Optical Crosstalk
in
WDM Fibre-Radio Networks

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M.Sc.

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the degree of Doctor of Philosophy

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Victoria 3010
Australia
To Annie and Nathalie,

and the rest of my family.

To all those who tried and didn’t make it,

remember life is a journey,

with many different roads,

that eventually lead to the same final destination!
Abstract

The predicted growth in mobile phone traffic and the move towards enhanced mobility will lead to a need for a wireless infrastructure that provides increasing bandwidth per user. It is envisaged that our world will become increasingly interconnected, with mobile communications enabling us to perform an increasing range of tasks.

Future wireless networks will require an optical network to provide antenna Base Stations with sufficient bandwidth to provide individual users with a larger bandwidth. The combined optical and wireless network is referred to as a “fibre-radio” or “radio-over-fibre” or “fibre-wireless” network. It is expected that such high-capacity networks will use Wavelength Division Multiplexing (WDM) to increase the total bandwidth transmitted over the optical access network. Such a high-capacity network would not be achievable using a single wavelength or using a copper or coax network.

Optical crosstalk is present in WDM optical networks and degrades the received signal quality, increasing the bit-error-rate. Two types of crosstalk occur, depending on whether the crosstalk channel is a different wavelength to the signal or at the same wavelength (out-of-band and in-band crosstalk, respectively). An important consideration for fibre-radio networks is whether or not the optical network transports data at baseband, using standard intensity modulation, or at an RF frequency, using subcarrier modulation. The nature of the optical modulation scheme has implications for the design of the Central Office and the Base Stations, and potentially for optical crosstalk.

This thesis investigates the effect of optical crosstalk in fibre-radio networks using subcarrier modulation, i.e. transmitting the modulated RF subcarriers over the optical network. We investigate ASK, BPSK and QPSK modulation, taking into account the wireless frequency domain. Analytical results on the impact of optical crosstalk are obtained and validated using simulation software and through experimental investigations, showing that the impact of optical crosstalk in WDM fibre-radio networks using subcarrier modulation is different to that for traditional baseband intensity modulated optical links.
The impact of these results for network design are assessed by showing the combined effect of both in-band and out-of-band optical crosstalk on the network capacity, i.e. the total number of WDM channels than can be transported. Results quantify the effect of different component crosstalk levels on the resulting optical power penalty and network dimension.

Finally, we investigate the implications of the results for a specific high-capacity wireless network operating at 20 GHz and providing 1 Gbps per radio cell, using ten RF subcarriers per wireless frequency band. Optical network topologies are compared based on optical power budgets and optical crosstalk levels, highlighting the difference between star and ring networks. More complex architectures are also explored. This leads to several important conclusions that affect the choice and design of future WDM fibre-radio networks.

This thesis provides tools to quantify the impact of optical crosstalk in WDM fibre-radio networks using subcarrier modulation and highlights the important differences between baseband and subcarrier modulated links. The choice of optical network topology is shown to be independent of modulation scheme once optical and crosstalk power budgets have been established.
Declaration

This thesis comprises only my original work towards the PhD and, except where acknowledged, includes no material previously published by any other person. I declare that none of the work presented in this thesis has been submitted for any other degree or diploma at any University. Furthermore, this thesis is less than 100,000 words in length, excluding figures, tables, bibliographies, appendices, and footnotes.

David Castleford
Acknowledgments

I would like to thank my three supervisors, Prof. Rod Tucker, Dr. A. Nirmalathas and Ass. Prof. Dalma Novak. I started this Ph.D. under the guidance of Dalma. Thank you for your support. Rod then took over, keeping me on track and encouraging me to start writing-up. I am grateful for the critical comments provided throughout this period – it has finally led to this thesis. I also appreciate ongoing encouragement during the final year. Thanks also to Thas for his technical advice and useful feedback during the writing stage. I would like to thank all of you for helping me publish, improve my presentation and writing skills, and finish my Ph.D.

I must also acknowledge the contribution of Dr. Sarah Dods, who helped me understand optical crosstalk and the associated analytical techniques. I also appreciate helpful comments on fibre-radio from Dr. Christina Lim. On the experimental side, our lab manager Dr. Henry Senko was very helpful with equipment and advice on experimental problems.

General thanks to fellow students, An, Elaine, Manik, Charlotte and Mark, for their friendship and help in general. Thanks to everyone who is part of the Photonics Research Laboratory and who all contributed towards helping me finish my Ph.D.

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I cannot overlook the financial aspect of my studies. I am very grateful to the Photonics Research Laboratory and the Australian Photonics CRC for providing me with a studentship, and for the University for providing me with a HECS-exemption scholarship. I would not have completed a Ph.D. without this financial assistance.
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<tr>
<td>3G</td>
<td>Third-generation mobile standard.</td>
<td>1.1</td>
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<tr>
<td>AC</td>
<td>Alternating Current.</td>
<td>2.3.2</td>
</tr>
<tr>
<td>AMP</td>
<td>Electrical amplifier.</td>
<td>4.3.1</td>
</tr>
<tr>
<td>ASE</td>
<td>Amplified Spontaneous Emission.</td>
<td>2.3</td>
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<td>ASK</td>
<td>Amplitude-Shift Keying.</td>
<td>1.2</td>
</tr>
<tr>
<td>ATT</td>
<td>Attenuator. Electrical or optical variable attenuator.</td>
<td>2.3.3(A)</td>
</tr>
<tr>
<td>AWG</td>
<td>Arrayed-Waveguide Grating.</td>
<td>2.3.3(A)</td>
</tr>
<tr>
<td>BER</td>
<td>Bit-Error Rate or Bit-Error Ratio.</td>
<td>2.3.1</td>
</tr>
<tr>
<td>BERT</td>
<td>Bit-Error-Rate Testset.</td>
<td>2.3.3(A)</td>
</tr>
<tr>
<td>BPF</td>
<td>Band-Pass Filter.</td>
<td>4.6.1</td>
</tr>
<tr>
<td>BPSK</td>
<td>Binary Phase-Shift Keying.</td>
<td>1.2</td>
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<tr>
<td>BS</td>
<td>Base Station.</td>
<td>1.1</td>
</tr>
<tr>
<td>CATV</td>
<td>CAble TeleVision.</td>
<td>6.3.1</td>
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<tr>
<td>CDMA</td>
<td>Code-Division Multiple Access.</td>
<td>2.2.2</td>
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<tr>
<td>CNR</td>
<td>Carrier-to-Noise Ratio.</td>
<td>6.2.5(A)</td>
</tr>
<tr>
<td>CO</td>
<td>Central Office.</td>
<td>1.1</td>
</tr>
<tr>
<td>CTB</td>
<td>Composite Triple-Beat.</td>
<td>6.2.1</td>
</tr>
<tr>
<td>DC</td>
<td>Direct Current.</td>
<td>3.2.1</td>
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<tr>
<td>DEMUX</td>
<td>DEMUltipleXing.</td>
<td>2.3.4(A)</td>
</tr>
<tr>
<td>DFB</td>
<td>Distributed FeedBack.</td>
<td>2.3.3(A)</td>
</tr>
<tr>
<td>EAT</td>
<td>Electro-absorption Transceiver.</td>
<td>2.2.3</td>
</tr>
<tr>
<td>EDFA</td>
<td>Erbium-Doped Fibre Amplifier.</td>
<td>4.3.1</td>
</tr>
<tr>
<td>EOM</td>
<td>Electro-Optic Modulator.</td>
<td>4.3.1</td>
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<tr>
<td>Abbreviation</td>
<td>Description</td>
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<tr>
<td>ETSI</td>
<td>European Telecommunications Standards Institute.</td>
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<tr>
<td>FBG</td>
<td>Fibre-Bragg Grating.</td>
<td>2.2.3</td>
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<tr>
<td>FDMA</td>
<td>Frequency-Division Multiple Access.</td>
<td>2.2.2</td>
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<tr>
<td>FEC</td>
<td>Forward-Error Correction.</td>
<td>3.2</td>
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<tr>
<td>FTTH</td>
<td>Fibre-To-The-Home.</td>
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<tr>
<td>FWM</td>
<td>Four-Wave Mixing.</td>
<td>6.2.5(A)</td>
</tr>
<tr>
<td>GOLD</td>
<td>Gigabit Optical Link Design. Optical software simulation package.</td>
<td>4.2.1</td>
</tr>
<tr>
<td>IF</td>
<td>Intermediate Frequency.</td>
<td>1.1.1</td>
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<tr>
<td>IM</td>
<td>Intensity Modulation.</td>
<td>1.1.1</td>
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<tr>
<td>ISO</td>
<td>Isolator.</td>
<td>4.3.1</td>
</tr>
<tr>
<td>LAN</td>
<td>Local Area Network.</td>
<td>1.1</td>
</tr>
<tr>
<td>LED</td>
<td>Light-Emitting Diode.</td>
<td>2.2.1</td>
</tr>
<tr>
<td>LMDS</td>
<td>Local Multipoint Distribution System.</td>
<td>1.1</td>
</tr>
<tr>
<td>LNA</td>
<td>Low-Noise Amplifier.</td>
<td>6.2.1</td>
</tr>
<tr>
<td>LO</td>
<td>Local Oscillator.</td>
<td>1.1.1</td>
</tr>
<tr>
<td>LPF</td>
<td>Low-Pass Filter.</td>
<td>4.2.1</td>
</tr>
<tr>
<td>MAN</td>
<td>Metropolitan Area Network.</td>
<td>1.1</td>
</tr>
<tr>
<td>MCB</td>
<td>Modified Chernoff Bound.</td>
<td>2.3.1</td>
</tr>
<tr>
<td>MGF</td>
<td>Moment Generating Function.</td>
<td>2.3.1</td>
</tr>
<tr>
<td>MPA</td>
<td>Medium-Power Amplifier.</td>
<td>6.2.1</td>
</tr>
<tr>
<td>MU</td>
<td>Mobile Unit.</td>
<td>1.1</td>
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<td>MUX</td>
<td>MULTipleXing.</td>
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<td>MZ</td>
<td>Mach-Zehnder modulator.</td>
<td>4.2.1</td>
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<tr>
<td>NF</td>
<td>Noise Figure.</td>
<td>6.2.1</td>
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<td>OADM</td>
<td>Optical Add-Drop Multiplexer.</td>
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<td>OPM</td>
<td>Optical Power Meter.</td>
<td>2.3.3(A)</td>
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<tr>
<td>Acronym</td>
<td>Definition</td>
<td>Page</td>
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<td>---------</td>
<td>------------</td>
<td>------</td>
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<tr>
<td>PD</td>
<td>PhotoDetector or PhotoDiode.</td>
<td>1.1.1</td>
</tr>
<tr>
<td>PDL</td>
<td>Polarisation-Dependent Loss.</td>
<td>2.3.4(B)</td>
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<tr>
<td>PIIN</td>
<td>Phase-induced interferometric noise.</td>
<td>1.1.2</td>
</tr>
<tr>
<td>PIN</td>
<td>Positive-Intrinsic-Negative.</td>
<td>2.3.2(F)(ii)</td>
</tr>
<tr>
<td>PLL</td>
<td>Phase-Locked Loop.</td>
<td>3.1</td>
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<tr>
<td>PMD</td>
<td>Polarisation Mode Dispersion.</td>
<td>2.3.4(B)</td>
</tr>
<tr>
<td>PON</td>
<td>Passive Optical Network.</td>
<td>6.3.1</td>
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<tr>
<td>PRBS</td>
<td>Pseudo-Random Bit Sequence.</td>
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<tr>
<td>QAM</td>
<td>Quadrature Amplitude Modulation.</td>
<td>2.3.1</td>
</tr>
<tr>
<td>QPSK</td>
<td>Quadrature Phase-Shift Keying.</td>
<td>1.2</td>
</tr>
<tr>
<td>RF</td>
<td>Radio Frequency.</td>
<td>1.1</td>
</tr>
<tr>
<td>RIN</td>
<td>Relative Intensity Noise.</td>
<td>5.3</td>
</tr>
<tr>
<td>SCM</td>
<td>SubCarrier Multiplexing.</td>
<td>1.1.1</td>
</tr>
<tr>
<td>SMF</td>
<td>Single-Mode Fibre.</td>
<td>2.2.1</td>
</tr>
<tr>
<td>SMS</td>
<td>Short Message System.</td>
<td>1.1</td>
</tr>
<tr>
<td>SNR</td>
<td>Signal-to-Noise-Ratio.</td>
<td>6.2.1</td>
</tr>
<tr>
<td>SPM</td>
<td>Self-Phase Modulation.</td>
<td>6.2.5(A)</td>
</tr>
<tr>
<td>SW</td>
<td>Optical switch.</td>
<td>2</td>
</tr>
<tr>
<td>TDMA</td>
<td>Time-Division Multiple Access.</td>
<td>2.2.2</td>
</tr>
<tr>
<td>TV</td>
<td>TeleVision.</td>
<td>2.2</td>
</tr>
<tr>
<td>WDM</td>
<td>Wavelength Division Multiplexing.</td>
<td>1.1.1</td>
</tr>
<tr>
<td></td>
<td>Wavelength Division Multiplexer.</td>
<td>2.2</td>
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<tr>
<td>XPM</td>
<td>Cross-Phase Modulation.</td>
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<tr>
<td>----------</td>
<td>-----------------------------------------------------------------------------</td>
<td>-----------</td>
</tr>
<tr>
<td>$\alpha$</td>
<td>Optical fibre attenuation (dB/km).</td>
<td>6.3.2(A)</td>
</tr>
<tr>
<td>$\alpha(t)$</td>
<td>Time-varying data envelope of the electrical signal.</td>
<td>3.2.1</td>
</tr>
<tr>
<td>$\alpha_c$</td>
<td>Signal amplitude in-phase component.</td>
<td>3.5.1</td>
</tr>
<tr>
<td>$\alpha_y$</td>
<td>Signal amplitude in quadrature component.</td>
<td>3.5.1</td>
</tr>
<tr>
<td>$\beta(t)$</td>
<td>Time-varying data envelope of the crosstalk electrical signal.</td>
<td>3.2.1</td>
</tr>
<tr>
<td>$\beta_c$</td>
<td>Crosstalk signal amplitude in-phase component.</td>
<td>3.5.1</td>
</tr>
<tr>
<td>$\beta_y$</td>
<td>Crosstalk signal amplitude in quadrature component.</td>
<td>3.5.1</td>
</tr>
<tr>
<td>$\delta$</td>
<td>Electrical crosstalk data bit.</td>
<td>3.7.1</td>
</tr>
<tr>
<td>$\Delta$</td>
<td>Optical power difference between crosstalk and signal bands.</td>
<td>6.4.2(B)</td>
</tr>
<tr>
<td>$\Delta\theta$</td>
<td>Optical phase difference between signal and crosstalk electric fields.</td>
<td>2.3.2(A)</td>
</tr>
<tr>
<td>$\Delta\phi$</td>
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Introduction

1.1 FIBRE-RADIO NETWORKS

Current second-generation wireless voice networks use carrier frequencies of 1-2 GHz and user data rates of 10 kbps for voice and SMS messages [1]. Third-Generation or 3G wireless networks can provide data rates varying from 100 kbps (in vehicles) to 2 Mbps indoors [2] whilst operating at similar RF (radio) frequencies. 3G offers the potential to distribute images and provide other low-data rate services such as maps, local information, timetables, etc. For higher data rates, Local Multipoint Distribution Service (LMDS), will operate at 28 GHz and can potentially deliver data rates of tens or hundreds of Mbps given the much higher carrier frequency [3]. While LMDS involves point-to-point links or indoor coverage, it offers sufficiently high data rates to be used for Internet access and to transfer large amounts of data, including video. One of the key factors in increasing data rates is to correspondingly increase carrier frequencies, potentially into the millimetre-wave range (30-300 GHz) [4]. While certain frequency bands suffer from very high loss due to OH absorption, this high loss allows very small cell-sizes to be used, re-using the same frequency in non-adjacent cells. This type of micro- or pico-cellular application will require numerous Base Stations (BS) [5-7].

A fibre-radio network is a network that can enable the provision of high data-rate wireless services, using many BSs covering small wireless cells [8-10]. A fibre-radio network is a hybrid network that uses an optical network to deliver wireless data from a Central Office (CO) to remote radio Base Stations (BSs) [11, 12], as shown in Figure 1.1.
Figure 1.1 shows that the CO provides the interface between an external network (typically a Metropolitan Area Network (MAN) or Local Area Network (LAN)) and a wireless network in which multiple BSs provide wireless coverage to Mobile Units (MUs). Each BS covers a specific radio cell, each providing wireless access, with multiple BSs forming a radio network. The terms “fibre-wireless” and “radio-over-fibre” are equivalent terms to “fibre-radio”. A fibre-radio network differs from a traditional fibre-to-the-home (FTTH) access network in that the transported data is at a wireless frequency and not at baseband, as discussed below.

### 1.1.1 Optical and Wireless Domains

A fibre-radio network comprises two distinct domains, one optical and one wireless. In the optical domain, Wavelength Division Multiplexing (WDM) can be used to combine several wavelengths together to send them through a fibre-optic network, greatly increasing the use of the available fibre bandwidth and maximising total data throughput [13-15]. Routing of signals based on wavelength is also possible and signals can be selected and separated in the wavelength domain. As discussed throughout this thesis, it is envisaged that in order to meet future wireless bandwidth requirements, a single CO feeds each remote radio BS and has access to a separate optical wavelength, as illustrated in Figure 1.2.
**Fig. 1.2 Optical spectrum usage in a WDM fibre-radio network.**

Figure 1.2 shows that each BS is allocated a separate optical wavelength in the downlink (from CO to BS), which can be re-used in the uplink direction (from BS to CO). Note that using the same wavelength in both directions is not a requirement, since a channel offset scheme can be used or downlink and uplink channels can be interleaved. In the latter case, half of the spectrum is allocated to each direction, whereas more efficient usage of the available spectrum can be made if a channel offset is used, as shown in Figure 1.3.

**Fig. 1.3 Offset downlink \((d)\) and uplink \((u)\) optical spectrum allocation.**

In the wireless domain, the electrical frequency spectrum allocated to the wireless service is typically split into multiple RF bands, so that several different frequency bands can be allocated to different BSs, as shown in Figure 1.4.
In Figure 1.4, adjacent radio cells are allocated different frequency bands, minimising interference. The limited radio spectrum is re-used in non-adjacent cells to fully utilise the limited RF spectrum available. Different uplink and downlink RF frequencies are presumed in the RF domain. Further information regarding wireless networks can be found in [16].

One of the main factors that determines the nature of the optical network and the equipment used at the CO and BSs is whether or not data destined for BSs is sent at baseband or at an intermediate- or radio-frequency (IF or RF) [17]. Figure 1.5 shows how baseband modulation is implemented.

Figure 1.5 shows the implementation for baseband modulation. In the downlink direction, the baseband data intensity modulates an optical carrier, which is transported over optical fibre to the BS. At the BS, the optical signal is converted into an electrical signal using a photodetector (PD). The baseband data is converted to the appropriate modulation format and wireless frequency using an RF modulator. Similarly for the
uplink, wireless data arriving at the BS must be converted back to baseband before being sent back to the CO via the optical network. Note that direct modulation of the lasers is possible at low data-rates, eliminating the need for an external modulator and hence providing a more cost-effective solution.

The implementation for IF modulation is shown in Figure 1.6.

![Figure 1.6 CO and BS configuration and RF and optical spectra for IF subcarrier modulation.](image)

In Figure 1.6, the baseband data is converted to the appropriate RF modulation format but the electrical carrier frequency is at an intermediate frequency (IF). The modulated IF subcarrier modulates the optical carrier and the optical signal is sent through the optical fibre. At the BS, the signal is detected and upconverted to the required RF frequency. The simplest implementation is to use a local oscillator (LO) and mixer to upconvert the data to the desired wireless frequency. While an electrical LO can be placed at each BS, optical transport of a remote LO is also possible [18-22]. For the uplink, the RF data is downconverted to IF using the same LO and sent through the optical link. An IF demodulator at the CO converts the data back to baseband. As discussed for baseband modulation, direct modulation of the lasers is possible for IF subcarrier modulation provided the IF frequency is sufficiently low, reducing equipment cost.

If the optical network transports data at the desired wireless frequency, no up- or down-conversion is required at the BS, as shown in Figure 1.7.
An RF modulator is required at the CO to convert baseband data to the appropriate modulation format at the desired RF frequency before subcarrier modulating the optical carrier in the downlink. In the uplink, RF data is modulated onto an optical carrier and sent to the CO via the optical network. After photodetection at the CO, the RF signal must be demodulated to baseband. Multiple IF or RF carriers can be transported over the optical access network, so that each BS has access to a given frequency band. This is referred to a subcarrier multiplexing or SCM [23-25]. Note that with RF modulation, direct modulation of the lasers may also be feasible, although this typically imposes restrictions on the carrier frequency and/or data bandwidth. Laser nonlinearities may also limit the use of direct modulation, particularly if there are multiple subcarriers.

This thesis considers WDM fibre-radio networks that transport data over the optical network using IF or RF subcarrier modulation.

### 1.1.2 WDM and Optical Crosstalk

WDM involves multiplexing multiple wavelengths and transporting them in a single fibre. Current technology allows one to two hundred channels to be transported in a single fibre, achieving Tb/s total capacity [26].

If WDM is used in a fibre-radio network, then each BS can be assigned a single wavelength. This means that each radio cell covered by a BS can have access to the full bandwidth available to each wavelength, allowing wireless users to have access to more
A WDM network requires wavelength selective optical components that can multiplex or demultiplex channels or that can drop or add channels [27]. These components are imperfect and can not fully remove unwanted channels, leading to optical crosstalk, i.e. the presence of an undesired optical signal. Although optical components can reject adjacent wavelength channels by up to 30 dB or more [28, 29], some residual signals will still be present, particularly if channels powers are unequal. This type of unwanted crosstalk is referred to as inhomodyne or heterodyne or inter-channel crosstalk, or simply as out-of-band crosstalk. The latter terminology will be used throughout this thesis. This type of crosstalk does not severely impair network performance as it is at a different wavelength as the desired signal and is simply added to the signal in the electrical domain [30, 31].

A much more detrimental type of crosstalk occurs when the crosstalk signal is at the same wavelength as the desired signal. This is referred to as intra-channel or homodyne or in-band crosstalk (this term is used throughout this thesis) and causes so-called phase-induced interferometric noise (PIIN) [32-34]. It is much more detrimental to the signal as optical mixing of the optical fields upon photodetection creates mixing terms that further degrade the signal compared to the out-of-band case. Furthermore, since it is at the same wavelength as the signal it can not be filtered. In-band crosstalk typically occurs between add and drop channels or along a link due to unwanted optical reflections, either due to the fibre itself, or to optical components or connectors or splices. The difference between the two types of crosstalk is illustrated in Figure 1.8.

![Fig. 1.8 Optical spectra illustrating the difference between (a) out-of-band crosstalk and (b) in-band optical crosstalk.](image-url)
The presence of optical crosstalk in WDM fibre-radio networks will affect network design performance. The nature of the modulation format if subcarrier multiplexing is used may also be important in determining the level of optical crosstalk that can be tolerated for a given system margin. The objective of this thesis is to study optical crosstalk in fibre-radio networks. The effect of modulation format on optical crosstalk is investigated to determine how power penalties and hence component crosstalk specifications may be affected. This may have consequences for future fibre-radio access networks if higher crosstalk levels can be tolerated, allowing cheaper components to be used. The combined effect of in-band and out-of-band optical crosstalk on the total number of channels in a WDM fibre-radio network is determined for various component crosstalk levels, highlighting the potential of crosstalk to limit network capacity. Finally, various optical network topologies are assessed from the point of view of a fibre-radio application, illustrating the effect of optical crosstalk.

1.2 Thesis Outline

The impact of optical crosstalk in WDM fibre-radio networks is investigated, demonstrating the importance of modulation format and revealing a new technique that can be used to minimise the impact of in-band optical crosstalk. The results are used to study various network architectures and highlight important design issues, in particular the importance of optical crosstalk in future WDM fibre-radio access networks.

Chapter 2 provides a review of the existing literature on WDM fibre-radio networks. The literature on optical crosstalk in digital subcarrier-modulated links is reviewed. The need for further research has motivated the work undertaken in this thesis. Research on optical crosstalk for baseband modulation is then reviewed, highlighting key analytical techniques and results, as well as experimental investigations.

In Chapter 3, an analytical model is presented that extends previous work and provides new insight into optical crosstalk for binary phase-shift keying (BPSK) modulated subcarriers, highlighting the significant difference in optical power penalties for in-band crosstalk. Furthermore, the importance in RF phase difference is highlighted where signal and crosstalk sources use the same RF carrier frequency. This leads to the important observation that the impact of in-band crosstalk, as well as out-of-
band crosstalk, can be significantly minimised by ensuring the in-band crosstalk term carries data at a different RF frequency. An extension to quadrature phase-shift keying (QPSK) modulation is presented, showing a higher sensitivity to crosstalk. The effect of the demodulator used to recover data from amplitude-shift keying (ASK) modulated carriers is also examined, as it affects the impact of crosstalk.

The analytical results presented in Chapter 3 are validated in Chapter 4. Two approaches are taken, one experimental, the other via the use of commercial simulation software. The experimental investigation confirms both in-band and out-of-band crosstalk results for BPSK modulation. The importance of RF frequency allocation to minimise in-band crosstalk is confirmed. Software simulations validate results both for BPSK and QPSK modulation. Measurements of electrical crosstalk are presented together with a theoretical discussion to clarify experimental results at very high crosstalk levels.

Following the theoretical and experimental investigation of optical crosstalk in Chapters 3 and 4, the implications of these results in the context of optical networks is quantified in Chapter 5. Theoretical results are presented from a network designer’s point of view, allowing component optical crosstalk levels to be determined for various modulation formats and optical power penalties. Out-of-band optical crosstalk is shown to potentially limit system capacity depending on out-of-band crosstalk levels, and the important trade-off between in-band and out-of-band crosstalk is highlighted. Results allow the impact of both in-band and out-of-band optical crosstalk to be fully quantified.

Chapter 6 investigates several potential WDM fibre-radio network architectures. A realistic high-capacity radio link operating at 20 GHz is presented, allowing optical specifications to be derived for subcarrier modulation. These parameters are used to provide design examples for star and ring networks, quantifying the component crosstalk requirements. This illustrates the important differences between these networks and identifies critical design issues that must be taken into account when designing future WDM fibre-radio networks. The chapter provides guidelines for network design and identifies key issues.

Chapter 7 gives a summary of the research undertaken for this thesis, together with main conclusions. Suggestions for future work based on the findings of this thesis are also included.
Chapter 1

1.3 ORIGINAL CONTRIBUTIONS

The following original contributions have originated from research undertaken for this thesis:

- Extension of analytical model for in-band crosstalk due to BPSK and QPSK modulation (Chapter 3).
- Identification of the importance of RF frequency for in-band crosstalk minimisation (Chapter 3).
- Confirmation of analytical model via simulation for BPSK and QPSK modulation (Chapter 4).
- Experimental validation of in-band and out-of-band crosstalk results for BPSK modulation (Chapter 4).
- Experimental demonstration of the importance of crosstalk RF frequency offset and electrical filtering for minimisation of in-band crosstalk for BPSK modulation (Chapter 4).
- Establishing the difference between out-of-band optical crosstalk power penalties and co-channel electrical crosstalk power penalties (Chapter 4).
- Quantifying the impact of in-band and out-of-band crosstalk on WDM fibre-radio network capacity (Chapter 5).
- Quantifying the trade-off between in-band and out-of-band crosstalk component specifications (Chapter 5).
- Identifying the effect of the RF frequency re-use plan on the impact of optical crosstalk (Chapter 5).
- Establishing a realistic wireless and optical power budget for a fibre-radio network operating at 20 GHz (Chapter 6).
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- Comparing WDM fibre-radio access network architectures using actual network parameters (Chapter 6).
- Identifying key design issues for various architectures (Chapter 6).

1.4 REFERENCES


2 Literature Review

2.1 INTRODUCTION

As introduced in Chapter 1, fibre-radio networks allow the delivery of high bandwidth wireless data to remote Base Stations (BSs) via an optical feed network. In such feed networks, wavelength division multiplexing (WDM) can provide significant advantages including a very high capacity. In such WDM fibre-radio networks, many WDM optical subsystems such as optical add-drop multiplexers (OADMs) and wavelength demultiplexers will be required to provide link connections. Such components usually do not have perfect isolation between wavelength channels, leading to optical crosstalk. An important consideration that arises in such optical networks is the potential impact of optical crosstalk on signal quality and system performance. The existing fibre-radio literature covering fibre-radio architectures will be reviewed in Section 2.2. Section 2.3.1 reviews research on optical crosstalk for RF subcarrier modulation, in which the RF subcarrier is modulated with the baseband binary data before being intensity modulated onto the optical carrier. The review shows the need for further research, which forms the basis of this thesis. Sections 2.3.2 to 2.3.4 review the literature that analyses optical crosstalk for baseband intensity modulated links. This area has been much more widely researched and provides several analytical techniques that may be used in analysing subcarrier-modulated links.

2.2 WDM FIBRE-RADIO NETWORKS

Wavelength division multiplexing (WDM) will allow future fibre-radio networks to use the large bandwidth capacity of optical fibre to deliver high-bandwidth wireless data services to multiple remote antenna BSs. While WDM technology is widely used in long-haul trunk networks and metro networks, it has yet to break into local access networks. Unlike cable TV optical networks, fibre-radio networks are not
distribution networks in which the same information is broadcast to all BSs, since individual BSs deliver different wireless data, which is digital and not analogue, to and from the CO. By using multiple optical wavelengths, different optical wavelengths are assigned to different BSs in a wavelength-routed fibre-radio network. The allocation of different optical wavelengths to different BSs allows each BS to have access to a much larger data bandwidth than for a network sharing one or more wavelengths. Future requirements for larger wireless data rates naturally leads to the use of WDM in the optical network. WDM minimises the fibre required as multiple wavelengths are transported over a single optical fibre. Network management is easier as different BSs are allocated different wavelengths. Future demand can be met by designing the network for more wavelengths, adding the appropriate equipment as demand increases.

This section reviews published work focusing on WDM fibre-radio networks. Section 2.2.1 reviews papers that use WDM to transmit multiple channels to multiple BSs along a single optical fibre link, focusing on the point-to-point link between the CO and the demultiplexer which feeds the BSs. Section 2.2.2 focuses on WDM star networks, either using a star coupler or using proper wavelength routing through the use of a wavelength division multiplexer (WDM). The literature that uses a ring network architecture for distribution of wireless signals is reviewed in Section 2.2.3, followed by bus networks in Section 2.2.4.

2.2.1 WDM Point-To-Point Links

A generic WDM fibre-radio point-to-point link is shown in Figure 2.1.

![Figure 2.1 Generic point-to-point WDM fibre-radio link.](image-url)
In Figure 2.1, data is modulated onto RF carriers within different frequency bands $f_i$ to $f_F$ before being modulated onto different optical wavelengths $\lambda_i$ to $\lambda_N$. The wavelengths are combined and coupled into a single optical fibre link via a Wavelength Division Multiplexer (WDM). Frequency re-use in the electrical domain is possible, so that all frequency bands $f_i$ to $f_F$ can be re-used and allocated to different wavelengths. This allows the limited wireless spectrum to be re-used in non-adjacent radio cells. The optical link distributes the wavelengths to multiple BSs, although only a single BS is considered in this section. A second WDM at a Remote Node (RN) demultiplexes individual wavelengths and distributes them to individual BSs. The uplink path uses the same multiplexing and demultiplexing technique as the downlink path. A single optical source is modulated at each BS with a frequency band different to that used in the downlink, and the $N$ wavelengths are multiplexed at a RN and sent via a single fibre optical fibre link back to the CO. All wavelengths are demultiplexed at the CO and detected separately.

Examples of the downlink implementation described in Figure 2.1 can be found in [1-4]. Experimental demonstrations are shown in [1, 3, 4] using two or more wavelengths in the 1550 nm band, although only [4] uses a demultiplexer to recover individual wavelengths. The remaining experimental demonstrations use a fibre Bragg grating to remove the unwanted channels.

Both downlink and uplink transmission were considered in [5-9]. Two different WDM schemes were used for the downlink and uplink wavelength allocation. In [5, 8, 9] downlink and uplink wavelengths were in the 1550 nm band, while in [6, 7] the downlink used the 1550 nm band and the uplink used the 1300 nm band. Figure 2.2 shows two possible implementations of downlink and uplink paths.

\begin{figure}[h]
\centering
\includegraphics[width=0.8\textwidth]{wavelength_division_multiplexer.png}
\caption{Two unidirectional links and Single bidirectional link.}
\end{figure}

\textit{Fig. 2.2 Downlink and uplink implementation using WDM multiplexers.}
Figure 2.2 (a) shows how both downstream and upstream paths can be implemented using two separate fibre links and only two WDM multiplexers / demultiplexers [5, 7]. At the CO, downlink wavelengths are multiplexed by the WDM, while arriving upstream WDM channels are demultiplexed by the same WDM. Similarly for the second RN WDM, which demultiplexes downstream wavelengths and sends them to individual BSs, whilst multiplexing wavelengths from different BSs in the uplink direction. This solution requires two $2N\times1$ WDMs (i.e. with $2N$ inputs and a single output) compared to four $N\times1$ WDMs if the approach in Figure 2.1 is used for both downlink and uplink. For a high-capacity WDM link, using two WDMs requires different wavelength bands, potentially halving total capacity. In Figure 2.2 (b) the two separate downlink and uplink unidirectional spans are replaced by a single bidirectional link. This approach is found in [6, 8].

Experimental demonstrations of downlink and uplink paths using the 1550 nm band can be found in [5, 8, 9]. In [5], Smith et al. used three wavelengths in the downlink and a single wavelength in the uplink, using two unidirectional links as shown in Figure 2.2 (a). A similar setup was implemented by Lim et al. [8] although two wavelengths were sent downstream and one upstream via a single bidirectional fibre link, as in Figure 2.2 (b). In [9], Noel et al. demonstrated a WDM link in the downstream direction. Two wavelengths were multiplexed in a WDM at the CO, although only one was modulated. A WDM at the CO demultiplexed the wavelengths. The unmodulated signal was used as the source for the upstream signal, which was transmitted over a separate fibre link. While the downlink is WDM, the uplink is single-wavelength.

Experimental demonstrations of downlink and uplink paths using the 1550 nm band and 1300 nm band can be found in [6, 7]. In [6], Smith et al. only considered a single BS using a single bidirectional fibre link. The 1550 nm band was used for the downlink, while the uplink used the 1300 nm band. A 1300 nm LED was directly modulated at the BS as fibre chromatic dispersion is zero at 1300 nm in single-mode fibre (SMF), enabling the use of broad spectral sources. Imai et al. [7] demonstrated a system comprising a CO and two BSs. Two wavelengths in the 1550 nm band were sent in the downlink, with optical couplers tapped off some of the optical power and
optical filters removed the unwanted wavelength at each BS. A separate uplink path for a single BS used a separate optical fibre and the 1300 nm band.

The literature reviewed above illustrates the potential offered by WDM, typically demonstrating the transmission of at least two wavelengths in the downlink, with a potential uplink path. The authors typically focus on the configuration of the CO or the BS, illustrating a modulation scheme or the implementation of a downlink and / or uplink path. They do not focus on network aspects and do not consider the effect of optical crosstalk. The following sections will review proposals of WDM fibre-radio architectures, although experimental demonstrations only consider a single BS, presumably due to equipment limitations.

2.2.2 WDM Star Networks

One of the common architectures used in access networks is the star topology, shown in Figure 2.3. A CO distributes multiple optical wavelengths to multiple BSs using a WDM link. A passive star coupler or a WDM at a RN splits the signal to multiple BSs, feeding them in a star architecture.

![Generic WDM fibre-radio star network topology.](image)

The WDM star topology shown in Figure 2.3 was proposed by Smith et al. [5]. The architecture is referred to as “star-tree”. Individual BSs were assigned individual wavelengths that were demultiplexed at a RN. The authors identify the possibility of sharing optical sources to feed different stars from the same CO, re-using the WDM optical spectrum in each separate star. The benefits of the star architecture include...
having similar paths from CO to BSs, ensuring a uniform quality of service at each BS. SCM was used within each optical wavelength to provide multiple RF frequencies within each radio cell.

Some proposals of fibre-radio networks using a star architecture consist of a single-wavelength source which is ‘broadcast’ to all BSs using a 1xN optical splitter instead of a WDM demultiplexer [10-13]. In this type of network each BS is allocated a specific subcarrier (for frequency-division multiple access or FDMA) [10], code (for code-division multiple access or CDMA) [11] or timeslot (for time-division multiple access or TDMA) [12, 13].

Multi-wavelength star network architectures using passive splitting are discussed in [14, 15], which discuss protocols and traffic capacity. The general architecture is shown in Figure 2.4.

![Star network topology using passive splitting](image)

**Fig. 2.4 Star network topology using passive splitting.**

Figure 2.4 incorporates two levels of passive splitting using star couplers. Firstly to split the WDM signal into separate branches, and secondly within each branch to cover individual BSs. This is referred to as a “double-star” or tree architecture. Lee et al. [14] described a general WDM/SCM access network which could be applied to fibre-radio in which individual wavelengths are assigned to a group of local BSs, with each one using different subcarrier frequencies. Other wavelengths cover different areas of the network. Similarly in [15], Sharony described how WDM can cover different groups of BSs and how TDMA for each wavelength is used to cover individual BSs. Xiang et al. [16] also proposed a tree architecture. The authors explain that an optical bandpass
filter is used to select wavelengths. Clearly using a WDM demultiplexer would be more suitable as only a single optical component would be needed rather than \( N \) filters, and the coupling loss would be much reduced to the use of a wavelength-selective component rather than passive splitting using a star coupler.

Demonstrations of star networks are given in [4, 5, 10, 11, 16]. Only Smith et al. [5] used multiple wavelengths, with three downstream wavelengths and one upstream wavelength. The remaining experimental demonstrations focused on the link between CO and BS using a single wavelength.

Several protocols have been proposed for fibre-radio star networks that use star couplers or WDM couplers. However, few realistic system demonstrations have been performed and the potential impact of optical crosstalk in such networks has not been discussed. The following sections reviews fibre-radio ring networks.

### 2.2.3 WDM Ring Networks

A generic ring network architecture is shown in Figure 2.5.

![Fig. 2.5 Generic WDM fibre-radio ring architecture.](image)

Multiple optical wavelengths are multiplexed at the CO using a WDM and are sent around the ring in one direction. Individual OADMs are used around the ring to drop individual wavelengths to individual BSs. The same wavelength can be added back into the ring in the uplink direction (note that a different wavelength can also be used). The main difference between ring and star networks lies in the fibre layout, which forms a loop for a ring network, terminating at the CO. A ring network also requires a larger
number of optical components, and has unequal CO-BS link distances. One of the main benefits of a ring architecture lies in its increased reliability if a second fibre is used, allowing a loop-back protection scheme to be used in case of link or node failure [17], as shown in Figure 2.6.

![Figure 2.6](image)

**Fig. 2.6 Single-wavelength bidirectional ring using SCM and CDMA and providing protection and using different wavelength bands for downlink and uplink.**

Fibre-radio ring architectures have been described in [17-19]. Kim et al. [17] considered a single-wavelength ring network using SCM and CDMA, as shown in Figure 2.6. Furthermore, since the same wavelength was destined for all BSs, optical couplers were used at the ring nodes instead of OADMs to tap off some of the optical power to each BS. While a 1550 nm source was used in the downlink, a 1300 nm LED was used for the uplink, requiring a 1550 / 1330 WDM at each BS. Experimental results demonstrated fault restoration in the ring.

A WDM ring was implemented by Heinzelmann et al. [18] using OADMs, as shown in Figure 2.7.
Fig. 2.7 WDM unidirectional ring network using separate downlink and uplink wavelengths.

As shown in Figure 2.7, each OADM comprised two fibre Bragg gratings (FBGs) and two optical circulators. While each BS was allocated a single wavelength, different wavelength bands were used for the downlink and the uplink. Interestingly the authors used an electroabsorption transceiver (EAT) at the BS, which detected the downlink wavelength and which simultaneously modulated the unmodulated wavelength originating from the CO for the uplink. This configuration simplified the BS architecture since all optical sources were at the CO, although it required an additional FBG at each OADM. An additional benefit of using different wavelengths is that in-band crosstalk is eliminated, although an additional Bragg grating is required.

Finally, Lim et al. [19] proposed a wavelength interleaving method that increases the spectral efficiency of a millimetre-wave fibre-radio network. The ring network topology is identical to that in Figure 2.5 with an OADM dropping a wavelength to each BS, with the same wavelength being added back into the ring in the upstream direction. The spectral interleaving technique requires a specially designed OADM and a three-wavelength link with a single ring node demonstrated the feasibility of the technique. A power penalty less than 0.5 dB was caused by the adjacent out-of-band wavelengths. The uplink path was not implemented.

The limited fibre-radio literature on the ring topology reflects the fact that traditional access networks use a star architecture as this minimises the optical component cost, while ring networks are used in metropolitan and longer-haul networks.
due to their greater reliability. This will be further discussed in Chapter 6. Bus network implementations are discussed in the next section.

### 2.2.4 WDM Bus Networks

A bus network can be though of as a variant of a ring network in which the end of the ring is not connected back to the CO, as shown in Figure 2.8.

![Diagram of WDM fibre-radio bus network](image)

Fig. 2.8 Schematic diagram of a WDM fibre-radio bus network.

A downlink fibre links the CO to all BSs, with an OADM at each BS node dropping the required wavelength to the BS. A separate fibre is used in the uplink, with the BS adding a wavelength to the fibre linking all BSs to the CO. The longest link is from the CO to the last BS, with equal uplink / downlink distances, unlike a unidirectional ring.

Jemison et al. [20] considered fibre-radio networks, including a bus network. The authors state that WDM is not viable for fibre-radio networks due to the cost and complexity of optical filtering required at each BS. Whilst this may be correct if each BS requires an optical filter, WDM can still be used in fibre-radio as discussed in Chapter 6. If a single demultiplexer is used as discussed in Section 2.2.1-2.2.3, then no additional optical filters are required at the BSs. Furthermore, WDM allows each BS to have access to a single wavelength, greatly increasing the available bandwidth per BS. Jemison et al. suggested SCM as an alternative, sharing a single wavelength over multiple BSs. Ogawa et al. [21] considered a fibre-radio bus network that uses optical power splitters instead ofOADMs at each BS as a single wavelength is used to transmit multiple RF frequencies. The uplink was not discussed, although presumably different
wavelengths would be used in order to avoid crosstalk. Experiments illustrated transmission techniques. Kojucharow et al. [22-24] considered a WDM bus network in which FBGs are used at each node to drop wavelengths to each BS in the downlink. A separate fibre was used in the uplink to add the same wavelength back into the uplink fibre using a second FBG. Experiments showed downlink transmission using two wavelengths and single-wavelength uplink transmission. Paolella et al. [25] implemented a system demonstration of a WDM bus network using four wavelengths in both downstream and upstream directions. A single wavelength is dropped at each BS, with the same wavelength being added to the separate uplink fibre.

Examples of fibre-radio bus networks demonstrate the functionality of such a type of network for a fibre-radio application. The experiments tend not to focus on the network application or discuss constraints imposed by the network or by optical crosstalk. Only two authors considered the use of WDM [22-25], whilst subcarrier multiplexing was preferred in [20, 21].

2.2.5 Conclusions

While fibre-radio network architectures are discussed in the literature, the focus is typically to present a novel transmission scheme or component. The implications of using a particular network architecture are not discussed, and potential constraints are not identified. Experiments are limited to typically one or two wavelengths at most, presumably due to the difficulty in implementing a larger-scale network.

This section has provided an overview of the literature concerning fibre-radio network architectures. A more detailed discussion of network architectures is presented in Chapter 6, including actual network design examples. As discussed in Chapter 1, optical crosstalk will occur in a WDM network due to the use of imperfect wavelength-selective components. Crosstalk can degrade network performance and must be taken into account when designing a network, so that crosstalk levels and power penalties are acceptable. The following section reviews the literature investigating the impact of optical crosstalk in fibre-radio networks using subcarrier-modulated links. As there is limited research for fibre-radio networks, the review is extended to traditional crosstalk analysis for baseband intensity modulated links.
2.3 Optical Crosstalk

Section 2.2 has provided a review of the literature covering fibre-radio network architectures. While providing insight into different technologies that may be used in WDM fibre-radio networks, optical crosstalk is rarely mentioned. This section presents an overview of the published literature covering the issue of optical crosstalk, i.e. leakage from other optical signal on a detected signal. Section 2.3.1 reviews the literature that considers RF subcarrier modulation, which can be used in a fibre-radio network. Given the limited scope of the existing research, Section 2.3.2 will present the techniques that have been used to analyse and quantify the effects of optical crosstalk for links using baseband intensity modulation. The analysis covers a single source, as well as multiple crosstalk sources. Section 2.3.3 reviews experimental results that quantify the impact of optical crosstalk, both for a single crosstalk term and for multiple crosstalk terms. Section 2.3.4 looks at optical network simulations, in which optical crosstalk is one of several sources of degradations in the link. Other sources of signal degradation include amplified spontaneous emission (ASE) noise due to optical amplifiers, fibre chromatic dispersion, fibre nonlinearities etc. This approach provides a more realistic evaluation of optical networks as multiple sources of degradation are combined. Due to the prohibitive cost of building experimental research networks, a numerical approach is used to simulate such networks. Sections 2.3.2 to 2.3.4 illustrate techniques that can potentially be applied to the analysis of optical crosstalk in WDM fibre-radio networks.

2.3.1 Optical Crosstalk in Fibre-Radio

As discussed in Chapter 1, fibre-radio networks typically transport RF data over an optical backbone. The modulated RF subcarriers can be modulated onto the optical carriers. This is in contrast to standard baseband modulation, in which the data intensity modulates an optical carrier. The difference in techniques is illustrated in Figure 2.9, which shows the two schematic diagrams for the modulators and the difference in modulated optical spectra.
The effect of optical crosstalk for non-baseband modulation was first investigated by Moura et al. in 1997 [26, 27]. The RF carrier is amplitude modulated, so that the modulation format is ASK (amplitude-shift-keying). The modulated RF carrier is then modulated onto an optical signal using a Mach-Zehnder external modulator. The expression for the interferometric noise term is obtained and the moment generating function (MGF) approach is used together with the modified Chernoff bound (MCB) to obtain an upper bound bit-error rate (BER). The moment generating function simplifies the analysis of crosstalk, in particular for multiple crosstalk terms, as a single integration is required, rather than $N$ integrations for $N$ crosstalk terms using other techniques. The solution is also very close to an exact solution. This approach will be further discussed in Section 2.3.2 (E).

Alternatively, the variance of the interferometric noise can be added to the receiver noise variance, the so-called Gaussian approximation, using the error function to evaluate BERs for both signal ones and zeros. A simulated fibre-radio link confirms the analytical results. For multiple crosstalk terms, the Gaussian approximation is shown to be inaccurate at low crosstalk numbers, with the MCB result providing accurate results. Comparing results from this paper to those in [28] for baseband modulation shows that power penalties are identical for a single crosstalk term, with lower power penalties for multiple crosstalk terms. The authors do not comment on this. Since the RF carrier phases are assumed to be random in the analysis, this is believed to be the cause of the lower power penalties at higher crosstalk numbers, as not
all crosstalk terms are in-phase with the signal RF carrier, reducing the effect of interferometric noise in the fibre-radio case.

Mitchell et al. in 2000 [29] used the same approach, namely using moment generating functions and the modified Chernoff bound for BER analysis, to investigate the case of an optical carrier modulated with a quadrature amplitude modulated (QAM) subcarrier. Only BER curves are presented, however, comparing QAM to standard amplitude modulation. QAM is shown to suffer more from crosstalk. The authors identify the fact that crosstalk affects both the amplitude and the phase of the QAM signal as explaining this, since a coherently detected signal is phase-sensitive. It is not clear from the paper how many data points are present in the QAM constellation. Since only a single interferer is considered, RF carriers are presumably aligned, although the potential difference between 4-QAM (QPSK), 8-QAM and higher-order QAM is not addressed.

While optical crosstalk in fibre-radio has been considered in the literature, only ASK modulation had been investigated at the start of this thesis. The analysis did not consider other modulation formats. The analysis only considered the case of multiple reflections where the RF carrier frequency was the same, without considering crosstalk arising between different channels where the RF carrier frequency can be different. Furthermore, no experimental investigation has been performed.

The next sections review research on optical crosstalk for baseband modulation, which has been investigated by many authors, using various techniques. Unlike subcarrier modulation, experimental results are also common.

2.3.2 Analytical Treatment of Optical Crosstalk

This section provides a brief overview of techniques used to analyse the effect of optical crosstalk. The reviewed techniques consider both single and multiple crosstalk sources. Section (A) describes the analytical treatment of a single crosstalk source. In Section (B) numerical techniques that can solve the problem are described, followed by a Gaussian approximation in (C). Analytical approximations are reviewed in Section (D) while Section (E) reviews techniques that provide an exact solution. Other parameters or effects that can change the effect of optical crosstalk are briefly discussed in Section (F). Section (G) shows how the effect of multiple crosstalk can be evaluated
using the Gaussian approximation, various other statistical methods, and the convolution method. A more comprehensive review of optical crosstalk can be obtained in Chapters 2 & 3 of Dr. Sarah Dods’s Ph.D. thesis [30]. Chapter 3 of this thesis provides a detailed analytical description of one of the techniques that can be used to quantify the impact of optical crosstalk, although Chapter 3 considers RF subcarrier modulation compared to baseband modulation. The method is an extension of the model used in [26] and reviewed in Section 2.3.1.

(A) Analytical Formulation

Although optical crosstalk has been widely researched in the literature, the main analytical techniques can be found in [28, 31-39]. The fundamental analysis of optical crosstalk due to a single crosstalk term can be found in [35]. If we consider a desired signal and a crosstalk term injected into a photodiode, the detected photocurrent $I$ at the output of a photodetector is expressed by:

$$I \propto P_1 + xP_2 + 2\sqrt{xP_1P_2} \cos(\Delta \theta)$$  \hspace{1cm} (2.1)

where $P_1$ and $P_2$ are the optical power of the signal and crosstalk respectively, $\Delta \theta$ is the optical phase difference between both fields, and $x$ is the optical crosstalk power ratio, i.e. the ratio of the crosstalk to the signal optical powers. The detected current comprises the signal current from the first term and the crosstalk current from the second term, and an additional term due to the interaction between the two fields. This last term has a maximum amplitude $\pm 2\sqrt{x}$ relative to the normalised signal term. The associated probability distribution of the last term is shown in Figure 2.10 and follows an arcsine distribution [32, 40].
Normalised Received Electrical Signal

*Fig. 2.10* Photocurrent probability distribution showing the effect of optical crosstalk.

At the receiver, the optical power is converted to a current and Gaussian thermal noise is added. This arcsine crosstalk distribution in Figure 2.10 is significantly different to a Gaussian distribution, which describes receiver thermal noise, as the crosstalk distribution is bounded. The optical phase difference $\Delta \theta$ in Equation 2.1 is fixed if there is a fixed relationship between the signal and crosstalk fields, resulting in coherent crosstalk. This occurs if the crosstalk signal is a delayed copy of the signal, provided the time delay is less than the laser coherence time; alternatively the path delay has to be less than the laser coherence length [35]. If the crosstalk signal is from a different laser source, or if the coherence time / length is exceeded, then the crosstalk is incoherent, i.e. the optical phase difference will vary randomly on a timescale related to the laser linewidth [35]. This type of interferometric crosstalk is referred to as “phase-induced interferometric noise” (PIIN) or simply interferometric noise [28, 34-36]. If both signal and crosstalk are intensity modulated with data “1”s and “0”s then there are four possible bit combinations, signal “0” + crosstalk “0”, signal “0” + crosstalk “1”, signal “1” + crosstalk “0” and signal “1” + crosstalk “1”. Assuming perfect extinction ratio (see Section 2.3.2 (F) (ii)), interferometric noise only occurs in the last combination, when both signal and crosstalk are transmitting “1”s. The complimentary error function can be used to evaluate the BER for the four possible cases [41]:

$$
BER = \frac{1}{4} \left[ \frac{1}{2} \text{erfc} \left( \frac{D - I_{00}}{2\sigma} \right) + \frac{1}{2} \text{erfc} \left( \frac{D - I_{01}}{2\sigma} \right) + \frac{1}{2} \text{erfc} \left( \frac{I_{10} - D}{2\sigma} \right) + \frac{1}{2} \text{erfc} \left( \frac{I_{11} - D}{2\sigma} \right) \right] \quad (2.2)
$$
where \( I \) is the detected current and the subscripts “00”, “01”, “10” and “11” denote the signal and crosstalk binary bits respectively, \( \sigma \) is the receiver thermal noise standard deviation, and \( D \) is the receiver decision threshold. The respective currents can be expressed in terms of the signal current \( I_{\text{sig}} \):

\[
\begin{align*}
I_{00} &= 0, \\
I_{01} &= x I_{\text{sig}}, \\
I_{10} &= I_{\text{sig}}, \\
I_{11} &= I_{\text{sig}} \left( 1 + x + 2 \sqrt{x \cos \Delta \theta} \right)
\end{align*}
\] (2.3)

where \( x \) is the crosstalk ratio. The decision threshold \( D \) is typically set to the average of the received signal, a so-called “AC-coupled” receiver, with \( D=0.5 \) [35], i.e.

\[
D = \frac{I_{00} + I_{01} + I_{10} + I_{11}}{4} = \frac{I_{\text{sig}} (1 + x)}{2}
\] (2.4)

Equation 2.2 can be simplified using Equations 2.3 and 2.4:

\[
\begin{align*}
\text{BER} &= \frac{1}{8} \text{erfc} \left( \frac{I_{\text{sig}} (1 + x)}{4 \sigma} \right) + \frac{1}{4} \text{erfc} \left( \frac{I_{\text{sig}} (1 - x)}{4 \sigma} \right) \\
&\quad + \frac{1}{8} \text{erfc} \left( \frac{I_{\text{sig}} (1 + x) + 4 \sqrt{x \cos \Delta \theta}}{4 \sigma} \right)
\end{align*}
\] (2.5)

Equation 2.5 shows that the BER depends on the crosstalk level \( x \) for all four bit combinations but also on the optical phase difference \( \Delta \theta \) for the “11” case. For the case of incoherent crosstalk, the final term must be averaged over the range \( \{0,2\pi\} \) [32]:

\[
\text{BER}(11) = \frac{1}{2\pi} \int_0^{2\pi} \frac{1}{2} \text{erfc} \left( \frac{I_{\text{sig}} (1 + x) + 4 \sqrt{x \cos \Delta \theta}}{4 \sigma} \right) d\Delta \theta
\] (2.6)

Once Equation 2.5 has been evaluated for incoherent crosstalk, BER curves versus received optical power can be plotted at different crosstalk values, allowing power penalties to be calculated [35].

Solving Equation 2.6 is the main difficulty encountered when analysing the impact of optical crosstalk, and various techniques have been used to deal with this. Numerical techniques are reviewed in Section 2.3.2 (B), the Gaussian approximation is reviewed in Section 2.3.2 (C), analytical approximations are reviewed in Section 2.3.2 (D) and exact analytical techniques reviewed in Section 2.3.2 (E). Section 2.3.2 (F) considers the importance of the decision threshold, the extinction ratio, and the effect of
ASE noise. The study of crosstalk due to multiple sources is reviewed in Section 2.3.2 (G).

(B) Numerical Evaluation

One possible technique that can be used to numerically solve Equation 2.6 is simply to perform the integral numerically. Legg et al. [35] use this approach as one of the techniques to evaluate the impact of optical crosstalk. This is a simple technique that gives an accurate solution unlike the Gaussian approximation discussed below.

Another approach is to consider the probability density functions of the interferometric noise and the Gaussian receiver thermal noise. The probability distribution function (pdf) for “00” (signal and crosstalk ‘0’ bits), “01” (signal ‘0’, crosstalk ‘1’) and “10” cases is Gaussian and for the “11” case is the convolution of the Gaussian noise distribution and the “arcsine” interferometric noise distribution. The resulting pdfs can be used to evaluate the probability of error. This can be performed numerically and also yields an accurate solution. Legg et al. [35], Cornwell et al. [36] and Moura et al. [28] all calculate power penalties using numerical convolutions.

(C) Gaussian Approximation

The Gaussian approximation is widely used in the literature and was one of the first approximations used to evaluate optical crosstalk. While the exact pdf of the interferometric noise term is an arcsine distribution [40] as shown previously in Figure 2.7, it can be approximated as a Gaussian distribution with the same mean and standard deviation. Although this clearly is not an accurate representation of a bounded distribution, it greatly simplifies the evaluation of BERs. Since both interferometric and thermal noise are treated as Gaussian, their convolution results in a Gaussian whose variance is the sum of the separate variances. The variance of the “11” case is thus:

\[ \sigma^2 = \sigma_{th}^2 + 2xI_{sig}^2 \]  

(2.7)
i.e. the sum of the thermal noise variance and the variance of the arcsine distribution. This approach is discussed in [28, 32, 35, 37]. As observed in the literature, the Gaussian approximation is accurate for a single crosstalk source at low crosstalk levels where receiver thermal noise dominates but not at larger crosstalk levels, where it overestimates power penalties [28, 32, 35, 37]. The Gaussian approximation is also typically used to evaluate the case of multiple crosstalk terms as the convolution of
arcsine distributions tends towards a Gaussian (central limit theorem). This will be discussed in Section 2.3.2 (G).

(D) **Analytical Approximation**

Approximate expressions for the power penalty as a function of crosstalk level can be found in [32, 33, 42, 43]. Gimlett *et al.* [32] in 1989 considered the dominant source of degradation as being due to the “11” condition and obtained the following expression for the power penalty *PP* as a function of crosstalk level *x*:

\[
PP(\text{dB}) = -5\log \left(1 - 4Q^2 x \right)
\]

(2.8)

where *Q* defines the probability of error \( P_e = \frac{1}{2} \text{erfc}(Q/\sqrt{2}) \). This is based on using the Gaussian approximation and finding the total variance of crosstalk and thermal noise. The same expression can be found in [33, 42]. The usefulness of the expression is limited by the Gaussian approximation, which overestimates power penalties at high crosstalk, as mentioned previously.

A more accurate approach involves using BER expressions as in Equations 2.2 and 2.3 and making suitable approximations. This typically involves simplifying the expression for the current due to the “11” case to always assume the worst-case optical phase difference, and obtaining a power penalty by taking the ratio of *Q* with crosstalk and *Q* without. This approach was taken by Goldstein *et al.* [33] and considers different decision thresholds. For an AC-coupled receiver, the worst-case power penalty is given by:

\[
PP(\text{dB}) = -10\log \left(1 - 6\sqrt{x} \right)
\]

(2.9)

Goldstein *et al.* do not explain the approximations made or justify them and the validity of the approximations can be questioned at large crosstalk. While the predicted power penalties are lower than those for a Gaussian approximation, they are actually larger than those predicted by an exact analysis. This is expected as the worst-case optical phase difference is used.

Although the above expressions allow power penalties to be easily obtained as a function of crosstalk level they are not accurate. Exact results can be used instead, either using a numerical approach as discussed in (B) or an exact analysis as discussed in (E) below.
(E) **Exact Analysis**

The exact analysis described in this section entails using a series expansion for the arcsine distribution pdf and using it to obtain an accurate expression for the BER at a given crosstalk level. This approach has been taken by Ho [37, 38] and Dods *et al.* [44].

Ho *et al.* [37, 38] use Hermite polynomials to obtain a series expansion of the interferometric noise pdf and obtain a closed-form BER expression that considers the four possible signal-crosstalk bit combinations. The difference between the exact analysis and the Gaussian approximation is shown and power penalties obtained using the exact analysis are shown to be lower than the Gaussian approximation and very close to experimental results [37].

Dods *et al.* [30, 44] extended the analysis of Ho *et al.* using Hermite polynomials, obtaining a closed-form BER expression considering the case of a signal corrupted by two crosstalk sources. The predicted power penalties are shown to be in excellent agreement with numerical techniques.

Another technique that is more approximate but still results in very accurate power penalties for a single crosstalk term is to use Gram-Charlier series to approximate the arcsine pdf [39]. Ho shows that provided enough terms are used the approximation is accurate even at large crosstalk levels.

Another approach that yields accurate results for single and multiple crosstalk terms is to use moment generating functions (MGFs) to obtain BERs [28, 45-49]. Moura *et al.* [28] show that even for a single crosstalk term the modified Chernoff bound used with MGFs gives very accurate results, indistinguishable from exact results using convolutions. Monroy *et al.* also use MGFs together with a saddlepoint approximation to provide similar results [45, 47]. Danielsen *et al.* use the same method [48] confirming previously published results.

(F) **Other Factors**

(i) **Decision threshold**

The importance of the decision threshold in determining the BER is shown schematically in Figure 2.11, which shows the Gaussian receiver noise pdfs of data ‘0’ s and ‘1’ s.
Figure 2.11 Decision threshold used to determine data ‘0’ and ‘1’.

Figure 2.11 does not show the non-Gaussian pdf for ‘11’ bits for clarity; the correct pdf is the convolution of the interferometric noise arcsine pdf and the Gaussian receiver noise pdf. The decision threshold is used to determine whether the received data is a ‘0’ (current below threshold) or ‘1’ (current above threshold). A standard decision threshold is an AC-coupled threshold that sets the threshold to the average of all received bits. In this case the decision threshold $D$ is equal to $0.5(1+x)$ [33, 35], as shown in the figure (this is usually referred to as 0.5 i.e. half the maximum value). This is the optimum threshold if only Gaussian noise is present at the receiver. The presence of optical crosstalk means that the ‘11’ signal-crosstalk combination has a different, non-Gaussian distribution. Hence when crosstalk is present, a more optimum threshold can be found, reducing the resulting power penalty. The importance of decision threshold is recognised in the literature [28, 33, 35], with some authors using optimised decision thresholds [28, 35, 42, 50-52]. The practical implementation of an optimum decision-threshold circuit may be challenging. Note that in BPSK and QPSK modulation, only the phase is modulated so the decision threshold is always zero, whilst for ASK modulation the amplitude of the RF carrier is modulated. For QAM, both amplitude and phase are modulated, so the decision threshold is important.

(ii) Extinction ratio

When analysing the impact of optical crosstalk in intensity modulated links, some of the literature mirrors that which was described in Section 2.3.1 by assuming that the ‘0’ level contains no optical power [32, 33, 35, 36]. This is referred to as
having a perfect or ideal extinction ratio, where the extinction ratio is either 0 or infinite depending on its definition. The extinction ratio is the ratio of optical power in ‘0’s to ‘1’s or vice-versa. If there is some residual signal transmitted for a data ‘0’ then this results in additional interferometric noise being present for the ‘00’, ‘01’ and ‘10’ signal-crosstalk binary combinations, increasing the resulting power penalty [35, 46, 47, 51-56].

(iii) Amplified receiver / ASE noise

The literature typically considers a photodetector that converts the signal and crosstalk optical powers into electrical current. The receiver contributes thermal noise that limits the sensitivity of the receiver. The effect of optical crosstalk is then evaluated as described in prior sections. Using optical amplifiers either in the optical link or as part of a pre-amplified receiver adds amplified spontaneous emission noise (ASE) which is broadband. Upon photodetection, beat products between the ASE noise and itself (spontaneous-spontaneous beat noise) and between the signal and ASE noise (signal-spontaneous beat noise) provide additional signal-level dependent noise.

Eskildsen et al. in 1997 [55] evaluated the BER due to the four possible signal-crosstalk bit combinations using arcsine statistics and considering the impact of signal-spontaneous beat noise (modeled as an additional variance). An optimised decision threshold is used and both theoretical and experimental results show that the power penalties are reduced relative to a non pre-amplified receiver. However it is not clear whether the comparison is justified as previous results seem to have been for a non-optimised decision threshold. Indeed, the effect of optimising the decision threshold is to improve performance and reduce the resulting power penalty. Rasmussen et al. in 1999 [53] use the exact arcsine pdfs for an amplified receiver and show that power penalties are worse for an amplified receiver, contradicting Eskildsen et al. Experimental measurements confirm their predictions, taking into account non-ideal extinction ratio. The authors do not mention the decision threshold, which is presumed to be non-optimised. As both experimental and theoretical predictions comparing a standard PIN receiver and a pre-amplified receiver show a higher penalty for the amplified case, these results are more credible. Zami et al. in 1999 [54] also confirm this prediction experimentally, showing an increasing penalty as the level of ASE noise
increases. The importance of decision threshold is highlighted, with an optimum threshold increasing the tolerance to crosstalk.

\section*{(G) Multiple Crosstalk Analysis}

The previous sections have considered the analytical treatment of crosstalk with a single crosstalk term. Depending on network topology, there may be more than one crosstalk term. This has been widely investigated in the literature. Two fundamental approaches can be found. A simple and popular technique is to use the Gaussian approximation as discussed in Section 2.3.1 (C) above. A much less widely used approach is to use other statistical techniques such as moment generating functions. Finally, the exact arcsine pdf can be convolved with itself to produce an exact pdf. The three different approaches are briefly discussed below.

\subsection*{(i) Gaussian approximation}

As discussed in Section (C), while the signal-crosstalk interferometric term depends on the relative optical phase difference, resulting in an arcsine pdf, the evaluation of BERs and power penalties is greatly simplified if the distribution is approximated as being Gaussian [32, 35]. The analysis is further simplified if the average variance $\sigma_t^2$ is calculated for $n$ crosstalk term as follows:

$$\sigma_t^2 = \sigma_{th}^2 + \frac{n}{2} \sigma_{stalk}^2$$

where $\sigma_{th}^2$ is the thermal noise variance and $\sigma_{stalk}^2$ is the signal-crosstalk interferometric noise variance. The factor of one half accounts for the fact that for binary data there are on average only $n/2$ crosstalk terms transmitting data ‘1’s at any time [32, 42]. The variance of the arcsine distribution is given by:

$$\sigma_{stalk}^2 = 2xI_{sig}^2$$

where $I_{sig}$ is the received signal current and $x$ is the linear crosstalk ratio.

Although the Gaussian approximation is not accurate for large crosstalk, for a single crosstalk term, or for a small number of crosstalk terms, it is applicable when there are sufficient number of crosstalk sources to render the resulting pdf Gaussian via the central limit theorem. This is because the convolution of the arcsine pdfs tends
towards a Gaussian distribution. The Gaussian approximation overestimates the effect of crosstalk for a small number of crosstalk terms [28, 35, 42, 57].

The effect of using a Gaussian approximation for a low number of crosstalk terms can also be observed when comparing the power penalty for a given total crosstalk for different number of crosstalk terms [36, 39, 49, 58]. Results show that whereas the Gaussian approximation predicts a decreasing power penalty as the number of interferers is increased for a given total crosstalk, an exact analysis predicts the opposite [39, 49]. The number of crosstalk terms required for the Gaussian approximation to predict power penalties that are close to an exact analysis is shown in the literature to be greater or equal to 10 [39] or greater than eight [28].

(ii) Other statistical approaches

As discussed in (E) above, other statistical techniques have been used to evaluate the impact of optical crosstalk. These include using Gram-Charlier series [39] to approximate the arcsine pdf, and using moment generating functions (MGFs) [28, 47-49]. Mitchell et al. confirm the predicted results made using MGFs using simulation software [49]. Moura et al. show agreement with exact analysis for one crosstalk term and convergence with the Gaussian approximation with a large number of terms [28]. Ho confirms the validity of the Gram-Charlier series approach using the same comparisons [39].

(iii) Convolution method

As interferometric noise is described by an arcsine pdf, the pdf can be convolved with itself multiple times to obtain the pdf describing the interferometric noise due to multiple crosstalk terms [36, 53], giving exact results.

2.3.3 Experimental Investigations

This section provides an overview of the literature that measures the effect of optical crosstalk experimentally. In order to simplify categorisation, Section 2.3.2 (A) will consider single crosstalk experiments, while Section 2.3.2 (B) will review experiments for multiple crosstalk terms.
(A) **Single Crosstalk Experiments**

Examples of experiments investigating the effect of a single crosstalk term can be found in [33-37, 40, 45, 51, 54, 55, 59, 60]. A schematic diagram of a typical experimental set-up used to measure the effect of crosstalk is shown in Figure 2.12.

![Schematic diagram](image)

**Fig. 2.12** Schematic diagram of experimental set-up used to measure the effect of optical crosstalk.

Figure 2.12 shows a single laser source that is externally modulated by a pseudo-random bit sequence (PRBS) and split into two different optical paths using an optical coupler. The two paths are optically decorrelated in order to obtain incoherent crosstalk using a length of optical fibre that exceeds the laser coherence length. The two paths are recombined, with a variable optical attenuator (ATT) used to set the crosstalk level. The combined signal is fed to a photodetector, typically a PIN diode. The optical power into the detector is controlled by a variable optical attenuator and monitored using an optical power meter (OPM). The electrical output of the photodetector is then fed to a bit-error-rate testset (BERT). BER curves versus optical power at a given crosstalk level are used to calculate power penalties as a function of crosstalk level.

Tur *et al.* in 1989 and 1990 [34, 40] measured the effect of optical crosstalk relative to the ratio of thermal Gaussian receiver noise, showing that the arcsine crosstalk statistics resulted in different BER characteristics. In 1994, both Goldstein *et al.* [33] and Legg *et al.* [60] measured power penalties for a single crosstalk term. Power penalties in [60] are for a directly modulated distributed-feedback laser (DFB) and are lower than in [33] in which external modulation is used. This is explained by
the effect of laser linewidth and electrical filtering which can lead to reduced penalties for directly modulated sources [45, 59]. Legg et al. [35] and Song et al. [51] show that a non-perfect extinction ratio results in larger power penalties (see Section 2.3.1 (F)(ii)), although in [51] the laser is directly modulated. Power penalties for an amplified receiver were provided by Eskildsen et al. in 1997 [55] with an optimised decision threshold. In 1999, Zami et al. [54] showed that the presence of ASE noise increased the measured power penalty. As mentioned previously, both Jin et al. [59] and Monroy et al. [45] in 1999 compared the effect of crosstalk for externally modulated and directly modulated sources. Power penalties are reduced for direct modulation due to the effect of electrical filtering at the receiver.

The above experiments investigate different effects and can not all be compared. Furthermore experimental uncertainties lead to measured power penalties that can differ from those predicted from theory, although this is not always clear in the literature. Note that the above experiments focus purely on optical crosstalk; other experimental investigations of optical crosstalk consider specific optical components such as arrayed-waveguide gratings (AWGs) [42, 61] and OADMs [62-65].

(B) Multiple Crosstalk Experiments

The impact of optical crosstalk can be evaluated experimentally using the same approach as shown in Figure 2.9 for a single crosstalk term by increasing the number of crosstalk paths, ensuring appropriate decorrelation using different lengths of fibre. This approach can be found in [42, 47, 52, 53, 66]. Gimlett et al. in 1989 [32] measured the effect of crosstalk due to single and multiple optical reflections. Large experimental uncertainties arise from the uncertainty in reflection coefficient. Polarisation states were not aligned for three reflections. Goldstein et al. in 1995 [66] measured the penalty due to up to eight crosstalk terms, comparing results with a Gaussian approximation. Results are slightly higher than theoretical predictions, even at low crosstalk where the Gaussian approximation is valid, suggesting experimental errors. In 1996, Takahashi et al. [42] measured the penalty due to a single crosstalk term and confirmed they were lower than those predicted by a Gaussian approximation. The validity of the Gaussian approximation for sixteen crosstalk terms was confirmed, showing the effect of decision threshold and random polarisation alignment. The authors used a 16x16 AWG to measure the power penalty due to 16 crosstalk terms,
showing the average total crosstalk predicted the resulting power penalty. In 1998, Monroy et al. [47] measured power penalties due to one, two and three crosstalk sources showing very good agreement with exact statistics. Power penalties are much lower than predicted using the Gaussian approximation. Rasmussen et al. [53] in 1999 performed similar measurements to Goldstein et al. [66] for one to six crosstalk sources. Both standard photodetector and pre-amplified detector are measured and compared. Unlike Goldstein et al., however, results are in excellent agreement with exact statistics, with ten crosstalk terms required to obtain Gaussian statistics.

Another technique used to study the effect of cascaded links is to use a recirculating loop. In this technique the signal is looped multiple times, with each loop adding crosstalk, as shown in Figure 2.13.

![Recirculating Loop Diagram](image)

**Fig. 2.13 Experimental investigation of optical crosstalk using a recirculating loop.**

Figure 2.13 shows how the setup shown in Figure 2.12 can be modified so that multiple circulations through a loop are possible, increasing the number of crosstalk terms at each loop. This technique can be found in [52, 57, 67]. In 1998, Liu et al. [52] used a recirculating loop with four crosstalk paths to verify the accuracy of an expression using Gaussian statistics taking into account extinction ratio. Results for up to four loops (16 crosstalk terms) show the importance of extinction ratio and the accuracy of the Gaussian approximation at low crosstalk and for multiple crosstalk terms. Caspar et al. in 2000 [67] consider the effect of an optical cross-connect in which in-band crosstalk is introduced. Each circulation totals 162 km of fibre. A transparency length, corresponding to the distance traversed for a 3 dB power penalty, is measured and compared to simulations using a Gaussian approximation. Results show good agreement at low crosstalk, although experiments give a lower transparency length at
high crosstalk. The authors state correctly that the Gaussian approximation should result in a lower transparency length. The pdf of data ‘1’s was seen to be non-Gaussian at high crosstalk. This behaviour is likely to be accounted for by the accumulation of amplifier ASE noise [53-55], although the authors do not mention this. Moon et al. [57] in 2001 used a slightly different experimental set-up in which a loop is used to produce multiple crosstalk terms, with each loop adding additional crosstalk. Experimental results at different crosstalk levels for multiple recirculations show good agreement at low crosstalk and with multiple recirculations when compared with the Gaussian approximation. Polarisation states are not controlled and assumed to be random.

Most experimental investigations show good agreement with theory, demonstrating the validity of the Gaussian approximation at low crosstalk for a small number of crosstalk terms and for a large number of crosstalk terms.

2.3.4 Optical Crosstalk in Networks

While the above experimental demonstrations demonstrate the effect of multiple crosstalk terms and the validity of the Gaussian approximation for multiple crosstalk terms, the results are not put in the context of a real network. This sections briefly reviews the published results on optical crosstalk from a network point of view. The nature of the problem means that the studies are numerical, using appropriate simulation software. Examples of this approach can be found in [31, 68-79].

The most popular approach is to use the Gaussian approximation to obtain analytical expression, which are then used to solve a given network configuration numerically [68, 70, 76-79]. Different statistical or analytical approaches are used in [69, 71-75, 80].

(A) Gaussian Approximation

Examples of investigations using the Gaussian approximation include the study of multiple OADMs [77, 78] and multiple optical cross-connects [68, 70, 76, 79]. Yu et al. [78] consider an \( N \times N \) multiplexer / demultiplexer (MUX / DEMUX) that is used to selectively add/drop wavelengths in a network. Fibre nonlinearities and ASE noise are also taken into account, showing that component crosstalk can limit transmission distance if such a component is cascaded. Pires et al. [77] perform a similar analysis for
a 32x32 AWG OADM considering both crosstalk and ASE noise. Results show that strict crosstalk requirements are needed to ensure 15 OADMs to be cascaded. Ramamurthy et al. [76] study the effect of cross-connect switches in both bidirectional ring and mesh networks, including the effect of ASE noise and crosstalk. Random polarisation is assumed. Different cross-connect configurations are compared by Gillner et al. [79] using a technique described in [68]. The effect of crosstalk on network size is quantified, limiting the number of nodes in a network. Shen et al. [70] consider cascaded cross-connects, including the effect of coherent crosstalk. As discussed in [68, 79], the probability that a certain power penalty is exceed is proposed as a way to relax otherwise stringent crosstalk requirements, allowing more cross-connects to be traversed in most cases.

(B) Exact Techniques

Numerical investigations involving exact formulae are found in [69, 71-75, 80]. Blumental et al. [69] study the effect of multiple crosstalk terms arising from space switches, comparing different switch configurations. Exact electric field expressions are used for 8x8 and 32x32 switches. The authors compare incoherent crosstalk with random binary data (binomially distributed), incoherent crosstalk and crosstalk data “1”s and coherent crosstalk with crosstalk terms $\pi$ out-of-phase (worst-case) and always “1”. Results indicate worst-case is more likely for 8x8 switches than 32x32 switches due to the lower number of crosstalk terms. Dods et al. [72, 73, 80] studied the combined effect of coherent and incoherent crosstalk in ring and bus networks [72, 80] and compared several OADM structures in [73]. Whereas power penalties are fixed for incoherent crosstalk for a given crosstalk level, if additional coherent crosstalk is present then the overall power penalty will depend on the phase of the coherent term. As the optical phase difference is unknown and may vary with time, the effect is to produce a distribution of power penalties. The combined effect of coherent and incoherent crosstalk is shown to impose strict requirements on crosstalk specifications. Shen et al. [71] perform a similar study of ring/bus networks although they assume worst-case coherent crosstalk. The authors evaluate power penalties for incoherent crosstalk only, and add a varying number of coherent crosstalk terms. As coherent crosstalk is introduced at each node, compared to only one node for incoherent crosstalk, results show that as the number of nodes increases coherent crosstalk can
rapidly dominate the overall power penalty. Antoniades et al. [74] consider two ring networks in which each node has OADMs that are interconnected using cross-connects. The effects of amplifier ASE noise and fibre nonlinearities are included. Crosstalk analysis based on the work of Ho using Gram-Charlier series [39] is used to determine an acceptable component crosstalk level for a 0.5 dB power penalty. In [75], Antoniades et al. describe the use of computer simulations for the design of a passive ring metropolitan network. The model uses simple expressions to include the effects of chirp and dispersion, polarisation mode dispersion (PMD) and model the effects of fibre nonlinearities and polarisation-dependent loss (PDL). Analytical expressions are used to derive signal-crosstalk pdfs using Hermite polynomials and moments. The dominant crosstalk term is found to come from the OADM at each node in which the same wavelength is dropped from the ring and added back into the ring. Channel power differences are found to be critical as the effective crosstalk can be higher than the component crosstalk.

Two different approaches have been used in the literature when studying the impact of optical crosstalk in networks. The Gaussian approximation is often used as it is valid when a sufficient number of crosstalk terms are present. More exact techniques can be used, however, given the accuracy of numerical techniques in computer simulations. The validity of results in the literature depends on the validity of the models and approximations used, so that care must be taken by the reader to understand the approaches taken. Network evaluations are useful in identifying limiting factors and comparing different sources of signal degradations. In all cases the accumulation of optical crosstalk due to cascaded optical components will ultimately limit network size.

2.4 CONCLUSIONS

Section 2.2 has presented an overview of the literature covering fibre-radio network architectures. While the main focus of the work has been on actual downlink and uplink implementations, several authors discuss the fibre-radio architecture. The standard star, ring and bus network architectures have all been proposed as potential architectures for future fibre-wireless networks. A more comprehensive review of architectures, including design examples, and implications of optical crosstalk will be
discussed in Chapter 6. Optical crosstalk arises in WDM optical networks and must be considered when designing a fibre-radio network.

From an optical crosstalk perspective, many techniques have been developed to study the impact of crosstalk for standard baseband intensity modulation. This has been reviewed in Section 2.3. The literature focusing on the analytical treatment of crosstalk due to single or multiple crosstalk terms was reviewed, including experimental investigations. Optical network simulations, in which optical crosstalk is one of many impairments considered, have also been reviewed.

The very limited literature considering the case of subcarrier modulation used in a fibre-radio network has been presented. Whilst power penalty results for ASK are identical to the standard crosstalk analysis, QAM modulation was found to be worse than ASK, although power penalties were not shown. It is not clear whether or not this applies to all QAM constellations. Furthermore, no experimental investigations have been performed. These limited investigations have motivated the research undertaken in this thesis. Chapter 3 will present the analytical model that is based on the analysis by Moura et al. [26, 27], providing results for BPSK and QPSK modulation. The importance of the RF phase and the potential for different RF frequencies to be used in networks is studied. Chapter 4 investigates crosstalk experimentally and by simulation, with Chapter 5 discussing the impact of optical crosstalk on network capacity. Chapter 6 concludes with an overview of potential fibre-radio architectures and gives design examples that illustrate the effect of optical crosstalk in WDM fibre-radio networks.

### 2.5 References


Chapter 2


[80] S. D. Dods, J. P. R. Lacey, and R. S. Tucker, “Homodyne crosstalk in WDM ring and
Analysis of In-band and Out-of-band Optical Crosstalk for Subcarrier Modulation

3.1 INTRODUCTION

The previous chapter has introduced key concepts in fibre-radio networks and highlighted the lack of research on the area of optical crosstalk in these networks. Optical crosstalk may be critical if low-cost optical components are used, so further research is required to quantify the significance of crosstalk in fibre-radio systems. The impact of optical crosstalk in fibre-radio networks depends on the nature of the modulation scheme used to transmit data via an optical network. Different transmission schemes have been proposed, as discussed in Chapter 1. The data can either be sent at baseband (intensity modulation or IM) or at an intermediate or RF frequency via subcarrier modulation.

The impact of optical crosstalk has been thoroughly analysed for baseband intensity modulation [1-3], as reported in Chapter 2. However, for subcarrier modulation in which the electrical data is modulated onto an electrical carrier before being modulated onto an optical carrier, limited research has been published. As reviewed in Chapter 2, Moura et al. have analysed in-band crosstalk for amplitude-shift-keying (ASK) [4] and Mitchell et al. have considered quadrature-amplitude modulation (QAM) [5]. Moura’s analysis considered the impact of incoherent in-band crosstalk due to multiple reflections along an optical link. Results show the optical power penalties are the same as for baseband intensity modulation (IM), assuming in-phase RF carriers. For multiple crosstalk terms, Moura assumes a random distribution of RF carrier phases. For QAM modulation, Mitchell analysed in-band crosstalk from a single source. Mitchell showed the effect was worse than for ASK modulation.

In Section 3.2 the model presented by Moura et al. is modified to analyse in-band optical crosstalk for Binary Phase-Shift-Keying (BPSK) modulation, providing insight into the importance of the RF carrier phase difference. We show that power
penalties are significantly different [6]. Section 3.3 shows that if in-band optical crosstalk arising from a different optical source, which carries an electrical subcarrier at a different RF frequency, the impact of a given level of optical crosstalk is reduced. This is achieved through electrical filtering, although this does not completely eliminate the optical power penalty. ASK modulation for in-band crosstalk is considered in Section 3.4 using both envelope detection of the carrier amplitude and the use of a phase-locked loop (PLL) and electrical downconversion as for BPSK demodulation. The analysis used for BPSK modulation is also extended to QPSK modulation in Section 3.5. Out-of-band optical crosstalk for ASK, BPSK and QPSK modulation is also analysed in Sections 3.2, 3.3, 3.4 and 3.5. The important case of out-of-band optical crosstalk where the RF carrier frequency is different is identified, allowing electrical filtering to remove the crosstalk signal in the electrical domain. The important issue of RF carrier phase variations is addressed in Section 3.6, since this defines the timescale over which the RF phase varies and determines whether or not a DC-coupled decision threshold can track these. ASK, BPSK and QPSK modulation formats are compared in terms of their tolerance to in-band and out-of-band optical crosstalk in Section 3.7. Section 3.8 briefly investigates whether electrical crosstalk in the RF signal used to modulate the optical carriers has an impact on the optical crosstalk results. Table 3.1 illustrates the types of optical crosstalk and modulation formats studied in this chapter, indicating the sections in which they are addressed.

Table 3.1 Modulation formats and types of optical crosstalk studied in this chapter – numbers indicate the section in which the investigation can be found.

<table>
<thead>
<tr>
<th></th>
<th>In-band Crosstalk</th>
<th>Out-of-band Crosstalk</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Frequency reuse</td>
<td>Adjacent frequency</td>
</tr>
<tr>
<td>BPSK</td>
<td>3.2.1</td>
<td>3.3.1</td>
</tr>
<tr>
<td>ASK</td>
<td>3.4.1</td>
<td>3.4.2</td>
</tr>
<tr>
<td>QPSK</td>
<td>3.5.1</td>
<td>3.5.2</td>
</tr>
</tbody>
</table>

Table 3.1 shows that both in-band and out-of-band optical crosstalk are analysed for BPSK, ASK and QPSK modulation. In each case, the impact of crosstalk is analysed
when the signal and crosstalk RF frequencies are the same (frequency reuse) and when they are different (adjacent frequency).

3.2 IN-BAND AND OUT-OF-BAND OPTICAL CROSSTALK FOR BPSK MODULATION WITH FREQUENCY RE-USE

This section presents the analytical expressions used to quantify the impact of in-band and out-of-band optical crosstalk for BPSK modulation. With frequency re-use both signal and crosstalk wavelengths transport data at the same RF carrier frequency. The importance of the relative phase difference between the two carriers is highlighted, and the behaviour of errors is investigated as this varies from 0° (in-phase) to 90° (out-of-phase). The potential benefits of forward-error-correction (FEC) [7] are also quantified.

3.2.1 In-band Optical Crosstalk with Frequency Re-use

A simple model of in-band crosstalk is developed considering the presence of a signal wavelength and one crosstalk wavelength, for an externally modulated optical link where a modulated subcarrier at frequency \( \omega_{RF} \) drives an external Mach-Zehnder modulator. The electric field associated with the desired and crosstalk optical signals that are at the same wavelength can then be expressed as:

\[
E_1(t) = |E_1| \cos(\alpha + \theta_1(t))[1 + m_1\alpha(t)\cos(\omega_{RF}t + \phi_1(t))]^{1/2},
\]
\[
E_2(t) = \sqrt{\lambda}|E_1| \cos(\alpha + \theta_2(t))[1 + m_2\beta(t)\cos(\omega_{RF}t + \phi_2(t))]^{1/2}
\]

(3.1)

where \( E_1, E_2 \) are the electric fields of the optical signals, \( \omega \) is the optical frequency, \( m_1 \) and \( m_2 \) are modulation indices, \( \alpha(t) \) and \( \beta(t) \) are the signal and crosstalk data, \( x \) is the optical crosstalk power ratio, \( \theta_1(t) \) and \( \theta_2(t) \) represent the optical phase, and \( \phi_1(t), \phi_2(t) \) are the electrical phases. The optical crosstalk ratio is defined as the ratio of the crosstalk optical power to the signal optical power. For simplicity, we consider a single wireless frequency, however the analysis also applies to multiple subcarriers assuming identical modulation indices, no electrical nonlinearities and perfect electrical filtering of undesired electrical frequencies.

If the in-band crosstalk is also optically coherent, then a fixed relationship exists between the two optical phase angles [2]. However, this would only occur if the
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crosstalk signal was a delayed version of the signal wavelength, i.e. if both signal and crosstalk wavelengths originated from the same laser. Furthermore, the delay would have to be less than the coherence length of the laser. In our analysis, we consider incoherent crosstalk in which signal and crosstalk optical fields originate from different lasers, so that there is no fixed phase difference between optical carriers. Incoherent crosstalk between different in-band channels can occur in optical networks and can result in large penalties. Coherent crosstalk, on the other hand, changes the optical power, without causing a degradation, provided the time delay is much less than a bit period [8].

The electrical phase difference depends on the RF carrier phase, which depends on the local oscillator (LO) phase as well as free-space [9] and optical fibre propagation delays [10, 11], as well as any effects due to the mobility of the mobile user. This will be addressed in Section 3.6 in more detail.

At the photodetector, the electric fields are squared, resulting in mixing terms arising from the beating between signal and crosstalk fields, as shown in Figure 3.1.

---

**Fig. 3.1 Optical and electrical spectra for in-band crosstalk with the same frequency bands.**

The resulting photocurrent, \( I \), will be proportional to the square of the signal and crosstalk electric fields:
Analysis of In-band and Out-of-band Optical Crosstalk for Subcarrier Modulation

\[ I \propto (E^2) = |E_1 + E_2|^2 \]
\[ = |E_1|^2 + |E_2|^2 + 2|E_1E_2| \]
\[ = E_1^2 \cos^2(\alpha + \theta_1)[1 + m_1\alpha(t)\cos(\omega_{RF}t + \phi_1)] \leftarrow \text{Signal} \]
\[ + xE_2^2 \cos^2(\alpha + \theta_2)[1 + m_2\beta(t)\cos(\omega_{RF}t + \phi_2)] \leftarrow \text{Crosstalk} \]
\[ + 2\sqrt{x}E_1^2 \cos(\alpha + \theta_1)\cos(\alpha + \theta_2) \times \leftarrow \text{Signal - Crosstalk} \]
\[ [1 + m_1\alpha(t)\cos(\omega_{RF}t + \phi_1)]^{1/2} [1 + m_2\beta(t)\cos(\omega_{RF}t + \phi_2)]^{1/2} \]

(3.2)

The first term in Equation 3.2 is due to the beating between the signal and itself, the second due to beating between the crosstalk channel and itself, and the last term due to beating between the signal and crosstalk electric fields.

Equation 3.2 can be simplified by using trigonometric identities to consider only the resulting electrical signals at the required RF frequency. These signals correspond to mixing between the modulated optical sidebands, which are separated by the required RF frequency from the optical carrier, and the two optical carriers. In addition, a first-order approximation can be used to evaluate the RF component due to mixing between \( E_1 \) and \( E_2 \):

\[ (1 + y)^{1/2} \approx \left(1 + \frac{y}{2}\right) \quad y \ll 1 \]

(3.3)

This assumes low optical modulation indices, linearising the modulator transfer function. Equation 3.3 can be used to simplify Equation 3.2 and obtain the current \( I(\omega_{RF}) \) that comprises the frequency components at \( \omega_{RF} \):

\[ I(\omega_{RF}) \propto m_1\alpha(t)\cos(\omega_{RF}t + \phi_1) \leftarrow \text{Signal} \]
\[ + xm_2\beta(t)\cos(\omega_{RF}t + \phi_2) \leftarrow \text{Crosstalk} \]
\[ + \sqrt{x} \cos(\theta_1 - \theta_2) \times \leftarrow \text{Signal - Crosstalk} \]
\[ [m_1\alpha(t)\cos(\omega_{RF}t + \phi_1) + m_2\beta(t)\cos(\omega_{RF}t + \phi_2)] \]

(3.4)

The first term in Equation 3.4 represents the recovered modulated signal RF carrier and the second term the crosstalk modulated RF carrier. The last two terms are due to the mixing between the signal optical sideband and crosstalk optical carrier, and crosstalk optical sideband and signal optical carrier, respectively. Since the optical fields have different phases, these terms depend on the relative optical phase difference \((\theta_1 - \theta_2)\).

For an RF subcarrier modulated by data in BPSK format, a phase-locked loop (PLL) must be used to recover the baseband data [7]. The PLL must be locked to the
phase of the signal electrical carrier, i.e. \( \phi_t \). The signal waveform \( I_{\text{down}} \) of the downconverted baseband data channel can then be represented by:

\[
I_{\text{down}} = I(\omega_{RF}) \cos(\omega_{RF} t + \phi_t) \\
\approx m_1 \alpha [1 + \sqrt{x} \cos(\theta_1 - \theta_2)] \\
+ m_2 \beta \sqrt{x} \cos(\phi_1 - \phi_2) [\sqrt{x} + \cos(\theta_1 - \theta_2)] \\
= m_1 \alpha [1 + \sqrt{x} \cos(\Delta \theta)] \\
+ m_2 \beta \sqrt{x} \cos(\Delta \phi) [\sqrt{x} + \cos(\Delta \theta)]
\]

(3.5)

where \( I(\omega_{RF}) \) is the amplitude of the RF carrier, \( \Delta \phi = \phi_1 - \phi_2 \) is the RF phase difference between the signal channel and crosstalk signal, and \( \Delta \theta = \theta_1 - \theta_2 \) is the optical phase difference between the two optical carriers. Equation 3.5 shows that due to the multiplication effect of the PLL, the mixing products arising from the crosstalk signal itself and intermixing terms due to the signal carrier and crosstalk sidebands beating are multiplied by \( \cos(\Delta \phi) \). This RF phase difference is clearly important since orthogonal RF carriers are rejected by the PLL. Hence the PLL output contains only the component of the crosstalk RF carrier which is in-phase with the signal carrier, rejecting the orthogonal component. This provides additional electrical filtering of the crosstalk signal if the RF carriers are not in-phase.

In the model we assume that the modulation indices are equal (\( m_1 = m_2 \)). If modulation indices are different, the analysis used in this section can be used to take this into account, changing the effective crosstalk level. The modulation format is BPSK, i.e. \( \alpha \) and \( \beta \) can take values of \( \pm 1 \). This means the RF carrier phase is changed by 180° between data ‘1’s and ‘0’s. Note that for ASK modulation \( \alpha \) and \( \beta \) take values of 1 and 0, so that the amplitude of the RF carrier is modulated. This is what Moura et al. considered in their paper, and will be briefly discussed in Section 3.4.

For Gaussian thermal receiver noise, bit-error-rates (BERs) can be calculated using the complementary error function [7]:

\[
\text{BER} = \frac{1}{2} \text{Erfc} \left( Q / \sqrt{2\sigma} \right)
\]

(3.6)

where \( Q = (I/\sigma) \) and \( \sigma \) is the standard deviation of the Gaussian thermal noise. This expression assumes only thermal noise is present [12]. This is a valid assumption for optical links in which no optical amplification is present. As explained in chapter 2, the presence of amplified spontaneous emission (ASE) noise changes the crosstalk penalty.
Equation 3.6 can be modified to include these effects by adding noise variances together [12, 13]. Since the binary data is bipolar the decision threshold is equal to 0. This corresponds to a DC-coupled threshold. For a receiver in which only thermal noise is present, this decision threshold is optimal, since the Gaussian noise impacts data ‘1’s and ‘0’s equally [14]. When both the signal and crosstalk are present, the PLL output current $I$ at the decision circuit will vary depending on the four equi-probable logic bit combinations 00, 01, 10 and 11:

$$I_{00} = -I_{11}$$
$$= +1 \times \left[ 1 + \sqrt{x} \cos(\Delta \theta) \right]$$
$$+ 1 \times \sqrt{x} \cos(\Delta \phi \left[ \sqrt{x} + \cos(\Delta \theta) \right])$$

$$I_{01} = -I_{10}$$
$$= +1 \times \left[ 1 + \sqrt{x} \cos(\Delta \theta) \right]$$
$$- 1 \times \sqrt{x} \cos(\Delta \phi \left[ \sqrt{x} + \cos(\Delta \theta) \right])$$

(3.7)

The average BER for a given crosstalk power ratio $x$, optical phase difference $\Delta \theta$ and RF phase difference $\Delta \phi$ as well as signal power, can then be calculated using Equations 3.6 and 3.7. For coherent crosstalk, $\Delta \theta$ will be constant. Meanwhile incoherent crosstalk will lead to a stochastically varying $\Delta \theta$ (uniformly distributed between 0° and 360°) which therefore requires the averaging of BER over $\Delta \theta$. This is performed numerically in steps of 1°, which was found to be sufficient to achieve an accurate BER.

By varying the signal power and crosstalk ratio, BER curves as a function of total optical power can be obtained, as shown in Figure 3.2, and optical power penalties for a BER of $10^{-9}$ to be calculated.
Figure 3.2 shows that by increasing the received optical power and thereby increasing the resulting received electrical signal, the impact of Gaussian receiver noise can be reduced, lowering (improving) the BER. This allows the degradation in BER due to optical crosstalk to be cancelled. The difference in received optical power between the situation of no crosstalk and with crosstalk being present at a BER of $10^{-9}$ corresponds to the optical power penalty. However, if the crosstalk level is too high then the intrinsic optical signal will be sufficiently degraded so that a BER of $10^{-9}$ can never be recovered, regardless of received optical power. This corresponds to a so-called “error floor” in the BER curve, as can be seen in Figure 3.2 and is indicated by an infinite power penalty.

Figure 3.3 shows the optical power penalty predicted using the analytical model for in-band incoherent crosstalk for BPSK modulation (Equations 3.6 and 3.7). The Gaussian receiver noise variance is set so as to give a BER of $10^{-9}$ with no crosstalk. The power penalty is plotted as a function of optical crosstalk level for RF phase differences of $\Delta \phi = 0$, $45^\circ$, and $90^\circ$. For BPSK modulation the remaining 3 quadrants (from $90^\circ$ to $360^\circ$) follow the same behaviour. For comparison, Figure 3.3 also shows the predicted power penalty for baseband data transmission, using an equivalent approach with a decision threshold equal to half the average total power [1].
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Figure 3.3 shows that the power penalties arising from in-band optical crosstalk in a fibre-radio WDM system depend on the RF phase difference between the crosstalk and signal carriers. For example at a crosstalk level of -12 dB, the power penalty is reduced by approximately 1.5 dB when $\Delta\phi$ is changed from 0 to 90°. This is to be expected as the full crosstalk signal amplitude is recovered by the PLL when $\Delta\phi = 0$. The difference in power penalties also increases with the crosstalk level $x$ (e.g. ~4 dB at a crosstalk level of $-4$ dB). Another important observation from Figure 3.3 is that the crosstalk-induced optical power penalty is significantly worse for baseband data transmission. The difference is due to the modulation format and the associated data recovery in the receiver and is explained in Section 3.7 when comparing modulation formats. It is also apparent that a large power penalty is still present for a 90° RF phase difference, even though no crosstalk electrical signal is present at the output of the receiver. The penalty is caused by the presence of the crosstalk optical carrier, which beats with the signal data sidebands thereby varying the resulting signal data amplitude. This variation in signal amplitude results in a worse BER, since the electrical crosstalk signal in this case is completely removed by the electrical receiver. As mentioned previously, the RF phase difference can vary, so that a given RF phase difference can not be guaranteed, as discussed in Section 3.6.

If we consider a 1 dB optical power penalty, then ~23 dB of in-band crosstalk can be tolerated for IM, and a much higher ~17.3 dB for BPSK modulation. The increased tolerance to in-band optical crosstalk means that optical component
specifications can be significantly relaxed, allowing cheaper optical components to be used. Alternatively, it means a lower power penalty will be observed for a given in-band crosstalk level.

### 3.2.2 Dependence of BER on RF Phase Difference and Crosstalk Level

The potential importance of the RF carrier phase difference between signal and crosstalk RF signals is evident from Figure 3.3. It is interesting to look at what is happening for the four possible bit combinations, in particular to look at the BER averaged over optical phase difference (i.e. incoherent BER) as a function of RF phase difference. This is shown in Figure 3.4.

![Fig. 3.4 BER vs. RF phase difference for –30 dB of BPSK in-band crosstalk.](image)

Figure 3.4 shows the BER for a crosstalk level of –30 dB as a function of RF phase difference. As expected, there is a symmetry about 180°. The difference between a both signal and crosstalk ‘1’s or ‘0’s and signal ‘1’ and crosstalk ‘0’ or signal ‘0’ and crosstalk ‘1’ is apparent, since in one case the crosstalk magnitude increases the signal, whilst in the other it decreases it. This produces the symmetrical behaviour about 90° between the ‘11’/‘00’ case and the ‘10’/‘01’ case, since beyond 90° the sign of the crosstalk contribution changes. The black line shows the overall average BER, showing the worst case at 0° and the best case at 90°. The change in BER at this low crosstalk level is between $10^{-9}$ and $3 \times 10^{-9}$, resulting in a low power penalty at all RF phase differences. For a higher crosstalk level, the minimum BER is higher as is the
difference in BER over RF phase difference, as can be seen in Figure 3.5 which shows an equivalent plot for –10 dB of optical crosstalk.

![Graph showing BER vs. RF phase difference for -10 dB of BPSK in-band crosstalk.](image)

*Fig. 3.5 BER vs. RF phase difference for –10 dB of BPSK in-band crosstalk.*

Figures 3.4 and 3.5 show the degradation in BER caused for the same received *signal* optical power as without any crosstalk. An important measure of the effect of optical crosstalk other than the optical power penalty is the BER as a function of the crosstalk level for no increase in signal optical power. This is shown in Figure 3.6.

![Graph showing BER as a function of optical crosstalk level for BPSK in-band crosstalk.](image)

*Fig. 3.6 BER as a function of optical crosstalk level for BPSK in-band crosstalk.*

Figure 3.6 can be used to determine the maximum permissible level of optical crosstalk that can be fully corrected using Forward Error Correction (FEC), assuming the worst-case of phase-aligned RF carriers. A typical Reed-Solomon code of RS(255,239) can correct a BER of up to $10^{-5}$ to $10^{-4}$ [15, 16], which corresponds to a maximum crosstalk...
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level of about −13 dB. Hence FEC would allow up to −13 dB of in-band optical
crosstalk with an optical power penalty solely due to optical power addition. This is a
gain of about 1.6 dB in power penalty compared to when there is no FEC (1.8 dB power
penalty − 0.2 dB power addition).

3.2.3 Out-of-band Optical Crosstalk with Frequency Re-use

We now consider optical crosstalk arising from a signal that is at a different
wavelength from the desired channel but at the same RF frequency as the desired signal
due to frequency re-use. The optical spectrum and RF spectrum after photodetection are
shown in Figure 3.7.

![Optical Spectrum and RF Spectrum](image)

*Fig. 3.7 Optical and electrical spectra for out-of-band crosstalk with frequency re-use.*

The signal and crosstalk electric fields can be represented as (following the previous
analysis):

\[
E_1(t) = |E_1| \cos(\omega_1 t + \phi_1(t))[1 + m_1 \alpha(t) \cos(\omega_{RF} t + \phi_1(t))]^{1/2}
\]

\[
E_2(t) = \sqrt{x} |E_1| \cos(\omega_2 t + \phi_2(t))[1 + m_2 \beta(t) \cos(\omega_{RF} t + \phi_2(t))]^{1/2}
\]

(3.8)

where \(\omega_1\) and \(\omega_2\) now represent the different optical frequencies of the desired channel
and crosstalk signal, respectively. Following the same analysis as described in Section
3.2.1 and neglecting all of the possible mixing terms at frequencies other than \(\omega_{RF}\), we
obtain:

\[
I(\omega_{RF}) = I(\omega_{1RF}) + I(\omega_{2RF})
\]

\[
\propto m_1 \alpha(t) \cos(\omega_{RF} t + \phi_1) \quad \leftarrow \text{Signal}
\]

\[
+ x m_2 \beta(t) \cos(\omega_{RF} t + \phi_2) \quad \leftarrow \text{Crosstalk}
\]

(3.9)
Equation 3.9 assumes that the frequency difference between signal and crosstalk wavelengths is larger than the signal and crosstalk RF frequency, otherwise the signal-crosstalk beat term can not be ignored.

As before, recovery of BPSK modulation using a PLL will result in the following downconverted signal:

\[ I_{\text{down}} = I(\omega_{RF}) \cos(\omega_{RF}t + \phi_1) \]

\[ \propto m_1 \alpha(t) + x m_2 \beta(t) \cos(\Delta \phi) \]  \hspace{1cm} (3.10)

A comparison of Equation 3.10 and Equation 3.5 shows that out-of-band crosstalk is only influenced by the RF phase difference, and therefore a potential reduction of crosstalk is possible when \( \Delta \phi = 90^\circ \). In addition, the signal and crosstalk data are recovered separately and no mixing terms between the signal and crosstalk are present at the required RF frequency. This is due to the frequency separation of the optical carriers that would normally be significantly greater than the RF modulation frequency in the wireless network. This then results in so-called linear or additive crosstalk, since both signals are recovered separately.

Figure 3.8 shows the predictions of the analytical model for out-of-band crosstalk for \( \Delta \phi = 0^\circ, 45^\circ \) and \( 90^\circ \), as a function of optical crosstalk ratio. Also shown in the diagram is the predicted power penalty for baseband transmission. As for in-band crosstalk, the optical power penalties for out-of-band optical crosstalk vary with \( \Delta \phi \), with the largest power penalty occurring for \( \Delta \phi = 0^\circ \). The actual power penalty for a particular crosstalk ratio however, is less for out-of-band crosstalk than for the in-band case. For example, at \(-12 \) dB of crosstalk the power penalty for all values of \( \Delta \phi \) is less than \( 0.5 \) dB. In addition, the variation of power penalty with \( \Delta \phi \) is reduced to less than \( 0.25 \) dB at the same crosstalk level. A \( 1 \) dB power penalty occurs at a crosstalk level of \(-9.3 \) dB for in-phase carriers. Figure 3.8 shows that for an RF phase difference of \( \Delta \phi = 90^\circ \), when the crosstalk signal is at the same optical power as the desired channel (i.e. \( 0 \) dB of optical crosstalk), the optical power penalty is \( 3 \) dB (the total power has doubled). This is as expected since the \( 90^\circ \) phase shift leads to a cancellation of the crosstalk contribution, so that although no electrical crosstalk signal is present at the PLL output, with no BER degradation, the total optical power creates an effective power penalty.
Fig. 3.8 Theoretical optical power penalties as a function of BPSK out-of-band crosstalk with frequency re-use use as a function of RF carrier phase difference.

In contrast to in-band crosstalk, where the optical power penalty for baseband data transmission is significantly higher than for RF subcarrier transmission (at all crosstalk ratios), Figure 3.8 shows that out-of-band crosstalk-induced power penalties for baseband data transmission are identical to those for RF subcarrier transmission when the carriers are in-phase. This is to be expected, as the crosstalk is linear irrespective of the modulation format or receiver structure.

### 3.3 **In-band and Out-of-band Optical Crosstalk with Adjacent Frequency Bands for BPSK Modulation**

The following section considers the more likely scenario in which the optical carriers transport data at different RF frequencies, allowing electrical filtering to be used. Both in-band and out-of-band crosstalk are analysed, showing the improvement in crosstalk tolerance.

#### 3.3.1 In-band Optical Crosstalk with Adjacent Frequency Bands

If we consider a crosstalk signal at the same wavelength as the signal channel but carrying data at a different wireless carrier frequency, the analysis presented above is slightly modified and results are different. Although signal and crosstalk wavelengths now carry data at different wireless frequencies, the optical carriers are at the same wavelength. Mixing terms are produced in the photodetector, resulting in the separate
signal and crosstalk RF carriers being received, together with mixing terms between signal and crosstalk carriers and sidebands (see Figure 3.9).

![Optical Spectrum and RF Spectrum](image)

*Fig. 3.9 Optical and electrical spectra for in-band crosstalk with different RF frequency bands.*

The result is similar to Equation 3.4, although the signal carrier beating with the crosstalk sidebands produces a signal at the crosstalk carrier frequency, as shown below:

\[
I(\omega_{RF}) \propto m_1 \alpha(t) \cos(\omega_{RF1} t + \phi_1) [1 + \sqrt{x} \cos(\Delta \theta)]
+ \sqrt{x} m_2 \beta(t) \cos(\omega_{RF2} t + \phi_2) [\sqrt{x} + \cos(\Delta \theta)]
\]  

(3.11)

Equation 3.11 shows that the presence of the crosstalk carrier produces a signal which is at the same RF frequency as the signal and whose amplitude depends on the relative optical phase difference $\Delta \theta$. The beating of the signal optical carrier with the crosstalk sidebands also produces a signal at the same RF carrier frequency as the crosstalk signal, with an amplitude depending on the relative optical phase difference.

Unlike the situation where both signal and crosstalk RF carrier frequencies are the same, we are now faced with the presence of two distinct electrical frequencies, allowing electrical filtering to be used to remove the unwanted carrier at $\omega_{RF2}$. However, even if we assume perfect filtering, the crosstalk optical carrier still ensures that the signal amplitude depends on the crosstalk ratio and optical phase difference, as shown below:

\[
I_{down} = I(\omega_{RF}) \cos(\omega_{RF1} t + \phi_1)
\propto m_1 \alpha(t) [1 + \sqrt{x} \cos(\Delta \theta)]
\]

(3.12)
The amplitude of the desired signal comprises an extra term due to beating between the crosstalk optical carrier and signal optical sidebands. The resulting optical power penalties as a function of optical crosstalk level are shown in Figure 3.10. This situation is in effect the same as that for crosstalk at the same RF carrier frequency but with an RF phase difference of 90°, since recovery of the wireless frequency also leads to elimination of the crosstalk electrical signal (see Figure 3.9). The benefit of the in-band crosstalk channel carrying data at a different RF frequency is clear since electrical filtering allows the effect of the crosstalk to be minimised.

![Graph](image)

**Fig. 3.10 Optical power penalties for in-band optical crosstalk with adjacent frequency bands for BPSK modulation.**

The ability to electrically filter the crosstalk RF frequency band results in a further increase in tolerance to in-band optical crosstalk relative to baseband intensity modulation. Whereas for a 1 dB optical power penalty –17.3 dB of optical crosstalk is tolerated with frequency re-use, this is relaxed to –13.4 dB if the crosstalk RF frequency can be electrically filtered. If we compare this to –23 dB of in-band crosstalk for IM, then the use of adjacent frequency bands for the same uplink / downlink wavelength for BPSK modulation allows an extra 9.4 dB margin in maximum in-band crosstalk tolerance. Clearly this indicates that in-band crosstalk in fibre-radio systems is not as serious as for baseband links, at least for BPSK modulation, allowing wavelength re-use in the uplink path, together with the use of cheaper optical components having less stringent specifications.
3.3.2 Out-of-band Optical Crosstalk with Adjacent Frequency Bands

As discussed in Section 3.2.3, out-of-band crosstalk results in a received electrical signal that is a sum of the electrical signals carried by each individual wavelength. If the crosstalk wavelength carries data at different electrical frequencies to the signal wavelength, then these can be filtered electrically, ensuring minimal impact on the signal quality. Hence the level of electrical crosstalk can be determined purely by the electrical channel separation and channel and filter bandwidths [17-19]. It should also be remembered that –10 dB of out-of-band optical crosstalk corresponds to a –20 dB level of electrical crosstalk, so that optical component specifications may be sufficient to ensure an adequate electrical carrier-to-noise ratio without the need for additional electrical filtering. Furthermore, electrical filtering of adjacent channels within the transmitted signal wireless band may be enough to ensure out-of-band optical crosstalk has a minimal effect.

If we consider the analysis presented in Section 3.2.3 for out-of-band crosstalk with frequency re-use, we can modify Equation 3.9 to reflect the fact that the crosstalk channel is at a different RF frequency, $\omega_{RF2}$:

$$I(\omega_{RF}) = I(\omega_{RF2}) + I(\omega_{RF2})$$

$$= m_1 \alpha(t) \cos(\omega_{RF1}t + \phi_1)$$

$$+ m_2 \beta(t) \cos(\omega_{RF2}t + \phi_2)$$

(3.13)

Electrical filtering of the unwanted RF frequency results in the following downconverted signal at the output of the PLL:

$$I_{\text{down}} = I(\omega_{RF}) \cos(\omega_{RF1}t + \phi_1)$$

$$= m_1 \alpha(t)$$

(3.14)

Equation 3.14 indicates that for out-of-band crosstalk electrical filtering allows the removal of the crosstalk channel, without any degradation in the signal. The resulting optical power penalty is due to the presence of additional optical power at the receiver, as shown in Figure 3.11.
Fig. 3.11 Optical power penalties for out-of-band crosstalk with adjacent frequency bands for BPSK modulation.

The optical power penalty due to optical power addition is given by the following equation [20]:

\[ PP(\text{dB}) = 10 \log_{10}(1 + x) \]

where \( x \) is the optical crosstalk power ratio. Alternatively, the maximum crosstalk for a given power penalty is given by re-arranging Equation 3.15:

\[ x = 10 \left( \frac{PP(\text{dB})}{10} \right) - 1 \]

where \( x \) is the linear crosstalk ratio. Hence doubling of total optical power at 0 dB crosstalk results in a maximum power penalty of 3 dB. If we consider a typical system penalty of 1 dB for optical crosstalk, then we can tolerate up to −5.9 dB of out-of-band crosstalk. This should be compared to Figure 3.8, which shows that about −9 dB of out-of-band crosstalk can be tolerated if the crosstalk RF frequency is the same as the signal frequency. Again, the benefits of electrical filtering are clear, even for out-of-band crosstalk.

### 3.4 In-band and Out-of-band Optical Crosstalk for ASK Modulation

The following sections follow the analysis of Moura et al. [4] to evaluate the photodetected current given the presence of signal and crosstalk optical fields, as derived in Equations 3.1 to 3.4 in Section 3.2, for an ASK modulated RF carrier. This
will illustrate the fact that ASK and BPSK modulation are very similar, although the nature of the modulation results in quite different results for in-band crosstalk, as has been highlighted in the previous sections. We extend the work of Moura et al. to show the importance of the RF carrier phase difference with frequency re-use, together with the benefits of electrical filtering where a different crosstalk carrier frequency is used. The issue of decision threshold will be addressed, together with the demodulation technique used, as these are factors that influence the expected power penalties. This was not investigated in [4], as coherent detection using a PLL was used at the receiver and a DC-coupled decision threshold was used. We also consider out-of-band crosstalk.

### 3.4.1 In-Band Optical Crosstalk for ASK Modulation with Frequency Re-use

The analytical model presented in Section 3.2 can be directly applied to ASK modulation. In both BPSK and ASK modulation an RF carrier is modulated by a data sequence. For BPSK the phase of the carrier is modulated, whereas for ASK the amplitude of the carrier is modulated.

Equation 3.4 can be directly used for in-band crosstalk, as shown below, although the signal and crosstalk RF frequencies are represented by \( \omega_{RF1} \) and \( \omega_{RF2} \), respectively:

\[
I(\omega_{RF}) \propto m_1 \alpha(t) \cos(\omega_{RF1} t + \phi_1) \\
+ x m_2 \beta(t) \cos(\omega_{RF2} t + \phi_2) \\
+ \sqrt{x} \cos(\Delta \theta) \times \\
[m_1 \alpha(t) \cos(\omega_{RF1} t + \phi_1) + m_2 \beta(t) \cos(\omega_{RF2} t + \phi_2)]
\] (3.17)

The only difference for ASK modulation is that \( \alpha(t) \) and \( \beta(t) \) now take values of +1 for a data ‘1’ and 0 for a data ‘0’ (compared to +1 for data ‘1’ and –1 for data ‘0’ for BPSK). The above expression shows that the amplitude of the RF carriers corresponds to the required data. An envelope detector (square-law) circuit can be used to recover the amplitudes, or a PLL as for BPSK demodulation [21]. For square-law detection, the signal is squared and low-pass filtered, to give:

\[
I(\text{env}) \propto [\alpha(1 + \sqrt{x} \cos(\Delta \theta))]^2 + [\sqrt{x} \beta(\sqrt{x} + \cos(\Delta \theta))]^2 + \\
2\alpha \beta \sqrt{x}(1 + \sqrt{x} \cos(\Delta \theta))(\sqrt{x} + \cos(\Delta \theta) \cos(\Delta \phi))
\] (3.18)
where \( I(env) \) is the output of the envelope detector, \( \alpha \) and \( \beta \) are signal and crosstalk data envelopes, \( x \) is the optical crosstalk ratio, and \( \Delta \theta \) and \( \Delta \phi \) are optical and RF phase differences respectively. Equation 3.18 assumes signal and crosstalk RF frequencies are identical. If we assume a DC-coupled decision threshold \( D_{env} \) that averages after the low-pass filter, we obtain the following time-averaged decision threshold:

\[
D_{env} = 0.5 \left( 1 + x + x^2 \right) \tag{3.19}
\]

since \( \alpha \) and \( \beta \) can take values of +1 and 0. Note that this is independent of optical and RF phase differences, since we assume these are averaged out. For the RF phase, the important factor is whether or not the RF phase difference varies quickly relative to both the bit duration and relative to the decision circuit time constant. This will be addressed in Section 3.6. If, on the other hand, the decision threshold circuit can track changes in RF phase difference, then the following decision threshold, \( D_{env}(\Delta \phi) \), is obtained:

\[
D_{env}(\Delta \phi) = \frac{1}{4} \left( 2 + 2x + 2x^2 + 3x \cos(\Delta \phi) \right) \tag{3.20}
\]

If the receiver uses a PLL to lock onto the signal RF carrier and downconvert the envelope to baseband, then the output current \( I(down) \) is given by Equation 3.5:

\[
I_{down} = I(\omega_{RF}) \cos(\omega_{RF} t + \phi_1) \\
\approx \alpha[1 + \sqrt{x} \cos(\Delta \theta)] + \beta\sqrt{x} \cos(\Delta \phi)[\sqrt{x} + \cos(\Delta \theta)] \tag{3.21}
\]

We assume both signal and crosstalk RF frequencies are the same, all other symbols are as defined previously. A similar argument to that above results in the follow time-averaged decision threshold \( D_{PLL} \):  

\[
D_{PLL} = 0.5 \tag{3.22}
\]

It can also be shown that if the decision threshold circuit can track changes in RF phase difference, the following expression for \( D_{PLL}(\Delta \phi) \) is obtained:

\[
D_{PLL}(\Delta \phi) = 0.5(1 + x \cos(\Delta \phi)) \tag{3.23}
\]

Equation 3.6 can be used to evaluate the BER as a function of optical power, crosstalk level, optical and RF phase difference:

\[
BER = \frac{1}{2} \text{Erfc}(Q / \sqrt{2} \sigma) \tag{3.24}
\]

where \( Q=I-D \), incorporating the fact that the decision threshold for ASK modulation is non-zero.
The associated optical power penalties for in-band crosstalk for ASK modulation are shown in Figures 3.12 and 3.13, for square-law envelope and PLL detection respectively, assuming identical signal and crosstalk RF frequencies. In each Figure, results for a decision threshold that is averaged over $\Delta \phi$ (independent of $\Delta \phi$) and for a decision threshold that is a function of $\Delta \phi$ are shown. Note that if envelope detection is performed using a diode (rectifier) and a low-pass filter, the worst-case result is very similar to that for PLL detection.

![Graph showing optical power penalties for in-band optical crosstalk with ASK modulation, for envelope detection with decision thresholds independent of (D(avg)) and a function of RF phase difference (D(\Delta \phi)), respectively.]

Figure 3.12 shows the same trend as for BPSK modulation (Figure 3.3), with the largest power penalty for in-phase RF carriers, with a reduction in power penalty as the RF phase difference increases. ASK modulation with square-law envelope detection shows less tolerance to crosstalk than BPSK modulation, with a maximum crosstalk level of –24.5 dB for a 1 dB optical power penalty (-17.5 dB for BPSK modulation). However, the region between 90° and 180°, i.e. the region in which $\cos(\Delta \phi)$ is negative shows very similar optical power penalties, irrespective of whether or not the decision threshold is able to track changes in RF phase difference. A 180° phase difference gives the minimum power penalty. In practical terms, there is no real benefit in having a decision threshold circuit that can track changes in RF phase difference, since maximum power penalties occurring for the worst-case of phase-aligned RF carriers are identical for realistic power penalties (up to 2-3 dB max.).
Fig. 3.13 Optical power penalties for in-band optical crosstalk with ASK modulation, for PLL detection with decision thresholds independent of \( (D=0.5) \) and a function of RF phase difference \( D(\Delta\phi) \), respectively.

Figure 3.13 shows the predicted optical power penalties for a receiver that demodulates ASK data using a PLL and a local oscillator for downconversion to baseband. Theoretical results for phase-aligned carriers agree with those published by Moura et al. [4], who do not consider the potential phase dependence. Power penalties are higher than those observed for BSKP modulation, but much lower than those for ASK modulation using square-law envelope detection. A PLL receiver is less sensitive to crosstalk than an envelope detector, since downconversion removes orthogonal carriers while an envelope detector takes the magnitude squared of the vector addition of carriers, without any filtering. Whilst a 1 dB power penalty in this region occurs at \(-24.5\) dB for square-law envelope detection, using a PLL with downconversion a 1 dB power penalty occurs between \(-17\) to \(-18\) dB of crosstalk depending on the RF phase difference and decision threshold response time. A fast decision threshold for PLL detection allows the power penalty to be reduced as the RF phase difference increases to \(180^\circ\), whilst a slow decision threshold with fixed (average) value shows a nearly fixed power penalty for the range \(90\) to \(180^\circ\). A fast decision threshold circuit is unhelpful for aligned RF carriers, resulting in slightly higher power penalties, although the difference is less than \(0.2\) dB at \(3\) dB optical power penalty. Hence a similar conclusion is reached for a PLL receiver; namely that having a fast decision threshold which can track RF phase difference changes is unnecessary.
3.4.2 In-Band Optical Crosstalk for ASK Modulation with Adjacent Frequency Bands

Equation 3.17 can be readily modified to incorporate the fact the crosstalk RF frequency is different and can be electrically filtered:

\[ I(\omega_{RF}) \propto m, \alpha(t)[1 + \sqrt{x} \cos(\Delta \theta)] \cos(\omega_{RF} t + \phi) \]

(3.25)

The following expressions for the envelope detector output and the PLL detector output are obtained:

\[ I_{env} \propto (\alpha[1 + \sqrt{x} \cos(\Delta \theta)])^2 \]

(3.26)

and

\[ I_{PLL} \propto \alpha[1 + \sqrt{x} \cos(\Delta \theta)] \]

(3.27)

For PLL detection, the decision threshold \( D_{PLL} \) is given by:

\[ D_{PLL} = 0.5 \]

(3.28a)

and for envelope detection the decision threshold \( D_{env} \) is given by:

\[ D_{env} = 0.5\left[1 + \frac{x}{2}\right] \]

(3.28b)

The corresponding optical power penalties as a function of crosstalk for both envelope detection and PLL detection are shown in Figure 3.14.

![Graph showing optical power penalties vs. optical crosstalk for ASK PLL and ASK Env.](image)

**Fig. 3.14** Optical power penalties for in-band crosstalk with adjacent frequency bands for ASK modulation, for envelope and PLL detection.

Figure 3.14 illustrates the clear benefits of using the RF frequency domain to minimise the impact of in-band optical crosstalk. Electrical filtering of the crosstalk carrier...
frequency results in a fixed optical power penalty and a better tolerance to crosstalk compared to when the crosstalk RF carrier frequency is the same as the signal frequency. Interestingly square-law envelope detection results in a lower power penalty compared to PLL detection. This is due to the dependence of decision threshold on crosstalk level, which reduces the impact of crosstalk for square-law envelope detection. The crosstalk at which a 1 dB optical power penalty is obtained is $-18$ dB for PLL detection and $-14.5$ dB for envelope detection, compared to $-23$ dB with frequency re-use (worst-case), a 5 dB improvement in crosstalk tolerance for PLL detection. As for in-band crosstalk, the difference in demodulation format is significant, although square-law envelope detection results in a greater tolerance to out-of-band crosstalk.

### 3.4.3 Out-of-Band Optical Crosstalk for ASK Modulation with Frequency Re-use

The analysis for out-of-band crosstalk is identical to that explained for BPSK modulation in Section 3.2.3, so that the received photocurrent is given by:

$$I(\omega_{RF}) = I(\omega_{1RF}) + I(\omega_{2RF})$$

$$\propto \alpha(t)\cos(\omega_{RF}t + \phi_1) + \beta(t)\cos(\omega_{RF}t + \phi_2)$$

(3.29)

The difference is that the envelope of the RF carriers is amplitude modulated. If we assume the use of a square-law envelope detector as described in the previous section, then the current $I_{env}$ is given by:

$$I_{env} = \left[\alpha^2 + x^2\beta^2 + 2\alpha\beta x\cos(\Delta \phi)\right]$$

(3.30)

where $\alpha$ and $\beta$ are the signal and crosstalk data amplitudes (1 or 0) and $\Delta \phi$ is the RF carrier phase difference. The corresponding ‘slow’ decision threshold $D_{env}$ is given by:

$$D_{env} = 0.5(1 + x^2)$$

(3.31)

and the decision threshold that is a function of $\Delta \phi$ is given by:

$$D_{env}(\Delta \phi) = 0.5\left[1 + x^2 + xc\cos(\Delta \phi)\right]$$

(3.32)

If a PLL is used to downconvert the data, then the PLL detector output $I_{PLL}$ is given by:

$$I_{PLL} = \alpha + \beta x\cos(\Delta \phi)$$

(3.33)

and the corresponding ‘slow’ decision threshold $D_{PLL}$ is given by:

$$D_{PLL} = 0.5$$

(3.34)
If the decision threshold circuit can track changes in RF phase difference, then the decision threshold is:

$$D_{PLL}(\Delta \phi) = 0.5[1 + x \cos(\Delta \phi)]$$

(3.35)

The resulting optical power penalties as a function of optical crosstalk for both square-law envelope detection and PLL detection are shown in Figures 3.15 and 3.16:

**Fig. 3.15 Optical power penalties for out-of-band crosstalk for ASK modulation with envelope detection with decision thresholds independent of (D(avg)) and a function of RF phase difference (D(\Delta \phi)), respectively.**

Figure 3.15 illustrates the complex dependence on RF phase difference. For both decision threshold types the lowest power penalty is observed for phase-aligned carriers (in contrast to in-band crosstalk), whilst the maximum power penalty occurs for $180^\circ$. The average decision threshold results in higher worst-case power penalties than the phase-tracking decision threshold, although the difference is only 0.2 dB around a 1 dB power penalty. A 1 dB power penalty occurs at $-10.3$ and $-11.2$ dB for the average and phase-tracking decision thresholds respectively. This compares to $-9.3$ dB for BPSK and baseband intensity modulation. The lowest power penalty curve at $0^\circ$ phase difference is worse than standard baseband IM results at high crosstalk ($\geq 6$ dB).
Fig. 3.16 Optical power penalties for out-of-band crosstalk for ASK modulation with PLL detection with decision thresholds independent of \((D=0.5)\) and a function of RF phase difference \((D(\Delta \phi))\), respectively.

Figure 3.16 shows the optical power penalties for out-of-band crosstalk for ASK modulation with PLL detection. Results show symmetry around 90°, with the highest power penalties at 0 or 180° and the lowest at 90°. Once again the RF phase-tracking decision threshold results in lower power penalties at a given crosstalk level, particularly at high crosstalk levels. For a 1 dB optical power penalty, this results in crosstalk levels of -9.1 and -10.7 dB respectively, phase-tracking providing a 1.6 dB increase in crosstalk tolerance. For BPSK and baseband IM, a 1 dB power penalty occurs at -9.3 dB of crosstalk.

Inspection of equations show that power penalties for a phase-dependent decision threshold are identical to those for baseband IM when \(\Delta \phi=0\). Power penalties at 90° are also identical for both decision thresholds. However, for other values of \(\Delta \phi\) power penalties are expected to be higher for an average decision threshold which is independent of the RF phase difference, as observed.

### 3.4.4 Out-of-Band Crosstalk for ASK Modulation with Adjacent Frequency Bands

The analysis of out-of-band crosstalk for ASK modulation with adjacent frequency bands mirrors that presented in Section 3.1.3 except that the crosstalk RF frequency is now different and can be electrically filtered. Hence the RF component at the signal frequency, \(I(\omega_{b\phi})\), is as follows:
Analysis of In-band and Out-of-band Optical Crosstalk for Subcarrier Modulation

\[ I(\omega_{RF}) \propto \alpha(t) \cos(\omega_{RF_1} t + \phi_1) \]  
\hspace{1cm} (3.36)

i.e. the filtered signal comprises the required signal only. Square-law envelope detection of the RF component results in the following signal:

\[ I_{\text{env}} = \alpha^2 \]  
\hspace{1cm} (3.37)

with a decision threshold:

\[ D_{\text{env}} = 0.5 \]  
\hspace{1cm} (3.38)

Similarly, use of a PLL and downconversion results in a current \( I_{PLL} \) given by:

\[ I_{PLL} = \alpha \]  
\hspace{1cm} (3.39)

with a decision threshold:

\[ D_{PLL} = 0.5 \]  
\hspace{1cm} (3.40)

Since in both cases only the signal is present, the only effective optical power penalty is due to the additional power at the receiver, and is given by Equation 3.15:

\[ PP(dB) = 10 \log_{10}(1 + x) \]  
\hspace{1cm} (3.41)

where \( x \) is the linear optical crosstalk ratio. This result is identical to that for BPSK modulation where the crosstalk frequency is also filtered electrically, removing the crosstalk signal in the electrical domain and resulting in a power penalty due solely to optical power addition (see Section 3.1.2).

### 3.4.5 Dependence of BER on RF Phase Difference and Crosstalk Level

This section presents the same results that were presented for BPSK modulation in Section 3.2.2, although in this case the modulation format is ASK. Results will be presented both for ASK modulation with a PLL receiver since it parallels the detection scheme for BPSK, and envelope detection which only works with ASK modulated carriers (envelope detection of BPSK data would result in a constant signal, since only the phase of the carrier is modulated). Results for PLL detection and envelope detection are shown for an average decision threshold, which is independent of RF phase difference.

For PLL detection, the change in BER with RF phase difference is shown in Figure 3.17 for a crosstalk level of −30 dB.
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Fig. 3.17 BER vs. RF phase for ASK modulation with PLL detection for −30 dB of in-band crosstalk.

Figure 3.17 shows the symmetry centered on 180° expected for ASK modulation. BERs for signal ‘1’ and ‘0’s with no crosstalk are phase-independent as expected, with the ‘11’ case being worst for most RF phase differences. The average BER is shown to be very nearly constant over the range between 100 and 260°, which explains why power penalties over that range are also nearly identical. Close inspection of Equation 3.20 shows the importance of both crosstalk level and RF phase difference in determining the overall impact of incoherent crosstalk. Recall that the above graph shows the BER averaged over optical phase difference Δθ. This is illustrated at a higher crosstalk level, namely −10 dB, as shown in Figure 3.18.

Fig. 3.18 BER vs. RF phase difference for ASK modulation with −10 dB of in-band crosstalk and PLL detection.

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Corresponding graphs for square-law envelope detection are shown in Figures 3.19 and 3.20 for –30 dB and –10 dB crosstalk respectively.

**Fig. 3.19** BER vs. RF phase difference for ASK modulation with –30 dB in-band crosstalk and envelope detection.

**Fig. 3.20** BER vs. RF phase difference for ASK modulation with –10 dB in-band crosstalk and envelope detection.

For envelope detection there is no change in BER over RF phase difference when there is only a signal ‘1’ or a crosstalk ‘1’, or both signal and crosstalk ‘0’ s. This is due to the nature of envelope detection that ignores the phase of the carrier, only recovering the amplitude squared, unlike for PLL downconversion. There is a dependence on RF phase difference when both signal and crosstalk ‘1’ s are transmitted since the envelope detection takes the magnitude squared of the vector addition of both signals, which is
dependent on the RF phase difference. Note that the above results assume a decision threshold that is averaged over RF phase difference, i.e. which is independent of $\Delta \phi$.

Of more relevance perhaps given the lack of control over RF phase difference is the variation of BER over crosstalk level at a fixed signal optical power, shown in Figures 3.21 and 3.22 for PLL and square-law envelope detection respectively.

![Graph 3.21](image1.png)

*Fig. 3.21 Change in BER with crosstalk level for ASK modulation with PLL detection at a constant signal level.*

![Graph 3.22](image2.png)

*Fig. 3.22 Change in BER with crosstalk level for ASK modulation with envelope detection at a constant signal level.*

Both graphs are different, showing that a BER of $10^{-5}$ arises at a crosstalk level of about $-19$ dB and $-25$ dB for PLL and square-law envelope detection, respectively. This compares to about $-13$ dB for BPSK modulation, which is expected since the actual
optical power penalties are lower for a given crosstalk level. The benefits of FEC are clear, since all errors below a BER of $10^{-5}$ can be corrected, so that the only optical power penalty would be due to the additional crosstalk optical power, given by Equation 3.42:

$$PP(dB) = 10\log_{10}(1 + x) \quad (3.42)$$

### 3.5 Extension of Model to QPSK Modulation

The analytical model used in previous sections has shown the important difference in sensitivity to in-band optical crosstalk between ASK and BPSK modulation. Whilst ASK modulation modulates the amplitude of the RF carrier, BPSK modulation modulates the phase of the RF carrier. The model used to analyse BPSK modulation can be extended to QPSK modulation, in which the phase of the RF can take four possible values, rather than only two for BPSK. This is described in the following sections, for in-band and out-of-band crosstalk with frequency re-use and with the use of different frequency bands.

#### 3.5.1 In-Band Optical Crosstalk for QPSK Modulation with Frequency Re-use

A typical QPSK constellation is shown in Figure 3.23, showing the four possible signal points.

![QPSK constellation](image)

*Fig. 3.23 QPSK signal constellation showing four possible signal states.*

The x-axis is the real axis and the y-axis is the imaginary axis, showing that the magnitude of the signal is unity whilst the phase can have four possible values, each separated by 90°. A QPSK signal is thus seen to comprise two orthogonal BPSK signals, one along the real axis (I or in-phase), and another along the imaginary axis (Q...
or quadrature). Another way of interpreting this is to realise that this corresponds to two BPSK signals, one corresponding to \( \cos(\omega_{RF}) \) and the other to \( \sin(\omega_{RF}) \).

If we consider the presence of an additional crosstalk QPSK signal offset from the signal constellation, as shown in Figure 3.24, then the additional signal contributions due to the crosstalk can be evaluated.

![Figure 3.24 Signal constellation showing QPSK signal and offset crosstalk signal.](image)

If we consider the extra noise signal present in the real axis, or the in-phase component \( I_n \), we obtain:

\[
I_n = X_x \cos(\Delta\phi) - X_y \sin(\Delta\phi)
\]  
\[3.43\]

where \( X_x \) and \( X_y \) correspond to the I and Q amplitudes in-phase with the crosstalk carrier, and \( \Delta\phi \) is the RF carrier phase difference. Similarly for the out-of-phase or quadrature component on the imaginary axis, \( Q_n \):

\[
Q_n = X_x \sin(\Delta\phi) + X_y \cos(\Delta\phi)
\]  
\[3.44\]

Expressions for \( X_x \) and \( X_y \) can be obtained from Equation 3.4:

\[
X_x = \beta_x [x + \sqrt{x} \cos(\Delta\theta)],
\]

\[
X_y = \beta_y [x + \sqrt{x} \cos(\Delta\theta)]
\]  
\[3.45\]

where \( \beta_x \) and \( \beta_y \) are crosstalk I and Q data amplitudes respectively, taking values of \( \pm 1 \), and \( \Delta\theta \) is the optical phase difference between signal and crosstalk optical carriers. Similarly, I and Q components in-phase with the signal carrier are given by:

\[
S_x = \alpha_s [1 + \sqrt{x} \cos(\Delta\theta)],
\]

\[
S_y = \alpha_y [1 + \sqrt{x} \cos(\Delta\theta)]
\]  
\[3.46\]

The resulting received in-phase component, \( I_{TOTAL} \) is given by:

\[
I_{TOTAL} = S_x + I_n
\]  
\[3.47\]
and the quadrature component, $Q_{TOTAL}$ is given by:

$$Q_{TOTAL} = S_y + Q_s$$  \hspace{1cm} (3.48)

Both I and Q expressions depend on the signal data $\alpha_i$ and $\alpha_j$ and both I and Q crosstalk data $\beta_i$ and $\beta_j$, so that there are eight possible bit combinations in each case. Symmetry arguments mean that four of those are identical to the other four but with opposite signs, so they can be ignored. Inspection of the equations reveals the following identical combinations:

$$
I_{00} = -I_{11} = Q_{010} = -Q_{101}, \\
I_{01} = -I_{10} = Q_{000} = -Q_{111}, \\
I_{00} = -I_{11} = Q_{011} = -Q_{100}, \\
I_{01} = -I_{10} = Q_{001} = -Q_{110}
$$

(3.49)

The above expressions can be used together with the error function to evaluate BERs as a function of optical crosstalk level and RF carrier phase difference, allowing the following optical power penalty curves to be obtained, as shown in Figure 3.25.

![Optical Power Penalty Curves](image)

**Fig. 3.25 Optical power penalties for in-band crosstalk for QPSK modulation.**

Figure 3.25 only considers a 45° area since QPSK points are separated by 90°, resulting in a symmetry about 45°, 135°, 225° and 315°. Contrary to ASK and BPSK modulation where the power penalty decreases as the RF phase difference increases, for QPSK modulation the minimum power penalty is obtained for in-phase RF carriers, with the power penalty increasing as the RF phase difference goes to 45°. This condition results in equal contributions of crosstalk I and Q components on received I and Q terms, which maximises the total ‘noise’ contribution (see Equations 3.39 and
3.40). Comparison with Figure 3.3 for BPSK modulation confirms that power penalties are identical to BPSK modulation when signal and crosstalk RF carriers are in-phase, since this results in two orthogonal BPSK signals. A 1 dB power penalty occurs at –18.5 dB for the worst-case, compared to –17.3 dB for the best case (same as BPSK worst-case). Hence QPSK modulation results in 1.2 dB tighter crosstalk tolerance relative to BPSK modulation. However, it is still better than baseband intensity modulation result of –23 dB.

### 3.5.2 In-band Crosstalk for QPSK Modulation with Adjacent Frequency Bands

If the crosstalk RF carrier frequency is different to the signal carrier frequency, then it can be electrically filtered. This greatly simplifies Equations 3.47 and 3.48 since the components in-phase with the crosstalk carrier are at a different frequency, so that:

\[
I_{\text{TOTAL}} = \alpha_s [1 + \sqrt{x} \cos(\Delta \theta)], \quad Q_{\text{TOTAL}} = \alpha_s [1 + \sqrt{x} \cos(\Delta \theta)]
\]

(3.50)

The corresponding optical power penalties for in-band crosstalk with adjacent frequency bands is shown in Figure 3.26.

*Fig. 3.26 Optical power penalties for in-band optical crosstalk for QPSK modulation with adjacent frequency bands.*

The graph also shows the results for frequency re-use, for 0° and 45° RF carrier phase difference, for reference. The benefits of being able to electrically filter the crosstalk signal are apparent, showing an increased tolerance to optical crosstalk or a reduced...
optical power penalty for a given crosstalk level. For example, a 1 dB optical power penalty occurs at −13.5 dB compared to −18.5 dB where the crosstalk carrier frequency is identical to the signal frequency. It should also be noted that optical power penalties are identical to those for BPSK modulation with adjacent frequency bands, as expected since QPSK modulation comprises two orthogonal BPSK signals.

### 3.5.3 Out-of-Band Crosstalk for QPSK Modulation

Out-of-band crosstalk expressions can easily be extracted from the two previous sections since any terms that depend on optical phase difference $\Delta \theta$ can simply be removed. The following expressions are obtained when carrier frequencies are the same:

$$ I_{TOTAL} = \alpha_x + \beta_x \cos(\Delta \phi) - \beta_y \sin(\Delta \phi) $$  \hspace{1cm} (3.51)

and

$$ Q_{TOTAL} = \alpha_y + \beta_x \sin(\Delta \phi) + \beta_y \cos(\Delta \phi) $$  \hspace{1cm} (3.52)

If the crosstalk carrier frequency is different and can be electrically filtered, the optical power penalty is due to optical power addition and is given by Equation 3.42:

$$ PP(dB) = 10 \log_{10}(1 + x) $$  \hspace{1cm} (3.53)

where $x$ is the linear crosstalk power ratio. Figure 3.27 shows the predicted optical power penalties for out-of-band optical crosstalk for QPSK modulation, both with the same and different frequency bands:

![Fig. 3.27 Optical power penalties for out-of-band optical crosstalk for QPSK modulation with frequency re-use and with adjacent frequency bands.](image-url)
The in-phase case is identical to BPSK modulation, with higher power penalties as the RF phase difference increases to 45°, as for in-band crosstalk. This increase is not observed for ASK or BPSK modulation, so that the worst-case power penalties are actually worse for QPSK modulation. A 1 dB optical power penalty occurs at crosstalk levels of $-9.7$ and $-9$ dB for 45° and 0° RF phase difference respectively. This is relaxed to $-5.9$ dB if the crosstalk signal is at a different RF frequency and can be filtered electrically. The degradation due to the presence of the crosstalk signal in the electrical domain can be quantified by taking the difference between the power penalty with frequency re-use and the power penalty due to optical power addition (adjacent frequency bands).

### 3.5.4 Dependence of BER on RF Phase Difference and Crosstalk Level

This section presents the same results that were presented for BPSK modulation and ASK modulation in Sections 3.2.2 and 3.4.5 respectively, although in this case the modulation format is QPSK.

Firstly, the variation in BER over RF phase difference for an in-band crosstalk level of $-30$ dB is shown in Figure 3.28.

![Fig. 3.28 BER vs. RF phase difference for $-30$ dB of in-band optical crosstalk with frequency re-use for QPSK modulation.](image)

Results show that although for individual bit combinations there is a large change in BER, the symmetrical nature of QPSK modulation results in an average BER which for
–30 dB is hardly varying over RF phase difference. A clearer picture is obtained for 
–10 dB of optical crosstalk, as can be seen in Figure 3.29.

![Graph showing BER vs. RF phase difference](image)

**Fig. 3.29** BER vs. RF phase difference for –30 dB of in-band optical crosstalk with 
frequency re-use for QPSK modulation.

The graph shows the complex behaviour of individual signal and crosstalk bit 
combinations, with the average BER clearly showing the expected symmetry about 45°, 
135°, 225° and 315°.

If we are concerned about the ability of FEC to correct for BERs less than 10⁻⁵, 
then Figure 3.30 shows the change in BER over optical crosstalk for a constant signal 
level.

![Graph showing change in BER with optical crosstalk](image)

**Fig. 3.30** Change in BER with optical crosstalk for QPSK modulation at a constant 
signal level.
Results indicate that up to −14 dB of in-band optical crosstalk can be tolerated if FEC can correct for BERs up to $10^5$. This should be compared to −19 dB for ASK modulation and −13 dB for BPSK modulation. These trends mirror the power penalty trends, as expected. Recall that the optical power penalty is the additional optical power required at the receiver to recover a BER of $10^9$.

3.6 Dependence on RF Carrier Phase Difference

Previous sections have demonstrated the importance of the crosstalk carrier phase relative to the signal phase for in-band and out-of-band crosstalk where the RF carrier frequencies are the same. Whilst there can be a large change in power penalty at a given crosstalk level, it is important to consider the rate at which the RF phase varies, since this will determine the effective ‘time averaged’ power penalty. The timescale over which changes in RF phase occur is also important when a non-zero decision threshold is used, or when a DC-average threshold is used, since the average threshold depends on the time constant of the circuit relative to RF phase changes. If the time constant is less than the RF phase variations, then the decision threshold will vary as a function of RF phase difference, allowing a more optimum threshold for all RF phase differences. On the other hand, if the decision threshold time constant is greater than the rate at which the RF phase varies, then the decision threshold will not be able to track the signal variation as the RF phase difference varies. In this case, the decision threshold will respond to the average signal level, averaged over the RF phase difference.

The actual rate at which the RF phase will typically change is explored in the following sections, considering both the wireless link and the optical link.

3.6.1 Changes in the Wireless Link

The main source of phase changes occurs due to free-space propagation of the RF carrier in the electrical domain. This is due to the physical change in propagation distance from the mobile user to the base station antenna, or from the BS to the user. The fractional number of RF carrier wavelengths covered per second can be plotted as a function of carrier frequency for different sources of movement, as shown in Figure 3.31.
Analysis of In-band and Out-of-band Optical Crosstalk for Subcarrier Modulation

**Fig. 3.31 Fractional number of RF carrier wavelengths covered per second for different carrier frequencies and mobility rates.**

Figure 3.31 indicates that for practical RF frequencies (1 GHz to 100 GHz) there are less than 1 wavelength changes per second due to head movements, whilst there are 1 or more wavelength changes per second for a mobile user moving at walking speed or more. This means that for a fixed user RF phase changes are slow (0.01 to 1 Hz), whilst moving users experience faster changes (1 Hz to 10 kHz).

The first observation is that these frequencies are much lower than data rates of MHz or more and the rate of change of optical phase difference due to the laser linewidth (~ MHz for DFB lasers), so that any given bit will experience a crosstalk level corresponding to the RF phase difference at that time. This means that instantaneous BERs will vary depending on the RF phase difference, at a rate ranging from 0.01 to 1 Hz for a fixed user to 1 to 1 kHz for mobile users, for 1 degree phase change variations. Secondly, a slow decision threshold corresponds to a frequency response less than 0.01 Hz, whilst a fast threshold response corresponds to 1 kHz or greater. Whilst we have noted that for BPSK modulation the decision threshold is always zero, for ASK modulation or a higher-order modulation which involves amplitude, a decision threshold setting circuit is required. As discussed in the introduction, one issue is to what extent a decision threshold can track RF phase changes can reduce the optical power penalty, considering the worst-case phase-aligned scenario.

An additional requirement has to be considered if FEC is used for error-correction, namely whether or not the RF phase difference varies over an FEC ‘window’. FEC codes typically work over around 1000 bits, so that the BER over 1000
Chapter 3

bits must be less than or equal $10^{-5}$ for full correction of errors. For example the Reed-Solomon code RS(255,239) operates on 255 bytes (2040 bits) and is used in telecommunications [15, 16]. Figure 3.32 shows the fractional number of 360° RF carrier phase changes per 1000 bits as a function of mobile user speed, for a bit-rate that is one-tenth of the carrier frequency.

![Graph showing fractional number of phase changes](image)

**Fig. 3.32 Fractional number of 360° phase changes per 1000 bits as a function of mobile user speed, assuming carrier frequency is ten times the data rate.**

The graph assumes a data rate corresponding to one tenth the RF carrier frequency. If the data rate were greater, a 1000 bit time-window would be shorter and hence the fractional change would also be less. In any case, the rate of change corresponds to at most 0.4° per 1000 bits for a user travelling in a car. Hence for an FEC window, the RF phase difference is fixed. This means the actual BER seen by the FEC code will vary as the RF phase difference varies, requiring the system designer to assume the worst-case, i.e. phase-aligned RF carriers.

Additional phase changes can occur due to LO drift within the mobile phone or at the BS, although these are much slower, occurring over several minutes or more.

### 3.6.2 Changes in the Optical Link

Whilst the preceding section has focused on the wireless link, in which the user is mobile and can be moving, a further source of RF phase changes between signal and crosstalk occurs in the optical domain, as the data is transported over the optical network. The issue becomes one of propagation distance, which is not totally fixed as
may initially be thought. Thermal and mechanical effects can affect the total length of fibre, as well as changing the optical polarisation state of light propagating down the fibre [10]. Variations in polarisation will translate in variations in propagation length due to Polarisation Mode Dispersion (PMD) [22]. Experimental evidence obtained from experiments undertaken for this thesis shows that these effects translate into changes of several degrees of phase change over the order of minutes, at a carrier frequency of 1 GHz. This rate of change is much slower than that associated with the wireless link itself, so that the dominant source of RF phase change will be due to free-space propagation length changes due to the physical movement of the mobile user, as described in the preceding section.

3.7 COMPARISON OF MODULATION FORMATS

Sections 3.2 to 3.5 have provided a detailed description of the analytical models used to investigate the impact of both in-band and out-of-band optical crosstalk for BPSK, ASK and QPSK modulation. The analysis considers the case where both signal and crosstalk optical channels carry data at the same RF carrier frequency, and also the case where the crosstalk signal is at a different RF carrier frequency and can be electrically filtered. In the first case the relative carrier phase difference is important in determining the effect crosstalk signal present at the receiver in the electrical domain. The variation in this relative phase difference was shown to be dependent mainly on the wireless link (Section 3.6.1), so that the worst-case has to be assumed, given that there is no control of this parameter. This is irrelevant in the second case, since the crosstalk carrier is at a different frequency and is filtered.

The following sections will summarise and compare the results for the 3 modulation formats considered both for in-band and out-of-band optical crosstalk. This will allow a comparison of modulation formats to be made, showing the relative merits of different schemes.

3.7.1 In-band Optical Crosstalk

Figure 3.33 shows the worst-case and best-case optical power penalties for in-band optical crosstalk with frequency re-use, for BPSK, ASK and QPSK modulation.
Fig. 3.33 Comparison of modulation formats for in-band optical crosstalk with frequency re-use.

For ASK modulation, best- and worst-case results are shown both for square-law envelope detection and for PLL detection for completeness. The worst-case and best-case decision threshold types were selected, providing an upper and lower bound on power penalties (see Section 3.4). Since zero decision threshold are used for BPSK and QPSK modulation, only two curves are shown for each modulation type. If we compare worst-case power penalty curves, ASK modulation is worst (and identical to baseband intensity modulation for PLL detection), followed by QPSK modulation and finally by BPSK modulation. Crosstalk tolerances at a 1 dB optical power penalty will be compared at the end of this section.

Similarly, Figure 3.34 shows the optical power penalties arising from in-band optical crosstalk where the crosstalk RF carrier frequency is at a different frequency and is electrically filtered, for BPSK, ASK and QPSK modulation.
Analysis of In-band and Out-of-band Optical Crosstalk for Subcarrier Modulation

Fig. 3.34 Comparison of modulation formats for in-band optical crosstalk with adjacent frequency bands.

When the crosstalk signal carries data at a different RF frequency, Figure 3.34 shows that power penalties for ASK modulation with PLL detection are larger than for BPSK and QPSK modulation. Power penalties are identical for BPSK and QPSK modulation, with those for ASK modulation with square-law envelope detection slightly higher.

The analytical formulae given in the previous sections can be used to explain the differences between the various modulation formats. Equation 3.7 leads to the following expressions for Q for BPSK:

\[ Q_{11} = [1 + \sqrt{x} \cos(\Delta \theta)] + \sqrt{x} \cos(\Delta \phi)\sqrt{x} + \cos(\Delta \theta), \]
\[ Q_{10} = [1 + \sqrt{x} \cos(\Delta \theta)] - \sqrt{x} \cos(\Delta \phi)\sqrt{x} + \cos(\Delta \theta)] \]  

while Equations 3.21 and 3.22 lead to the following expressions for ASK:

\[ Q_{11} = [0.5 + \sqrt{x} \cos(\Delta \theta)] + \sqrt{x} \cos(\Delta \phi)\sqrt{x} + \cos(\Delta \theta), \]
\[ Q_{10} = [0.5 + \sqrt{x} \cos(\Delta \theta)], \]
\[ Q_{01} = 0.5 - \sqrt{x} \cos(\Delta \phi)(\sqrt{x} + \cos(\Delta \theta)) \]
\[ Q_{00} = 0.5 \]  

If we normalise the expressions so that \( Q \) is unity for no crosstalk and take the difference between the expressions and unity, we obtain the following expressions:

\[ \Delta Q_{11} = \sqrt{x} \cos(\Delta \theta)[1 + \cos(\Delta \phi)] + x \cos(\Delta \phi), \]
\[ \Delta Q_{10} = \sqrt{x} \cos(\Delta \theta)[1 - \cos(\Delta \phi)] - x \cos(\Delta \phi) \]  

for BPSK and for ASK:
\[
\Delta Q_{11} = 2 \sqrt{x} \cos(\Delta \theta)[1 + \cos(\Delta \phi)] + 2x \cos(\Delta \phi), \\
\Delta Q_{10} = 2 \sqrt{x} \cos(\Delta \theta), \\
\Delta Q_{01} = -2 \sqrt{x} \cos(\Delta \phi)[\sqrt{x} + \cos(\Delta \theta)] \\
\Delta Q_{00} = 0 
\]  
\text{(3.57)}

A positive difference improves the BER, whilst a negative difference worsens the BER. The symmetry exhibited by BPSK around \(\Delta \phi = 90^\circ\) is evident as \(\Delta Q_{11}\) and \(\Delta Q_{10}\) are equal and opposite to one another, with \(Q_{11}\) dominating errors for \(\cos(\Delta \phi)\) positive. Expressions for ASK show that \(Q_{11}\) dominates errors. Points at which Qs are equal can be obtained by inspection, but only explain the RF phase difference at which this occurs. The more important observation is that \(\Delta Q_{11}\) for ASK is twice that for BPSK, explaining why the resulting power penalties for BPSK are lower than for ASK. This arises from the bipolar nature of BPSK, which sets a zero decision threshold, effectively decreasing the relative impact of crosstalk relative to the signal, as indicated by the expressions above. The equations for ASK modulation can also be compared to those for baseband IM (3.58), confirming that expressions are identical when signal and crosstalk carriers are in-phase.

### 3.7.2 Out-of-band Optical Crosstalk

As explained previously, the situation for out-of-band optical crosstalk is analytically simpler since the crosstalk is linear or additive, with no mixing between the signal and crosstalk optical channels. However, there is still a dependence on modulation format and decision threshold type as shown in Figure 3.35.
Fig. 3.35 Comparison of modulation formats for out-of-and optical crosstalk with frequency re-use.

The approach taken for ASK PLL and envelope detection is the same as for the graph for in-band crosstalk. Best- and worst-case results are taken from the power penalty curves of Figures 3.15 and 3.16 i.e. considering both average and phase-dependent decision thresholds. The worst-case crosstalk curve for ASK square-law envelope detection is higher than the worst-case curve for ASK PLL detection. Similarly, the minimum power penalty curve for ASK modulation occurs for PLL detection, although results for square-law envelope detection are very similar up to around 1 dB penalty. These results indicate the importance of detection method for ASK modulation. For QPSK modulation the worst-case power penalty curve is also above the traditional baseband IM curve, also indicating that the modulation format is relevant.

When signal and crosstalk RF carrier frequencies are different, the crosstalk signal can be electrically filtered, resulting in an optical power penalty which is entirely due to the additional crosstalk optical power, so that there is no dependence on modulation scheme, as shown in Figure 3.36.
Fig. 3.36 Comparison of modulation formats for out-of-and optical crosstalk with adjacent frequency bands.

The potential benefits arising from using the electrical domain to electrically filter crosstalk signals is evident.

### 3.7.3 Summary of Results

Results can be summarised by considering the crosstalk level at which a 1 dB optical power penalty occurs, since this represents a typical system margin that might be allowed for degradations due to optical crosstalk. Table 3.2 summarises these crosstalk levels for the worst-case RF phase difference for ASK, BPSK and QPSK modulation, for in-band and out-of-band crosstalk.
Table 3.2 Optical crosstalk at which a 1 dB optical power penalty occurs for in-band and out-of-band crosstalk for various modulation formats, with the same and different crosstalk RF carrier frequency.

<table>
<thead>
<tr>
<th></th>
<th>In-Band Optical Crosstalk</th>
<th>Out-of-Band Optical Crosstalk</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Same frequency</td>
<td>Adj. Frequency</td>
</tr>
<tr>
<td>IM</td>
<td>-23.0 dB</td>
<td>-17.9 dB</td>
</tr>
<tr>
<td>ASK  Env.</td>
<td>-24.5 dB</td>
<td>-14.5 dB</td>
</tr>
<tr>
<td>ASK PLL</td>
<td>-23.0 dB</td>
<td>-17.9 dB</td>
</tr>
<tr>
<td>BPSK</td>
<td>-17.3 dB</td>
<td>-13.4 dB</td>
</tr>
<tr>
<td>QPSK</td>
<td>-18.3 dB</td>
<td>-13.4 dB</td>
</tr>
</tbody>
</table>

The results in Table 3.2 are based on the worst-case power penalty for a specific modulation format, which for ASK may occur for an average decision threshold or a phase-tracking decision threshold. The dependence on modulation format is clear, with BPSK the least sensitive to crosstalk, followed by QPSK modulation and then ASK modulation. The difference between ASK modulation and baseband modulation for out-of-band crosstalk occurs because the worst-case for ASK corresponds to envelope detection with an RF phase-tracking decision threshold. The most significant result for all modulation formats is the additional margin achieved when the crosstalk RF frequency is different to the signal RF frequency and can be electrically filtered. For a 1 dB power penalty, this corresponds to an extra 4 to 5 dB of in-band crosstalk, and an extra 3 to 5 dB of out-of-band crosstalk.

The benefit of forward error-correction (FEC) has been discussed in previous sections, allowing the tolerance to crosstalk to be greatly increased. Table 3.3 compares the crosstalk level at which a 1 dB power penalty occurs with no FEC and the crosstalk level at which a BER of $10^{-5}$ occurs, which can be corrected with FEC.
Table 3.3 Optical crosstalk at which a 1 dB optical power penalty occurs for in-band crosstalk for various modulation formats, with the same and different crosstalk RF carrier frequency and corresponding FEC limit of $10^{-5}$ BER.

<table>
<thead>
<tr>
<th></th>
<th>In-Band Optical Crosstalk</th>
<th>Same RF frequency</th>
<th>Adj. Frequency</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td>1 dB penalty</td>
<td>BER $10^{-5}$</td>
</tr>
<tr>
<td>IM</td>
<td>-23.0 dB</td>
<td>-19.2 dB</td>
<td>N/A</td>
</tr>
<tr>
<td>ASK Env.</td>
<td>-24.5 dB</td>
<td>-25.0 dB</td>
<td>-19.8 dB</td>
</tr>
<tr>
<td>ASK PLL</td>
<td>-23.0 dB</td>
<td>-19.2 dB</td>
<td>-14.9 dB</td>
</tr>
<tr>
<td>BPSK</td>
<td>-17.3 dB</td>
<td>-13.0 dB</td>
<td>-8.7 dB</td>
</tr>
<tr>
<td>QPSK</td>
<td>-18.3 dB</td>
<td>-13.9 dB</td>
<td>not calculated</td>
</tr>
</tbody>
</table>

Table 3.3 shows that error-free operation can be achieved with relatively high levels of in-band crosstalk using FEC even when the RF crosstalk frequency is the same as the signal RF frequency. Comparing Tables 3.3 and 3.2 shows the crosstalk level at which error-free operation is achieved using FEC is very similar to the crosstalk at which a 1 dB penalty occurs when no FEC is used and signal and crosstalk RF frequencies are different, except for ASK square-law envelope detection. Table 3.3 shows that ensuring the crosstalk RF frequency is different and having FEC further increases the tolerance to in-band crosstalk.

3.8 **Impact of Electrical Crosstalk on Optical Crosstalk**

The analytical technique presented in the previous sections is one of the main techniques used to analyse the impact of optical crosstalk. It allows the impact of an unwanted optical signal to be quantified in terms of the degradation in the received electrical waveform. However, both optical signals are assumed to be transporting a pure electrical signal, i.e. an electrical signal that is free from unwanted electrical crosstalk, or co-channel interference. The following section demonstrates how the model used can be readily modified to assess the impact of additional electrical
crosstalk. While the analysis is shown for baseband intensity modulation, this should be valid for ASK modulation as well since both have the same in-band and out-of-band crosstalk characteristics.

3.8.1 Modified Model used to Assess the Impact of Additional Electrical Crosstalk

The received photocurrent for a baseband intensity modulated optical signal beating with an in-band crosstalk channel produces the following current \( I \):

\[
I = \alpha + \beta x + 2\alpha \beta \sqrt{x} \cos(\Delta \theta)
\]  
(3.58)

where \( \alpha \) and \( \beta \) correspond to signal and crosstalk data bits, taking values +1 and 0, \( x \) is the linear optical crosstalk ratio, and \( \Delta \theta \) is the optical phase difference between the two optical carriers. The presence of additional electrical crosstalk is incorporated by modifying the expressions for \( \alpha \) and \( \beta \) as follows:

\[
\alpha \rightarrow \frac{\alpha + \delta z}{1 + z},
\]

\[
\beta \rightarrow \frac{\beta + \epsilon z}{1 + z}
\]  
(3.59)

where \( \delta \) and \( \epsilon \) are the electrical crosstalk data bits, and \( z \) is the electrical current ratio, i.e. \( z^2 \) is the electrical crosstalk power ratio. Note that we assume the electrical signal is amplified and kept at a constant output power, so that the maximum electrical current is normalised to 1 for all electrical crosstalk values. The error function can be used to derive BER for the 16 possible bit combinations for all possible combinations of \( \alpha, \beta, \delta \) and \( \epsilon \). A standard DC-coupled decision threshold \( D \) is used:

\[
D = 0.5(1 + x)
\]  
(3.60)

Figure 3.37 shows the predicted optical power penalties as a function of optical crosstalk for different levels of electrical power crosstalk ratio for in-band crosstalk:
Fig. 3.37 Effect of electrical crosstalk on in-band optical crosstalk for baseband intensity modulation.

The results indicate that up to –40 dB of electrical co-channel interference can be present without any noticeable impact on optical power penalties. A level of –30 dB produces an additional 0.2 dB penalty, increasing to 0.4 dB for –25 dB. If –20 dB of electrical crosstalk is present, then the resulting additional optical power penalty is about 1 dB. This indicates that a relatively high level of electrical co-channel interference can be tolerated without increasing optical power penalties.

A similar analysis for out-of-band optical crosstalk produces results shown in Figure 3.38.

Fig. 3.38 Effect of electrical crosstalk on out-of-band optical crosstalk for baseband intensity modulation.
Figure 3.38 shows that up to $-40 \text{ dB}$ of co-channel electrical crosstalk can be tolerated without any measurable excess penalty. Excess optical power penalties of 0.1, 0.2 and 0.8 dB for electrical crosstalk levels of $-30$, $-25$ and $-20 \text{ dB}$ respectively. These results are very similar to those for in-band crosstalk, indicating that the main degradation lies in the signal wavelength itself, with minor additional degradation due to in-band effects. Both results indicate that co-channel specifications in the electrical domain will ensure no additional impact is observed due to optical crosstalk. Applying the above analysis to BPSK modulation is thus unnecessary, as very similar results are expected.

## 3.9 Conclusions

This chapter has provided a detailed description of the analytical model used to investigate optical crosstalk in fibre-radio networks. Analytical expressions for ASK, BPSK and QPSK modulation have been presented, allowing the impact of both in-band and out-of-band optical crosstalk to be assessed. Optical power penalties were shown to depend on modulation format. This must be considered when designing fibre-radio networks, as discussed in Chapters 5 and 6.

The importance of the RF phase difference between signal and crosstalk has been highlighted when there is frequency re-use, as this varies optical power penalties at a given crosstalk level. Since RF phase difference cannot be controlled in a fibre-radio environment, worst-case must be assumed as this guarantees operation in all conditions. Analysis for ASK modulation showed that the decision threshold is also important. Different power penalties are observed depending on whether the decision threshold responds to the average RF phase difference (i.e. is independent of RF phase difference) or is able to track the RF phase and adjust the decision threshold accordingly. The effect of receiver detection method for ASK modulation, for envelope detection and PLL detection, also influences the resulting power penalties. These factors are important as they can affect the tolerance to crosstalk and hence modify component crosstalk specifications.

The dependence of power penalties on modulation format is important, both for in-band and out-of-band crosstalk. The potential for reducing the impact of crosstalk has been demonstrated for subcarrier modulation, by ensuring the crosstalk signal is at a different RF frequency to the signal and can be electrically filtered. This has the effect...
of either reducing the optical power penalty at a given crosstalk level, or of relaxing optical component specifications if a given system margin is used. This has to potential to enable the use of low-cost optical components or increase network size, as will be discussed in Chapter 5. Results derived in this chapter will be used in Chapter 6 to illustrate how optical crosstalk can be evaluated for different WDM fibre-radio networks, comparing different topologies. Theoretical results also allow the potential of FEC to be assessed for the various modulation formats, indicating that optical component specifications can be further relaxed if FEC is used to correct errors resulting from higher levels of optical crosstalk. The effect of electrical crosstalk in the wireless domain on optical power penalty was also investigated and shown to be insignificant for reasonable levels of electrical crosstalk.

The following chapter will present experiments undertaken to confirm the analytical models used and predicted optical power penalties. Computer simulations performed using commercial software packages will also be presented, further confirming theoretical results.

3.10 REFERENCES


Analysis of In-band and Out-of-band Optical Crosstalk for Subcarrier Modulation


4 Simulation and Measurement of In-band and Out-of-band Optical Crosstalk

4.1 INTRODUCTION

Chapter 3 has presented the analytical models used to numerically evaluate the impact of optical crosstalk in fibre-radio networks, considering three modulation formats: ASK, BPSK and QPSK. It has been shown that the BER and power penalty depend on the phase difference between signal and crosstalk RF carriers when the crosstalk and signal RF carrier frequencies are the same. Furthermore, optical power penalties depend on modulation format, so that the tolerance to in-band optical crosstalk will depend on the RF modulation scheme. The importance of the crosstalk RF carrier frequency has also been shown, since electrical filtering of the crosstalk carrier when the crosstalk RF frequency is different to the signal carrier frequency allows the impact of crosstalk to be minimised.

The present chapter provides confirmation of the theoretical results. Two approaches have been taken in order to achieve this. Firstly, software simulation programs have been used to run ‘virtual’ experiments, allowing both in-band and out-of-band optical crosstalk to be assessed for ASK, BPSK and QPSK modulation formats. Confidence in the software was ensured by reproducing baseband intensity modulation crosstalk results in the optical domain, and electrical crosstalk results for ASK and BPSK modulation. A software simulation package was used in a similar way by Moura et al. for an ASK modulated link [1]. Secondly, an experimental investigation of optical crosstalk for BPSK modulation was performed, both for in-band optical crosstalk and out-of-band optical crosstalk. As discussed in Chapter 2, many experimental investigations of optical crosstalk have been undertaken, particularly for in-band crosstalk with baseband IM [2-8]. An experimental investigation of out-of-band crosstalk can be found in [9]. No experimental results have been presented for systems using digital subcarrier modulation.
Section 4.2 presents simulations of in-band optical crosstalk with frequency re-use. Experimental results for BPSK modulation will be presented in Sections 4.3 and 4.4 for the cases of frequency re-use and adjacent frequency bands respectively. Similarly, Sections 4.5 and 4.6 present simulation and experimental results for out-of-band optical crosstalk. Section 4.7 focuses on electrical crosstalk measurements, in order to clarify results obtained for optical crosstalk, since BER measurements are performed electrically. As electrical power penalties are different to optical power penalties, both theoretical and experimental results will be discussed.

4.2 Simulation of In-band Optical Crosstalk for ASK, BPSK and QPSK Modulation

This section discusses simulation results obtained for ASK and BPSK modulation for in-band optical crosstalk. Both signal and crosstalk channels use the same RF carrier frequency. The states of polarisation of both optical signals are assumed to be aligned as this ensures maximum crosstalk. States of polarisation are not controlled in a network and it has been shown that the worst-case alignment should be considered in network design [10]. This ensures that the network can operate for all polarisation states.

4.2.1 Simulation of In-band Crosstalk for ASK Modulation

A commercial optical link simulation software package, GOLD [11], was used to run a virtual experiment. The software allows the models of various optical components to be linked together within Labview [12]. Optical and electrical components are available and the Labview environment can be used to display results and vary system parameters. Figure 4.1 shows a representation of the optical link used to simulate in-band optical crosstalk, while Figure 4.2 shows a representation of the electrical link used to modulate and demodulate data in ASK modulation format.
Figure 4.1 shows that a single laser diode is split into two separate paths, one representing the signal laser, the other the crosstalk laser. A separate Mach-Zehnder (MZ) modulator modulates each optical signal. Each modulator is driven by electrical data in the appropriate format, in this case ASK modulation [13]. Two distinct pseudo-random bit sequences, each comprising 32 bits, are used for the simulation, ensuring an equi-probable distribution of 1s and 0s. The crosstalk signal level is controlled using a variable optical attenuator (ATT), before being combined with the signal. The combined optical signals are then detected using a PIN diode. Since the same laser diode is used the relative optical phase difference between signal and crosstalk fields is constant. In order to produce results for in-band incoherent optical crosstalk [2, 3, 5], an optical phase delay is used in the crosstalk path ($\Delta \theta$). This allows the phase of the crosstalk optical carrier to be offset by a specified amount. Varying the phase delay between 0° and 356° in 4° steps allows different BERs to be obtained and averaged over the range. Hence results are obtained for incoherent crosstalk. Note that this is equivalent to using a length of optical fibre so as to decorrelate the two paths; this approach was taken experimentally as described in the experimental section. The simulation method used to obtain incoherent crosstalk is preferred over using two separate laser sources, as the degree of incoherence would depend on the way in which the software varies the optical phase. The laser power is increased to increase the received optical power so as to compensate for the degradation due to optical crosstalk, allowing an optical power penalty to be obtained for a given level of in-band crosstalk.
Figure 4.2 shows the electrical set-up used to produce data in ASK modulation format, together with the receiver demodulator structure:

![Diagram of Modulator and Demodulator](image)

Fig. 4.2 Schematic representation of the electrical modulator/demodulator used for ASK modulated data.

A local oscillator (LO) and mixer are used to upconvert the baseband data to the appropriate RF carrier frequency, 10 GHz in this case. The data rate was 1 GHz. The Mach-Zehnder modulator was biased at its 3 dB transmission point. An optical modulation depth of 1 % was used. Note that an electrical phase shifter ($\Delta\phi$) allows the relative phase between the signal and crosstalk RF carriers to be adjusted (two separate LOs could have been used if their initial phase could have been controlled and had their frequencies been identical). The receiver uses a square-law envelope detector [14], obtained using Labview tools by squaring the electrical signal ($X^2$) and then removing RF components with a low-pass filter (LPF). In order to obtain BERs, Labview was used to process the electrical output of the PIN diode, sampling the output bit sequence at bit intervals centered in the middle of the bit period. Q values were calculated for each bit, evaluating the BER for individual bits using the error function [13]. The average BER for the whole bit-sequence was then obtained and plotted against total received optical power. A DC-averaged decision threshold was used, as derived in Section 3.4.1 (Equation 3.19). The appropriate choice of bit sequences ensured all four signal-crosstalk combinations (00, 01, 10, 11) were equally represented. Figure 4.3 shows the recovered electrical 32 bit sequence obtained for in-band crosstalk levels of $-50$ dB.
and −10 dB respectively, for in-phase RF carriers (Δϕ=0) and an optical phase difference (Δθ) of 90°.

![Electrical bit sequence for ASK modulation at −50 dB and −10 dB crosstalk for in-phase RF carriers and an optical phase difference of 90°.](image)

**Fig. 4.3** Electrical bit sequence for ASK modulation at −50 dB and −10 dB crosstalk for in-phase RF carriers and an optical phase difference of 90°.

Figure 4.3 shows the effect of electrical filtering (rounding of bit transitions), together with the four possible signal-crosstalk bit combinations: signal zero and crosstalk zero (00), signal zero and crosstalk one (01), signal one and crosstalk zero (10) and signal one and crosstalk one (11). The corresponding decision thresholds, D, at −50 dB and −10 dB are also shown. Note that for this particular optical phase difference, signal-crosstalk beat terms are equal to zero, so that the noise is additive (i.e. the same as for out-of-band crosstalk). Hence the only degradation in BER occurs for ‘01’ bit combinations, as the ‘11’ level is increased.

The simulated optical power penalty for in-band optical crosstalk at different RF carrier phase differences is shown in Figure 4.4.
4.2.2 Simulation of In-band Crosstalk for BPSK Modulation

The investigation of in-band optical crosstalk with RF frequency re-use for BPSK modulation was also performed with GOLD using a procedure which is nearly identical to that described above. The optical link is identical since the crosstalk is also in-band and incoherent. The only difference is the modulation format, which is changed by modifying the electrical modulation set-up slightly, as can be seen in Figure 4.5.
Since BPSK modulation changes the phase of the RF carrier [13], the incoming baseband data has to be biased appropriately so the data becomes bipolar, i.e. ±0.5. Upconversion to RF via the mixer produces the required BPSK modulated signal. This is fed to the Mach-Zehnder modulator, as before. The electrical receiver is different, however, since envelope detection of the RF carrier would not discriminate between 1s and 0s. The modulator LO is used for downconversion, producing an output whose polarity varies depending on the bit sequence. Numerical calculation of Q is then possible using a decision threshold equal to zero, allowing BERs to be calculated and averaged over optical phase difference. Optical power penalties are then obtained by varying the received optical power. Software simulation results are shown in Figure 4.6, together with analytical results (from Chapter 3).
4.2.3 Simulation of In-band Crosstalk for QPSK Modulation

The simulation set-up described in Section 4.2.2 for BPSK modulation was modified to allow the investigation of in-band optical crosstalk for QPSK modulation. The QPSK data format is achieved by combining two orthogonal BPSK signals, as shown in Figure 4.7.
Simulation and Measurement of In-band and Out-of-band Optical Crosstalk

Fig. 4.7 Schematic representation of the electrical modulator used for QPSK modulated data.

A single LO signal is used for both signal and crosstalk data, as before, using a variable RF phase shifter ($\Delta\phi$) to control the relative phase difference between signal and crosstalk RF carriers. The corresponding LO signal is mixed with a 32 bit PRBS to produce a BPSK modulated RF carrier. A 90° phase shifter allows the signal and crosstalk LO signals to be mixed with separate PRBS data to produce the desired BPSK channel orthogonal to the original carriers. The two orthogonal carriers are combined to form a QPSK modulated carrier, one for the signal path and one for the crosstalk path. Four different 32-bit sequences were used in the simulation, ensuring an equi-probable distribution of signal and crosstalk ones and zeroes. The receiver structure is identical to that used for BPSK modulation.

Figure 4.8 shows the simulation results obtained for in-band crosstalk, together with analytical results from Section 3.5.1.
4.3 **Experimental Investigation of In-band Optical Crosstalk for BPSK Modulation with Frequency Re-use**

This section provides experimental confirmation [15] of the analytical model for BPSK modulation where the signal and crosstalk RF frequencies are the same.

4.3.1 **Experimental Set-up and Results**

Figure 4.9 shows the experimental set-up used for the measurement of optical power penalties at varying RF phase differences due to in-band optical crosstalk:
Fig. 4.9 Experimental set-up for the measurement of in-band optical crosstalk for BPSK modulation.

A single DFB laser at wavelength $\lambda_1$ (1547 nm) provides both the desired optical signal and the in-band crosstalk source. The two optical signals are each externally modulated by an RF carrier at 3.5 GHz carrying 155 Mb/s data in BPSK modulation format. In order to ensure adequate statistics and full signal-crosstalk bit combinations, two separate data generators provide the 155 Mb/s radio data, with a pseudo-random bit sequence length of $2^{23}-1$. A single local oscillator (LO) provides a 3.5 GHz RF carrier that is mixed with the baseband data to provide an output in BPSK format. An electrical phase shifter allows the relative phase between the signal and crosstalk RF carriers to be varied. Electrical amplifiers are used to obtain an electrical signal that can drive the two separate electro-optic modulators (EOMs). Any harmonics of the 3.5 GHz carrier were attenuated by the 4.2 GHz bandwidth of the electrical amplifier and the bandwidth of the EOMs, which was less than 4 GHz.

In Figure 4.9 a 2-km long spool of single-mode fibre ensures that the optical signals are rendered incoherent, so that the optical phase differences between the desired and crosstalk signals are random and time-averaged over a bit-period (i.e. incoherent [2]). An optical isolator (ISO) ensures that reflections and Rayleigh backscatter do not affect the laser or the crosstalk optical path. A variable optical attenuator (ATT) in the crosstalk path enables the level of optical crosstalk to be set accurately. After the optical channel and crosstalk signals are modulated, they are combined and amplified to compensate for losses due to the EOMs and other optical components. An erbium-doped fibre amplifier (EDFA) operating at constant current
provides a gain of around 20 dB. A variable optical attenuator at the input of the amplifier allows the input to the amplifier to be kept constant throughout the experiment, as the level of crosstalk is varied. This ensures the EDFA provides a constant gain and a constant level of amplified spontaneous emission (ASE) noise (-32 dB below the signal at 0.1-nm resolution). The optical amplifier is followed by an optical attenuator that allows the total received optical power incident on a PIN photodiode (PD) to be varied. An optical filter with a bandwidth of 1.2 nm is centered on the optical signal, allowing most of the ASE to be optically filtered. Polarisation controllers are not shown in the figure for clarity. Polarisation controllers were used before the Mach-Zehnder modulators, which are highly polarisation sensitive, and before the signal and crosstalk paths were combined via the 3 dB coupler as optical polarisation states had to be matched before being recombined and detected at the receiver. In-band crosstalk relies on identical wavelengths and states of polarisation [10]; use of a single laser source ensures the first condition is automatically met, otherwise two separate laser sources can never be matched sufficiently closely. Polarisation states were matched by monitoring the detected signal on an RF spectrum analyser, maximising the noise level. Additional optical isolators were used along the path to minimise the impact of optical reflections.

The detected electrical RF carrier is amplified before being downconverted using a mixer driven by the signal LO. The recovered 155 Mb/s data is filtered using a low-pass filter (LPF), removing unwanted higher-frequency components. Further electrical amplification is required to obtain enough RF power. A bit-error-rate testset (BERT) is used to obtain bit-error-rates (BERs) at different optical powers at a given crosstalk level. This is repeated at various crosstalk levels, producing a series of BER curves as shown in Figure 4.10.
Figure 4.10 shows the degradation in BER caused by higher levels of optical crosstalk. The BER curves shift towards higher received optical powers. Only two points per BER curve were taken as thermal effects caused the RF phase difference to vary quickly [16, 17], limiting the total measurement time over which readings could be taken at different crosstalk levels. In all cases BERs better than $10^{-9}$ could have been obtained, as no indication of a BER floor is apparent from Figure 4.10 (this was confirmed experimentally). However, this would have required additional measurement time, leading to a changing RF phase difference. The measurements were taken for RF phase differences of 0°, 45° and 90°, allowing optical power penalties as a function of optical crosstalk for different RF carrier phase differences to be obtained, as shown in Figure 4.11.
4.3.2 Discussion of Results

The measurements show that the largest power penalties are obtained for no RF phase difference (where the RF carriers are aligned), whereas orthogonal carriers produce the lowest penalties. Experimental results are in very good agreement with theory up to around 4 dB of optical power penalty, beyond which there is an increasing discrepancy, for all three sets of results.

The potential effect of ASE at high levels of optical power were investigated using a Gaussian approximation, adding a signal-dependent noise term to the thermal noise variance used in numerical evaluations of power penalties. The effects of spontaneous-spontaneous beat noise and signal-spontaneous beat noise were found to produce at most 0.5 dB of extra power penalty at a power penalty of 9 dB. Whilst this decreases the difference between theory and experiment, it does not fully account for
the discrepancy. Optical reflections were minimised using isolators in the link, and mismatched optical polarisation states would have reduced measured penalties.

In the electrical domain, nonlinearities caused by electrical amplifiers and the EOMs could potentially degrade the received signal. However, care was taken to ensure harmonics of the RF carrier were removed; an RF spectrum analyser confirmed harmonics were at least 40 dB below the signal carrier. Distortions caused by the non-linear EOM response were minimised by using a low optical modulation depth (10%), by controlling the RF power driving the modulators. The main source of degradations is believed to be due to the use of two separate data generators (different brands) and differences in electrical amplification and mixing in the generation of BPSK modulated data. A difference in EOM characteristics may produce a slight discrepancy, although $V_\pi$ and extinction ratios were similar. Different clock synchronisation between data generators can also result in a degradation in BER at high crosstalk, as will be shown in Section 4.7. An increase in measured power penalty can be caused by an imperfect synchronisation by the BERT between its own clock and the incoming data, resulting in a non-optimum sampling time, which is not in the middle of the eye. Electrical BER measurements of electrical crosstalk were undertaken to investigate any potential effects in the BER measurements themselves, and will be discussed in Section 4.7.

The experimental results presented in this section confirm that power penalties for BPSK modulation are indeed much lower than for ASK and baseband IM modulation. The dependence on relative signal-crosstalk RF carrier phase difference has also been confirmed, in good agreement with theory. Experimental differences at high levels of optical crosstalk are partly explained by the presence of ASE, although electrical effects are believed to account for most of the differences.

4.4 Experimental Investigation of In-band Optical Crosstalk for BPSK Modulation with Adjacent Frequency Bands

4.4.1 Experimental Results and Discussion

The experimental set-up described in the previous section was also used to investigate the impact of in-band crosstalk as a function of the frequency separation
between the wireless signal and crosstalk electrical carrier frequencies [15]. A separate local oscillator was used for the crosstalk carrier, allowing the crosstalk carrier frequency to be varied.

Figure 4.12 shows the experimental results obtained for in-band crosstalk where the crosstalk carrier frequency is offset by 75, 150 and 300 MHz (the data rate is 155 MHz).

![Graph showing optical crosstalk and Power Penalty vs RF offset](image)

*Fig. 4.12 Experimental and analytical optical power penalties for in-band crosstalk for BPSK modulation with crosstalk RF carrier offset from signal carrier.*

Also shown are analytical predictions for perfect electrical filtering of the crosstalk carrier, which is the best case, and for identical RF frequencies that are in-phase, which is the worst case. Optical power penalties at a given crosstalk level are shown to decrease as the RF frequency offset increases. The importance of the electrical frequency separation is evident showing the importance of electrical filtering [18-21]. Good agreement is observed between theory (different RF frequency) and experiment for a 300 MHz offset, which corresponds to twice the data rate, up to a crosstalk level of approximately −8 dB. Beyond this power penalties rapidly increase, as observed for identical RF frequencies (Section 4.3). Since the same experimental set-up was used, the same measurement issues exist, as previously discussed. If we consider a maximum power penalty of 1 dB, the allowed maximum crosstalk varies from −13 dB to −17 dB depending on whether the signal and crosstalk carrier frequencies are identical or different (and filtered).
4.5 Simulation of Out-of-band Optical Crosstalk for ASK and BPSK Modulation

This section describes simulation results obtained for ASK and BPSK modulation for out-of-band optical crosstalk. A brief description of the simulation software will be given.

4.5.1 Simulation of Out-of-band Crosstalk for ASK Modulation

*VPITransmissionMaker* [22], an improved version of the *GOLD* software package developed by the same authors, was used to run a virtual experiment allowing the impact of out-of-band optical crosstalk to be evaluated. The simulation set-up is an idealised version of the experiment described in Section 4.3, and closely matches that described for *GOLD* in Section 4.2.

Two separate laser sources are used, separated by 50 GHz. A PIN diode detects the combined signal and crosstalk, adding Gaussian receiver thermal noise. In the electrical domain, standard ASK modulation is achieved by mixing data at 1 Gb/s with an RF carrier at 8 GHz. Two separate LOs are used, and two separate data generators provide different pseudo-random bit sequences ($2^7$-1). The simulation is run over a time window of 1024 bits. The receiver uses a separate LO, downconverting the detected RF signal to baseband. Low-pass filtering removes unwanted RF components, before going through a clock-recovery circuit and a BER module. The software modules allow plots of BER versus total received optical power to be plotted for different crosstalk levels, as shown in Figure 4.13.
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Fig. 4.13 BER curves for out-of-band crosstalk with in-phase carriers for ASK modulation obtained using VPITransmissionMaker.

Optical power penalties at different crosstalk levels are obtained from Figure 4.13, and are plotted in Figure 4.14 for varying RF phase differences.

Fig. 4.14 Simulated optical power penalties for out-of-band crosstalk for ASK modulation as a function of RF phase difference.

Figure 4.14 also shows analytical results for RF phase differences of $0^\circ$, $45^\circ$ and $90^\circ$. Simulated results for out-of-band crosstalk are in perfect agreement with analytical results.

4.5.2 Simulation of Out-of-band Crosstalk for BPSK Modulation

VPITransmissionMaker software was also used to confirm out-of-band crosstalk predictions for BPSK modulation. The simulation schematic is identical to that
described in the previous section, except that the electrical modulation format is BPSK modulation. The receiver operates with a zero decision threshold. Simulation results are identical to those obtained for ASK modulation using downconversion, as shown above. This is in agreement with theory.

4.6 **Experimental Investigation of Out-of-band Optical Crosstalk for BPSK Modulation with Frequency Re-use**

This section provides an experimental investigation of out-of-band crosstalk for BPSK modulation [15]. The experimental set-up is described, together with experimental results.

4.6.1 **Experimental Set-up and Results**

Figure 4.15 shows the experimental set-up used to verify the theoretical predictions of out-of-band crosstalk in a fibre-radio WDM system using BPSK modulation.

![Experimental set-up for the measurement of out-of-band optical crosstalk for BPSK modulation.](image)

The experimental set-up is similar to that used for the measurement of in-band optical crosstalk described in Section 4.3. However, two different laser sources are used. Two separate optical paths implement the desired channel and crosstalk signal, each comprising a distributed feedback (DFB) laser (λ_1 = 1549.4 nm, λ_2 = 1550 nm) and an electro-optic modulator (EOM). Separate data channels at 155 Mb/s modulate an RF signal at 2 GHz in BPSK modulation format. This is achieved by mixing the
baseband data with the LO. An RF phase shifter was used to control the RF phase difference between the signal channel and crosstalk signal RF carriers. The two data generators generate pseudo-random bit sequences (PRBS) of length $2^{23} - 1$. Electrical amplifiers are used to amplify the upconverted signal so as to drive the EOMs. The EOMs are biased at the 3 dB transmission point.

The two optical signals are coupled together and amplified using an erbium-doped fibre amplifier (EDFA) with a gain of 23 dB in order to compensate for losses in the optical link. The EDFA operates in constant current mode with a fixed input power, controlled via the input optical attenuator. This ensures a constant amplified spontaneous emission (ASE) noise contribution during measurements. An optical bandpass filter (BPF) with a bandwidth of 2 nm was included after the EDFA in order to filter out excess ASE noise. Since out-of-band crosstalk is being measured, a large filter bandwidth was required as both signal and crosstalk wavelengths have to be present at the photodetector (PD). Polarisation controllers are used before the polarisation-dependent modulators. However, unlike in-band crosstalk, no further polarisation control is necessary as there is no mixing between the optical signals which results in a component at the required RF frequency.

At the receiver the optical signals are detected using a PIN diode and the same 2 GHz signal LO is used to downconvert the detected modulated RF signals and recover the data. Electrical amplifiers are used in the receiver together with a 150 MHz low-pass filter that removes any high frequency components at the output of the mixer. A bit-error-rate testset (BERT) measures the BER of the recovered 155 Mb/s data as a function of optical crosstalk ratio (varied via the optical attenuator in the optical path of $\lambda_2$). A BER of $10^{-9}$ is obtained for a received optical input power at the PD input of $-19$ dBm.

By measuring a BER curve at a particular crosstalk ratio and carrying out a back-to-back system measurement, the crosstalk-induced optical power penalty could be determined. Figure 4.16 shows the BER curves at different crosstalk levels for phase-aligned carriers.
Fig. 4.16 Experimental BER curves for out-of-band crosstalk with in-phase carriers for BPSK modulation.

Figure 4.16 shows the BER against total received optical power at various crosstalk levels. It clearly shows the degraded sensitivity as the crosstalk level is increased. Note that multiple BER measurements were taken at each crosstalk level. This was possible as no optical fibre was present in the experimental set-up, eliminating the RF phase drifts observed for in-band crosstalk, which seriously limited measurement times.

Figure 4.17 shows the measured optical power penalty at a BER of $10^{-9}$ as a function of optical crosstalk ratio, for $\Delta \phi = 0$, 45° and 90°, plotted alongside the analytical predictions for out-of-band crosstalk.

Fig. 4.17 Experimental and analytical optical power penalties versus out-of-band crosstalk as a function of RF carrier phase difference.
4.6.2 Discussion of Results

The experimental measurements in Figure 4.17 show very good agreement with theory, with a small discrepancy between the two sets of results appearing at higher crosstalk levels (> -6 dB). There is better agreement for orthogonal RF carriers compared to in-phase carriers. However, there is still a discrepancy at > -2 dB crosstalk at 90° RF phase difference. Hence an additional penalty was measured electrically, possibly due to a residual crosstalk RF carrier component observed in the time-domain due to imperfect rejection by the mixer. Another potential source of error is imperfect clock-data synchronisation by the BERT. Experimental measurement error in the optical domain should not exceed 0.2 dB. Signal-dependent noise due to the presence of ASE due to the large optical filter bandwidth can account for a proportion of the deviation, particularly at high crosstalk levels. For the in-phase case the difference between theory and experiment at high crosstalk levels is of the order of 1 dB, 0.5 dB of which may be due to ASE. The remaining error could be assigned to different data generator signal qualities and differing electrical modulator characteristics (amplifiers and mixers). Different EOM characteristics can possibly also contribute to the difference, although this should be minimised by the fact that modulator characteristics were similar.

Experimental results are in good agreement with theory up to 2 to 3 dB of optical power penalty, showing the importance of RF phase difference in determining the optical power penalty at a given crosstalk level. When RF carriers are orthogonal results are identical to the case where different RF frequencies are used and electrical filtering removes the crosstalk signal (Section 3.3.2). Although this was not performed experimentally for out-of-band crosstalk, experiments for in-band crosstalk with different crosstalk RF frequencies described in Section 4.4 confirm that results are the same as for identical RF frequencies but with a 90° RF phase difference.

4.7 Electrical Crosstalk Measurements

The experimental results described in Sections 4.3, 4.4 and 4.6 have been in good agreement with theoretical predictions at low crosstalk levels and therefore low power penalties (up to 3-5 dB). However, results show an increasing deviation at high
crosstalk levels / high power penalties, with differences of 1 to 1.5 dB around 8-10 dB of power penalty. Whilst several possible sources of increasing experimental error have been identified, the nature of BER measurements can be investigated experimentally. In order to do this, a simple electrical crosstalk experiment was undertaken in order to assess the ability of experimental results undertaken in a purely electrical domain to match theoretical predictions. Although the experimental set-up was simplified as much as possible, experimental results provide further insight into potential sources of error in BER measurements and incorrect power penalty readings.

This section will describe the experimental set-up and results, and will provide a brief theoretical analysis of electrical crosstalk that is slightly different from optical crosstalk, predicting different power penalties.

4.7.1 Experimental Set-up and Results

The aim of the electrical crosstalk experiment is to remove optical effects from the experiment and simplify the electrical set-up as much as possible. The experimental set-up used is shown in Figure 4.18.

![Diagram of experimental set-up]

6 dB = electrical coupler  
ATT = variable attenuator  
Noise = white noise source  
BERT = BER test-set

*Fig. 4.18 Experimental set-up used for the measurement of electrical crosstalk.*

Figure 4.18 shows that two separate pattern generators are used to provide electrical data for a signal and crosstalk path. The data format was bipolar data i.e. of normalised amplitude ±0.5, simulating a downconverted BPSK signal, allowing a zero decision threshold to be used at the receiver. Variable electrical attenuators in the crosstalk path allow the crosstalk level to be adjusted, before being coupled with the signal using a power combiner. The combined signal level is then controlled via electrical attenuators, before being combined with an electrical noise source via a power combiner. The
electrical noise source is a 0-18 GHz white noise source and provides the reference receiver noise that determines the receiver sensitivity. The combined electrical signal, which comprises the signal data, the crosstalk data and electrical noise is then fed to a BERT for BER analysis.

The experimental procedure was identical to that for the measurement of optical crosstalk: BER curves are obtained as a function of electrical crosstalk level by varying the total signal plus crosstalk electrical power using the variable attenuators. An RF power meter was used to measure the total electrical power at the ‘receiver’ before electrical noise was added. A 700 MHz data rate was used so as to facilitate signal and crosstalk bit alignment using electrical delay lines, with a PRBS of $2^{23}-1$. Figure 4.19 shows the electrical eye diagram traces at various crosstalk levels showing the effect of electrical crosstalk on the received signal bits.

![Eye Diagrams](image)

**Fig. 4.19** Eye diagrams of the electrical signal with various levels of electrical crosstalk using different data generators.

As the data format is bipolar, the effect of the crosstalk signal is to increase or decrease the signal level depending on whether the crosstalk signal is a logical ‘1’ or ‘0’. The blurred nature of the traces is due to the fact that a different data generator is used for the crosstalk signal, so that clock synchronisation is not perfect. Hence, over the measurement period the crosstalk signal walks through the signal in the time-domain. Whilst this has an effect on the BER measurement, it provides a more realistic measurement as crosstalk signals usually originate from completely different sources.
The resultant BER curves for various levels of electrical crosstalk are shown in Figure 4.20.

Fig. 4.20 BER curves for electrical crosstalk for BPSK modulation using different data generators.

Figure 4.20 shows BER curves for electrical crosstalk levels ranging from –78 dB to –7 dB. BER curves are all parallel and show no sign of an error floor down to 10^9. The receiver sensitivity is determined by the level of electrical noise in the system and is not important for power penalty estimation since relative power levels are used. The BER curves allow electrical power penalties to be obtained as a function of electrical crosstalk, as shown in Figure 4.21.

Fig. 4.21 Experimental and analytical power penalties for electrical crosstalk for BPSK modulation. Experimental results were obtained using different data generators.
Experimental results show the same trend as for optical crosstalk, i.e. that the power penalty increases slowly at low crosstalk levels, before higher levels of crosstalk are reached and power penalties increase more rapidly. Analytical results are explained in Section 4.7.2. Close inspection of results compared to optical crosstalk shows that there is a difference in power penalties, even when taking into account the fact that optical power is proportional to electrical current, not electrical power. This is further detailed in the following Section.

The eye diagram traces and the difference in power penalties between theory and experiments indicated that the discrepancy is perhaps due to the lack of synchronisation between the signal and crosstalk data generator clocks. In order to investigate this, the signal generator was used to provide the crosstalk data by using the inverted output and delaying the bit sequence so as to decorrelate the signal and crosstalk data. Figure 4.22 shows the eye diagram waveform for synchronised signal and crosstalk data clocks, showing a clear eye.

![Eye diagrams](image)

(a) 10 dB crosstalk  (b) 6 dB crosstalk  (c) 3 dB crosstalk

*Fig. 4.22 Eye diagrams of the electrical signal with various levels of electrical crosstalk using the same data generator.*

The effect of a synchronised crosstalk channel on BER measurements is shown in Figure 4.23.
Fig. 4.23 Experimental BER curves for electrical crosstalk for BPSK modulation using the same data generator.

BER measurements shown in Figure 4.23 show that only two to three points were taken for each curve – this was done in order to make a full set of measurements rapidly so as to minimise equipment drift over time. Note that the improved sensitivity compared to Figure 4.20 is due to a calibration offset only. BER curves were extrapolated so as to measure the power penalty at $10^{-7}$, although the same results would apply at $10^{-9}$. Power penalties obtained from Figure 4.23 are shown in Figure 4.24.

Fig. 4.24 Experimental and analytical power penalties for electrical crosstalk for BPSK modulation. Experimental results were obtained using the same data generator.

Figure 4.24 shows two sets of experimental results. The experimental uncertainty in the measurement is about $\pm 0.2$ dB. However, very good agreement is now observed for power penalties up to 5 dB, beyond which the experimental results
deviate from theory, as observed previously for optical measurements. The effect of signal and crosstalk data synchronisation is to provide much better agreement between experimental results and theory. Note that several measurements were performed when two different data generators were used, and all were consistent with Figure 4.21. It remains unclear whether the reason for the increased power penalties is the lack of clock synchronisation, or whether the actual quality of the signal at the same power level was worse. This could have been investigated by using the crosstalk generator for the signal and vice-versa, but was not performed at the time.

4.7.2 Theoretical and Simulation Results

This section provides a brief description of the theoretical analysis of electrical crosstalk for BPSK modulation assuming phase-aligned carriers [23, 24]. The analysis is identical to that used for optical crosstalk as described in Chapter 3, but with appropriate expressions for electrical crosstalk.

If we consider an electrical signal power, $P_{\text{sig}}$, an electrical crosstalk power ratio $\varepsilon$, then the following expressions for the 4 possible signal-crosstalk bit combinations are obtained:

$$I_{00} = -\sqrt{P_{\text{sig}}} \left(1 + \sqrt{\varepsilon}\right),$$
$$I_{01} = -\sqrt{P_{\text{sig}}} \left(1 - \sqrt{\varepsilon}\right),$$
$$I_{10} = \sqrt{P_{\text{sig}}} \left(1 - \sqrt{\varepsilon}\right),$$
$$I_{11} = \sqrt{P_{\text{sig}}} \left(1 + \sqrt{\varepsilon}\right).$$

(4.1)

Equation 4.1 considers bipolar data, as would be the case for a downconverted BPSK signal with in-phase signal and crosstalk RF carriers. The decision threshold in this case is always zero, allowing the use of the error function for BER analysis (see Chapter 3). BER curves can be obtained by varying the signal power $P_{\text{sig}}$ and the crosstalk power ratio $\varepsilon$, from which electrical power penalties can be extracted. BER curves are plotted against total electrical power, $P_{\text{sig}}(1+\varepsilon)$. This is the source of discrepancy between optical and electrical power penalties, as total optical power corresponds to $I_{\text{sig}}(1+x)^{1/2}$, not $I_{\text{sig}}(1+\sqrt{\varepsilon})$. Figure 4.25 shows analytical results for electrical crosstalk using the analysis outlined above, together with simulation results.
Simulation results were obtained using *VPITransmissionMaker*, using electrical components only, in a similar experiment to that described for optical out-of-band crosstalk (Section 4.5). Perfect agreement between simulation and analytical results is observed.

### 4.7.3 Discussion of Results

The experimental investigation of electrical crosstalk provided excellent agreement with the analytical results for electrical crosstalk when signal and crosstalk data originated from the same clock source. Discrepancies were observed at high levels of crosstalk, with an increasing difference between experiment and crosstalk as power penalties exceeded 4 to 5 dB. Higher power penalties were obtained when a different data generator was used for the crosstalk signal, with an unsynchronised data clock. In this case, a larger difference was obtained between theory and analytical results. This can explain the experimental results for optical crosstalk, where a similar behaviour was observed. For both electrical and optical results the deviation between predicted power penalties and experimental power penalties increased at high power penalties. For out-of-band optical crosstalk a 1 dB discrepancy occurs at –3 dB optical crosstalk, while for electrical crosstalk a 2 dB difference occurs at –8 dB electrical crosstalk (-4 dB optical). A direct comparison is not possible, however, as explained in Section 4.7.2 as the analytical formulation is different. As the experimental discrepancies were observed for both optical and electrical crosstalk, they are not due to an optical effect but due to the
nature of electrical BER measurements. As the electrical crosstalk experiment was simplified, amplifier nonlinearities and mixer effects were removed, indicating that the BER measurements themselves could be the source of the errors. One additional potential source of errors occurs if the BERT does not sample bits exactly in the middle of the eye. This was confirmed experimentally for electrical crosstalk measurements, showing that as the decision threshold moves towards the eye edges the measured power penalties are increased, particularly at higher crosstalk levels.

4.8 CONCLUSIONS

This chapter has provided simulation and experimental results that confirm the analytical models presented in Chapter 3. Simulation results were undertaken using two commercial software packages (GOLD and VPITransmissionMaker), allowing both in-band and out-of-band optical crosstalk to be simulated, as well as electrical crosstalk. Experimental results were presented for both in-band and out-of-band crosstalk for BPSK modulation, and were in good agreement with analytical results up to high levels of crosstalk. The effect of electrical filtering on the reduction of optical power penalties for in-band crosstalk was also confirmed experimentally. An experimental investigation of electrical crosstalk was undertaken in order to evaluate the ability to match electrical crosstalk power penalty measurements with theory. A brief description of the analytical model for electrical crosstalk was presented, and experimental results indicate that the nature of experimental BER measurements in the electrical domain lead to a discrepancy between experimental and theoretical results at high crosstalk and power penalties. This is caused by the use of different data generators whose clocks are unsynchronised and is also believed to account for some of the errors in optical crosstalk measurements.

The reduction in optical power penalties for in-band crosstalk using BPSK modulation was confirmed. The importance of the RF carrier phase difference between signal and crosstalk was also verified. More importantly, the ability to minimise the impact of in-band optical crosstalk using electrical filtering, by ensuring the crosstalk signal carries data at a different RF frequency, was also measured experimentally.

Chapter 3 and Chapter 4 have focused on the theoretical and experimental investigation of optical crosstalk in fibre-radio networks. The following chapter
investigates the impact of in-band and out-of-band optical crosstalk on fibre-radio networks. Results are presented that allow component specifications for both types of crosstalk to be chosen so as to meet a specific system margin. A theoretical analysis of network capacity is given assuming typical in-band and out-of-band component crosstalk levels, showing the potential impact of crosstalk in limiting the total number of WDM channels transported in such networks. The impact of an RF frequency re-use plan is assessed.

4.9 REFERENCES


Chapter 4


5 Impact of Optical Crosstalk on WDM Fibre-Radio Network Capacity

5.1 INTRODUCTION

The previous chapters have provided a theoretical and experimental investigation of both in-band and out-of-band optical crosstalk in fibre-radio systems using ASK, BPSK [1] and QPSK modulation schemes. Several important observations were made. Firstly, the change of optical power penalty with the RF phase difference between signal and crosstalk RF carriers was demonstrated, to illustrate such phase dependence and to highlight the importance of the worst-case situation for network design purposes. Secondly, the importance of the modulation format was established, with BPSK modulation showing an increased tolerance to in-band crosstalk, followed by QPSK modulation and ASK modulation. Finally, it was shown that optical crosstalk in subcarrier multiplexed systems can be minimised due to the use of different RF channels frequencies. This is achieved through electrical filtering of the unwanted crosstalk RF frequencies at the receiver [1-4].

In this chapter, the result obtained in Chapter 3 is further extended to investigate the impact of optical crosstalk on the design of large-scale networks. Section 5.2 presents total optical power penalties for various levels of in-band and out-of-band crosstalk levels, allowing component crosstalk specifications to be determined for a given system margin. These results highlight the importance of using adjacent RF frequency bands for the in-band crosstalk signal, as well as the existence of a trade-off between in-band and out-of-band crosstalk levels. Section 5.3 considers the effect of multiple out-of-band crosstalk sources arising from a network incorporating WDM. The impact of multiple WDM channels on allowable component crosstalk specifications is also investigated. Section 5.4 extends the theoretical model used in Section 5.3 to include the possibility of using multiple RF frequencies/frequency bands in the RF domain [5-8], allowing electrical filtering to reduce the impact of optical crosstalk. The
impact of a frequency re-use plan on component crosstalk specifications is quantified. In Section 5.5, results from Sections 5.3 and 5.4 are used to quantify the total number of WDM channels that can be used in a WDM fibre-radio network for various typical levels of in-band and out-of-band component crosstalk. These results highlight the trade-off between in-band and out-of-band crosstalk specifications and the impact of using adjacent RF frequencies in a fibre-radio network. The important conclusions arising from results presented in the chapter will be discussed in Section 5.6.

5.2 SINGLE CROSSTALK CONSTRAINT

This section presents contour plots of total optical power penalties for various levels of in-band and out-of-band optical crosstalk levels, illustrating the dependence on modulation format and the effect of the crosstalk signal being at a different RF carrier frequency. These results provide the basic tool for network design by providing the direct link between system penalties and crosstalk specifications.

5.2.1 Power Penalty Contour Plots with Frequency Re-use

The power penalty curves presented in Chapter 3 for in-band and out-of-band crosstalk can be used to provide contour plots showing the total optical power penalty due to a combination of in-band and out-of-band crosstalk. The total optical power penalty, $PP_{\text{total}}$, is given by:

$$PP_{\text{total}} = PP_{\text{in-band}} + PP_{\text{out-of-band}} (dB)$$

(5.1)

where $PP_{\text{in-band}}$ and $PP_{\text{out-of-band}}$ are the power penalties in dB due to in-band and out-of-band crosstalk, respectively. Since both in-band and out-of-band crosstalk show increasing power penalties at increasing crosstalk levels, contour plots of total power penalty will show the range of possible in-band and out-of-band crosstalk combinations which lead, for example, to a fixed total power penalty. Figure 5.1 shows contour plots for BPSK modulation with RF frequency re-use, i.e. where the signal and crosstalk RF frequencies are the same.
Figure 5.1 shows the change in power penalty as the level of in-band or out-of-band crosstalk is varied. RF carriers are assumed to be phase-aligned as this results in the worst-case performance. A 1 dB system margin, for example, can be used to allow for the power penalty due to optical crosstalk. The 1 dB power penalty contour shows the range of possible in-band and out-of-band crosstalk component specifications. There is a fixed in-band crosstalk level for a 1 dB penalty (-17.3 dB) with no out-of-band crosstalk, and a fixed level of out-of-band crosstalk for the same penalty (-9 dB) with no in-band crosstalk. If the same margin is used and both types of crosstalk are present, then the maximum tolerable crosstalk limit will be reduced relative to the case where the margin is due to one type of crosstalk only. For example, -20 dB of in-band crosstalk and about –12.5 dB of out-of-band crosstalk gives a 1 dB total power penalty. The contour plots also show that contours become closer to one another at high power penalties, so that a small change in crosstalk level can result in a large change in power penalty.

Similar plots can be obtained for ASK and QPSK modulation using results from Chapter 3. Contour plots will reflect the fact that in-band crosstalk produces a higher power penalty for ASK and QPSK modulation, for a given component crosstalk level.
This is summarised in Figure 5.2 which shows 1 dB power penalty contours for ASK, BPSK and QPSK modulation, respectively.

![Contour plot of total power penalty for ASK, BPSK and QPSK modulation with frequency re-use.](image)

Figure 5.2 shows the impact of modulation format on the contour plots. For simplicity results for ASK modulation are taken for a PLL receiver. Chapter 3 provides further information on the impact of the receiver structure for ASK demodulation. Higher power penalties are obtained for envelope detection, and are also dependent on the decision threshold circuit (see Section 3.4). Out-of-band results are different for QPSK modulation since a 45° RF carrier phase difference produces the highest power penalty. In-band crosstalk results are quite different, confirming the importance of modulation format.

### 5.2.2 Power Penalty Contours with Adjacent Frequency Bands (In-band)

As highlighted in Chapters 3 and 4, one of the important results arising from the analysis of in-band crosstalk for subcarrier-modulated links is the fact that different RF carrier frequencies may be used in a network. This has the real potential of either reducing the optical power penalty for a given level of component crosstalk, or of
relaxing in-band crosstalk specifications if a given power penalty margin is used in the network [1].

The contour plots in Figure 5.3 illustrate this for BPSK modulation.

![Contour Plot](image)

*Fig. 5.3 Contour plot of total power penalty for BPSK modulation with adjacent frequency bands (in-band).*

Figure 5.3 shows that a higher level of in-band crosstalk can be tolerated if the in-band crosstalk signal carries data at a different RF frequency, allowing electrical filtering of the unwanted RF carrier. This should be compared to Figure 5.1 where the same RF frequency is used. The vertical parts of the curves for in-band crosstalk in Figure 5.3 have shifted to the right, showing that a higher level of in-band crosstalk can be tolerated for the same power penalty. This means that for an out-of-band crosstalk level of about −12.5 dB, an in-band crosstalk level of around −15 dB can be tolerated for a total power penalty of 1 dB, compared to −20 dB for frequency re-use. Alternatively, the same in-band crosstalk level of −20 dB can be considered, increasing the acceptable level out-of-band crosstalk to about −10.5 dB (up from −12.5 dB). This highlights the fact that the use of a different RF frequency results in a lower impact of in-band crosstalk, allowing a relaxation in the out-of-band crosstalk component specification (provided there is a power penalty due to out-of-band crosstalk).
As before, similar plots can be obtained for ASK and QPSK modulation showing the improvement in in-band crosstalk tolerance as an adjacent RF frequency is used for the in-band crosstalk signal. Figure 5.4 summarises this for ASK, BPSK and QPSK modulation showing the 1 dB power penalty contours.

![Contour plot of total power penalty for ASK, BPSK and QPSK modulation with adjacent frequency bands (in-band).](image)

**Fig. 5.4** Contour plot of total power penalty for ASK, BPSK and QPSK modulation with adjacent frequency bands (in-band).

Figure 5.4 shows the fact that in-band crosstalk curves are the same for BPSK and QPSK modulation when the RF crosstalk can be filtered electrically. Power penalties are higher for ASK modulation. Comparing Figures 5.4 and 5.2 allows the benefits of RF filtering to be assessed for the three modulation formats.

### 5.2.3 Power Penalty Contours with Adjacent Frequency Bands (In-band & Out-of-band)

Contour plots can also be obtained for in-band with systems employing frequency re-use and out-of-band crosstalk where signal RF frequencies are different. In the case of in-band crosstalk, this may arise from the use of frequency re-use in neighbouring cells. In the case of out-of-band crosstalk, it may arise from different frequency bands being used. However, the margin gained from using different RF frequencies for out-of-band crosstalk is small. This is because most of the power
penalty is due to the additional optical power present at the receiver [9, 10], with a smaller impact in the electrical domain compared to in-band crosstalk. If the tolerance to in-band crosstalk is to be increased or power penalties reduced, then the network design must ensure that the in-band crosstalk signal transports data at a different RF frequency to that of the signal. Hence this section presents contour plots showing the optimum situation where both in-band and out-of-band crosstalk channels use adjacent frequency bands. This would ensure that the crosstalk electrical signals could be filtered electrically at the receiver.

Figure 5.5 shows the combined effect of removing the crosstalk signal in the electrical domain by electrically filtering the unwanted crosstalk RF frequency, both for in-band and out-of-band crosstalk, for BPSK modulation.

![Contour plot of total power penalty for BPSK modulation with adjacent frequency bands (in-band and out-of-band).](image)

Figure 5.5 shows that the 1 dB power penalty limits occur at −13.4 dB and −5.9 dB for in-band and out-of-band crosstalk respectively (compared to −17.3 dB and −9.1 dB with frequency re-use). However, if both in-band crosstalk and out-of-band crosstalk are present at the same time, the component crosstalk levels are reduced. For example, with an in-band crosstalk of −15 dB, the out-of-band crosstalk requirement is reduced to about −12 dB, for a total power penalty of 1 dB.
Similar contour plots can be obtained for ASK and QPSK modulation, but are not presented for brevity. However, contour plots for ASK, BPSK and QPSK modulation are shown for a 1 dB total power penalty in Figure 5.6.

![Contour plot of total power penalty for ASK, BPSK and QPSK modulation with adjacent frequency bands (in-band and out-of-band).](image)

Figure 5.6 shows the effect of ensuring both in-band and out-of-band crosstalk channels carry data at a different RF frequency, which can be electrically filtered at the receiver. The effect of modulation format is apparent. Results can be compared to Figures 5.4 and 5.1 in order to observe the benefits of electrical filtering, both for in-band and out-of-band crosstalk. Note that out-of-band crosstalk results are identical for all three modulation formats when a different RF frequency is used, as the optical power penalty is due to optical power addition alone.

### 5.3 Analysis for Multiple Out-of-band Crosstalk Terms

Section 5.2 provided the network designer with an easy tool to specify component crosstalk specifications for a given total optical power penalty or to assess the optical power penalty for given levels of in-band and out-of-band optical crosstalk. The importance of both modulation format and RF frequency allocation can be observed visually from the contour plots provided.
This section considers the fact that a multiple WDM channels will be present in future WDM fibre-radio networks [7, 11-13], resulting in multiple out-of-band channels. As discussed in [14-19] for in-band optical crosstalk, the statistics involved when there are multiple crosstalk terms means that for a given level of total crosstalk, the resulting power penalty depends on the number of crosstalk terms. This effect is analysed in this section for out-of-band optical crosstalk, showing the increased tolerance to crosstalk as total crosstalk is spread over multiple WDM channels. The effect of multiple out-of-band crosstalk terms has been considered in [20, 21]. The analysis used below follows that of Buckman et al. [20]. However, only thermal noise is considered here with the assumption of very low relative-intensity noise (RIN) from laser sources available now. Hence RIN will not have a significant impact. Also note that power penalties in [20] exclude the penalty due to optical power addition.

5.3.1 Analytical Model

Chapter 3 presented analytical expressions allowing the BER for a single source of out-of-band crosstalk. In this section, the same analytical approach is taken for in-phase RF carriers where there are multiple crosstalk terms. The analysis produces results which are identical for baseband IM, ASK (with PLL detection and phase-tracking decision threshold, see Section 3.4.3) and BPSK modulation. The worst-case scenario of phase-aligned RF carriers is used in the following analysis as the analytical treatment can be simplified. There is also a finite probability that all crosstalk terms will have RF carriers that are in-phase, particularly in the downlink which only contains optical fibre (no wireless link) and uses stable local oscillators at the central office (see Section 3.6 for further details). It would be undesirable to have a degraded BER in this situation. Furthermore, the case of RF frequency re-use where different RF frequency bands are used is considered in Section 5.4, allowing the benefits of electrical filtering to be assessed. This is equivalent to orthogonal RF carriers for ASK/BPSK modulation with frequency re-use.

As discussed in Chapter 3, out-of-band optical crosstalk is much simpler to analyse than in-band crosstalk as optical carriers are at different wavelengths, so that the signal and crosstalk modulated RF carriers are recovered separately upon photodetection [22]. The detected current can be represented as:
where $I_{RF}$ is the current at the RF frequency, and $I_{\text{sig}}$ and $I_{\text{crosstalk}}$ are the signal and crosstalk RF contributions respectively. $P_{\text{sig}}$ and $P_{\text{crosstalk}}$ are the signal and crosstalk optical powers respectively. As before, the optical crosstalk ratio is $x$, so that:

$$P_{\text{crosstalk}} = xP_{\text{sig}}$$  

(5.3)

Assuming in-phase RF carriers, the downconverted signal, $I_{\text{down}}$, can be represented as:

$$I_{\text{down}} = \alpha + x\beta$$  

(5.4)

where $\alpha$ and $\beta$ are the signal and crosstalk data values, respectively. For BPSK modulation [23], $\alpha$ and $\beta$ can take values of $\pm 1$. As before, we consider a signal ‘1’ and crosstalk ‘0’ or ‘1’, and a signal ‘0’ with crosstalk ‘0’ or ‘1’. For $n$ crosstalk terms, if we consider $j$ crosstalk terms to be transmitting ‘1’s, we obtain the following expressions:

$$I_0 = -1 + jx - (n - j)x,$$

$$I_j = 1 + jx - (n - j)x.$$  

(5.5)

Equation 5.5 gives expressions for the detected signal when a signal data ‘0’ is transmitted ($I_0$) and when a signal data ‘1’ is transmitted ($I_j$). Since there are $n$ total crosstalk terms with $j$ of them transmitting data ‘1’s, there are $n-j$ transmitting ‘0’s. Using a decision threshold equal to zero allows the following expression for $Q$ to be obtained from Equation 5.5:

$$Q_0 = 1 - jx + (n - j)x = 1 + (n - 2j)x,$$

$$Q_j = 1 + jx - (n - j)x = 1 - (n - 2j)x.$$  

(5.6)

It can be seen from Equation 5.6 that when a signal ‘0’ is sent the total signal amplitude is increased provided $j < n/2$ (i.e. provided more crosstalk ‘0’s are transmitted than crosstalk ‘1’s). The opposite is true when a signal ‘1’ is transmitted (i.e. the signal level is increased provided $j > n/2$). Since the transmitted data is random, each crosstalk term can transmit a data ‘0’ or ‘1’, so that we must consider the whole range of possible bit combinations, i.e. $j$ ranging from 0 to $n$. The following BER expression is obtained for each value of $j$, weighted by the number of cases for which there are $j$ crosstalk ‘1’s transmitted [14, 20-22, 24]:

$$BER(j) = 0.5 \left\{ \frac{1}{2} \text{erfc}\left( \frac{Q_0}{\sqrt{2}\sigma} \right) + \frac{1}{2} \text{erfc}\left( \frac{Q_j}{\sqrt{2}\sigma} \right) \right\} \frac{n!}{j!(n-j)!}.$$  

(5.7)
where $\sigma$ is the standard deviation of Gaussian receiver noise. The average BER is obtained as follows:

$$BER = \frac{1}{2^n} \sum_{j=0}^{2^n} BER(j)$$

(5.8)

The resulting BER is obtained by considering all possible signal and crosstalk bit combinations and taking the average. While the BER for individual bits will vary, measuring the BER over a sufficiently large number of bits ($2^n$) will ensure that the average BER will be measured.

5.3.2 Analytical Results

Whilst Equations 5.4 to 5.6 do not show the dependence on the optical power $P_{\text{sig}}$, the expressions are all linearly proportional to it since the detected current is proportional to the optical power. This allows BERs as a function of optical power to be obtained for a given crosstalk level. By taking the difference between the total optical power at which a reference BER of $10^{-9}$ is obtained, optical power penalties for different crosstalk levels can be obtained. This corresponds to the additional received optical power required to achieve the original error rate in the presence of optical crosstalk.

The calculated optical power penalty with out-of-band component crosstalk is shown in Figure 5.7 as a function of the number of crosstalk terms.

![Figure 5.7](image_url)  
*Fig. 5.7 Optical power penalty vs. component out-of-band crosstalk with multiple crosstalk terms.*
Figure 5.7 shows that for as the number of crosstalk terms increases, the power penalty increases, as expected. Alternatively, these results provide the network designer with a means to assess the out-of-band crosstalk component specification if a certain system margin is to be met. The total number of WDM channels in the network is an important parameter, not only in terms of total channel capacity, but also in terms of the resulting power penalty. These results agree with those presented by Buckman et al. [20] provided the power penalty due to power addition is taken into account, with a slight discrepancy believed to be due to the addition of RIN in [20].

If we consider the total optical crosstalk rather than component crosstalk, then we obtain the plots shown in Figure 5.8.

![Graph showing optical power penalty vs. total out-of-band crosstalk with multiple crosstalk terms.](image)

*Fig. 5.8 Optical power penalty vs. total out-of-band crosstalk with multiple crosstalk terms.*

Figure 5.8 shows the expected effect of statistics as a given total crosstalk level is shared between multiple crosstalk channels. This is particularly noticeable at higher crosstalk levels. For 128 out-of-band crosstalk terms and a 0 dB total crosstalk level, which corresponds to a component out-of-band crosstalk level of –21 dB, the resulting power penalty is 3.7 dB. Since the penalty due to optical power addition is 3 dB, the remaining 0.7 dB is the impairment caused by the reduction of signal level in the electrical domain.

An alternative way to visualise the above results is to consider, once again, a typical system margin of 1 dB, which can be allocated in this case to out-of-band
crosstalk. The total crosstalk for a 1 dB power penalty as a function of out-of-band channel number can be obtained from the results and are shown in Figure 5.9:

![Figure 5.9 Total crosstalk for a 1 dB power penalty as a function of the number of crosstalk terms.](image)

Figure 5.9 clearly shows the improvement in crosstalk tolerance, with a 1 dB power penalty occurring at –9.1 dB for a single crosstalk term, and close to –6 dB for 128 crosstalk terms. There appears to be an asymptote at around –6 dB. A similar shift for a large number of crosstalk terms was identified in [23] for in-band crosstalk, as for an infinite number of terms half the channels are transmitting ‘1’ s and the other half are transmitting ‘0’ s. Since the out-of–band crosstalk arises from an optical component, it is useful to present the above information in terms of the actual component crosstalk, as shown in Figure 5.10.
Fig. 5.10 Component crosstalk for a 1 dB power penalty as a function of the number of crosstalk terms.

Figure 5.10 includes a curve which corresponds to the component crosstalk required to give a 1 dB optical power penalty due to optical power addition, i.e. solely due to the additional power in the crosstalk terms. The excess margin corresponds to the actual degradation in BER due to the effect of out-of-band crosstalk on the received electrical signal. This is seen to converge towards the power addition curve as the number of out-of-band terms increases, confirming the presence of an asymptote. Indeed, a 1 dB penalty due to power addition occurs for a total crosstalk level of $-5.9$ dB. A similar graph can be obtained for a different power penalty using the results in Figure 5.7, allowing the network designer to quantify the effect of multiple WDM channels in the network in terms of the optical power penalty due to out-of-band optical crosstalk.

5.3.3 Discussion of Results

Results presented in the previous section confirm that the optical power penalty due to out-of-band optical crosstalk depends on the number of crosstalk terms. For a given component out-of-band crosstalk level, the level of total crosstalk increases as the number of channels in the network increases. This increase must be compensated by a tighter component specification if a certain power penalty is to be maintained. Results indicate that as the number of WDM channels increases the power penalty approaches that due to optical power addition only. This leads to a 3 dB relaxation in total crosstalk tolerance for a 1 dB optical power penalty as the number of crosstalk terms exceeds 100 channels, compared to a single crosstalk channel.
5.4 Impact of RF Frequency Re-use Plan

Whilst Section 5.3 analysed the impact of multiple WDM channels on the out-of-band crosstalk component specification, it was assumed that all channels transported data at the same RF frequency and that all RF carriers were in-phase. This section considers the more realistic scenario where several different RF frequency bands are allowed in the network [8], so that WDM channels do not all carry data at the same RF frequency [5-7]. As stated previously, a frequency re-use plan in the wireless domain maximises network capacity by re-using limited bandwidth in non-adjacent radio cells [8]. In terms of optical crosstalk, the benefits this offers has been demonstrated in Chapters 3 and 4, as well as in previous sections in this chapter, since electrical filtering of unwanted RF frequency bands is possible at the receiver. This section quantifies the impact that this has on results given in Section 5.3 for multiple out-of-band crosstalk terms. Since using adjacent frequency bands is equivalent to using orthogonal RF carriers where the same RF frequency is used for all channels, this analysis effectively considers the case where RF carriers are not all aligned (as per Section 5.3).

5.4.1 Modified Analytical Model

The analytical model presented in Section 5.3 can be modified to consider the fact that multiple RF frequencies can be assigned to different WDM channels. In the following analysis, we will consider the effective number of out-of-band crosstalk terms that carry data at the same RF frequency as the signal. The effective number of crosstalk terms, denoted as \( n_{\text{eff}} \), depends on the frequency re-use factor. This is defined as the total number of different RF frequencies or frequency bands that are used in the wireless network. Hence a frequency re-use factor of 1 corresponds to a single frequency/frequency band, so that all WDM channels use the same RF frequency (or RF frequencies if multiple subcarriers are used). This corresponds to the condition analysed in Section 5.3. For a frequency re-use factor of \( F \), the following expression relates the total number of crosstalk terms, \( n \), to the effective number of crosstalk terms at the same RF frequency as the signal, \( n_{\text{eff}} \):

\[
 n_{\text{eff}} = \text{int} \left( \frac{n}{F} \right) \quad (5.9)
\]
where the ‘int’ operator takes the integer number only, ignoring the decimal places. Equation 5.9 shows that as the frequency re-use factor increases, the effective number of crosstalk channels that are at the same frequency as the signal decreases. All other crosstalk signals are filtered electrically at the receiver, and are removed. For example, if there are 64 out-of-band crosstalk channels and a frequency re-use factor of 8 is used in the network, then there are only 8 channels that carry data at the same frequency as the signal. Also, if there are less than 8 channels, then there are no channels at the same RF frequency as the signal.

The mathematical analysis is otherwise the same as that presented in Section 5.3, so that Equations 5.7 and 5.8 can be used with $n_{\text{eff}}$ replacing $n$:

$$
\text{BER}(j) = 0.5 \left\{ \frac{1}{2} \text{erfc} \left( \frac{Q_0}{\sqrt{2\sigma}} \right) + \frac{1}{2} \text{erfc} \left( \frac{Q_1}{\sqrt{2\sigma}} \right) \right\} \frac{n_{\text{eff}}^j}{j! (n_{\text{eff}} - j)!} \tag{5.10}
$$

and

$$
\text{BER} = \frac{1}{2^{n_{\text{eff}}}} \sum_{j=0}^{n_{\text{eff}}} \text{BER}(j) \tag{5.11}
$$

The important fact that can be overlooked is that the total optical power incident on the receiver is unaffected, and is given by:

$$
P_{\text{total}} = P_{\text{sig}} (1 + nx) \tag{5.12}
$$

where $P_{\text{total}}$ is the total received optical power, $P_{\text{sig}}$ is the signal optical power, $n$ is the total number of out-of-band channels, and $x$ is the linear crosstalk ratio of each crosstalk term (or component crosstalk ratio). Equation 5.12 represents the effect of optical power addition due to the presence of the crosstalk terms. Since BER curves are plotted against total received optical power, the resulting power penalty comprises the effect of optical power addition and the actual degradation in signal quality in the electrical domain. The fact that WDM channels can transport data at different RF frequencies does not modify the total received optical power, since only the electrical signal is affected by electrical filtering.

### 5.4.2 Analytical Results

The power penalty for a given component crosstalk level for different numbers of out-of-band crosstalk terms is shown in Figure 5.11 for a frequency re-use factor of 2, i.e. when half the out-of-band channels are filtered electrically.
Impact of Optical Crosstalk on WDM Fibre-Radio Network Capacity

Fig. 5.11 Optical power penalty vs. component out-of-band crosstalk with multiple crosstalk terms for a frequency re-use factor of 2 i.e. half of the channels are filtered electrically.

Figure 5.11 illustrates clearly the impact of electrical filtering when there is a single crosstalk term. The optical power penalty at 0 dB crosstalk is solely due to optical power addition and is therefore equal to 3 dB. Compared to the case where there is frequency re-use in all WDM channels, Figure 5.11 shows that the optical power penalty is reduced when electrical filtering can remove some of the crosstalk channels in the electrical domain. If the same data is plotted against total crosstalk, the graphs in Figure 5.12 is obtained.

Fig. 5.12 Optical power penalty vs. total out-of-band crosstalk with multiple crosstalk terms for a frequency re-use factor of 2.
Chapter 5

Comparing Figure 5.12 with Figure 5.8 shows the improvement due to electrical filtering, particularly for a low number of crosstalk terms. The single crosstalk curve corresponds to optical power addition since the crosstalk RF frequency is filtered. Power penalties are also reduced for a larger number of crosstalk terms. However, the difference is less noticeable as the number of crosstalk terms increases.

While a frequency re-use factor of 2 is rather unlikely in a network, a more realistic re-use factor is 8 [5], meaning that only 1 in 8 crosstalk terms are at the same RF frequency as the signal. Figure 5.13 shows the reduction in power penalty for a given component crosstalk level.

![Optical Power Penalty vs. Component Out-of-Band Crosstalk](image)

*Fig. 5.13 Optical power penalty vs. component out-of-band crosstalk with multiple crosstalk terms for a frequency re-use factor of 8 i.e. 7/8 of the channels are filtered electrically.*

Figure 5.13 shows that power penalties for 1 to 7 crosstalk terms correspond to optical power addition only [9, 10], i.e.

\[
PP(\text{power addition}) = 10 \log_{10}(1 + nx) \quad (\text{dB}) \quad (5.13)
\]

where \(PP(\text{optical power})\) is the power penalty due to optical power addition caused by \(n\) crosstalk terms each with a linear crosstalk ratio of \(x\). This condition is satisfied provided

\[
n_{\text{eff}} < F \quad (5.14)
\]

i.e. provided all out-of-band channels are using different RF frequencies. This is further highlighted when optical power penalties are plotted against total optical crosstalk, as shown in Figure 5.14.
Figure 5.14 shows that the range of power penalties at 0 dB total crosstalk is much narrower than for lower frequency re-use factors, with a difference of less than 0.5 dB. Note also that a single curve corresponding to optical power addition is shown for $n$ ranging from 1 to 7, a condition that is stated in Equations 5.13 and 5.14. Hence 8 crosstalk terms are required before a single electrical crosstalk signal is present ($n_{eff}=1$) at the same RF frequency as the signal, causing an additional power penalty (less than 0.5 dB extra penalty). As the same total crosstalk is spread over more crosstalk terms (really $n_{eff}$) the impact of binary data statistics starts to lower the additional penalty occurring in the electrical domain. This is clearly observed for 128 crosstalk channels, with the power penalty curve being extremely close to that for optical power addition.

Results presented in this section can be used to characterise the out-of-band crosstalk requirement for a fixed power penalty as a function of the number of crosstalk terms, as presented in Section 5.3. Firstly, the total crosstalk at which a 1 dB power penalty occurs for out-of-band crosstalk can be plotted as a function of $n$, the total number of crosstalk channels, as shown in Figure 5.15.
Figure 5.15 shows the total crosstalk at which a 1 dB power penalty occurs for different frequency re-use factors. As different frequency bands are introduced in the network, allowing electrical filtering to remove some of the out-of-band crosstalk signals, the tolerance to crosstalk is increased. In particular, for $n<F$ the power penalty is due to optical power addition alone (no actual degradation in BER), so that $-5.9$ dB of total crosstalk is acceptable (for a 1 dB power penalty). As soon as $n\geq F$, which corresponds to $n_{eff}>0$, there is a jump in the curve. This is caused by the introduction of a crosstalk signal in the electrical domain, creating an excess penalty in addition to that due to additional optical power at the receiver. These jumps occur at $n=2, 4$ and 8 for frequency re-use factors of 2, 4 and 8 respectively. The impact of binary data statistics as the crosstalk is distributed over an increasing number is also seen, so that curves converge in all cases to the crosstalk level at which power addition alone gives a 1 dB power penalty, i.e. at $-5.9$ dB of total crosstalk. It is also apparent that as the frequency re-use factor $F$ increases, the jump in total crosstalk tolerance is reduced. This means that electrical filtering reduces the excess penalty due to the presence of a crosstalk signal in the electrical domain, so that the constraint tends towards that due to power addition alone. This limit is approached in all cases at a large number of crosstalk channels. For greater frequency re-use factors (>8), the optical power penalty is very close to that due to optical power addition, independent of the number of WDM channels.
This behaviour can be clearly visualised in Figure 5.16 which shows the component crosstalk at which a 1 dB power penalty occurs as a function of the number of crosstalk terms and frequency re-use factor.

Figure 5.16 shows the effect of various frequency re-use factors. Full frequency re-use refers to the case when \( F=1 \), i.e. when the same frequency band is used throughout the network. This is clearly the worst-case situation as all the crosstalk signals fall within the signal RF frequency band, so that none can be removed via electrical filtering. Curves are also shown for frequency re-use factors of 2, 4 and 8. In each case, the component crosstalk, which gives a 1 dB power penalty, reduces as the number of crosstalk terms increases, tending towards a penalty defined by optical power addition. The additional margin in component crosstalk clearly improves (reduces) as different RF frequencies are used for different WDM channels. While the improvement is greatest at a low number of crosstalk terms (i.e. WDM channels in the network), the improvement becomes much less significant when there are more than 8 channels in the optical network. This is due to the statistics involved in signal-crosstalk bit combinations which reduces the degradation due to the presence of crosstalk signals in the electrical domain. For example, there is a 2 dB improvement in component crosstalk specification with 8 crosstalk terms if a frequency re-use factor of 8 is employed. However, the difference between full frequency re-use and a frequency re-
use factor of 8 is reduced to less than 0.5 dB in component specification for 32 out-of-band channels.

5.4.3 Discussion of Results

While Section 5.3 has discussed the effect of multiple out-of-band crosstalk terms, this section has assessed the impact of a frequency re-use plan on those results. The importance of electrical filtering has been highlighted for in-band crosstalk and also out-of-band crosstalk for a single crosstalk channel (Chapter 3), allowing either a relaxation of optical component crosstalk levels for a given power penalty or a reduction in power penalty for a given component crosstalk level [1]. This section has demonstrated the importance of electrical filtering for a low number of out-of-band channels, allowing a relaxation in the out-of-band crosstalk specification. However, the impact is much less significant as a large number of WDM channels in the network contribute to a total crosstalk level. In this situation the statistics described in Section 5.3 dominate, meaning almost all the optical power penalty is due to the additional optical power present at the receiver, not due to an actual degradation in signal in the electrical domain.

Results show that for a 1 dB power penalty a total crosstalk level of $-9.1$ dB is acceptable for a single crosstalk channel with full frequency re-use. This is relaxed to $-5.9$ dB of out-of-band crosstalk if all WDM channels are electrically filtered, i.e. when the number of crosstalk terms is less than the frequency re-use factor. In all cases the worst-case occurs when the number of crosstalk terms is equal to the frequency re-use factor. However, the worst-case crosstalk level for a 1 dB penalty increases as the frequency re-use factor increases, indicating a greater tolerance to crosstalk. In all cases the total crosstalk level of $-5.9$ dB, indicating a power penalty due to optical power addition only, is approached asymptotically for a large number of crosstalk terms. For example, for full frequency re-use ($F=1$), a 1 dB excess component crosstalk margin is required for 10 crosstalk channels, compared to around 0.5 dB for 40 crosstalk channels. Results therefore indicate that electrical filtering has little impact on the penalty due to out-of-band crosstalk if more than 40 WDM channels are present in the fibre-radio network. However, for frequency re-use factors $>8$, optical power penalties are very
close to being solely due to optical power addition, even for low WDM channel numbers, given the impact of electrical filtering at low channel numbers.

### 5.5 Impact of Optical Crosstalk on Network Dimensioning

The previous sections in this chapter have provided an insight into the constraints imposed by optical crosstalk, in particular those due to multiple out-of-band channels. The effect of using several RF frequencies or frequency bands in the network can improve the tolerance to out-of-band crosstalk arising from a low number of WDM channels. However, for larger networks (around 40 channels or more), the statistics of crosstalk bit combinations lead to a power penalty which tends towards that due to optical power addition only. For in-band optical crosstalk, only one crosstalk channel is usually expected leading to a situation whereby the use of different RF frequencies for the crosstalk channel can significantly improve the crosstalk tolerance.

This section uses the results presented in the previous sections to quantify the total number of WDM channels that can be used in a fibre-radio WDM network for a total system margin of 1 dB, i.e. assuming total crosstalk produces a 1 dB optical power penalty. Typical levels of in-band and out-of-band optical crosstalk are considered, showing the potential trade-off between both types of crosstalk. The effect of using multiple RF frequencies in the network is highlighted in terms of the potential gain in network dimension or capacity (total number of WDM channels).

Similar studies have been undertaken which quantify the impact of optical crosstalk due to cascaded OADMs or optical cross-connects [24-31]. However, these studies typically consider the impact of in-band crosstalk, either incoherent or coherent, and can also include effects such as ASE noise due to optical amplifiers, nonlinearities etc. A more fundamental analysis of the impact of multiple in-band crosstalk channels can be found in [14-19]. The following section quantifies the effect of multiple out-of-band crosstalk channels only.

#### 5.5.1 System Penalties for Typical In-band & Out-of-band Crosstalk Levels

If a WDM fibre-radio network is to be designed one of the important considerations is the system margin allocations to different kinds of impairments or
component drifts. The network is designed so that certain design criteria are met for all channels in the network. Additional optical power is budgeted in the link so as to ensure that degradations in the link can be compensated by increasing the received optical power, thereby increasing the received electrical signal relative to receiver noise. A 1 dB margin is assumed in the following sections for degradations due to optical crosstalk. Hence the combined penalties due to both in-band and out-of-band optical crosstalk are equal to 1 dB. Since optical crosstalk arises from the use of imperfect optical components, a range of optical crosstalk levels are considered. Table 5.1 shows optical power penalties due to in-band optical crosstalk for BPSK modulation, together with the corresponding margin allocated for out-of-band crosstalk, where:

\[ PP(\text{in-band}) + PP(\text{out-of-band}) = 1 \text{ dB} \]  

(5.15)

\(PP(\text{in-band})\) and \(PP(\text{out-of-band})\) are power penalties due to in-band and out-of-band crosstalk, respectively.

**Table 5.1 Power penalties due to in-band and out-of-band crosstalk for BPSK modulation, with and without frequency re-use, for various in-band crosstalk levels.**

<table>
<thead>
<tr>
<th>In-band Crosstalk (dB)</th>
<th>Same RF frequency</th>
<th>Different RF frequency</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>PP(in-band) (dB)</td>
<td>PP(out-of-band) (dB)</td>
</tr>
<tr>
<td>-10</td>
<td>3.27</td>
<td>1.79</td>
</tr>
<tr>
<td>-15</td>
<td>1.46</td>
<td>0.76</td>
</tr>
<tr>
<td>-20</td>
<td>0.63</td>
<td>0.37</td>
</tr>
<tr>
<td>-25</td>
<td>0.25</td>
<td>0.75</td>
</tr>
<tr>
<td>-30</td>
<td>0.09</td>
<td>0.91</td>
</tr>
<tr>
<td>-35</td>
<td>0.03</td>
<td>0.97</td>
</tr>
</tbody>
</table>

Table 5.1 shows that an in-band crosstalk level of \(-15\) dB exceeds the 1 dB power penalty margin if the crosstalk channel carries data at the same RF frequency as the signal. A \(-10\) dB in-band crosstalk level exceeds the margin even if the crosstalk RF frequency is at a different RF frequency and is filtered. A shaded box for the out-of-band crosstalk penalty margin denotes that no margin can be allocated as the total power
penalty margin is exceeded by in-band crosstalk alone. As in-band crosstalk levels are reduced, so are the power penalties due to in-band crosstalk, while corresponding out-of-band crosstalk penalty margins increase, maintaining a 1 dB total optical power penalty. The general observation that as in-band crosstalk levels are reduced the impact of RF filtering reduces can also be seen, so that −35 dB of in-band crosstalk in both cases produces an insignificant power penalty (0.02 and 0.03 dB respectively). This condition means that there is effectively no in-band crosstalk, with all the 1 dB margin allocated to out-of-band optical crosstalk.

5.5.2 WDM Channel Capacity

The power penalty margins for out-of-band crosstalk given in Table 5.1 can be used in conjunction with results obtained using the analytical model explained in Section 5.3 to establish the maximum number of out-of-band crosstalk terms that can be present in a fibre-radio network. This maximum number will depend both on the in-band and out-of-band component crosstalk levels, on the modulation format (BPSK is considered here) and on the RF frequency allocation. For example, Figure 5.17 shows the optical power penalty against out-of-band channel number for various levels of out-of-band component crosstalk levels.

![Graph showing optical power penalty against number of out-of-band channels](image)

*Fig. 5.17 Power penalties due to multiple out-of-band channels for different levels of component out-of-band crosstalk.*

Figure 5.17 shows that for a given component crosstalk level the optical power penalty increases as the number of out-of-band channels increases. The curves show that the
gradient of the curves depends on the component out-of-band crosstalk level, with high levels of crosstalk clearly limiting network capacity for a given penalty. Using these results in conjunction with out-of-band crosstalk margins in Table 5.1 allows the network capacity to be determined for a 1 dB total optical power penalty. Table 5.2 shows the maximum network capacity (out-of-band channels+1) for various levels of in-band and out-of-band component crosstalk levels.

Table 5.2 Total WDM network channel capacity for a 1 dB total optical power penalty for various levels of component in-band and out-of-band crosstalk levels for BPSK modulation with frequency re-use (F=1).

<table>
<thead>
<tr>
<th>In-band Crosstalk (dB)</th>
<th>Out-of-band Component Crosstalk (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>-15</td>
</tr>
<tr>
<td>-15</td>
<td></td>
</tr>
<tr>
<td>-20</td>
<td>2</td>
</tr>
<tr>
<td>-25</td>
<td>4</td>
</tr>
<tr>
<td>-30</td>
<td>5</td>
</tr>
<tr>
<td>-35</td>
<td>5</td>
</tr>
</tbody>
</table>

Table 5.2 shows that a 1 dB power penalty is exceeded for an in-band crosstalk of –15 dB. Network capacity increases as in-band and / or out-of-band component crosstalk levels are reduced. A capacity of more than 200 channels is guaranteed for in-band and out-of-band crosstalk levels of –20 dB and –35 dB respectively. The out-of-band crosstalk requirement can be relaxed to –30 dB provided the in-band crosstalk specification is tightened to –30 dB. Results show that tightening out-of-band component specifications provides a greater increase in network capacity than keeping the same out-of-band crosstalk level and tightening in-band crosstalk levels by the same amount. For example, at –25 dB of out-of-band crosstalk, network capacity is increased from 56 to 69 WDM channels as the in-band crosstalk level is reduced from –25 to –30 dB. If, on the other hand, the in-band crosstalk level is kept at –25 dB, network capacity can be increased from 56 to 184 by decreasing the out-of-band crosstalk level from –25 dB to –30 dB.
While Table 5.2 shows the network capacity for a 1 dB total power penalty when the same RF frequency or frequency band is used for all optical channels. This is highly unlikely, since radio cells typically use different RF frequencies in adjacent cells in order to minimise co-channel interference. Frequencies are re-used in non-adjacent cells in order to re-use the limited RF spectrum. The benefit this offers in the optical domain is particularly apparent for in-band crosstalk and is shown in Table 5.3, which shows the network capacity when the in-band crosstalk channel carries data at a different RF frequency, allowing the power penalty to be reduced. This increases the margin allocated to out-of-band crosstalk and hence increases the total network capacity.

<table>
<thead>
<tr>
<th>In-band Crosstalk (dB)</th>
<th>Out-of-band Component Crosstalk (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>-15</td>
</tr>
<tr>
<td>-15</td>
<td>2</td>
</tr>
<tr>
<td>-20</td>
<td>4</td>
</tr>
<tr>
<td>-25</td>
<td>5</td>
</tr>
<tr>
<td>-30</td>
<td>5</td>
</tr>
<tr>
<td>-35</td>
<td>5</td>
</tr>
</tbody>
</table>

Table 5.3 shows that more than 200 channels can be used in the network for an in-band crosstalk level of $-15$ dB if the out-of-band crosstalk level is $-35$ dB. Alternatively, in-band and out-of-band crosstalk levels of $-25$ dB and $-30$ dB, respectively, can be used. The improvement in network capacity provided by minimising the impact of in-band crosstalk is shown in Table 5.4, which shows the ratio of network capacities with same and different RF frequencies for the in-band crosstalk channel.
Table 5.4 Factor by which network capacity is increased by ensuring the in-band crosstalk channel carries data at a different RF frequency, for a total optical power penalty of 1 dB.

<table>
<thead>
<tr>
<th>In-band Crosstalk (dB)</th>
<th>Out-of-band Component Crosstalk (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>-15</td>
</tr>
<tr>
<td>-15</td>
<td>∞</td>
</tr>
<tr>
<td>-20</td>
<td>2.0</td>
</tr>
<tr>
<td>-25</td>
<td>1.3</td>
</tr>
<tr>
<td>-30</td>
<td>1.0</td>
</tr>
<tr>
<td>-35</td>
<td>1.0</td>
</tr>
</tbody>
</table>

Table 5.4 shows an infinite factor for −15 dB of in-band crosstalk since operation is only possible when the in-band crosstalk channel uses a different RF frequency. Blanks indicate that the factor is unavailable as more than 200 channels can be used in the network in both cases. However, results show a clear trend. The factor by which capacity can be increased is almost constant for a given level of in-band crosstalk. More significantly, the factor varies from two to one as the in-band crosstalk level is decreased from −20 dB to −30 dB. Hence the network capacity can be doubled by ensuring the in-band crosstalk channel carries data at a different RF frequency at relatively high levels of in-band crosstalk. However, no benefit occurs for an in-band crosstalk level of −30 dB. This can easily be understood by examining Table 5.1, which shows that at −20 dB of in-band crosstalk the power penalties are 0.63 dB and 0.32 dB, respectively, for the same/different in-band RF frequencies. At −30 dB of in-band crosstalk, these values are reduced to 0.09 dB and 0.05 dB, so that almost all of the 1 dB power penalty margin is allocated to out-of-band crosstalk. Hence a factor of greater than two improvement is possible at in-band crosstalk levels greater than −20 dB, although in this case the network capacity when the same RF frequency is used will be very low. This means that at high in-band crosstalk levels the in-band crosstalk channel must use a different RF frequency in order to allow a reasonable network size, whereas this is irrelevant for low in-band crosstalk, in which the total power penalty is dominated by out-of-band crosstalk.
The observations made in the previous paragraph are illustrated in Figure 5.18, which shows the total number of out-of-band channels (total capacity-1) for a 1 dB total power penalty for various levels of in-band and out-of-band crosstalk.

**Fig. 5.18** Total number of out-of-band channels as a function of component out-of-band crosstalk for various levels of in-band crosstalk, with the in-band crosstalk signal at the same and different RF frequency as the signal.

Figure 5.18 can be plotted versus in-band component crosstalk, showing the effect of reducing in-band crosstalk at various levels of out-of-band crosstalk, as shown in Figure 5.19.

**Fig. 5.19** Total number of out-of-band channels as a function of component in-band crosstalk for various levels of out-of-band crosstalk, with the in-band crosstalk signal at the same and different RF frequency as the signal.
Figure 5.19 confirms that improving in-band crosstalk improves the total network capacity when out-of-band crosstalk is low, since in this situation most of the 1 dB total power penalty is due to in-band crosstalk. This also explains why ensuring the in-band crosstalk channel uses a different RF frequency is more effective at low out-of-band crosstalk. The additional capacity, however, reduces towards zero as the in-band crosstalk level is reduced, since the power penalty due to in-band crosstalk becomes less and less significant.

5.5.3 Impact of Frequency Re-use on WDM Channel Capacity

Section 5.4 provided analytical results of out-of-band power penalties for various frequency re-use factors. Results indicated that using different RF frequencies was effective in reducing power penalties when the number of WDM channels was low. In particular, the power penalty corresponded to optical power addition provided the number of crosstalk channels was less than the frequency re-use factor $F$. The same approach as that described above can be used to determine the total WDM network capacity for a 1 dB total optical power penalty. Table 5.5 shows the network capacity for in-band crosstalk with the same RF frequency as the signal, for a frequency re-use factor of 8.

Table 5.5 Total WDM network channel capacity for a 1 dB total optical power penalty for various levels of component in-band and out-of-band crosstalk levels for BPSK modulation with a frequency re-use factor $F = 8$.

<table>
<thead>
<tr>
<th>In-band Crosstalk (dB)</th>
<th>Out-of-band Component Crosstalk (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>-15</td>
</tr>
<tr>
<td>-15</td>
<td></td>
</tr>
<tr>
<td>-20</td>
<td>3</td>
</tr>
<tr>
<td>-25</td>
<td>7</td>
</tr>
<tr>
<td>-30</td>
<td>8</td>
</tr>
<tr>
<td>-35</td>
<td>8</td>
</tr>
</tbody>
</table>
The trend in Table 5.5 is identical to that of Table 5.2 for $F=1$. An increase in capacity of between 1 to 5 channels is observed.

The effect of ensuring the in-band crosstalk signal carries data at the same RF frequency is shown for completeness in Table 5.6, showing the same gain in capacity, particularly at high in-band crosstalk levels.

*Table 5.6 Total WDM network channel capacity for a 1 dB total optical power penalty for various levels of component in-band and out-of-band crosstalk levels for BPSK modulation with in-band crosstalk at a different RF frequency, for a frequency re-use factor $F=8$."

<table>
<thead>
<tr>
<th>In-band Crosstalk (dB)</th>
<th>Out-of-band Component Crosstalk (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>-15</td>
</tr>
<tr>
<td>-15</td>
<td>2</td>
</tr>
<tr>
<td>-20</td>
<td>6</td>
</tr>
<tr>
<td>-25</td>
<td>8</td>
</tr>
<tr>
<td>-30</td>
<td>8</td>
</tr>
<tr>
<td>-35</td>
<td>8</td>
</tr>
</tbody>
</table>

Again, a comparison of Tables 5.6 and 5.3 shows that a frequency re-use factor of 8 provides a gain in capacity ranging from 0 to 6 channels. The ratio of channel capacity with same and adjacent RF frequency bands for the in-band crosstalk signal is nearly identical to that for $F=1$ and is not shown for brevity. The conclusions stated in Section 5.4 relating to the benefits of using multiple RF frequencies/frequency bands can be summarised in Table 5.7 which shows the ratio of channel capacity for $F=8$ to the channel capacity for $F=1$. 


Table 5.7 Section 5.5.4 Discussion of Results

Table 5.7 Factor by which network capacity is increased by using a frequency re-use factor = 8 compared to F=1, for a total optical power penalty of 1 dB.

<table>
<thead>
<tr>
<th>In-band Crosstalk (dB)</th>
<th>Out-of-band Component Crosstalk (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>-15</td>
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<tr>
<td>-15</td>
<td></td>
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<tr>
<td>-20</td>
<td>1.5</td>
</tr>
<tr>
<td>-25</td>
<td>1.3</td>
</tr>
<tr>
<td>-30</td>
<td>1.6</td>
</tr>
<tr>
<td>-35</td>
<td>1.6</td>
</tr>
</tbody>
</table>

Table 5.7 re-iterates the fact that channel capacity is only increased when the channel number is below or close to the frequency re-use factor, which is equal to 8 in this case. RF filtering of out-of-band crosstalk provides no benefit when there are a large number of WDM channels in the network.

5.5.4 Discussion of Results

The capacity of a WDM fibre-radio network based on typical levels of in-band and out-of-band crosstalk results has been determined. More than 200 WDM channels can be accommodated in a fibre-radio network if in-band and out-of-band crosstalk levels are –20 dB or better. The importance of RF frequency allocation for the in-band crosstalk channel is particularly significant at high in-band crosstalk levels where RF filtering of the in-band crosstalk signal can significantly reduce the in-band crosstalk penalty. Hence using a different RF frequency for the in-band crosstalk channel increases the penalty allocated to out-of-band crosstalk and increases network capacity. This allows a doubling of network capacity at large in-band crosstalk, reducing to no gain in capacity at low crosstalk, i.e. where out-of-band crosstalk dominates and is allocated most of the 1 dB power penalty margin. Ensuring different RF frequencies are used for between signal and crosstalk channels at the same wavelength is critical in network design as it allows either component crosstalk specifications to be relaxed or
the number of WDM channels in the network to be increased for a given total optical power penalty.

The ability to select different levels of in-band and out-of-band crosstalk component levels has also been mentioned. Improving the out-of-band crosstalk specification provides a greater increase in network capacity since more channels contribute to the same out-of-band penalty margin, improving capacity due to the effect of statistics as the number of out-of-band channels increases. Finally, the use of multiple RF frequency bands has been quantified, showing that network capacity is only improved for a small WDM network (of the order of the frequency re-use factor or less). No improvement is seen for large WDM channel numbers where the out-of-band crosstalk is mainly due to optical power addition (additional optical power at the receiver).

5.6 CONCLUSIONS

This chapter shows how the impact of optical crosstalk in large WDM fibre-radio networks can be assessed quantitatively. Results in Section 5.2 provide an initial estimate of component crosstalk specifications for both in-band and out-of-band crosstalk for a given power penalty margin, or the means to quantify optical power penalties for given levels of in-band and out-of-band component crosstalk. The importance of allocating a different RF frequency to the in-band crosstalk signal was highlighted. Section 5.3 provided an analysis of the power penalty due to multiple out-of-band crosstalk terms arising in a WDM network. Results indicate that the power penalty tends towards optical power addition as the number of WDM channels present in the network increases, allowing more crosstalk to be tolerated. The impact of a frequency re-use plan was analysed in Section 5.4 and was found to be important for low channel numbers (relative to the frequency re-use factor), but insignificant for large WDM networks. Using a frequency re-use factor >8 ensures out-of-band penalties are due to optical power addition alone, independent of the number of WDM channels. Section 5.5 quantified the capacity of a fibre-radio WDM network for a 1 dB total optical power penalty, confirming the importance of RF filtering of the in-band crosstalk channel. Using different RF frequencies for different WDM channels is only
useful in increasing network capacity for low channel numbers, as mentioned previously.

This chapter provides the network designer with a comprehensive set of results, allowing a WDM fibre-radio network to be designed for maximum capacity for a given optical crosstalk power penalty margin. The following chapter will review typical WDM network architectures and assess their impact in terms of optical crosstalk. More complex architectures will also be presented, showing how in-band and out-of-band crosstalk levels can vary. Both unidirectional and bidirectional networks will be discussed, highlighting similarities and differences. More importantly, a set of network design rules will be given based on the results presented in Chapters 3 to 5. These will allow an optimum network design, in which the impact of optical crosstalk is minimised, allowing network capacity to be maximised. The importance of signal and crosstalk optical powers will also be discussed, since this affects the effective level of optical crosstalk and can significantly affect network size or performance. Thus Chapter 6 will offer the network designer a choice of architectures and guidelines for network design taking into account optical crosstalk.

5.7 References


6

WDM Fibre-Radio Networks

6.1 INTRODUCTION

As explained in Chapter 1, WDM fibre-radio networks have the potential to provide high-bandwidth wireless access to a large number of users, using the high capacity offered by WDM optical networks [1-4]. In a WDM network, a single or multiple wavelengths can be allocated to each Base Station (BS) and this requires the use of optical components for wavelength routing, which can introduce optical crosstalk between WDM channels in the network. The analysis of in-band and out-of-band optical crosstalk in fibre-radio networks using subcarrier multiplexing has been presented in Chapter 3, with experimental verification and simulation validation in Chapter 4. Chapter 5 demonstrated how the analysis could be used to obtain requirements on component crosstalk levels to achieve a specific power penalty, taking into account the presence of multiple WDM channels. This study highlighted the importance of RF frequency allocation for situations where in-band crosstalk is present, allowing the component crosstalk requirement to be relaxed by ensuring a different RF frequency allocation to interfering optical in-band channels. In the case of out-of-band crosstalk with a low number of wavelengths, it was found that wireless frequency re-use planning could potentially minimise the effects of crosstalk. For a large number of out-of-band crosstalk signals, the resulting power penalty was found to be influenced by the additional power present at the receiver only (crosstalk power addition of large number of signals).

This chapter considers specific fibre-radio network architectures and shows how the results from the studies presented in the previous chapters can be incorporated into the design fibre-radio networks, allowing the comparison of network topologies from the perspective of optical crosstalk and component specifications. We will show the main advantages and disadvantages of different network topologies and provide the
network designer with specific techniques and recommendations that can be used when designing a fibre-radio network. In order to provide a realistic assessment, network design examples will be given for a high-capacity fibre-radio network operating at millimetre-wave frequencies [2]. This will highlight the main engineering considerations by providing quantitative results.

Section 6.2 presents wireless and optical power budgets for a fibre-radio network operating at 20 GHz with each wavelength transporting 10 subcarriers, each providing a 100 Mbps data rate. The section also summarises the impact of in-band and out-of-band crosstalk in a network. The difference in crosstalk between unidirectional and bidirectional optical fibre links is discussed, as well as other factors that must be considered when designing an optical network. Sections 6.3, 6.4 and 6.5 consider the star, ring and bus network architectures, respectively, comparing in-band and out-of-band crosstalk requirements. Crosstalk values are derived based on the optical power budget derived in Section 6.2.1 for both star and ring networks. More complex architectures are presented and analysed in Section 6.6. Final conclusions comparing the different architectures are made in Section 6.7.

6.2 NETWORK DESIGN CONSIDERATIONS

This section provides a detailed analysis of the wireless and optical power budgets in a fibre-radio system that is used to study the impact of optical crosstalk and power budgets on different network architectures. Other important factors that must be taken into account when designing WDM fibre-radio networks are also discussed.

Section 6.2.1 presents a complete wireless link power budget for both downlink and uplink directions for a system operating at 20 GHz. The conclusions resulting from work presented in previous chapters relating to in-band and out-of-band optical crosstalk are summarised in Section 6.2.2 and 6.2.3 respectively. A brief discussion on unidirectional and bidirectional networks follows in Section 6.2.4. Section 6.2.5 briefly discusses optical powers and fibre chromatic dispersion.

6.2.1 Wireless & Optical Link Power Budget

Before discussing the design of a fibre-radio optical network, it is important to clarify the requirements of the wireless link. In order to provide an interesting and
realistic scenario, we consider a system operating at 20 GHz. This is close to millimetre-wave frequencies (30 GHz to 300 GHz) and allows small cell sizes (< 1 km) and high data rates per radio cell [5, 6]. The 20 GHz carrier frequency will be shown to allow a 100 m cell radius (higher frequencies suffer from higher loss, limiting the RF power budget). The 20 GHz carrier carries data within a certain RF frequency band, with each frequency band carrying ten RF subcarriers, as shown in Figure 6.1. A binary data rate of 100 Mbps per subcarrier is used, providing a total capacity of 1 Gbps per radio cell. This is around 50 times the capacity of 3G wireless networks, as discussed in Section 1.1. QPSK modulation is used for each subcarrier, so that the baud rate is 25 MHz per subcarrier, reducing the impact of thermal noise. A tight channel spacing of 50 MHz allows the ten subcarriers to take up 500 MHz of bandwidth. The number of separate RF bands is not specified, but is typically between five and ten, depending on the frequency re-use factor, so as to minimise co-channel interference between adjacent wireless cells.

![Diagram of WDM Fibre-Radio Networks](image)

**Fig. 6.1** Wireless frequency bands and modulated subcarriers per band centered at 20 GHz.

The numbers used above have been chosen so as to provide a realistic scenario for a future fibre-radio network, providing a much greater capacity than current networks. The approach taken in this section can be used to evaluate power budgets for different operating frequencies. European standards for 24 to 29 GHz point-to-multipoint radio systems ETSI EN 301 213-1/2/3 [7] and ETSI EN 301 215-2/3 [8] have been used as guidelines for transmit power levels. The main parameters that define the limits of the wireless link are:
1) Maximum power radiated by the antenna is +35 dBm total [7] i.e. +25 dBm per subcarrier.

2) The minimum RF power per subcarrier before electrical amplification is −75 dBm, which corresponds to 25 dB above the thermal noise level \( k_B T \Delta f \), where \( k_B \) is Boltzmann’s constant, \( T \) is the temperature in Kelvin, and \( \Delta f \) is the channel bandwidth, in this case 25 MHz.

A schematic diagram of the downlink and uplink paths from the BS to the Mobile Unit is shown in Figure 6.2, for a system using RF subcarrier modulation in the optical link.

![Diagram](image_url)

**Fig. 6.2 (a) Downlink from Base Station to Mobile Unit and (b) uplink from Mobile Unit to Base Station.**

Considering the downlink, Equation 6.1 relates the RF power after photodetection, \( P_{PIN} \), to the RF power received at the mobile unit, \( P_{Mobile} \):

\[
P_{Mobile} = P_{PIN} + G_{BS} - L_{Ant} + G_{AntBS} - L_{WirelessLink} + G_{AntMobile} - L_{Ant} \tag{6.1}
\]

where \( G_{BS} \) is the electrical gain at the BS, \( L_{Ant} \) is the antenna loss due to the feeder, \( G_{AntBS} \) is the gain of the BS antenna relative to isotropic radiation, \( L_{WirelessLink} \) is the wireless link loss and \( G_{AntMobile} \) is the mobile antenna gain, which is equal to zero as it is not directional. Values used for various parameters are shown in Table 6.1.
Table 6.1 BS and Mobile Unit parameters and values.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$P_{PIN}$</td>
<td>-75 dBm</td>
</tr>
<tr>
<td>$G_{BS}$</td>
<td>105 dB</td>
</tr>
<tr>
<td>$L_{Ant}$</td>
<td>5 dB</td>
</tr>
<tr>
<td>$G_{AntBS}$</td>
<td>20 dBi</td>
</tr>
<tr>
<td>$G_{AntMobile}$</td>
<td>0 dBi</td>
</tr>
</tbody>
</table>

$P_{PIN}$ is 25 dB above the noise floor, $G_{BS}$ is determined by the antenna transmit power and other parameters are realistic values.

The wireless link loss can be further separated into the following components:

$$L_{WirelessLink} = L_{Free} + L_{Atm} + L_{Rain} + L_{Foliage} + Margin$$  \hspace{1cm} (6.2) 

where $L_{Free}$ is the loss due to isotropic free-space radiation, $L_{Atm}$ is the loss at the millimetre-wave frequency due to atmospheric (gaseous) absorption, $L_{Rain}$ is the loss due to rainfall, $L_{Foliage}$ is the loss due to absorption by leaves / trees, and Margin is an additional margin. The free-space loss, $L_{Frees}$ can be evaluated using the following expression [9]:

$$L_{Free} = \left( \frac{4\pi L}{\lambda} \right)^2 = 92.4 + 20 \log F + 20 \log L \text{ (dB)}$$  \hspace{1cm} (6.3) 

where $\lambda$ is the wavelength of the electrical carrier, $F$ is the frequency in GHz and $L$ is the propagation distance in km. At a 20 GHz frequency and for a propagation distance of 0.1 km, $L_{Free}$ is equal to 98.4 dB. As can be seen, the free space loss increases by a factor of four for a doubling in length or frequency, which explains why the propagation distance is limited to only 100 m. This allows a pico-cellular network architecture to be used, with many small cellular coverage zones each providing a large bandwidth.

The gaseous absorption due to water is given by [10]:

$$L_{Atm} = 0.1 \text{ dB} / \text{ km} @ 20 \text{ GHz}$$  \hspace{1cm} (6.4) 

so $L_{Atm}$ is equal to 0.01 dB. For a rainfall rate of 75 mm/h, the absorption is given by [9]:

$$L_{Rain} = 10 \text{ dB} / \text{ km.}$$  \hspace{1cm} (6.5)
Thus $L_{\text{rain}}$ is equal to 1 dB. A more significant problem is due to the higher absorption of millimetre-waves by vegetation. The following empirical formula is taken from [9] and is valid for frequencies in the range 200 MHz to 95 GHz and for foliage depths $L$ less than 400 m:

$$L_{\text{foliage}} = 0.2F^{-0.3}L^{0.6} \text{ dB} \quad (6.6)$$

where $F$ is the carrier frequency in MHz, $L$ is the distance traversed through foliage. At a frequency of 2 GHz and with 5 m of foliage, $L_{\text{foliage}}$ is equal to 10.3 dB. This is much greater than the losses due to rainfall and signifies that near line of sight transmission is required. Antenna diversity is one solution that allows areas that would otherwise be in a “shadow” area to be covered. Unexpected loss due to the presence of thick vegetation can render the wireless link inoperable. Finally, a margin of about 5 dB is possible given the requirements stated in points 1) and 2) above.

$$\text{Margin} = 5 \text{ dB} \quad (6.7)$$

Equation 6.1 and the parameter values given above mean that $P_{\text{Mobile}}$ is equal to $-74.7$ dBm, i.e. the same electrical power as that received at the BS after photodetection.

The uplink path can also be analysed, assuming a transmit power of +25 dBm at the mobile unit, also resulting in a value of $-74.7$ dBm for $P_{\text{BS}}$, the power at the input of the electrical amplifiers at the BS. In both cases the total loss from antenna to antenna feeder output is 99.7 dB.

We now consider both electrical to optical and optical to electrical conversions, as shown in Figure 6.3, for a fibre-radio system using RF subcarrier modulation for the optical link.
Fig. 6.3 (a) Downlink optical to electronic conversion at the Base Station and (b) uplink electrical to optical conversion at Base Station.

In the downlink, an external Mach-Zehnder (MZ) modulator can be used to modulate the optical carrier with the desired RF subcarriers. The output of a Mach-Zehnder modulator can be expressed by [11]:

$$P_{out} = P_{in} \sin^2 \left( \frac{\pi V}{2V_{\pi}} \right)$$

where $P_{in}$ and $P_{out}$ are the input and the output to the modulator, $V$ is the applied voltage and $V_{\pi}$ is the voltage that produces a $\pi$ phase-shift in the two arms of the modulator, as shown in Figure 6.4.

Fig. 6.4 Mach-Zehnder modulator transfer function. The modulator is biased at the 3 dB transmission point.
As can be seen in Figure 6.4, the modulator is biased at the 3 dB transmission point, as the modulator transfer characteristic is almost linear in that region, minimising distortion. The required RF modulation is fed to the modulator, producing a corresponding modulation of the amplitude of the optical carrier. The voltage driving the modulator including bias is given by:

\[
V = \frac{V_r}{2} \left[ 1 + \sum_{n=1}^{N} m \alpha_n(t) \cos(\omega_n t + \phi_n) \right]
\]  

(6.9)

where \( m \) is the modulation depth of the RF subcarriers, \( \alpha_n \) is the data amplitude and phase of the \( nth \) subcarrier, and \( \omega_n \) and \( \phi_n \) are the carrier frequencies and phases, respectively. Although the modulator response is relatively linear around the 3 dB point, some distortion will be experienced. If we transmit 10 subcarriers centered on 20 GHz, only third-order composite triple beat (CTB) terms will fall within the signal band. A standard technique based on counting CTB products [12] was used to choose a modulation depth of 13% per carrier for an equivalent electrical CTB of -25 dBC. This corresponds to a linear Q of 8.9 or a BER better than \( 10^{-17} \) assuming that there are enough CTB products to treat the interference from CTB products as Gaussian. While the maximum total modulation depth is \( 10 \times 13\% = 130\% \), the fact that carriers are at different frequencies means this an acceptable modulation depth per subcarrier.

The modulator output can be approximated as follows [11]:

\[
P_{out} = P_{in} \sin^2 \left( \frac{\pi}{4} + \frac{\pi}{2} m \sum \cos(\alpha + \phi) \right)
= \frac{P_{in}}{2} \left[ 1 + \sin(\pi m \sum \cos(\alpha + \phi)) \right]
\]

(6.10)

where indices for multiple RF carriers have been left out for clarity. Equation 6.10 is the small-signal output of the modulator, valid for low modulation depths. This can be used to evaluate the RF power at the output of a photodiode given an incident optical power:

\[
P_{RF} = \frac{(\Re P_{opt})^2}{2R}
\]

(6.11)

where \( \Re \) is the photodiode responsivity, \( R \) is the circuit resistance, and \( P_{opt} \) is the average optical power incident on the photodiode. Using Equation 6.10 the optical
power containing the modulated RF signal, which is equal to $P_{\text{opt}}$ can be found, and is given by:

$$P_{\text{opt}} = \frac{P_{\text{in}}\pi m}{2}$$  \hspace{1cm} (6.12)

where $P_{\text{in}}$ is the optical power at the input of the modulator and $m$ is the carrier modulation depth. The optical power that corresponds to an RF power of $-75$ dBm at the photodiode output can now be evaluated. Rearranging Equation 6.11 and using log scales:

$$P_{\text{opt}} \text{ (dBW)} = \frac{P_{\text{RF}} - 10\log\left(\frac{R}{2}\right) - 20\log(\Re \pi m)}{2}$$  \hspace{1cm} (6.13)

The parameter values are shown in Table 6.2 for a 50 $\Omega$ resistance, using the modulation depths and RF power that were derived previously. A conservative value of 0.5 A/W is used for the PIN detector responsivity at 20 GHz.

**Table 6.2 Parameters and values used to evaluate minimum received optical power.**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$R$</td>
<td>50 $\Omega$</td>
</tr>
<tr>
<td>$\Re$</td>
<td>0.5 A / W</td>
</tr>
<tr>
<td>$m$</td>
<td>0.13</td>
</tr>
<tr>
<td>$P_{\text{RF}}$</td>
<td>-105 dBW</td>
</tr>
</tbody>
</table>

Using Equation 6.13 and values from Table 6.2, $P_{\text{opt}}$ is equal to $-52.6$ dBW. Hence the minimum optical power incident on the BS photodiode is $-22.6$ dBm, resulting in an average RF power of $-75$ dBm. Note that a rather conservative responsivity of 0.5 A/W was used. Using a figure of 0.8 A/W gives a 2 dB improvement in $P_{\text{opt}}$, i.e. it is reduced to $-24.6$ dBm. This means designing a photodetector with a high responsivity at millimetre-wave frequencies can relax the optical power budget.

For the uplink, we require the amount of RF power driving the modulator to ensure a 13% modulation depth of the optical carrier. This required power is also the same as that needed at the CO in order to modulate the RF data onto the optical carrier. Recall Equation 6.9,
\[ V = \frac{V_\pi}{2} \left[ 1 + \sum_{n=1}^{N} m \alpha_n(t) \cos(\phi_n) \right] \] (6.14)

which gives the expression for the voltage applied to the modulator. The RF term has amplitude \( V_\pi m/2 \). Hence the average RF power required to drive the modulator, \( P_{RF\, Mod} \), is given by:

\[ P_{RF\, Mod} = \frac{(V_\pi m/2)^2}{2R} \] (6.15)

Parameter values shown in Table 6.3 can be used to evaluate \( P_{RF\, Mod} \). The same resistance and modulation depths are used as before, together with a conservative value for \( V_\pi \).

**Table 6.3 Parameters and values used to evaluate modulator RF power.**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>( V_\pi )</td>
<td>5 V</td>
</tr>
<tr>
<td>( R )</td>
<td>50 Ω</td>
</tr>
<tr>
<td>( m )</td>
<td>0.13</td>
</tr>
</tbody>
</table>

Using Equation 6.15 and values from Table 6.3, \( P_{RF\, Mod} \) is equal to 0.3 dBm. The critical factor here is the value of \( V_\pi \) since a factor of 2 increase produces a 6 dB increase in the RF power required to drive the modulator. Recall that for the uplink, the output of the antenna after the antenna feeder loss was −75 dBm, so only 75 dB of electrical gain is required in the uplink direction at the BS compared to 105 dB gain at the BS for the downlink. This would have to be increased if \( V_\pi \) were greater than 5 V.

Table 6.4 provides a summary of link parameters and shows the actual power budget used for the downlink and for the uplink.
Table 6.4 Wireless and Optical Link Budget parameters and values.

<table>
<thead>
<tr>
<th>Downlink</th>
<th>PIN Diode</th>
<th>BS Gain</th>
<th>BS Ant. feed loss</th>
<th>BS Ant. Gain</th>
<th>Wireless link</th>
<th>Mob. Ant. Gain</th>
<th>Mob. Ant. feed loss</th>
</tr>
</thead>
<tbody>
<tr>
<td>Gain / Loss (dB)</td>
<td>-</td>
<td>105</td>
<td>-5</td>
<td>20</td>
<td>-114.7</td>
<td>0</td>
<td>-5</td>
</tr>
<tr>
<td>RF power (dBm)</td>
<td>-75</td>
<td>+30</td>
<td>+25</td>
<td>+45</td>
<td>-69.7</td>
<td>-69.7</td>
<td>-74.5</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Uplink</th>
<th>Mob. Ant. Gain</th>
<th>Ant. Gain</th>
<th>Wireless link</th>
<th>BS Ant. Gain</th>
<th>BS Ant. feed loss</th>
<th>BS Gain</th>
</tr>
</thead>
<tbody>
<tr>
<td>Gain / Loss (dB)</td>
<td>-</td>
<td>0</td>
<td>-114.7</td>
<td>+20</td>
<td>-5</td>
<td>+75</td>
</tr>
<tr>
<td>RF power (dBm)</td>
<td>+25</td>
<td>+25</td>
<td>-89.7</td>
<td>-69.7</td>
<td>-74.7</td>
<td>+0.3</td>
</tr>
</tbody>
</table>

| Optical Budget | MZ modulator $V_\pi$  
Modulation depth per subcarrier  
RF power to modulator per subcarrier  
PIN diode responsivity  
Min. Received Optical Power for $-75$ dBm RF subcarrier power | $V_\pi = 5V$  
$m = 0.13$  
$P_{RF_{Mod}} = 0.3$ dBm  
$\Re=0.5$ A/W  
$P_{opt}=-22.6$ dBm |

In order to provide a more realistic assessment of the values used above and in order to check that noise does not significantly degrade the electrical SNR, an evaluation of noise floors at each successive stage of the link was performed using realistic noise figures (NFs) for amplifiers. Table 6.5 follows the evolution of electrical SNR though the link, taking into account gain and NF.
Table 6.5 Noise calculation for both downlink and uplink (Gain, NF and SNR in dB and output powers in dBm).

<table>
<thead>
<tr>
<th>Downlink</th>
<th>PIN Output</th>
<th>LNA</th>
<th>MPA</th>
<th>MPA</th>
<th>BS Ant.</th>
<th>Wireless link</th>
<th>Mob. Ant.</th>
<th>Mob. LNA</th>
<th>Mob. Mixer</th>
</tr>
</thead>
<tbody>
<tr>
<td>Gain</td>
<td>-</td>
<td>20</td>
<td>40</td>
<td>45</td>
<td>15</td>
<td>-114.7</td>
<td>-5</td>
<td>20</td>
<td>-10</td>
</tr>
<tr>
<td>NF</td>
<td>-</td>
<td>3</td>
<td>10</td>
<td>10</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>3</td>
<td>10</td>
</tr>
<tr>
<td>Total Gain</td>
<td>-</td>
<td>20</td>
<td>60</td>
<td>105</td>
<td>120</td>
<td>5.3</td>
<td>0.3</td>
<td>20.3</td>
<td>10.3</td>
</tr>
<tr>
<td>Total NF</td>
<td>-</td>
<td>3.0</td>
<td>3.19</td>
<td>3.19</td>
<td>3.19</td>
<td>3.19</td>
<td>4.8</td>
<td>4.92</td>
<td></td>
</tr>
<tr>
<td>Pout</td>
<td>-75</td>
<td>-55</td>
<td>-15</td>
<td>+30</td>
<td>+45</td>
<td>-69.7</td>
<td>-74.5</td>
<td>-54.5</td>
<td>-64.5</td>
</tr>
<tr>
<td>Noise out</td>
<td>-100</td>
<td>-77</td>
<td>-36.8</td>
<td>+8.2</td>
<td>23.2</td>
<td>-91.5</td>
<td>-96.5</td>
<td>-74.9</td>
<td>-64.7</td>
</tr>
<tr>
<td>SNR out</td>
<td>25</td>
<td>22</td>
<td>21.8</td>
<td>21.8</td>
<td>21.8</td>
<td>21.8</td>
<td>20.2</td>
<td>20.1</td>
<td></td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Uplink</th>
<th>Mob. Ant.</th>
<th>Wireless link</th>
<th>BS Ant.</th>
<th>LNA</th>
<th>MPA</th>
<th>MPA</th>
<th>Opt. Link</th>
<th>CO LNA</th>
<th>CO Mixer</th>
</tr>
</thead>
<tbody>
<tr>
<td>Gain</td>
<td>-</td>
<td>-114.7</td>
<td>15</td>
<td>20</td>
<td>35</td>
<td>20</td>
<td>-75.3</td>
<td>15</td>
<td>-10</td>
</tr>
<tr>
<td>NF</td>
<td>-</td>
<td>-114.7</td>
<td>-99.7</td>
<td>-79.7</td>
<td>-44.7</td>
<td>-24.7</td>
<td>-100</td>
<td>-85</td>
<td>-95</td>
</tr>
<tr>
<td>Total Gain</td>
<td>-</td>
<td>-114.7</td>
<td>-99.7</td>
<td>-76.9</td>
<td>-41.7</td>
<td>-21.7</td>
<td>-95.2</td>
<td>-79</td>
<td>-88.7</td>
</tr>
<tr>
<td>Total NF</td>
<td>-</td>
<td>0</td>
<td>0</td>
<td>99.7</td>
<td>100.1</td>
<td>100.1</td>
<td>103.1</td>
<td>104.8</td>
<td>105.2</td>
</tr>
<tr>
<td>Pout</td>
<td>-</td>
<td>-89.7</td>
<td>-74.7</td>
<td>-54.7</td>
<td>-19.7</td>
<td>0.3</td>
<td>-75</td>
<td>-60</td>
<td>-70</td>
</tr>
<tr>
<td>Noise Out</td>
<td>0</td>
<td>-114.7</td>
<td>-99.7</td>
<td>-76.9</td>
<td>-41.7</td>
<td>-21.7</td>
<td>-95.2</td>
<td>-79</td>
<td>-88.7</td>
</tr>
<tr>
<td>SNR out</td>
<td>25</td>
<td>25</td>
<td>25</td>
<td>22.1</td>
<td>21.9</td>
<td>21.9</td>
<td>20.2</td>
<td>19</td>
<td>18.7</td>
</tr>
</tbody>
</table>

As pointed out earlier, the receiver level of –75 dBm is 25 dB above the thermal noise floor. For the downlink, it is assumed that the electrical SNR upon photodetection is 25 dB, i.e. the noise level of the optical signal is the same as the thermal noise floor. The final electrical SNR is 20 dB at the output of the mixer which downconverts the RF data to baseband. For uplink, again the input electrical SNR is assumed to be 25 dB, with thermal noise being added by the PIN diode at the CO, resulting in a 19 dB electrical SNR at the output of the CO mixer. This is more than adequate to ensure a high signal quality. Since the BER for QPSK modulation is given by $0.5 \text{erfc}(\sqrt{\text{SNR}})$ [13], this corresponds to $<10^{-17}$ without error correction. This suggests there is the potential for an additional margin. For example, an SNR of 12.6 dB corresponds to a BER of $10^{-9}$. Hence the power budget may be reduced by an additional 5 dB in the electrical domain, which corresponds to 2.5 dB of optical power.
One aspect that has not considered is the linearity requirements of the electrical amplifiers, since they will have to amplify a whole RF frequency band. As each BS is assigned ten subcarriers, the electrical amplifiers must be designed so as to minimise electrical distortion. The above analysis assumes no nonlinear distortion due to electrical amplification.

6.2.2 In-band Optical Crosstalk

Chapters 3, 4 and 5 have shown that for subcarrier modulation the modulation determines the resulting power penalty [14-16]. For BPSK, in-band crosstalk is actually reduced compared to baseband modulation, but we are considering QPSK modulation, which is slightly worse than BPSK. We have shown that the impact of in-band crosstalk can be minimised if signal and crosstalk RF carriers are at different wireless frequencies. If in-band crosstalk occurs between uplink and downlink wavelengths, FDMA will ensure that different frequencies are used in both directions, ensuring the power penalty is minimised. In this chapter, we will assign a 0.5 dB system margin to in-band optical crosstalk, allowing us to determine the maximum in-band crosstalk level acceptable, which is −18 dB for QPSK modulation with different RF carrier frequencies (see Section 3.7.3).

6.2.3 Out-of-band Optical Crosstalk

Chapter 5 showed how in-band and out-of-band crosstalk together impose rather tight requirements on component crosstalk specifications, particularly if many WDM channels are present [17]. However, one of the important results arising from the analysis of out-of-band crosstalk is that for over a dozen crosstalk terms the resulting power penalty is due to optical power addition alone (see Section 5.3) i.e. is equal to the total crosstalk optical power. The power penalty in this case, for an out-of-band crosstalk level \( x \) and for \( n \) crosstalk terms, is given by:

\[
PP(dB) = 10 \log[1 + nx]
\]

(6.16)

where \( x \) is the linear crosstalk ratio, i.e. \( x = 10^{L_x(dB)/10} \). For example, 99 out-of-band crosstalk terms, at −30 dB crosstalk each, produces an extra 0.4 dB of power. For 199 channels, 0.8 dB. With a system margin of 1 dB typically allocated to optical crosstalk, we can allocate 0.5 dB each to in-band and out-of-band crosstalk. If a higher total
received optical power can be tolerated, maintaining the signal power level constant, there will be no degradation in BER due to out-of-band optical crosstalk. In this case the effective power penalty due to out-of-band crosstalk is zero.

If a margin is allocated to both in-band and out-of-band crosstalk, it may be advantageous to allocate a greater proportion of a 1 dB total power penalty to in-band crosstalk or to out-of-band crosstalk. This depends on actual crosstalk levels and potential cost savings due to a higher crosstalk tolerance. In the network design examples discussed in this chapter, a 0.5 dB system margin will be allocated to out-of-band crosstalk.

6.2.4 Unidirectional versus Bidirectional Links

An important decision that has to be made when designing an optical link or network is whether individual links are unidirectional or bidirectional [18] as this affects network design. The difference between these two types of links is illustrated in Figure 6.5.

![Diagram of unidirectional and bidirectional links](image)

(a) 2 unidirectional links  
(b) 1 bidirectional link

Fig. 6.5 (a) Unidirectional and (b) bidirectional links.

For unidirectional links, the signals propagate down the optical fibre in one direction only. For a fibre-radio network comprising both a downlink and an uplink, two separate links are required, one for each direction, as shown in Figure 6.5 (a). This has the benefit of eliminating both in-band and out-of-band crosstalk between downlink and uplink channels.
On the other hand, bidirectional links have information travelling in both directions along the fibre, so that downlink and uplink data can be transported in the same fibre, as shown in Figure 6.5 (b). This eliminates the need for duplication of fibre spans, although optical components have to modified / added relative to a unidirectional link. In-band crosstalk is possible between downlink and uplink channels if they use the same wavelength bands. A further problem associated with bidirectional links is Rayleigh backscattering in optical fibre, which typically causes a reflection of $-30$ to $-33$ dB [19]. Depending on whether or not the same wavelengths are used in both directions, in-band crosstalk may be significant. In terms of spectral efficiency, if separate wavelength bands are used for the two different directions, then only half of the optical bandwidth can be used per direction compared to a unidirectional link.

While bidirectional links reduce the amount of optical fibre used, it is at the cost of additional complexity. Potential problems with optical reflections and fibre Rayleigh backscattering [20] may occur, and the downlink and uplink traffic must be separated at both CO and individual BSs. The focus of this chapter is to provide design examples showing how optical crosstalk can impact on network performance. Subsequent sections comparing different architectures briefly discuss the issues associated with having bidirectional links.

### 6.2.5 Other Factors

#### (A) Optical Power

One of the key parameters that defines the potential size of an optical network or the reach of a network is the transmit optical power per channel at the output of the transmitter, i.e. at the output of the CO or the BS in a fibre-radio network. Since the minimum received optical power derived in Section 6.2.2 is $-22.6$ dBm, this is increased to $-19$ dBm for a 3.5 dB margin. The total maximum loss, Optical Link Loss, experienced by an optical channel is therefore:

$$\text{Optical Link Loss (dB)} = P_{Tx} - P_{Rx} \quad (6.17)$$

where $P_{Tx}$ is the optical power at the output of the transmitter and $P_{Rx}$ is the received optical power, equal to $-19$ dBm in our case. If we assume a conservative $0$ dBm per channel, the total link loss is $19$ dB. If this could be increased to $+5$ dBm per channel,
the total link loss would increase to 24 dB. Two factors limit the maximum optical power per channel, nonlinear effects arising in the fibre and optical amplifier saturation if optical amplification is required.

The effect of nonlinearities is well known for amplitude modulated optical carriers, and has been quantified in terms of carrier-to-noise ratio (CNR) levels for subcarrier modulated analogue links [21-27]. However, these studies usually consider a single optical wavelength with multiple (low-frequency) subcarriers. There is limited research on crosstalk arising between different wavelengths and results indicate that there is a limit to the channel power [25, 26]. Two factors can reduce nonlinear effects in fibre-radio network operating at 20 GHz. Firstly, the effect of self-phase modulation (SPM) between the optical carrier and subcarriers and between different subcarriers can be reduced by fibre chromatic dispersion. Similarly, for cross-phase modulation (XPM) and four-wave mixing (FWM) between different wavelength channels, the high dispersion of SMF in the 1550 nm region will reduce the impact of fibre nonlinearities. These nonlinear effects are illustrated in Figure 6.6.

![Nonlinear effects within a single wavelength](image)

**Fig. 6.6 Nonlinear effects due to fibre for single-sideband subcarrier modulation.**

Secondly, fibre-radio networks have a limited maximum fibre length, around 40 km at most, but more likely distances are 5 to 10 km from CO to BS, so that the effects of nonlinearities will be limited. Digitally modulated links are also much more tolerant to nonlinear effects than analogue links as much lower SNR ratios are acceptable, making them more tolerant to nonlinear effects.

If optical amplification is required at the CO, for example, to compensate for high multiplexing / demultiplexing losses etc., then the total output power of the EDFA would be around +20 dBm. This is distributed over the number of WDM channels, so that the channel power would be +0 dBm for 100 channels. If a lower number of channels were present, then channel powers may be too high and nonlinear effects could
degrade transmission. This is not the case in passive networks, however, as component losses due to external modulation and multiplexing will limit the actual maximum channel power at the output of the CO as typical maximum laser output powers are in the order of +10 dBm.

(B) Dispersion

If a subcarrier-modulated optical carrier is transmitted through optical fibre, a relative phase-shift between both sidebands will accumulate due to dispersion. The two RF tones present at the photodetector due to each sideband beating with the carrier will thus be out-of-phase, which can lead to a significant loss in received RF power [28]. Several techniques can be used to overcome this, one of which is to transmit a single-sideband only [29]. Dispersion is ignored in this chapter, as it is expected that its effects in terms of RF power fading are eliminated. Note that if the effect is not fully compensated, then an additional margin will have to be taken from the available optical link budget.

6.3 Star Network Architecture and Design Example

This section describes the star architecture and shows how it can be used in fibre-radio networks. A network design is presented using the power budget from Section 6.2.1.

6.3.1 Star Network Architecture

The main access network architecture is the star, which is used in Passive Optical Networks (PONs) [30-32] and CATV access networks [33-35]. A generic star network architecture is shown in Figure 6.7.
Fig. 6.7 Generic WDM star network architecture. (a) Unidirectional links (only downlink shown for clarity) and (b) bidirectional links.

As shown in Figure 6.7, if we consider the downlink from CO to BS in a fibre-radio network, multiple WDM channels can be multiplexed at the CO before being sent through a WDM link to a remote node (RN). At the RN, an optical demultiplexer (WDM) routes different wavelengths to individual BSs. This is different to PONs or CATV networks in which passive splitting is usually used, i.e. the networks are distributive [34]. For the unidirectional network shown in (a), a single WDM link is required together with a single optical component, $N$ single-wavelength links are required to reach $N$ Base Stations. The same architecture can be used for the uplink, duplicating the downlink. For a bidirectional network as shown in (b) a single WDM link and $N$ single-wavelength links are required together with a single WDM multiplexer / demultiplexer (MUX / DEMUX).

(A) Coverage Area and WDM Channel Number

Figure 6.7 is simplified, as in reality a single CO will distribute data to several different areas, each using a separate WDM star network, as shown in Figure 6.8.
Hence the length of the WDM link (from CO to RN) may vary from 5 to 20 km or more, with the actual RN to BS length less than 1-2 km. Note that if both CO to WDM and WDM to BS link lengths are of similar magnitude, it may be cost-effective to remove the RN demultiplexer and do away with the WDM link altogether. This means that all $N$ links from CO to BSs are single-wavelength, and the star originates at the CO, as shown in Figure 6.8.

If a single BS provides radio coverage to a 100 m radius, the number of BSs required, $N_{BS}$, for an area $m$ by $m$ km$^2$ is given by

$$N_{BS} = \frac{m^2}{\pi 0.1^2} = 31.8m^2$$  \hspace{1cm} (6.18)

This means that about 30 BSs are required for a 1 km$^2$. Hence about 120 WDM channels would be needed for a 2x2 km$^2$ area, and 286 for a 3x3 km$^2$ area. This confirms that for a small radio cell size, RN to BS distances will be small. If we assume no RN is used, the furthest BS would be 2 km away, so that adjacent COs would be 4 km apart. If, however, the CO to RN distance can be 10 to 20 km or more, then adjacent COs would be 20 to 40 km away from each other. Note that in this case $N_{BS}$ is equal to 3180 (10 km) and 12720 (20 km) if the whole area is to be covered, assuming no overlap of cells is required. It may be physically impossible to fit such a large amount of equipment into a single location. Table 6.6 shows the number of BSs or WDM channels required to cover different sized areas.
Table 6.6 Number of BSs / WDM channels required as a function of fibre-radio coverage area assuming each BS covers a 100 m wireless cell radius.

<table>
<thead>
<tr>
<th>Wireless coverage area</th>
<th>BSs / WDM channels</th>
</tr>
</thead>
<tbody>
<tr>
<td>1x1 km²</td>
<td>30</td>
</tr>
<tr>
<td>2x2 km²</td>
<td>120</td>
</tr>
<tr>
<td>3x3 km²</td>
<td>286</td>
</tr>
<tr>
<td>5x5 km²</td>
<td>795</td>
</tr>
<tr>
<td>10x10 km²</td>
<td>3180</td>
</tr>
<tr>
<td>20x20 km²</td>
<td>12720</td>
</tr>
</tbody>
</table>

The greatest advantage of the star architecture is its inherent simplicity. The power budget of the longest link should be considered during design, so that if RN to BS distances are shorter the network is still operational. Since RN to BS distances would not vary by more than 1 to 2 km, the actual difference in received optical power would be less than 0.5 dB anyhow.

(B) Unidirectional vs. Bidirectional Links

The choice between unidirectional and bidirectional networks depends not only on cost but also on network design specifications. While bidirectional links allow the number of fibre spans to be halved, this is not really relevant given the low cost of fibre and the short distances involved in access networks. A more important consideration is the fact that in bidirectional networks the signals travelling in opposite directions have to be separated. Figure 6.9 shows two separate unidirectional links that allow downlink and uplink transmission together with various bidirectional link configurations.
For unidirectional links, two completely separate links can be used for the downlink and uplink, as shown in (a). Alternatively, a single RN can be used to multiplex and demultiplex downlink and uplink wavelengths as shown in (b), although this requires different wavelength bands to be used for the two directions.

For bidirectional links, Figure 6.9 shows that both optical circulators and multiplexers / demultiplexers are needed. Although cheaper optical couplers can replace circulators, circulators are less lossy (~1 dB), while optical couplers automatically imply a 3- to 4 dB loss in power budget. Using an additional component at each BS, however, is uneconomical. A potential solution is to use separate wavelengths for both directions and use unidirectional links between the RN WDM MUX / DEMUX and individual BS, as illustrated in Figure 6.9 (d).

The significant advantage of using two separate unidirectional networks is that in-band optical crosstalk between downlink and uplink channels is eliminated. In bidirectional networks, this can be eliminated by ensuring different wavelength bands are used for both directions, also minimising the impact of out-of-band crosstalk due to Rayleigh backscatter in the fibre. These issues are demonstrated in the following section.

### 6.3.2 Star Network Design Example

The star network is shown in Figure 6.10, which follows the topology of the first unidirectional link shown in Figure 6.9.
Figure 6.10 shows that the CO comprises a set of N lasers and external modulators and an arrayed-waveguide grating (AWG) that multiplexes the signals together for the downlink path. A remote AWG demultiplexes the wavelengths, which are then distributed to individual BSs in a star architecture. At the BS, the downlink fibre is directly connected to a PIN diode, where the optical signal is converted to an electrical signal, allowing the RF subcarriers to be recovered and amplified before being sent to the BS antenna. The uplink path is similar, with each BS comprising a laser and external modulator. The modulated output is connected to a separate AWG that multiplexes all N wavelengths together. The WDM output of the AWG is sent back to the CO, where the signals are demultiplexed by a second AWG in the CO before being detected by N PIN diodes.

(A) Optical Power Budget

Figure 6.10 includes component losses for the external modulator, including 3 dB bias, AWG loss, together with an unspecified loss for both fibre lengths, $L_1$ and $L_2$. $L_1$ is the length of the link from the CO to the remote AWG and $L_2$ is the length of the longest link from the AWG to the furthest BS. The link budget for both downlink and uplink are given by:

$$P_{R_{BS}} = T_{x_{CO}} - L_{Mod} - L_{AWG} - \alpha L_1 - L_{AWG} - \alpha L_2$$  \hspace{1cm} (6.19)

and

$$P_{R_{CO}} = T_{x_{BS}} - L_{Mod} - \alpha L_2 - L_{AWG} - \alpha L_1 - L_{AWG}$$  \hspace{1cm} (6.20)
where $P_{RxCO}$ and $P_{RxBS}$ are the received optical powers at the CO and BS PIN diodes respectively, $L_{Mod}$ is the modulator loss including 3 dB bias, $L_{AWG}$ is the loss of the AWGs, $\alpha$ is the fibre loss in dB/km and $L_1$ and $L_2$ are the two fibre lengths in km, as defined above. Using the component values from Figure 6.10, and the minimum received optical power derived in Section 6.2.1 of −19 dBm, we obtain:

$$P_{RxBS} = Tx_{CO} - \alpha(L_1 + L_2) - 17 = -19 \text{ dBm} \quad (6.21)$$

and

$$P_{RxCO} = Tx_{BS} - \alpha(L_4 + L_2) - 17 = -19 \text{ dBm} \quad (6.22)$$

Rearranging for $L_1+L_2$, we obtain:

$$L_1 + L_2 = L_{Total} = \frac{Tx + 2}{\alpha} \quad (6.23)$$

Using a value for $\alpha$ of 0.2 dB/km, $L_{Total}$ is equal to 60 km for a laser output power of +10 dBm. Since the combined loss of the CO modulator and AWG is 12 dB, this means the actual CO channel output power is −2 dBm, while for the BS a laser output power of +10 dBm means the BS channel output power is +3 dBm. For the CO to AWG link, however, multiple optical channels are present, whereas the BS to AWG link is single-channel, so that a higher channel power can be tolerated for that link. In Section 6.3.1 (A), the total fibre span length was limited to around 10 km for a pico-cellular radio cell size of 100m radius. If a remote location is to be provided with wireless coverage, the 60 km maximum distance provides this opportunity. The above link budget uses realistic component values and includes a 3.5 dB system margin. The minimum received optical power was derived in Section 6.2.1 for a specific wireless link operating at 20 GHz with a 100 m maximum wireless propagation length. It is important to realise that component losses total 19 dB, compared to 12 dB for 60 km of fibre, so that for shorter distances component losses dominate. The optical power budget for both downlink and uplink is summarised in Table 6.7.
Table 6.7 Optical power budgets for downlink and uplink paths for a star network.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Downlink</th>
<th>Parameter</th>
<th>Uplink</th>
</tr>
</thead>
<tbody>
<tr>
<td>CO Laser Output</td>
<td>+10 dBm</td>
<td>BS Laser Output</td>
<td>+10 dBm</td>
</tr>
<tr>
<td>MZ modulator</td>
<td>-7 dB</td>
<td>MZ modulator</td>
<td>-7 dB</td>
</tr>
<tr>
<td>AWG MUX</td>
<td>-5 dB</td>
<td>AWG MUX</td>
<td>-5 dB</td>
</tr>
<tr>
<td>Max. 60 km fibre</td>
<td>-12 dB</td>
<td>Max. 60 km fibre</td>
<td>-12 dB</td>
</tr>
<tr>
<td>AWG DEMUX</td>
<td>-5 dB</td>
<td>AWG DEMUX</td>
<td>-5 dB</td>
</tr>
<tr>
<td>BS PIN diode Rx</td>
<td>-19 dBm</td>
<td>CO PIN diode Rx</td>
<td>-19 dBm</td>
</tr>
<tr>
<td>Min. Rx</td>
<td>-22.5 dBm</td>
<td>Min. Rx</td>
<td>-22.5 dBm</td>
</tr>
<tr>
<td>System Margin</td>
<td>3.5 dB</td>
<td>System margin</td>
<td>3.5 dB</td>
</tr>
</tbody>
</table>

(B) Optical Crosstalk Calculation

From an optical crosstalk point of view, two types of crosstalk can arise: out-of-band and in-band crosstalk. Out-of-band crosstalk is present at each BS in the downlink and at the CO due to the imperfect demultiplexing, as shown in Figure 6.11.

Fig. 6.11 Out-of-band optical crosstalk in a star network.

The power penalty due to out-of-band crosstalk can be calculated using a power addition formula (see Section 6.2.3). From Equation 6.16 the total crosstalk level \( x(dB) \) for a given power penalty \( PP(dB) \) is given by:
\[ x(\text{dB}) = 10\log\left(10^{PP(\text{dB})/10} - 1\right) - 10\log(N - 1) \]  

(6.24)

where \( x \) and \( PP \) are in dB and \( N \) is the total number of WDM channels. For a power penalty of 0.5 dB,

\[ x(\text{dB}) = -9.1 - 10\log(N - 1) \]  

(6.25)

which is equal to \(-29\) dB for 100 WDM channels (-32 dB for 200). Hence the out-of-band crosstalk requirement is around \(-30\) dB, which is achievable in current AWGs [36-39].

There is no in-band crosstalk in unidirectional networks unless two reflections occur from optical components or fibre, so that unless there is a faulty component in-band crosstalk can be ignored. In-band crosstalk is in bidirectional networks, however, as shown in Figure 6.12.

![Diagram](Image)

**Fig. 6.12 In-band optical crosstalk in a bidirectional star network.**

Figure 6.12 shows how in-band crosstalk can arise between the downlink and uplink channels when they are coupled together at the BS or CO. The in-band crosstalk from downlink to uplink, \( x(\text{down-up}) \) and uplink to downlink, \( x(\text{up-down}) \) are given by:

\[ x(\text{down-up}) = P_{\text{down}} - P_{\text{up}} - x(\text{in-band}) \]  

(6.26)

and

\[ x(\text{up-down}) = P_{\text{up}} - P_{\text{down}} - x(\text{in-band}) \]  

(6.27)

where \( P_{\text{down}} \) and \( P_{\text{up}} \) are the downstream and upstream channel powers respectively, and \( x(\text{in-band}) \) is the actual optical component in-band crosstalk level. For a 0.5 dB
power penalty, the effective in-band crosstalk level is –18 dB for QPSK. At the CO, the in-band crosstalk \(x(\text{down-up})\) is given by:

\[
x(\text{down-up}) = (T_{xCO} - L_{Mod} - L_{AWG}) - (T_{xBS} - L_{Mod} - \alpha L_2 - L_{AWG} - \alpha L_1) - x(\text{in-band})
\]

\[
= T_{xCO} - T_{xBS} + \alpha L_{Total} - x(\text{in-band})
\]

(6.28)

and at the BS, the in-band crosstalk \(x(\text{up-down})\) is given by:

\[
x(\text{up-down}) = (T_{xBS} - L_{Mod}) - (T_{xCO} - L_{Mod} - L_{AWG} - \alpha L_1 - L_{AWG} - \alpha L_2) - x(\text{in-band})
\]

\[
= T_{xBS} - T_{xCO} + (2L_{AWG} + \alpha L_{Total}) - x(\text{in-band})
\]

(6.29)

Equations 6.28 and 6.29 show that the relative uplink / downlink powers determine the effective in-band crosstalk [40-42], so that component losses and fibre lengths are important. Using a 20 km fibre length,

\[
x(\text{down-up}) = T_{xCO} - T_{xBS} + 4 - x(\text{in-band})
\]

(6.30)

and

\[
x(\text{up-down}) = T_{xBS} - T_{xCO} + 14 - x(\text{in-band})
\]

(6.31)

Equations 6.30 and 6.31 show that for equal transmit powers the effective in-band crosstalk is 4 dB higher at the CO and 14 dB higher at the BS than the actual component in-band crosstalk. This means that for an effective maximum in-band crosstalk level of –18 dB (0.5 dB PP) the required in-band crosstalk component specifications are –22 dB and –32 dB at the CO and BS respectively. This is achievable using a circulator, whose isolation is typically 40 dB or better [43, 44]. Note that the crosstalk specification decreases to –30 and –40 dB for a 60 km total fibre length, which are much more stringent requirements.

Figure 6.12 also shows Rayleigh backscatter in the fibre that also results in in-band crosstalk as analysed in Equations 6.28 to 6.33, except that \(x(\text{in-band})\) is replaced by the Rayleigh backscatter coefficient. Hence the in-band crosstalk due to Rayleigh backscatter is equal to –26 dB at the CO and –16 dB at the BS, for –30 dB of backscatter. This is above the –18 dB limit at the BS for a 0.5 dB power penalty. This can be improved by reducing the BS laser power relative to the CO laser power. A 2 dB reduction would achieve the –18 dB limit, increasing the in-band crosstalk at the CO to –24 dB. The situation would be worse for a longer fibre length. However, if separate wavelength bands are used for downlink and uplink directions, the crosstalk between
both directions is no longer in-band, but out-of-band, so that in-band crosstalk is no longer a concern. This technique is systematically used when designing bidirectional networks. If this is the case, then the out-of-band crosstalk due to Rayleigh backscatter has to be included in the out-of-band crosstalk analysis. Figure 6.12 also shows the out-of-band crosstalk occurring due to imperfect demultiplexing, as for unidirectional networks.

6.3.3 Summary

Section 6.3.1 has introduced the star architecture and discussed the main factors that had to be considered when designing a star network. Section 6.3.2 gave a design solution based on the optical and RF power budgets presented in Section 6.2.1. The simplicity of the star network allows for very large transmission distances, up to 60 km for our example. The star network can be designed for the longest link, guaranteeing operation for the shorter RN to BS links.

Out-of-band crosstalk due to optical power addition requires component crosstalk levels of –32 dB for 200 WDM channels and a 0.5 dB power penalty. In-band crosstalk occurs only in bidirectional networks where downlink and uplink traffic has to be separated using an optical component at the CO and the BS. The difference in downlink and uplink power levels results in effective in-band crosstalk levels that are 4 dB and 14 dB higher than the component crosstalk at the CO and BS respectively. This means component in-band crosstalk levels of –22 dB and –32 dB respectively are required for a 0.5 power penalty. While this may seem high, it is easily achievable using an optical circulator. Rayleigh backscattering, however, can lead to unacceptable in-band crosstalk at the BS. This can be overcome by reducing the BS transmit power by 2 dB for a 20 km total fibre length. However, if bidirectional links are used the simplest strategy is to use different wavelength bands for the two directions, so that crosstalk is out-of-band – this eliminates any in-band crosstalk.

In conclusion, the star architecture provides a total link loss that allows for up to 60 km separation between CO and BS. The increasing difference between upstream and downstream channels can lead, however, to problems with in-band crosstalk in a bidirectional network, not only due to the optical components, but also due to Rayleigh backscatter. Hence in-band crosstalk can be a limiting factor in bidirectional star
networks, so that two separate wavelength bands are typically used for the two directions.

6.4 **RING NETWORK ARCHITECTURE AND DESIGN EXAMPLE**

This section discusses the merits of a ring architecture and illustrates some of the problems associated with it using a design example.

6.4.1 **Ring Network Architecture**

One of the main network architectures that is used in metropolitan and long-haul networks is the ring network [45-47], which is shown schematically in Figure 6.13 for a fibre-radio network.

![Fig. 6.13 Generic WDM fibre-radio ring network topology.](image)

Comparing Figure 6.13 to Figure 6.7 for a star network shows that the ring architecture is very different. The architecture allows the CO to distribute wavelengths to remote BSs that are placed along the ring, with an optical add-drop multiplexer (OADM) routing the relevant wavelength to the relevant BS. The ring architecture has unidirectional WDM traffic throughout the ring, with add (uplink) and drop (downlink) wavelengths going from the OADMs to the BSs typically on two separate unidirectional links. One of the main benefits of a ring architecture compared to a star is the potential for fault restoration using a second protection ring, allowing a fibre break between nodes or a node failure to be bypassed [48, 49]. A fault along the WDM link of a star network, on the other hand, leads to all BSs losing traffic.
Compared to a star network that uses only a single optical component for multiplexing / demultiplexing, the ring network uses as many OADMs as there are BSs. This has immediate implications for a passive fibre-radio network in which individual BSs are sent one wavelength each. Since the optical power budget is limited and the OADM loss is typically 3 dB or more [39, 50], having one OADM per BS is impossible. Not only would the received optical power quickly drop below the minimum required (-19 dBm in our case), but the cost associated with having one OADM per BS would be excessive. One of the most important observations is that if a passive fibre-radio ring network is to be feasible, groups of wavelengths rather than individual wavelengths must be dropped and added by individual OADMs. Two options are available to the network designer, as illustrated in Figure 6.14. Firstly, to add and drop a wavelength band at each OADM and then use a subsequent remote AWG to demultiplex / multiplex (MUX / DEMUX) individual wavelengths to / from individual BSs. Secondly, an AWG at the ring node itself can be used, to combine the add/drop functionality and the individual wavelength MUX / DEMUX function.

Fig. 6.14 Two potential options for adding & dropping wavelength bands.

Unless ring nodes are physically located in areas that are too far from the actual area in which the wavelength band BSs are located, it is logical and cost-effective to use the latter approach, i.e. to use a single optical component.

There is no real advantage in having a bidirectional arrangement, as shown in Figure 6.15, which is essentially a bus network as described in Section 6.5.
Fig. 6.15 Ring topology showing bidirectional transmission along the ring.

In the unidirectional configuration the total number of channels in the ring remains constant, with the uplink channels replacing the downlink channels at each node, whereas in a bidirectional ring there are only $N-n+1$ channels left in the fibre span to the $n$th node. The comments regarding additional component requirements and problems due to Rayleigh backscatter made in the previous section also apply here.

The main problem with a passive ring network is the cumulative component loss and channel power differences along the ring lead to potentially high crosstalk levels. This will be demonstrated in the following section, which uses the power budget presented in Section 6.2.2 to design a WDM fibre-radio ring network.

### 6.4.2 Ring Network Design Example

A WDM fibre-radio star network is shown in Figure 6.16, including CO and BS optical components and realistic component losses.
Fig. 6.16 CO and BS bloc diagram for a WDM ring network.

The ring network shown in Figure 6.16 uses AWGs in the ring to drop and add multiple wavelengths at each node, distributing them to multiple BSs. The CO is identical to that described for the star network, comprising multiple lasers and external modulators and an AWG to multiplex all wavelengths into the one fibre. Each node within the ring is separated by a distance $L_1$, and the longest distance from AWG node to BS is $L_2$. Each BS comprises a laser and external modulator, using a separate fibre for the uplink path to the AWG. The uplink channels are added back into the ring by the AWG at each node, and go through the remaining nodes before reaching the CO. At the CO, upstream wavelengths are demultiplexed by an AWG and individual wavelengths detected by individual PIN diodes. It may be possible to demultiplex wavelengths into groups in which different subcarrier frequencies are used, so that one PIN diode would be used per group. If 10 different sets of frequencies were used, this would allow a reduction in the number of required PIN diodes by a factor of 10, which may be a significant cost saving.

(A) Optical Power Budget

The link budget for downlink and uplink paths are given by:

$$P_{BS}(n) = T_{x_{CO}} - L_{Mod} - L_{AWG_1} - n\alpha L_1 - nL_{AWG_2} - \alpha L_2$$  \hspace{1cm} (6.32)

and

$$P_{CO}(n) = T_{x_{BS}} - L_{Mod} - \alpha L_2 - (NN - n + 1)L_{AWG_2} - (NN - n + 1)\alpha L_1 - L_{AWG_1}$$  \hspace{1cm} (6.33)
where $P_{BS}(n)$ is the optical power at the BS served by the $nth$ ring node, $P_{CO}(n)$ is the optical power at the CO from the BS served by the $nth$ node, $Tx_{CO}$ and $Tx_{BS}$ are the laser powers at the CO and BS respectively, $L_{Mod}$ is the modulator loss including 3 dB bias, $L_{AWG1}$ is the AWG loss at the CO, $L_{AWG2}$ is the loss of the node AWGs, $\alpha$ is the fibre attenuation in dB/km and $L_1$ and $L_2$ are node to node and AWG to BS fibre lengths respectively, in km, and $NN$ is the total number of nodes in the ring. Equations 6.34 and 6.35 show the dependence of both downlink and uplink powers on the node number $n$ and the total number of nodes $NN$.

AWG losses are assumed to be 5 dB [36, 38, 39] and both $L_1$ and $L_2$ are set at 5 km. The output power per channel at the CO is set at 0 dBm, so that $Tx_{CO}$ is +12 dBm. $L_{Mod}$ is taken to be equal to 7 dB. The received optical power at the furthest BS served by the $nth$ node is given by:

$$P_{BS}(n) = -1 - 6n$$
$$= -19 \text{ dBm} \quad (6.34)$$

and is equal to −19 dBm, the minimum received optical power, for the last node. Equation 6.34 shows that in order to meet the minimum received optical power at the BS the number of nodes in the ring network is limited to three. This number depends on fibre lengths and node losses, but further reducing component losses and fibre lengths would still severely restrict the maximum number of nodes in the network.

The lowest uplink channel power is for the channels added at the first node in the ring, $P_{CO}(1)$. This is given by:

$$P_{CO}(1) = Tx_{BS} - 13 - 6NN$$
$$= -19 \text{ dBm} \quad (6.35)$$

If $Tx_{BS}$ is equal to $Tx_{CO}$ i.e. +12 dBm, then the worst-case downlink and uplink are identical, so that $NN$ equal to three is also the maximum number of nodes in the ring.

In the downlink, the power of the dropped channels depends on the node number, with the last node having the lowest drop power. Similarly for the uplink, the weakest channels will be those that were added into the ring at the first node, with the strongest channels coming from the last node. This is shown in Figure 6.17.
Fig. 6.17 Three node ring network and add, BS and CO received channel powers for different nodes.

Figure 6.17 shows the reduction in BS received optical power as more nodes are traversed. The optical power received at the CO is lowest for the channels added at the first node, and strongest from the third and final node. The power level of added channels at the output of each node is also shown. It is constant as the BSs are using the same output powers.

(B) Optical Crosstalk Calculation

While it is important to meet the optical power budget when designing a network, optical crosstalk in a ring network is also critical, particularly as channel powers are unequal. This leads to potentially high effective crosstalk levels. Figure 6.18 shows where out-of-band crosstalk occurs in a ring network.

Fig. 6.18 Out-of-band optical crosstalk in a ring network.
If we first consider the downlink, out-of-band crosstalk at the BS is determined by the AWG out-of-band crosstalk and the total number of WDM channels in the ring, taking into account their respective powers. The optical power of drop channels, $P(d\text{rop } n)$ at the input of the $n$th node is given by:

$$P(d\text{rop } n) = Tx_{CO} - L_{Mod} - L_{AWG1} - n\alpha L_1 - (n - 1)L_{AWG2}$$

$$= -6n + 5 \quad (6.36)$$

and the power of the add channels from the $m$th node at the input to the $n$th node, $P(\text{add } m,n)$ is given by:

$$P(\text{add } m,n) = Tx_{BS} - \alpha L_2 - (n - m)(L_{AWG2} + \alpha L_1)$$

$$= 11 - 6(n - m) \text{ for } n > m \quad (6.37)$$

As we have three nodes in our example, we can find the power levels of the three wavelength bands at the input to the three nodes, as shown in Figure 6.19.

---

**Fig. 6.19 Optical powers of each band for the three node ring network. Optical powers are wavelength band powers at the input of the nth node.**

Figure 6.19 shows that at the input of node 1, all wavelength bands have equal powers. At the input of node 2, bands 2 and 3 have experienced an additional 6 dB loss, whilst band 1 has a higher power level as the band was added at node 1. Similarly at node 3, where band 3 still hasn’t been dropped, experiencing an additional 6 dB loss as for band 1, whereas band 2 has a higher power as it was added at node 2.

For a 0.5 dB power penalty, the out-of-band crosstalk at node $n$, $PP(n)$, for a linear crosstalk ratio of $x$ and $N$ WDM channels for the BSs served by each node satisfies:

$$PP(1) = 1 + 2x \frac{N}{3} = 10^{0.5/10}, \quad (6.38)$$
\[ PP(2) = 1 + (10^{+12/10} + 1) \frac{N}{3} = 10^{0.5/10}, \]  
\[ PP(3) = 1 + (10^{12/10} + 10^{18/10}) \frac{N}{3} = 10^{0.5/10} \]  

where \( 10^{N/10} \) is the ratio of the crosstalk to signal power where \( \Delta \) is the difference between crosstalk and signal bands in dB. Equations 6.38 to 6.40 show that the effective out-of-band crosstalk is different at different nodes. For a 0.5 dB power penalty, the crosstalk requirement at node \( n \), \( x(n) \) in dB is given by:

\[ x(1) = -7.4 - 10 \log N, \]  
\[ x(2) = -16.6 - 10 \log N, \]  
\[ x(3) = -23.3 - 10 \log N. \]  

The required out-of-band crosstalk requirements given by Equations 6.41 to 6.43 are given in Table 6.8 for 100 and 200 WDM channels respectively.

**Table 6.8 Out-of-band crosstalk requirements at BSs in nodes 1 to 3.**

<table>
<thead>
<tr>
<th>Node</th>
<th>BS out-of-band crosstalk requirement for ( N=100 )</th>
<th>BS out-of-band crosstalk requirement for ( N=200 )</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>-17.4 dB</td>
<td>-20.4 dB</td>
</tr>
<tr>
<td>2</td>
<td>-26.6 dB</td>
<td>-29.6 dB</td>
</tr>
<tr>
<td>3</td>
<td>-33.3 dB</td>
<td>-36.3 dB</td>
</tr>
</tbody>
</table>

Table 6.8 show that the out-of-band crosstalk specification for the node AWG becomes much more stringent at further nodes, as the difference between the dropped channel and the other channels increases, with the previously added channels having higher relative powers.

If the node OADM only adds and drops the wavelength band without demultiplexing / multiplexing channels, then an additional MUX / DEMUX would follow the OADM. This would increase the effective out-of-band crosstalk of other wavelength bands, so that the dominant source of out-of-band crosstalk may only be due to the node wavelength band, allowing individual component crosstalk levels to be relaxed.
Chapter 6

If we now consider out-of-band crosstalk at the CO, as shown in Figure 6.18, received channel powers vary according to the originating node (see Figure 6.19). The power difference between adjacent uplink channel bands is equal to the combined loss of the AWG and the fibre span, i.e. it is equal to 6 dB. Hence for our three-node ring, the three separate wavelength bands will differ by a maximum of 12 dB. The total out-of-band power penalty is equal to the total optical power of out-of-band channels. The worst-case occurs for the wavelength band with the lowest optical power, from node 1, whereas the best case occurs for the last wavelength band, from node 3, which has the highest optical power. Similarly to Equations 6.41 to 6.43, the total power penalty $PP(n)$ due to out-of-band crosstalk for the three wavelength bands is:

$$PP(1) = 1 + \left(10^{6/10} + 10^{12/10}\right)x \frac{N}{3} = 10^{0.5/10},$$

(6.44)

$$PP(2) = 1 + \left(10^{-6/10} + 10^{6/10}\right)x \frac{N}{3} = 10^{0.5/10},$$

(6.45)

$$PP(3) = 1 + \left(10^{-6/10} + 10^{-12/10}\right)x \frac{N}{3} = 10^{0.5/10}$$

(6.46)

where $PP$ is the linear power penalty, $x$ is the linear component out-of-band crosstalk level, $N$ is the total number of WDM channels, and $10^{0.05}$ is the linear value corresponding to a 0.5 dB power penalty. Rearranging for $x$,

$$x(1) = -17.3 - 10\log N,$$

(6.47)

$$x(2) = -10.6 - 10\log N,$$

(6.48)

$$x(3) = 0.7 - 10\log N.$$  

(6.49)

Table 6.9 shows the required out-of-band crosstalk at the CO for the three different wavelength bands coming from the three nodes in the ring.

Table 6.9 Out-of-band crosstalk requirements at CO for bands 1 to 3.

<table>
<thead>
<tr>
<th>Band</th>
<th>CO out-of-band crosstalk requirement for $N=100$</th>
<th>CO out-of-band crosstalk requirement for $N=200$</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>-37.3 dB</td>
<td>-40.3 dB</td>
</tr>
<tr>
<td>2</td>
<td>-30.6 dB</td>
<td>-33.6 dB</td>
</tr>
<tr>
<td>3</td>
<td>-19.3 dB</td>
<td>-22.3 dB</td>
</tr>
</tbody>
</table>
Table 6.9 show that out-of-band crosstalk levels at the CO AWG must be less than -37 dB for 100 WDM channels and less than -40 dB for 200 WDM channels. The maximum crosstalk levels assuming equal powers would be –27.4 dB and –30.4 dB for 100 and 200 channels respectively. The 6 dB power difference between each node results in an 18 dB difference between the three wavelength bands’s maximum out-of-band crosstalk requirements. This is three times the 6 dB power difference and is a general result. This can be proven as follows.

If we consider the optical powers from consecutive nodes, where $P$ is the transmit power, $L$ is the loss between nodes and $NN$ is the number of nodes, we have the following geometric series:

$$P = PL, PL^2, PL^3, \ldots, PL^n, \ldots, PL^{NN}$$  \hspace{1cm} (6.50)

The total power, $P_{(total)}$ is given by the sum of the geometric series, which is given by:

$$P_{(total)} = \frac{PL[1-L^{NN}]}{(1-L)}$$ \hspace{1cm} (6.51)

If we are detecting channel $n$, then the power penalty $PP(n)$ due to power addition for an out-of-band crosstalk ration $x$ is given by:

$$PP(n) = \frac{PL^n + x(P_{(total)} - PL^n)}{PL^n} = 1 + x \left( \frac{L^{1-n}(1-L^{NN})}{(1-L)} - 1 \right)$$ \hspace{1cm} (6.52)

The ratio of best-case to worst-case crosstalk level $x$ for a given power penalty is given by:
\[
\begin{aligned}
    x(\text{best}) &= \left( \frac{P(\text{total}) - PL^{NN}}{PL^{NN}} \right) = \left( \frac{P(\text{total}) - PL}{P(\text{total}) - PL} \right) L^{NN-1} \\
    &= \frac{PL(1 - L^{NN})}{(1 - L)} - PL^{NN} \\
    &= \left( \frac{PL(1 - L^{NN})}{(1 - L)} - PL \right)^{L^{NN-1}} \\
    &= \frac{PL - PL^{NN+1} - PL^{NN} + PL^{NN+1}}{(PL - PL^{NN+1} - PL + PL^2) L^{NN-1}} \\
    &= \frac{L - L^{NN}}{L^{NN+1} - L^2NN} \\
    &= \frac{L}{L^{NN+1}} = L^{-NN}
\end{aligned}
\]

Equation 6.53 shows that the difference between the best case and the worst case is \(-NN\).L where \(NN\) is the total number of nodes and \(L\) is the loss of node plus fibre in dB. Even if the optical power budget could be met for more than three nodes, the out-of-band crosstalk specification would become very stringent. Although this could be achieved by cascading multiple AWGs, this would increase the loss and the cost at the CO.

We now consider in-band crosstalk in the ring network, as shown in Figure 6.20.

**Fig. 6.20 In-band optical crosstalk in a ring network.**
Figure 6.20 shows that in-band crosstalk at the BS can arise due to coupling of the ‘add’ wavelength back into the ‘drop’ wavelength path via the AWG [40, 41]. As for the star network, the difference in channel power between the ‘drop’ and ‘add’ channels determines the effective crosstalk level. The effective in-band crosstalk at the BS, \( x(\text{up} - \text{down}) \) is given by:

\[
x(\text{up} - \text{down}) = (T_{BS} - L_{\text{Mod}} - \alpha L_2) \\
- (T_{CO} - L_{\text{Mod}} - L_{\text{AWG1}} - n\alpha L_1 - nL_{\text{AWG2}}) + x(\text{in-band}) \\
= 4 + 6n + x \\
= -18 \text{ dB maximum}
\]

(6.54)

and is equal to \(-18 \text{ dB}\) for a 0.5 dB power penalty. Equation 6.54 shows that as the node number increases, the drop power reduces and hence the effective in-band crosstalk level increases. For the three-node ring, the last node has an effective in-band crosstalk level that is 22 dB above the AWG in-band crosstalk level. This means that the maximum component in-band crosstalk level should be \(-40 \text{ dB}\). This is a rather stringent requirement. Currently available devices have in-band crosstalk levels of \(-30\) to \(-40 \text{ dB}\) [36-39], although lower crosstalk AWGs have been reported [51-53].

If we now consider in-band crosstalk at the CO, we must focus on what occurs at each node in the ring. Recall that a wavelength band is dropped at the ring in the downlink and the same wavelength band is added back into the ring in the uplink. As the optical component that performs this will not be perfect, there is some residual ‘drop’ signal that continues through the ring and is combined with the desired ‘add’ signal. Since the power of the added channels at the output of the AWG, \( P(\text{add}) \) is given by:

\[
P(\text{add}) = T_{BS} - L_{\text{Mod}} - \alpha L_2 - L_{\text{AWG2}}
\]

(6.55)

and the power of the drop channel at the \( n \)th node at the input of the AWG, \( P(\text{drop}) \) is given by:

\[
P(\text{drop}) = T_{CO} - L_{\text{Mod}} - L_{\text{AWG1}} - n\alpha L_1 - (n-1)L_{\text{AWG2}}
\]

(6.56)

the effective in-band crosstalk for the channels coming from the \( n \)th node, \( x(\text{down-up}) \) is given by:
\[ x(\text{down-up}) = P(\text{drop}) - P(\text{add}) + x(\text{in-band}) \]
\[ = T{x}_{CO} - L_{AWG1} - n\alpha L_1 - (n - 1)L_{AWG2} - T{x}_{BS} + \alpha L_2 + x \]  
(6.57) 
\[ = 6 - 6n + x \]
\[ = -18 \text{ dB maximum} \]

Equation 6.57 shows that as the node number \( n \) increases, the effective crosstalk decreases. Hence the maximum AWG in-band crosstalk for nodes 1, 2 and 3 is equal to –18 dB, -12 dB and –6 dB respectively. This should be compared to corresponding values of –28 dB, -34 dB and –40 dB for the uplink-downlink in-band crosstalk. Figure 6.21 illustrates the varying effective crosstalk levels for two different fixed component in-band crosstalk levels: -18 dB for downlink-uplink crosstalk and –40 dB for the uplink to downlink crosstalk.

![Effective in-band crosstalk levels at different nodes.](image)

### 6.4.3 Summary

The first problem encountered with a ring architecture is the cumulative loss incurred by channels as the number of nodes traversed increases. This limits the number of nodes in a passive fibre-radio network and immediately suggests that wavelength bands / groups must be added and dropped at each node, not individual wavelengths. Given the optical power of Section 6.2.1 this limits the total number of nodes to three.

The second consequence of having multiple nodes is that the power of dropped bands decreases as the node number increases, while the power received at the CO
increases as the node number increases. These differences in optical power have potentially catastrophic consequences for optical crosstalk if this is not taken into account. Out-of-band crosstalk at the BS varies according to the node number, due to the combination of channel bands which have yet to be dropped and channels which have already been dropped and added back into the ring channels, being worst at the last node. Similarly in-band crosstalk due to leakage of the ‘add’ channels to the downlink path increases as the ‘drop’ power decreases with node number, limiting the number of nodes that can be traversed. At the CO, the level of out-of-band crosstalk depends on which band of wavelengths is considered, so that more stringent component crosstalk levels are required. In-band crosstalk arising at the node is worst at the first node, since the power of dropped channels is highest. Hence in-band crosstalk at the CO can be limited by ensuring appropriate downlink-uplink crosstalk at the first node. Table 6.10 shows the required component crosstalk levels for our design example.

Table 6.10 Optical component crosstalk levels for the three node ring network.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Crosstalk (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>CO out-of-band</td>
<td>-37.3 for 100 channels</td>
</tr>
<tr>
<td>BS out-of-band</td>
<td>-33.3 for 100 channels</td>
</tr>
<tr>
<td>In-band downlink-uplink</td>
<td>-18</td>
</tr>
<tr>
<td>In-band uplink-downlink</td>
<td>-40</td>
</tr>
</tbody>
</table>

In-band crosstalk specifications assume that different downlink and uplink wireless frequencies are used within each wireless cell, minimising the effect of in-band crosstalk. Note that while out-of-band component crosstalk specifications depend on the number of WDM channels, in-band crosstalk is due to a single optical channel only. The above values are specific to our design and would be worse if the number of nodes and / or the loss between nodes was increased. Absolute values also depend on relative ‘add’ and ‘drop’ channel powers.

The final conclusion is that if more nodes are required to increase coverage then optical amplifiers must be placed in the network to boost signal levels within the ring back to acceptable levels. Although optical amplifiers are typically placed at the output
of each node so as to compensate for the previous node and fibre losses, a cheaper alternative is to place an optical amplifier at the output of the last node in which optical power levels are acceptable. These two options are shown in Figure 6.22.

![Diagram of two configurations of optical amplifiers in a ring network: (a) 1 EDFA per node and (b) 1 EDFA every 3 nodes.]

**Fig. 6.22 Optical amplifiers in a ring network: (a) amplifiers at each node and (b) amplifiers every three nodes.**

In our example, which for a passive network is limited to three nodes, this means placing amplifiers at the output of every third node in the ring, i.e. at nodes 3, 6, 9 etc. as in Figure 6.22 (b). The only limiting factor that has to be considered is the relative level of amplifier ASE noise, taking into account the relative channel power levels. However, this is not a real concern as no more than 10 to 20 nodes at most would be present in a fibre-radio network. A more relevant question perhaps is whether or not using optical amplifiers in a fibre-radio network is cost-effective. This is not further discussed as this is not an engineering question, but it can not be ignored when designing real networks.

While the ring network can offer redundancy in case of failure if designed with a loopback feature, this is not the case for a star network. In terms of engineering design, the star architecture is much simpler.

### 6.5 Bus Network Architecture Considerations

This section very briefly discusses the bus network architecture. As a bus network suffers from the same problems associated with cascaded nodes as a ring network, no design example is necessary.
6.5.1 Bus Network Architecture

A generic bus network architecture is shown in Figure 6.23.

The bus topology shown in Figure 6.23 is very similar to the ring network discussed in the previous section. Wavelengths or wavelength bands are dropped at various nodes along the bus in the downlink direction, with wavelengths being sent back to the CO either along the same fibre using a bidirectional arrangement, or using a separate fibre [4, 54-56]. Whereas the ring network starts and finishes at the CO, the end of the bus network can terminate anywhere, offering more potential flexibility. If separate unidirectional links are used, then the number of WDM channels along a link varies as wavelengths / wavelength bands are dropped / added along the link.

(A) Optical Power Budget

The bus network suffers from the same problem as the ring when it comes to the optical power budget, since at each successive node an additional fibre loss and component loss is incurred, which quickly limits the number of nodes in the network. This problem can be eliminated using optical amplification, although it increases the cost of the network significantly.

(B) Optical Crosstalk

If two separate unidirectional bus links are used for the downlink and the uplink, then in-band crosstalk is not present, so that only out-of-band crosstalk must be considered, as shown in Figure 6.23. At the BS, all remaining channel powers are
equal, so calculating the crosstalk specification is simple. Since the first node will have most WDM channels, it will suffer from the worst out-of-band crosstalk level. At the CO, however, as for the ring, channels originating from different nodes will have different received optical powers, which can quickly increase the effective crosstalk seen by lower channels. If a bidirectional link is used, then both uplink and downlink channels are dropped and added at the node, so that in-band crosstalk can occur. Rayleigh backscattering is also likely to be a problem, as discussed in Section 6.3.2. As for the ring, downlink-uplink crosstalk is worst at the first node, while uplink-downlink crosstalk is worst at the final node. Actual component crosstalk requirements and power budgets can be calculated following the same procedure used for the ring network in Section 6.4.2.

6.5.2 Summary

The bus network architecture is similar to the ring architecture in that channel powers vary along the link. Physically, however, the end of the bus network can be placed anywhere, whereas a ring terminates at the CO. This means the bus network node layout can be more flexible, with the potential to add additional nodes at the end of the bus. The bus network can be unidirectional if two separate links are used for the downlink and uplink, or bidirectional if a single link is used. As for a ring network, optical amplifiers may be used to eliminate the constraint due to optical power and optical crosstalk, provided ASE noise level is sufficiently low so as not to degrade signal quality.

6.6 COMPLEX ARCHITECTURES

This section illustrates some of the more complex network architectures that may be considered for fibre-radio networks. Their relative merits are discussed briefly.

6.6.1 Multiple Star

If we extend the concept of a star network to include further stars we obtain a network schematic diagram as shown in Figure 6.24 [57], which is also called a tree topology [32].
(A) Optical Power Budget

Figure 6.24 shows that a link from the CO to a remote WDM allows traffic to be routed into three separate wavelength bands. Each output of the WDM is then connected to a remote AWG where the wavelengths in the band are demultiplexed to individual BSs. Optical amplifiers are shown, compensating additional component and fibre losses. Given the additional requirement of the WDM and EDFAs, it may be more cost effective to use three separate star networks, each originating at the CO. This would also allow each separate star network to fully utilise the available optical spectrum, although physical space may limit the total number of WDM channels used at the CO. Note that $N$ WDM lasers would be required for this case, with a total of $3N$ modulators for the three separate star networks. The optical power budget can be expressed as follows:

$$P_{RxB} = P_{TxCO} + L_{Mod} + L_{AWG1} + L_{MUXTotal} + aL_{Total}$$  \hspace{1cm} (6.58)

and

$$P_{RxCO} = P(BS) + P_{TxBS} - P_{TxCO}$$  \hspace{1cm} (6.59)

where $P_{RxB}$ and $P_{RxCO}$ are the received optical powers at the BS and CO respectively, $L_{MUXTotal}$ is the total loss due to WDM MUX / DEMUX in the link (i.e. WDM+AWG above) and $L_{Total}$ is the total fibre length between CO and the furthest BS and other symbols are as defined previously. Using the same component values as in Section 6.3.2 we obtain:
\[ P_{Rx} = -12 + 10 \log_{10} d_{\text{Total}} = -19 \text{ dBm} \] (6.60)

for two AWGs in the link, so that for our specific power budget we are limited in this case to a total fibre length of 35 km. This compares to 60 km maximum for a single AWG in the link.

(B) Optical Crosstalk

From an optical crosstalk point of view, the only benefit of a multiple star network is the fact that two stages of filtering are occurring for unwanted channels, firstly at the wavelength band stage, then on individual wavelengths. This means that the effective out-of-band crosstalk for channels from other bands can potentially be ignored. This approach is sometimes referred to as “wavelength dilation” [32] or “waveband routing” [58].

6.6.2 Multiple Ring

If reliability is a concern, a ring network may be favoured, with the option of having secondary rings for each wavelength band, as shown in Figure 6.25. This is referred to as hierarchical rings in the literature [59].

![Multiple ring fibre-radio network](image_url)

Fig. 6.25 Multiple ring fibre-radio network.

(A) Optical Power Budget

If the multiple ring topology has \( X \) nodes in the main ring, and \( Y \) nodes within each secondary ring, \( XY \) OADMs are required. The losses associated with the ring means that optical amplification would be required, although optical amplifiers are not
shown in the figure. These could be incorporated into the main ring OADMs, but may also be required within secondary rings depending on losses. While this topology is popular in metropolitan and larger networks [45, 60], it is not suitable for a low-cost fibre-radio network.

(B) Optical Crosstalk

Optical crosstalk is a problem in rings if unequal channel powers are present, as discussed in Section 6.4. This quickly results in unrealistic component crosstalk requirements, limiting the size of the network. If optical amplifiers are used in the network, however, channel powers differences can be minimised, with realistic component specifications.

6.6.3 Hybrid Ring-Star

Figure 6.26 shows a more favourable architecture that combines the benefits of a main ring in which fault restoration is possible, together with a star architecture for each wavelength band.

![Hybrid Ring-Star Diagram](image)

*Fig. 6.26 Hybrid ring-star fibre-radio network.*

Figure 6.26 shows two possible alternatives, depending on whether a wavelength band is dropped then demultiplexed, or directly demultiplexed at the ring node. The latter minimises component requirements and total loss. The problems associated with power budget and optical crosstalk within the main ring remain, however, implying that optical amplification may be required if more than a few ring nodes are present. Refer to
Section 6.4 for further details. Note that this was the effective topology discussed in Section 6.4 since dropping and adding a single wavelength in a fibre-radio network is impractical, due to the large number of components and optical amplifiers required in such a situation.

6.6.4 Summary

This section has briefly discussed three different types of network architecture, which are more complex than the simple star / ring / bus networks. The simplest topology involves multiple stars (also referred to as a star-tree or tree architecture), which is a natural extension to a star, maintaining the simplicity in terms of power budget and optical crosstalk. A second topology involving multiple rings exhibits the same disadvantages as a ring network, mainly due to the high cumulative losses due to the multiple OADMs in the rings, requiring optical amplification. Potential crosstalk problems due to channel power differences have also been highlighted. Finally, if the benefits of a ring network are required (namely in terms of fault restoration), then a ring-star topology can be adopted, as discussed in Section 6.4. This minimises the losses as a single AWG is used at each node to drop and demultiplex channels to individual BSs. However, problems inherent to the main ring remain.

6.7 CONCLUSIONS

This chapter has analysed traditional star, ring and bus networks in a fibre-radio context in Sections 6.3, 6.4 and 6.5 respectively. This has been done not only from a theoretical point of view but also with practical network design examples based on optical and wireless power budgets presented in Section 6.2. More complex architectures that may be considered for fibre-radio networks were briefly discussed in Section 6.6. This has highlighted the relative merits of the different topologies when designing a WDM fibre-radio network.

The following conclusions have been obtained:

- The star architecture is the simplest and most cost-effective architecture:
  1. A single WDM MUX / DEMUX is required in the link.
  2. Losses are kept to a minimum.
3. Optical amplifiers are not required.
4. Design for the longest link guarantees other link budgets are met.

- A ring architecture provides potential fault restoration but:
  1. Cumulative loss due to multiple nodes is a problem.
  2. Channel powers vary along the ring, increasing the effective crosstalk.
  3. Optical bands and not individual wavelengths must be dropped / added for a cost-effective and passive ring.
  4. Optical amplifiers will be required for rings with more than a few nodes in order to restore optical powers and avoid excessive crosstalk.
  5. In-band crosstalk between uplink and downlink becomes excessive, limiting the number of nodes in a passive ring.
  6. Out-of-band crosstalk specifications must be correctly calculated as channel powers vary.
  7. A single AWG at each node can drop and demultiplex wavelengths simultaneously to BSs served by each node.

- A bus architecture is similar to a ring architecture due to the cumulative effect of component losses and the problems associated with power budgets and crosstalk.

- More complex architectures are possible but:
  1. Component losses become more significant due to the use of more components.
  2. Network engineering complexity increases.
  3. Optical amplifiers are required to compensate for additional component losses.
  4. Network cost increases due to the extra components and optical amplifiers required.
  5. The relevance of using such networks in fibre-radio is questionable, both from an engineering and cost point of view. Simpler architectures can be passive and are simpler to design.
These comments are well-established for traditional baseband optical networks. The same comments apply when considering a WDM fibre-radio network.

The following observations about optical crosstalk can be made:

1. In-band crosstalk in fibre-radio is minimised by using different uplink / downlink RF frequencies.
2. In-band crosstalk can be eliminated by using separate downlink / uplink networks.
3. In-band crosstalk between downlink and uplink can be eliminated by using different downlink / uplink wavelengths, but this may increase component complexity / cost.
4. In-band crosstalk due to Rayleigh backscatter in bidirectional links can dominate over component-induced crosstalk.
5. Out-of-band crosstalk produces a penalty due to excess optical power only, which can potentially be tolerated without changing signal power levels (leading to no real power penalty).

Point number 1 is specific to fibre-radio networks as the impact of in-band crosstalk can be minimised through appropriate frequency allocation. The remaining points apply equal well to baseband optical networks.

Table 6.11 summarises the main conclusions for different network architectures, including unidirectional and bidirectional networks. ‘Down’ and ‘Up’ refer to downlink and uplink respectively.
Table 6.11 Comparison of network architectures.

<table>
<thead>
<tr>
<th></th>
<th>UNIDIRECTIONAL</th>
<th></th>
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<th></th>
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</thead>
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<tr>
<td></td>
<td>STAR</td>
<td>RING</td>
<td>BUS</td>
<td>STAR</td>
</tr>
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<td>Downlink / uplink λs</td>
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<td>N / N</td>
<td>N / N</td>
<td>N / N</td>
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<td>N</td>
<td>N-1</td>
<td>1</td>
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<tr>
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<td>-</td>
<td>-</td>
<td>N+1</td>
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<td>YES</td>
<td>NO</td>
</tr>
<tr>
<td>λ bands</td>
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<td>YES</td>
<td>NO</td>
</tr>
<tr>
<td>In-band xtalk CO</td>
<td>NO</td>
<td>Down / Up Unequal</td>
<td>NO</td>
<td>Down / Up Unequal</td>
</tr>
<tr>
<td>In-band xtalk BS</td>
<td>NO</td>
<td>Up / Down Unequal</td>
<td>NO</td>
<td>Up / Down Unequal</td>
</tr>
<tr>
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<td>NO</td>
<td>NO</td>
<td>YES</td>
</tr>
<tr>
<td>Out-of-band xtalk</td>
<td>YES</td>
<td>YES Unequal</td>
<td>YES Unequal</td>
<td>YES</td>
</tr>
</tbody>
</table>

This chapter has applied results and conclusions obtained for all previous chapters to provide a comprehensive overview of practical design considerations for WDM fibre-radio networks. The calculation of downlink and uplink channel powers at various points in the networks has been explained, highlighting the importance of cumulative losses in ring networks. Analytical techniques have been presented which allow both in-band crosstalk and out-of-band crosstalk component specifications to be calculated. Practical observations and recommendations have been made for the design of future WDM fibre-radio networks. One important observation is that provided optical power budgets and crosstalk specifications have been calculated for a fibre-radio network using subcarrier modulation, the design implications of various optical networks apply equal well to fibre-radio and baseband networks. The following chapter provides final conclusions and suggests future work.
6.8 REFERENCES


7

Conclusions

7.1 SUMMARY

An investigation into optical crosstalk in WDM fibre-radio networks has provided insight into important considerations that must be taken into account when designing such networks. Chapter 2 provided a literature review of optical crosstalk and fibre-radio networks, highlighting the need for research into the analysis of optical crosstalk in fibre-radio networks. In Chapter 3, an analytical model was used to evaluate in-band and out-of-band crosstalk for BPSK, ASK and QPSK modulation schemes, taking into account the RF frequency domain. The experiments discussed in Chapter 4 confirmed analytical results, which were also validated by computer simulation. The implications of these results on the design of fibre-radio networks were illustrated in Chapters 5 and 6, through the consideration of tolerable component crosstalk specifications and their impact on different network topologies. Chapter 6 also illustrated the design of a fibre-radio network by considering different topologies using realistic optical and wireless power budgets for a high-capacity wireless system operating at 20 GHz. The outcomes of the research reported provides insight into the nature of optical crosstalk impairments, analytical tools for simplified analysis of such networks, as well as valuable information for those who wish to design and implement future WDM fibre-radio networks.

The main contributions that have arisen as a result of the research undertaken are as follows:

- Demonstration of the relationship between the RF modulation schemes and the system power penalties arising from in-band and out-of-band crosstalk (Chapters 3 & 4).
• Experimental investigation of in-band and out-of-band crosstalk for BPSK modulation, including the effect of crosstalk RF frequency allocation (Chapter 4).

• Validation of analytical results for ASK, BPSK and QPSK modulation using simulation software (Chapter 4).

• Investigation of the component crosstalk requirements for a WDM fibre-radio network comprising both in-band and out-of-band crosstalk (Chapter 5).

• Evaluation of WDM fibre-radio network capacity for typical component crosstalk levels for a 1 dB system power penalty (Chapter 5).

• Investigation of the impact of different wireless frequency re-use plans on component crosstalk specifications (Chapter 5).

• Crosstalk and power budget analysis of various WDM fibre-radio network topologies, based on a realistic high-capacity wireless link budget (Chapter 6).

The following conclusions have arisen as a result of the work summarised above:

• Optical power penalties for subcarrier-modulated links are dependent on modulation format, receiver structure, and the RF frequency allocation (Chapters 3 & 4).

• If in-band crosstalk is at the same RF frequency as the crosstalk signal, the power penalty will depend on the RF phase difference between signal and crosstalk carriers, so that the network must be designed for worst-case (Chapters 3 & 4).

• The effect of in-band crosstalk can be significantly reduced if the crosstalk channel carries data at a different RF frequency to the signal – this will occur in practice between downlink and uplink channels (Chapters 3 & 4).
Conclusions

• For networks in which the RF frequency re-use factor is comparable to the number of WDM channels, the out-of-band crosstalk penalty will be reduced compared to a network using a single frequency band (Chapter 5).

• Out-of-band crosstalk for a WDM network is equal to optical power addition alone and is therefore unaffected by the RF frequency allocation (Chapter 5).

The following conclusions on network topologies (Chapter 6) are general conclusions that are already established for baseband optical networks and was verified in this thesis as applicable to WDM fibre-radio networks:

• The star architecture is the simplest from an engineering perspective.

• The ring architecture severely limits the size of a passive network due to optical losses.

• Cumulative losses in a ring result in high effective crosstalk due to channel power differences, both for in-band and out-of-band crosstalk.

• Ring, bus and more complex architectures will require optical amplification and careful design if unequal channel powers are present.

For a WDM fibre-radio application, the following conclusions apply:

• Ring nodes must add / drop wavelength bands, otherwise OADM component costs will be excessive.

• A main ring, with each node feeding local BSs in a star, offers the benefits of fault-restoration at minimum cost.

• Cost considerations imply that the star architecture is preferred for a cheap WDM fibre-radio access network.
7.2 SUGGESTIONS FOR FUTURE WORK

The work described in this thesis [1-3] as well as previous work [1-3] demonstrated that the different modulation formats were found to have different levels of tolerance to optical crosstalk. Given that most future wireless systems will deploy higher-order modulation schemes such as QAM [3], it will be beneficial to quantify the crosstalk-induced power penalties for such modulation schemes in subcarrier modulated fibre-radio networks.

Another related issue is the potential for multiple in-band crosstalk terms. While this is not likely to occur in fibre-radio systems in which a wavelength is dropped and added only once within the network, perhaps future architectures involving interconnected networks may result in more than one in-band crosstalk source. While the Gaussian approximation can be applied for a large number of terms, it may be relevant to investigate the case of two to eight crosstalk terms using a more exact analysis. Recently, much more elaborate WDM schemes such as wavelength re-use [4] and wavelength interleaving [5, 6] have been proposed. It would be important to study the effect of optical crosstalk in such schemes, where multiple interferers may be present.

One of the factors that has been mentioned in Chapter 6 is the issue of channel power and optical nonlinearities, particularly when a large number of WDM channels are present. While some work has been done for analogue subcarrier modulation, most considered low frequencies (up to 1 GHz) [7-10], although frequencies up to 18 GHz are considered in [11]. The potential for nonlinear effects between WDM channels or between multiple RF subcarriers within a single wavelength to degrade performance remains uncertain. While the research in this thesis was focused on optical crosstalk effects, it will be of great importance to develop much more complete link models including electrical and optical nonlinearities.

Future research is likely to focus on integrating current knowledge and demonstrating the feasibility of fibre-radio networks. A network demonstrator, however, requires a large amount of equipment, so this may not be achievable within a research environment. Another potential alternative is to focus on software simulations, including all potential sources of signal degradation.
Conclusions

7.3 REFERENCES


Appendix

A List of Supporting Publications

Journal Articles [1]

Conference Proceedings [2-7]


