>
<thead>
<tr>
<th>$\xi$ [$s^{-3}$]</th>
<th>GD (peak) [ps]</th>
<th>Selectivity</th>
<th>-0.5dB BW [GHz]</th>
<th>-3dB [GHz]</th>
<th>-20dB BW [GHz]</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>0</td>
<td>0.74</td>
<td>18.0</td>
<td>22.6</td>
<td>30.7</td>
</tr>
<tr>
<td>$5.0 \times 10^{-32}$</td>
<td>1.40</td>
<td>0.74</td>
<td>18.0</td>
<td>22.6</td>
<td>30.7</td>
</tr>
<tr>
<td>$1.0 \times 10^{-31}$</td>
<td>2.80</td>
<td>0.74</td>
<td>18.0</td>
<td>22.6</td>
<td>30.7</td>
</tr>
<tr>
<td>$1.5 \times 10^{-31}$</td>
<td>4.20</td>
<td>0.74</td>
<td>17.9</td>
<td>22.6</td>
<td>30.7</td>
</tr>
<tr>
<td>$2.0 \times 10^{-31}$</td>
<td>5.60</td>
<td>0.74</td>
<td>17.8</td>
<td>22.6</td>
<td>30.7</td>
</tr>
<tr>
<td>$2.5 \times 10^{-31}$</td>
<td>7.00</td>
<td>0.73</td>
<td>17.8</td>
<td>22.6</td>
<td>30.7</td>
</tr>
<tr>
<td>$3.0 \times 10^{-31}$</td>
<td>8.40</td>
<td>0.73</td>
<td>17.6</td>
<td>22.6</td>
<td>30.7</td>
</tr>
<tr>
<td>$3.5 \times 10^{-31}$</td>
<td>9.81</td>
<td>0.73</td>
<td>17.5</td>
<td>22.6</td>
<td>30.7</td>
</tr>
<tr>
<td>$4.0 \times 10^{-31}$</td>
<td>11.22</td>
<td>0.73</td>
<td>17.3</td>
<td>22.6</td>
<td>30.6</td>
</tr>
<tr>
<td>$4.5 \times 10^{-31}$</td>
<td>12.64</td>
<td>0.73</td>
<td>17.2</td>
<td>22.6</td>
<td>30.6</td>
</tr>
</tbody>
</table>

Figure 3.15: Group delay (peak) dependence on $\xi$, with $B = 25$ GHz and $B_{OTF} = 5.2$ GHz.

The relation between $\xi$ and the peak group delay $\tau_{g,p}$, is shown in figure 3.15 and table 3.3, and is given by

$$
\xi \approx 3.563 \times 10^{-20} \cdot \tau_{g,p} \ [s^3].
$$

The above expressions can be generalized for different values of $B_{OTF}$ as

$$
\tau_{g,p-p} = m_\xi \left( \frac{B}{B_{OTF}} \right) \zeta = \frac{1}{K_{g,p-p}} \zeta \ [ps]
$$
Dependence of the predicted WSS amplitude response and group delay on the model parameters

\begin{equation}
\tau_{g,p} = m_\xi \left( \frac{B}{B_{\text{OTF}}} \right) \xi = \frac{1}{K_{g,p}} \xi \quad [\text{ps}]
\end{equation}

where \( K_{g,p} = 1/[m_\zeta(B/B_{\text{OTF}})] \ [s^{-3}] \) and \( K_{g,p} = 1/[m_\xi(B/B_{\text{OTF}})] \ [s^{-4}] \) are slopes such as the ones that were included in equations 3.15 and 3.16.

The parameters \( m_\zeta \) and \( m_\xi \) are plotted in figure 3.16 as function of \( B/B_{\text{OTF}} \).

\begin{figure}[h]
\centering
\includegraphics[width=\textwidth]{figure3_16.png}
\caption{\( m_\zeta \) dependence on \( B/B_{\text{OTF}} \), for \( B = 25 \) GHz (a), \( m_\xi \) dependence on \( B/B_{\text{OTF}} \), for \( B = 25 \) GHz (b).}
\end{figure}

The expressions for \( m_\zeta \) and \( m_\xi \) can be obtained by doing a polynomial fitting of the curves in figure 3.16

\begin{equation}
m_{\zeta,25\text{GHz}} = 1 \times 10^{21} \left[ -0.4430 \left( \frac{B}{B_{\text{OTF}}} \right)^2 + 1.4030 \frac{B}{B_{\text{OTF}}} - 0.4280 \right] \ [s^3]
\end{equation}

\begin{equation}
m_{\xi,25\text{GHz}} = 1 \times 10^{30} \left[ -0.0849 \left( \frac{B}{B_{\text{OTF}}} \right)^2 + 9.2310 \frac{B}{B_{\text{OTF}}} - 6.1647 \right] \ [s^4]
\end{equation}

These expressions allow to relate the peak-to-peak and peak group delay to the \( \zeta \) and \( \xi \) parameters, respectively, for different values of \( B/B_{\text{OTF}} \). The parameters of the model can then be matched to experimental measurements of group delay introduced by the WSS to evaluate the performance of a chain of ROADMs.
3.6 Conclusions

In this chapter, the structures of the MB-OFDM metro network and of the MB-OFDM metro network nodes were described. Various ROADM architectures that can be implemented with the MB-OFDM network were studied. The ROADM architecture employing a passive splitter and a mux WSS was identified as the most adequate, since it balances the flexibility of the WSS and the amount of spectral filtering that is performed at the node.

The analytical model of the WSS, which is the enabling technology of the ROADM, was thoroughly described. An extension of the model was proposed to take into account the group delay introduced by the WSS. The dependence of the amplitude response and the group delay on the model parameters was analysed. Expressions were obtained to adjust the model parameters and match the group delay to experimental measurements. The impact of the group delay on the transmission performance along a set of ROADMs can then be evaluated.
Chapter 4

Performance degradation along a concatenation of MB-OFDM metro network nodes

In this chapter, the transmission performance of the virtual carrier-assisted MB-OFDM system is evaluated along a set of concatenated metro network nodes. The MB-OFDM signal is affected by various impairments during transmission: nonlinearities due to the optical transmitter and optical receiver, attenuation due to fibre losses and insertion losses of the optical components, fibre dispersion, crosstalk, ASE noise introduced by the EDFAs, electrical noise of the receiver and passband narrowing due to successive filtering actions at the ROADMs. In this work, the BER is estimated using an Exhaustive Gaussian Approach (EGA) \[59\]. The above stated impairments are considered in the numerical model used to evaluate the transmission performance, though the analysis is focused on the impact of the passband narrowing, also known as the filter concatenation effect. The optical noise is distributed along the network.

In section 4.1, the MB-OFDM system parameters used in the analysis of the transmission performance are indicated. The 3-band and 4-band MB-OFDM signals employing virtual carriers are specified. In section 4.2, the performance is evaluated without considering the fibre dispersion. In this case, the spans are characterized only by the fibre loss. In section 4.3, the performance is evaluated considering that the spans are modelled by the loss and chromatic dispersion introduced by the optical fibre. In section 4.4, the impact of the electrical noise of the receiver on the transmission performance is analysed. In section 4.5, the impact of the WSS delay distortion the transmission performance is analysed, using the model proposed in chapter 3. Section 4.6 presents the main achievements and conclusions of the chapter.

The performance results given in this chapter were presented in papers [60] and [61].
4.1 MB-OFDM system parameters

This work uses optimized parameters that were obtained in back-to-back operation for the virtual carrier-assisted MB-OFDM system [38]. A Gaussian and 2nd order super-Gaussian band-selector are considered. In order to suppress the SSBI that appears after photodetection, an approach similar to the technique demonstrated in [42] is used.

Figure 4.1 shows the signal spectra of the 42.8 Gb/s 3-band and 4-band SSB MB-OFDM signal at the output of the electro-optical modulator (DP-MZM). The spectra were obtained using the numerical simulator and considering optimized parameters when a 2nd order super-Gaussian BS is used. The bands are numbered from 1 (lowest frequency) to 3 or 4 (highest frequency). The virtual carrier is located at a higher frequency than its corresponding band. The band with the lowest frequency is centered at 3.18 GHz for the 3-band MB-OFDM system and 2.54 GHz for the 4-band MB-OFDM system due to the passband of the hybrid coupler (HC). The HC is employed to generate the Hilbert Transform of the signal that is fed to one of the branches of the DP-MZM. The amplitude and phase responses of the HC were shown in chapter 2.

In table 4.1, the parameters used in the MB-OFDM system are summarized. The use of 2nd order super-Gaussian BS allows for a narrower band gap in comparison with the Gaussian BS. This is due to the higher selectivity of the 2nd order super-Gaussian BS. The bandwidth of the MB-OFDM signal depends on the system parameters used to generate the OFDM bands. In any case, the 42.8 Gb/s SSB MB-OFDM signal does not exceed a bandwidth of 20 GHz.

A modulation index of 5% is used, as mentioned in the description of the electro-optical modulator in chapter 2.
Table 4.1: Parameters of the MB-OFDM system.

<table>
<thead>
<tr>
<th>System</th>
<th>4-band</th>
<th>3-band</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of subcarriers</td>
<td>128</td>
<td>128</td>
</tr>
<tr>
<td>Mapping scheme</td>
<td>16-QAM</td>
<td>16-QAM</td>
</tr>
<tr>
<td>Line bit-rate [Gb/s]</td>
<td>42.8</td>
<td>42.8</td>
</tr>
<tr>
<td>Bandwidth of each band [GHz]</td>
<td>2.68</td>
<td>3.57</td>
</tr>
<tr>
<td>Virtual carrier-to-band power ratio [dB]</td>
<td>7</td>
<td>7</td>
</tr>
<tr>
<td>Super-Gaussian BS order</td>
<td>1/2</td>
<td>1/2</td>
</tr>
<tr>
<td>-3 dB bandwidth of the BS [GHz]</td>
<td>3.6/3.6</td>
<td>4.8/4.8</td>
</tr>
<tr>
<td>Band gap [GHz]</td>
<td>2.8</td>
<td>2.1</td>
</tr>
<tr>
<td>Virtual carrier-to-band gap [MHz]</td>
<td>20/20</td>
<td>20/20</td>
</tr>
<tr>
<td>Bandwidth of the MB-OFDM signal [GHz]</td>
<td>19.2/17.0</td>
<td>18.2/15.7</td>
</tr>
</tbody>
</table>

Figure 4.2: Schematic diagram of the MB-OFDM network model.

Figure 4.2 shows the schematic diagram of the MB-OFDM network model used to evaluate the transmission performance along a set of concatenated nodes. The average power at the optical fibre input is 0 dBm. Nonlinear effects in the optical fibre are not considered. EDFAs are used prior and after each ROADM. At the input of the node, an EDFA with a gain of 10 dB is used to compensate for the 40 km-long fibre span loss. At the output of the node, an EDFA with a gain of 10.5 dB is used to compensate for node losses. These losses comprise 6.5 dB from the WSS and 4 dB from the passive splitter. The EDFAs noise figure is 5 dB. The square-root of the power spectral density of the current noise of the receiver electrical circuit is 25 pA/√Hz. It is considered that all bands composing the MB-OFDM signal are generated in the same node.

The ROADM architecture and the WSS model were described in chapter 3. The parameters of the WSS model were adjusted to a 25 GHz WSS. The bandpass filter created by the WSS has a selectivity of 0.74 and -3 dB bandwidth of 22.6 GHz.
Figure 4.3: Amplitude response of a cascade of ROADMs with 1, 10 and 20 WSSs. $\xi = 0$, $\zeta = 0$.

Figure 4.3 shows the amplitude response of a cascade of ROADMs with 1, 10 and 20 WSSs. A shift in frequency was performed in order to coincide the central frequency of the bandpass filter created by the WSS with the central frequency of the MB-OFDM signal. The central frequency is 9.7 GHz which corresponds to the 4-band MB-OFDM signal employing a 2nd order super-Gaussian BS (SG-BS).

4.2 Results without fibre dispersion

The BER was assessed for 3-band and 4-band MB-OFDM signals with the parameters shown in table 4.1. In this section, the spans are modelled only by the loss introduced by the optical fibre. Fibre dispersion is not considered.

Figure 4.4: BER after a cascade of nodes for the 3-band MB-OFDM signal employing a) a Gaussian BS and b) a 2nd order SG-BS. Fibre dispersion was not considered.
Figure 4.4 shows the degradation of the BER after a cascade of nodes for the 3-band MB-OFDM signal. The signal is being affected by the optical noise originated in the EDFAs and the electrical noise of the receiver, as well as the distortion caused by the passband narrowing at the ROADMs. In figure 4.4 a), a Gaussian BS is used. Band 3 is the most affected as its virtual carrier located at a higher frequency is being attenuated by the cascade of ROADMs leading to a reduced signal power after photodetection. With a small number of ROADMs, band 1 suffers from less degradation, because its virtual carrier is less attenuated in comparison with Band 3 due to the location of the virtual carrier with regard to the bandwidth narrowing. Band 2 is only affected by the accumulation of optical noise from the EDFAs and the electrical noise of the receiver. Using a Gaussian BS, the 3-band MB-OFDM signal can traverse up to 10 ROADMs with a BER<$10^{-3}$.

In figure 4.4 b), a 2nd order SG-BS is used. The use of 2nd order SG-BS allows a MB-OFDM signal with a narrower bandwidth. For this reason, the signal is less affected by the passband narrowing and can traverse up to 26 ROADMs with BER<$10^{-3}$. We notice that bands 1 and 2 have similar performance even after 38 ROADMs, which hints that the virtual carrier of band 1 is not being attenuated. The overall filtering response of the ROADM cascade becomes more selective as the number of ROADMs increases (after 20 ROADMs, the selectivity is 0.8).

![Signal spectrum of the 3-band MB-OFDM signal after a cascade of ROADMs.](image)

Figure 4.5: Signal spectrum of the 3-band MB-OFDM signal after a cascade of ROADMs.

Figure 4.5 shows the signal spectrum of the 3-band MB-OFDM signal after a cascade of ROADMs. Parameters corresponding to the system employing a Gaussian BS were used. Bands 1 and 3 show considerable distortion. The virtual carrier corresponding to band 3 is being attenuated due to the passband narrowing.
Figure 4.6: BER after a cascade of nodes for the 4-band MB-OFDM signal employing a 2nd order SG-BS. Fibre dispersion was not considered.

Figure 4.6 shows the degradation of the BER after a cascade of nodes for the 4-band MB-OFDM signal using a 2nd order SG-BS. The signal can traverse up to 18 ROADMs with BER<10^{-3}. Although the 4-band signal is advantageous because it has finer granularity, the 3-band signal is more suited for transmission through cascades with a considerable number of ROADMs.

4.3 Results with fibre dispersion

In this section, the spans are modelled by the loss and chromatic dispersion introduced by the optical fibre. The BER was assessed for 3-band and 4-band MB-OFDM signals with the parameters shown in table 4.1.

Figure 4.7: BER after a cascade of nodes for the 3-band MB-OFDM signal employing a) a Gaussian BS and b) a 2nd order SG-BS.
Figures 4.7 and 4.8 show the degradation of the BER after a cascade of nodes for the 3-band and 4-band MB-OFDM signal, respectively. In comparison with figures 4.4 and 4.6, it can be seen that the impact of the fibre dispersion on the BER is very small (<1 dB). This indicates that the DP-MZM is able to adequately suppress one sideband of the MB-OFDM signal, and the CDIPF is negligible.

### 4.4 Analysis of the impact of the electrical noise on the MB-OFDM system performance

The BER was assessed for 3-band and 4-band MB-OFDM signals with the parameters shown in table 4.1. Fibre dispersion was not considered. In this section, the electrical noise of the receiver was not included in the model.

Figure 4.9: BER after a cascade of nodes for the 3-band MB-OFDM signal employing a) a Gaussian BS and b) a 2nd order SG-BS. Electrical noise of the receiver was not considered.
Figure 4.9 shows the degradation of the BER after a cascade of nodes for the 3-band MB-OFDM signal, when the electrical noise of the receiver is not considered. Using a Gaussian BS, the 3-band MB-OFDM signal can traverse up to 12 ROADMs with BER<10^{-3}. The attenuation of the virtual carrier due to the passband narrowing causes the signal×carrier beating term to have less power after photodetection and be more affected by the electrical noise of the receiver. In this case, since no electrical noise is present, band 3 is able to traverse an additional 2 ROADMs, when compared with the results shown in figure 4.4. However, the BER of band 3 is still greater than band 1 and 2. This occurs because the beat between the signal and the optical noise after photodetection is severely affecting band 3, which has less power than the other bands.

Comparing figures 4.9 b) and 4.4 b), it can be seen that the BER of band 3 shows only a small improvement when electrical noise of the receiver is not present, using a 2nd order SG-BS. This is explained by the fact that the 3-band MB-OFDM signal using a 2nd order SG-BS occupies the least amount of bandwidth of all the cases analysed. Hence, the virtual carrier is less affected by the passband narrowing and band 3 has only slightly lower power level than the other bands after photodetection.

Figure 4.10 shows the degradation of the BER after a cascade of nodes for the 4-band MB-OFDM signal using a 2nd order SG-BS, when the electrical noise of the receiver is not considered.

Figure 4.10 shows the degradation of the BER after a cascade of nodes for the 4-band MB-OFDM signal using a 2nd order SG-BS, when the electrical noise of the receiver is not considered. For the same reasons given for the 3-band MB-OFDM signal, the BER shows a small improvement when compared with the case where electrical noise is present.
4.5 Analysis of the impact of the group delay of the WSS on the MB-OFDM system performance

In this section, the BER results for the 3-band and 4-band MB-OFDM systems, when group delay is introduced by the WSS, are presented. A performance comparison is made among systems employing WSSs with null, linear and parabolic group delay.

![Group delay introduced by a cascade of ROADMs with 1, 10 and 20 WSSs. a) $\xi = 0, \zeta = 1.61 \times 10^{-21} \text{s}^2$; b) $\xi = 2.5 \times 10^{-31} \text{s}^3, \zeta = 0$.](attachment:image.png)

Figure 4.11: Group delay introduced by a cascade of ROADMs with 1, 10 and 20 WSSs. a) $\xi = 0, \zeta = 1.61 \times 10^{-21} \text{s}^2$; b) $\xi = 2.5 \times 10^{-31} \text{s}^3, \zeta = 0$.

Figure 4.11 shows the group delay introduced by a cascade of ROADMs with 1, 10 and 20 WSSs. The group delay was obtained by calculating the integral in equation 3.5 and using relations 3.15 and 3.16 to set a peak-to-peak or a peak group delay of 7 ps for each WSS. It can be observed that the group delay increases proportionally to the number of WSSs traversed in the network. After 20 WSS, the peak group delay reaches 140 ps in figure 4.11 b).

![BER as a function of the number of ROADMs for a 4-band MB-OFDM system employing WSSs with null delay (continuous lines) and WSSs with linear delay in the passband (dashed lines). Band 1 (circles), band 2 (squares), band 3 (triangles), band 4 (stars).](attachment:image.png)

Figure 4.12: BER as a function of the number of ROADMs for a 4-band MB-OFDM system employing WSSs with null delay (continuous lines) and WSSs with linear delay in the passband (dashed lines). Band 1 (circles), band 2 (squares), band 3 (triangles), band 4 (stars).
Figure 4.12 shows the BER after a cascade of ROADMs for a 4-band MB-OFDM system using a SG-BS. The continuous lines correspond to the system employing WSSs described by equation 3.6, i.e., with null group delay ($\xi = 0$, $\zeta = 0$). The dashed lines correspond to the system employing WSSs with a linear delay inside the passband ($\xi = 0$, $\zeta = 1.61 \times 10^{-21} \text{s}^3$). The performance of the two systems is similar. After 30 ROADMs, the peak-to-peak group delay introduced by the cascade of WSSs is around 210 ps, which is a small amount compared to the OFDM symbol duration ($47.85 \times 10^{-9} \text{s}$). The cyclic prefix in the OFDM symbols is compensating for the group delay effect. As shown previously, bands 1 and 4 have a higher BER than bands 2 and 3 due to the distortion caused by the passband narrowing after a cascade of ROADMs. Band 4 is more affected because its virtual carrier is being more attenuated than the virtual carrier of band 1.

![Figure 4.13: BER as a function of the number of ROADMs for a 4-band MB-OFDM system employing WSSs with null delay (continuous lines) and WSSs with parabolic delay in the passband (dashed lines). Band 1 (circles), band 2 (squares), band 3 (triangles), band 4 (stars).](image-url)

Figure 4.13 shows the BER after a cascade of ROADMs for the 4-band MB-OFDM system employing WSSs with null delay (continuous lines) and WSSs with parabolic delay in the passband (dashed lines). Band 1 (circles), band 2 (squares), band 3 (triangles), band 4 (stars).

The additional degradation is not caused by the WSS delay distortion, but from a reduction of the WSS channel passband caused by the group delay of the micromirror. The introduction of a frequency-dependent phase in equation 3.1 influences both the amplitude response and phase response of the WSS. This effect is more noticeable when the group delay has a parabolic shape inside the passband. This behaviour is in qualitative agreement with the experimental results.
presented in [55] where it was shown that a curvature of the micromirror (switching element) causes a reduction of the passband in MEMS-based devices.

![Graphs showing BER vs. number of ROADMs for 3-band MB-OFDM system](image)

Figure 4.14: BER as a function of the number of ROADMs for a 3-band MB-OFDM system employing WSSs with null delay (continuous lines) and WSSs with parabolic delay in the passband (dashed lines). Band 1 (circles), band 2 (squares), band 3 (triangles).

Fig. 4.14 shows the BER after a cascade of ROADMs for the 3-band MB-OFDM system using a) a Gaussian BS and b) a SG-BS. The continuous lines correspond to the system with null group delay ($\xi = 0, \zeta = 0$). The dashed lines correspond to the system employing WSSs with a parabolic delay inside the passband ($\zeta = 0, \xi = 2.5 \times 10^{-31} \text{s}^3$). It can be seen that the group delay distortion is also being compensated for in the 3-band MB-OFDM system. Therefore the impact of group delay distortion on the MB-OFDM system performance is reduced.

4.6 Conclusions

In this chapter, the BER of the virtual carrier-assisted MB-OFDM system was evaluated along a set of concatenated nodes. It was shown that the 3-band MB-OFDM signal can traverse up to 10 ROADMs using a Gaussian BS and 26 ROADMs using a 2nd order SG-BS, with a BER $< 10^{-3}$. The 4-band MB-OFDM signal can traverse up to 18 ROADMs using a 2nd order SG-BS, with a BER $< 10^{-3}$. Fibre dispersion was shown to have a very small impact on the BER, due to the use of SSB transmission. The results with no electrical noise show a slightly lower BER.

It was shown that the typical group delay introduced by the WSS has a negligible impact on the BER performance. The peak group delay is too small to cause enough distortion that cannot be compensated for at the OFDM receiver. From the WSS extended model, it was shown that a phase response introduced at the switching element reduces the passband of the WSS. This is noticeable in the BER performance, but the impact is still very small to reduce the number of ROADMs that can be traversed.
Chapter 5

Conclusions and future work

In this chapter, the final conclusions of the work developed in this dissertation are presented. In addition, suggestions for future work are provided.

5.1 Final conclusions

In this dissertation, the transmission of virtual carrier-assisted MB-OFDM signals along a set of concatenated MB-OFDM metro network nodes was studied. The impact of the node filtering concatenation effect on the BER was assessed.

In chapter 2 the virtual carrier-assisted MB-OFDM system was described. The fundamentals of OFDM were presented, including the advantages and disadvantages of this type of modulation. The models of the electro-optical modulator and the optical receiver were described. The beating components that appear after photodetection at the receiver were identified and the SSBI was recognized as a limiting factor of the transmission performance. The necessity of employing SSB transmission due to the CDIPF was recognized. The MB-OFDM optical signal was characterized and various related definitions were introduced.

In chapter 3 the MB-OFDM metro network is characterized and its operation principles are described. Various possible ROADM architectures that can be employed in the MB-OFDM metro network were analysed. The architecture employing a passive splitter and a mux WSS was identified as being the solution that reduces the number of successive filtering actions, while maintaining the flexibility provided by the WSS. In addition, this architecture also provides power level equalization of the wavelengths exiting the node, without employing additional components, such as VOAs. A thorough analytical description of the WSS, which is the enabling technology of the ROADMs, was provided. An extension of the physical model of the WSS was developed to take into account the group delay introduced by MEMS-based WSS. The dependence of the
predicted WSS amplitude response and group delay on the model parameters was analysed in
detail and expressions were obtained for the relevant relations. A linear and parabolic group
delay shapes inside the passband were obtained from the extended model. The predicted group
delay is in qualitative agreement with experimental measurements of MEMS-based WSS devices.

In chapter 4, the transmission performance of the MB-OFDM metro network was analysed. A
numerical simulator was implemented to study the transmission performance of virtual carrier-
assisted MB-OFDM signals along a set of concatenated nodes. The optical noise (originated by
the EDFAs) is distributed along the network. The figure of merit is the BER, estimated using
an exhaustive Gaussian approach. It was shown that the 3-band MB-OFDM signal can traverse
up to 10 ROADMgs using a Gaussian BS, and 26 ROADMgs using a 2nd order SG-BS, with a
BER<10^{-3}. The 4-band MB-OFDM signal can traverse up to 18 ROADMgs using a 2nd order
SG-BS, with a BER<10^{-3}. The band with the highest frequency in the MB-OFDM has the
highest BER, due to the virtual carrier being attenuated by the passband narrowing caused by
the successive filtering actions at the nodes. The attenuation of the virtual carrier causes the
signal×carrier beating term to have less power after photodetection and be more affected by the
signal×noise beating term and the electrical noise of the receiver. Fibre dispersion was shown to
have a very small impact on the BER, due to the use of SSB transmission. The results with no
electrical noise show a slightly lower BER. It was shown that the typical group delay introduced
by the WSS has a negligible impact on the BER performance. From the WSS extended model,
it was shown that the phase response introduced at the switching element reduces the width of
the passband of the WSS. This is noticeable in the BER performance, but the impact is still
very small to reduce the number of ROADMgs that can be traversed.

5.2 Future work

From the work performed in this dissertation, the following topics are suggested as future work:

- to develop a model for the WSS based on a spatial light modulator [62] and study its
  impact on the transmission performance of direct detection MB-OFDM networks,

- to analyse the BER performance of direct detection MB-OFDM metro networks with fibre
  spans longer than 40 km, alternative EDFA placement and EDFA noise figure other than
  5 dB,

- to study the transmission performance of MB-OFDM signals with a number of OFDM
  bands other than 3 or 4, channel spacing other than 25 GHz (within the ITU grid) and
  investigate alternative virtual carrier placement,
• to implement an adaptive modulation technique on the subcarriers comprising the OFDM bands and study its potential ability to mitigate the impact of the filter concatenation effect on the transmission performance of direct detection MB-OFDM networks,

• to investigate the use of optical wave-shaping \cite{63} at the network nodes to mitigate the impact of the filter concatenation effect on the transmission performance of direct detection MB-OFDM networks,

• to implement a DSP algorithm to reconstruct and remove the SSBI after photodetection and assess the impact of the filter concatenation effect on the quality of the estimation of the SSBI,

• to study the impact of in-band crosstalk and fibre nonlinearities on the transmission performance of direct detection MB-OFDM networks,

• to experimentally implement the MB-OFDM system described in this dissertation and compare the performance measurements with the results obtained through numerical simulation.
References


[38] A. Cartaxo, T. Alves and L. Mendes, “42.8 Gb/s SSB DD MB-OFDM metro networks assisted by virtual carriers: system parameters optimization”, IEEE International Conference on Transparent Optical Networks, paper We.A1.1, Graz, Austria, July 2014.


REFERENCES


Appendix A

MB-OFDM system details

A.1 Electrical noise

Electrical noise is present in every electrical component of a system. In the receiver, resistive and active elements generate electrical noise. In this work, it is assumed, for simplicity, that the electrical noise present in the system is generated at the receiver, after the photodetector.

The one-sided power spectrum density of the electrical circuit noise is given by

\[ S_c(f) = \frac{4k_BT_r}{R_{bias}} f_{n,e} \]  

(A.1)

where \( k_B \) is the Boltzmann constant, \( T_r \) is the room temperature, \( R_{bias} \) is the bias resistance and \( f_{n,e} \) is the noise figure of the active components of the electrical circuit of the receiver.

In this work, the square-root of the power spectral density of the current noise of the electrical circuit of the receiver is \( \sqrt{S_c(f)} = 25 \text{ pA}/\sqrt{\text{Hz}} \).

A.2 Optical noise

EDFAs generate amplified spontaneous emission (ASE) noise, which is the dominant form of optical noise in the system.

The power spectral density of the ASE noise \( (S_{ASE}) \) for each polarization mode in the optical fibre is given by

\[ S_{ASE}(\upsilon_0) = \frac{f_{n,o}}{2} (g_{EDFA} - 1) h\upsilon_0 \]  

(A.2)

where \( f_{n,o} \) is the noise figure of the EDFA, \( g_{EDFA} \) is the amplifier gain, \( h \) is the Planck constant and \( \upsilon_0 \) is the optical frequency. In this work, a noise figure of 5 dB is considered.
The optical fibre has two polarizations, i.e. the signal travelling in the optical fibre is composed of two orthogonal electrical vector field components, parallel (∥) and perpendicular (⊥), that vary in amplitude and frequency with directions $\hat{e}_∥$ and $\hat{e}_⊥$, respectively. In addition, the optical noise has an in-phase component $n_I$ and a quadrature component $n_Q$. The ASE noise power is equally divided by both components in each polarization.

Assuming that the signal optical field travels in the parallel direction, and the optical noise travels in both parallel and perpendicular directions, the optical field at the EDFA output $e_{o,\text{EDFA}}(t)$ is given by

$$e_{o,\text{EDFA}}(t) = [e_{i,\text{EDFA}}(t) + e_{n,∥}(t)]\hat{e}_∥ + e_{n,⊥}(t)\hat{e}_⊥$$  \hspace{1cm} (A.3)

where $e_{i,\text{EDFA}}(t)$ is the optical field at the EDFA input, and $e_{n,∥}(t) = n_{I,∥}(t) + jn_{Q,∥}(t)$ and $e_{n,⊥}(t) = n_{I,⊥}(t) + jn_{Q,⊥}(t)$ are the optical noise fields at the parallel and perpendicular directions, respectively.

The ASE noise power at the EDFA output ($p_{\text{ASE}}$) for each polarization mode is given by

$$p_{\text{ASE}} = S_{\text{ASE}(\nu_0)}B_0$$  \hspace{1cm} (A.4)

where $B_0$ is the optical reference bandwidth, typically 0.1 nm.