Modelling Spurious Signals in Fibre Networks; Applied to HFR Systems

Thesis submitted in candidature for the degree of
Doctor of Philosophy

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October 1997

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Acknowledgements

I would like to thank many people who have given time and effort to help me in completion of this work from my first days at UCL. I would first like to thank my supervisor John O'Reilly for overseeing the project and for providing direction, advice and encouragement. Similarly, thanks to Phil Lane for technical guidance and help with preparation of the Thesis. Several people at GEC-Marconi Research Centre have also given up time to help me on many occasions, particularly Claire Moss, last of my industrial supervisors, and Ian Alstom from the Communications and Computing Laboratory.

I would particularly like to thank Luis Moura his joint work on the interferometric noise analysis, at which he has considerable skill and experience. Also for their invaluable day-to-day help, from the most simple questions to lengthy discussions, a huge thank-you to Sen Lin Zhang and Assaad Borjak.

Financial support was in the form of a CASE studentship, jointly funded by the Engineering and Physical Sciences Research Council (EPSRC) and GEC-Marconi Research Centre, Great Baddow, for which I am very grateful.
Abstract

The project work undertaken in this thesis has been concerned with the characterisation and modelling of spurious optical signals which may be generated in optical fibre communication networks. In particular the aim of the modelling work was to develop software tools which could be easily manipulated to represent different fibre systems and be used in the design, development and analysis of any desired network structure. The work can be divided into two studies, that of feedback travelling backwards through the fibre network, and of spurious signals propagating forwards and arriving with the signal at the detector, causing the generation of interferometric noise. Hybrid fibre radio (HFR) systems have provided the focus for the work, both for analysis and as illustrative examples of simulation techniques.

In order to simulate optical feedback effects a series of modules were developed for use within the Signal Processing Worksystem simulation environment. The package and developed tools are graphically interactive, providing the user with an ergonomic windowed environment for system design, re-configuration and optimisation. The developed software includes a model for a semiconductor laser and EDFA, together with feedback generating modules which simulate the effects of back-reflections and Rayleigh back-scatter from a fibre network. Some illustrative results using the software are also presented.

A mathematical analysis of interferometric noise (IN) in heterodyning HFR systems is presented. The analysis considers two frequency displaced optical carriers, one of which is on-off keyed. The signal is detected together with an interfering
sum of time-delayed and attenuated reflectors which cause unwanted beat signals between signal and crosstalk components. The moment generating functions for the noise distributions on one and zero symbols are evaluated and the modified Chernoff bound and a Gaussian approximation used to determine the impact on the system error rate. Simulation results confirm the analytical predictions and go on, through a series of system studies, to show that angle modulation schemes are more resilient to IN generation.
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<tbody>
<tr>
<td>ACTS</td>
<td>Advanced Communications Technologies and Services</td>
</tr>
<tr>
<td>ADSL</td>
<td>Asymmetric Digital Subscriber Line</td>
</tr>
<tr>
<td>AM</td>
<td>amplitude modulation</td>
</tr>
<tr>
<td>ASE</td>
<td>amplified spontaneous emission</td>
</tr>
<tr>
<td>ASK</td>
<td>amplitude-shift keting</td>
</tr>
<tr>
<td>ATM</td>
<td>Asynchronous Transfer Mode</td>
</tr>
<tr>
<td>BB</td>
<td>broadband</td>
</tr>
<tr>
<td>BDE</td>
<td>Block Diagram Editor</td>
</tr>
<tr>
<td>BER</td>
<td>bit error rate</td>
</tr>
<tr>
<td>BISDN</td>
<td>Broadband Integrated Services Digital Network</td>
</tr>
<tr>
<td>BSON</td>
<td>broadcast and select optical network</td>
</tr>
<tr>
<td>CAD</td>
<td>computer aided design</td>
</tr>
<tr>
<td>CaTV</td>
<td>common antenna television</td>
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<tr>
<td>CCB</td>
<td>Custom Coded Block</td>
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<tr>
<td>CDMA</td>
<td>code division multiple access</td>
</tr>
<tr>
<td>DFB</td>
<td>distributed feedback</td>
</tr>
<tr>
<td>DSL</td>
<td>digital subscriber loop</td>
</tr>
<tr>
<td>DSP</td>
<td>digital signal processing</td>
</tr>
<tr>
<td>EDFA</td>
<td>Erbium doped fibre amplifier</td>
</tr>
<tr>
<td>FDM</td>
<td>frequency division multiplex(ing)</td>
</tr>
<tr>
<td>FFT</td>
<td>fast Fourier transform</td>
</tr>
<tr>
<td>FRANS</td>
<td>Fibre Radio ATM Networks and Services</td>
</tr>
<tr>
<td>Abbreviation</td>
<td>Full Form</td>
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<tr>
<td>--------------</td>
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<tr>
<td>FSK</td>
<td>frequency-shift keying</td>
</tr>
<tr>
<td>FTTC</td>
<td>fibre-to-the-curb</td>
</tr>
<tr>
<td>FTTH</td>
<td>fibre-to-the-home</td>
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<tr>
<td>GA</td>
<td>Gaussian approximation</td>
</tr>
<tr>
<td>GSM</td>
<td>Global System for Mobile Communications</td>
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<tr>
<td>HDTV</td>
<td>high definition television</td>
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<tr>
<td>HDWDM</td>
<td>high density wavelength division multiplex(ing)</td>
</tr>
<tr>
<td>HFC</td>
<td>hybrid fibre-coax</td>
</tr>
<tr>
<td>HFR</td>
<td>hybrid fibre-radio</td>
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<tr>
<td>IM</td>
<td>intensity modulation</td>
</tr>
<tr>
<td>IN</td>
<td>interferometric noise or intelligent network(ing)</td>
</tr>
<tr>
<td>ISDN</td>
<td>Integrated Services Digital Network</td>
</tr>
<tr>
<td>ISP</td>
<td>internet service provider</td>
</tr>
<tr>
<td>LMDS</td>
<td>Local Multipoint Distribution Service</td>
</tr>
<tr>
<td>LO</td>
<td>local oscillator</td>
</tr>
<tr>
<td>MAN</td>
<td>metropolitan area network</td>
</tr>
<tr>
<td>MM</td>
<td>multimedia</td>
</tr>
<tr>
<td>MMDS</td>
<td>Multipoint Multichannel Distribution Service</td>
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<tr>
<td>mmw</td>
<td>millimeter wave</td>
</tr>
<tr>
<td>MODAL</td>
<td>Microwave Optical Duplex Antenna Link</td>
</tr>
<tr>
<td>MPEG</td>
<td>Motion Picture Experts Group</td>
</tr>
<tr>
<td>MZ</td>
<td>Mach-Zehnder</td>
</tr>
<tr>
<td>LAN</td>
<td>local area network</td>
</tr>
<tr>
<td>MCB</td>
<td>modified Chernoff bound</td>
</tr>
<tr>
<td>MGF</td>
<td>moment generating function</td>
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<tr>
<td>MUX</td>
<td>multiplexer</td>
</tr>
<tr>
<td>NRZ</td>
<td>non return-to-zero</td>
</tr>
<tr>
<td>OFLL</td>
<td>optical frequency locked loop</td>
</tr>
<tr>
<td>OO</td>
<td>object-orientation (or object-oriented)</td>
</tr>
<tr>
<td>OOK</td>
<td>on-off keying</td>
</tr>
<tr>
<td>OPLL</td>
<td>optical phase locked loop</td>
</tr>
<tr>
<td>OS</td>
<td>operating system</td>
</tr>
<tr>
<td>PCS</td>
<td>personal communications services</td>
</tr>
<tr>
<td>Acronym</td>
<td>Definition</td>
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<tr>
<td>PDA</td>
<td>personal digital assistant</td>
</tr>
<tr>
<td>PDF</td>
<td>probability density function</td>
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<tr>
<td>PON</td>
<td>passive optical network</td>
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<tr>
<td>POP</td>
<td>point of presence</td>
</tr>
<tr>
<td>POTS</td>
<td>plain old telephone service</td>
</tr>
<tr>
<td>PP</td>
<td>power penalty</td>
</tr>
<tr>
<td>QAM</td>
<td>quadrature amplitude modulation</td>
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<tr>
<td>RACE</td>
<td>Research and Development for Advanced Communications for Europe</td>
</tr>
<tr>
<td>RF</td>
<td>radio frequency</td>
</tr>
<tr>
<td>RZ</td>
<td>return-to-zero</td>
</tr>
<tr>
<td>SDH</td>
<td>Synchronous Digital Hierarchy</td>
</tr>
<tr>
<td>SDSL</td>
<td>Symmetric Digital Subscriber Line</td>
</tr>
<tr>
<td>SMF</td>
<td>standard mono-mode fibre</td>
</tr>
<tr>
<td>SNR</td>
<td>signal to noise ratio</td>
</tr>
<tr>
<td>SONET</td>
<td>Synchronous Optical Network</td>
</tr>
<tr>
<td>SPW</td>
<td>Signal Processing Worksystem</td>
</tr>
<tr>
<td>TDM</td>
<td>time division multiplexing</td>
</tr>
<tr>
<td>TWP</td>
<td>twisted wire pair</td>
</tr>
<tr>
<td>VOD</td>
<td>video-on-demand</td>
</tr>
<tr>
<td>WAN</td>
<td>wide area network</td>
</tr>
<tr>
<td>WDM</td>
<td>wavelength division multiplex(ing)</td>
</tr>
<tr>
<td>WLL</td>
<td>wireless local loop</td>
</tr>
<tr>
<td>WRON</td>
<td>wavelength-routed optical network</td>
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</table>
Chapter 1

Introduction

Telecommunications is facing a paradigm shift in the nature of teletraffic and associated applications. Traditional voice transmission is giving way to heterogeneous traffic types that encompass all aspects of multimedia communication, from images to video and from high quality audio to large computer data files. In order to realise the full potential of multimedia applications, a huge increase in bandwidth is required over existing telecommunications infrastructure.

Optical fibre, since its invention in 1966 [1], has revolutionised the potential capacity of telecommunication networks. Research has delivered continually lower attenuation and tapped proportionally more of the enormous bandwidth available using fibre. Though much of the core network has now migrated to optical and high bandwidth technologies, local access networks remain an obstacle to delivering true broadband services to customer premises. Upgrading existing infrastructure or deploying new access networks can prove the most costly capital expenditure faced by telcos. This was highlighted recently by MCI Communications Corporation, which announced losses of US$1 billion attributed to the costs of breaking into local access markets in the United States [2]. Wireless access networks offer advantages over wired alternatives in many market positions: network deployment time, cost and maintenance can all be vastly reduced using wireless technology.
The next generation of wireless networks - fixed or mobile - will have to operate at higher frequencies than existing systems, extending to millimetre wavelengths. Development of these higher frequencies will be necessitated by congestion in the lower frequency spectrum and simultaneously driven by the demand for increased bandwidth [3]. Optical fibre is an attractive medium for carrying millimetre-wavelength signals and a great deal of research has been undertaken in developing optical technologies to provide the transmission infrastructure for the next generation of cellular networks. The term hybrid fibre/radio (HFR) is commonly used to describe these systems.

Modelling and analysis of emerging systems is an important aspect of the design and development process. Aspects of optical fibre systems need to be investigated with respect to HFR technology and system design tools developed to allow performance assessment of proposed networks. In particular this Thesis is concerned with the effect of spurious signals which may occur in an HFR network and their impact on system performance.

1.1 Thesis Organisation

Following this introductory chapter, a common structure is observed through the Thesis. Each chapter is opened with an introductory section which sets out its aims and provides an overview of the chapter's content. A summary closes each chapter, which reviews the main points and relates the material to other material in the Thesis.

An overview of the main Thesis content follows:

Chapter 2, following this introduction, aims to provide a review of relevant aspects of telecommunication technology. Various access networks are described and wireless access systems discussed, with millimetre-wave systems identified as the likely next generation technology. Aspects of optical transport network technology are then discussed with a view to presenting the current state of HFR system development. The problem of spurious signals propagating in a fibre net-
work is presented and interferometric noise introduced as a potentially limiting problem generated by such signals. Applications and services likely to be provided using HFR networks are proposed and two experimental systems, MODAL and FRANS, are described.

Alternative approaches and software packages for computer simulation of optical systems are discussed in Chapter 3. Signal Processing Worksystem (SPW) is described in some detail, highlighting the aspects which led to its adoption: in particular the object-oriented handling of system design and facility for embedding custom code into the simulator interface.

Modelling of spurious signals as feedback into network devices is tackled in Chapter 4, which describes models for a semiconductor laser and EDFA. Custom-coded SPW modules which generate optical feedback in the from of unwanted reflections in the network and Rayleigh backscatter are described and the effect of back-reflections on DFB laser and EDFA operation is simulated.

Chapter 5 is concerned with spurious signals propagating forwards through the network and arriving at the receiver. Specifically the effect of these signals in producing interferometric noise (IN) is addressed in the context of systems using two-carrier heterodyne mixing to produce RF sub-carrier frequencies. An analysis of IN generation in these systems is presented by treating the crosstalk as a sum of delayed and attenuated replicas of the original signal with uncorrelated phase. The moment generating functions (MGFs) of the resulting IN noise distributions are found and system impact is estimated using a Gaussian approximation (GA) of the noise and by use of the modified Chernoff bound (MCB). The MCB is shown to give a much tighter upper bound on the BER than the GA, particularly for a small numbers of interferers or single dominant term. Simulation results are also presented which confirm the predicted IN distributions and show that FSK modulation would be more robust than ASK in the presence of IN.

The software and techniques developed in the previous chapters were used in some exemplar system studies in Chapter 6. A directly modulated system is simulated and the impact of IN and feedback into the laser shown in received eye-diagrams.
MODAL and a hybrid FRANS system employing multiple frequency channels are then studied in terms of IN impairment. The Thesis is concluded with Chapter 7 which summarises the main contributions of the project and suggests some areas for further research.

Two appendices are included at the end of the Thesis covering the full derivation of IN moment generating functions and software documentation produced for GEC-Marconi Research Centre, covering the modules described in Chapter 4.

1.2 Contributions

The main contributions made by this research project can be summarised as

- The capacity to simulate linear optical feedback was introduced as a series of CCBs for use with an existing optical network modelling facility, developed within SPW.

- Study of interferometric noise was completed for systems using optical heterodyning of two optical signals to produce an RF carrier. Linear crosstalk, primary, secondary and tertiary beating terms were identified in the received signal where primary beating forms the ‘wanted’ signal component and secondary beating forms the salient IN contribution.

- MGFs of the IN components were derived and used to predict system impact in terms of BER and PP.

- Simulation of IN was carried out, confirming analytical predictions and suggesting the use of phase modulation to minimise system impact.

The following research papers were published as a result of the work undertaken

45(8):1398-1402, August 1997, Part II Special Issue on Microwave and Millimetre-wave Photonics.


- L. Moura, M. Darby, P. Lane and J. J. O'Reilly. “The Effect of Interferometric Crosstalk on Optically Generated Microwave Signals”. International Topical Meeting: Microwave Photonics, Kyoto, Japan, 1996


Chapter 2

Optical Network Technologies and HFR Systems

2.1 Introduction

This chapter reviews the current state of telecommunications technologies influencing the work in this thesis. The project work is centred around optical networks, with an emphasis on the effect of unwanted reflections occurring in these systems. In particular, optically supported wireless access networks have provided the focus for much of the work. In view of these subject areas, optical transport and wireless access technologies are presented in this chapter, which describes their context, current state of development and probable future evolution.

Section 2.2 briefly describes the current alternatives for broadband access via the local loop, before concentrating on wireless access technologies. The potential of millimetre-wave systems is presented and identified as the likely development of wireless technology towards a third generation mobile systems and true broadband wireless access networks. The current state of millimetre-wave technology is described and likely future developments and applications proposed.
Key issues relating to the project with respect to optical networks are detailed in Section 2.3. The main system and device technologies are described with emphasis on the effect of reflections in an optical network. Interferometric noise is identified as a particularly important reflection-induced effect and a general analysis is developed in Section 2.3.4 to emphasise its potential impact.

Convergence of optical transport and wireless access technologies is described in Section 2.4 which focuses on optically supported wireless millimetre-wave networks. The limitations of optical components in supporting millimetre-waves is discussed. In view of these limitations, current methods of optical millimetre-wave generation are reviewed and an analysis of coherent mixing presented to illustrate this mechanism fully. Potential applications are then proposed, followed by a description of two experimental systems MODAL and FRANS in sections 2.4.3 and 2.4.4 respectively. The chapter concludes with a summary of the review material.

2.2 Broadband Wireless Access Networks

2.2.1 Broadband Access Alternatives

The access network is the key to deployment of true broadband communications services. The current state of wired access network deployment almost completely comprises switched telephony and broadcast CaTV networks. In addition, mobile access is provided by wireless cellular networks. These systems have evolved to deliver their services efficiently, but are not necessarily optimal for emerging broadband applications. Access infrastructure must thus be enhanced or re-deployed in the future, driven by the demand for new services and their cost of provision.

Wireless networks are just one of several broadband access technologies, representing various stages of development and deployment. Figure 2.1 is an illustration of the architecture of a public broadband network. The core distribution network comprises service nodes, which interface the distribution and access
2.2. BROADBAND WIRELESS ACCESS NETWORKS

networks and media servers, which provide storage and connectivity to broadband information and other networks. This core network is connected to subscribers by several digital broadband alternatives in the local loop:

**Digital 'wideband' access** is provided over existing twisted wire pairs (TWP) by several technologies, most importantly including the integrated services digital network (ISDN) and asymmetric digital subscriber line (ADSL) systems. These achieve various data rates to the subscriber and combined with compression techniques such as MPEG encoding can provide limited video distribution. Generally however, these systems offer only small incremental improvements of the existing cable infrastructure and do not provide sufficient bandwidth for true broadband access [4].

---

Figure 2.1: Network architecture for various broadband access systems
Hybrid fibre / coax (HFC), systems provide a full broadband duplex link and can transmit broadcast digital video and broadband digital services together with analogue television signals over the same medium using FDM. Coaxial cable is used for the final connection from an optical fibre feed, similar to existing cable TV networks [5].

Switched digital video, also known as fibre to the curb (FTTC), systems distribute base-band digital signals to curb-side cabinets using fibre. This arrangement shares the cost of optoelectronic components between a number of users: a single fibre would typically connect tens of homes to the core network. Final connection to the subscriber could be provided by a number of media. Existing copper infrastructure might be used but radio distribution may also find deployment, forming a wireless tail to the system.

Passive optical networks offering the prospect of fibre to the home (FTTH), potentially provide the largest available bandwidth in the local loop, with fibre connected directly to subscriber premises. Passive configuration removes powered electronic and optoelectronic components from distribution cabinets and local exchanges to give high reliability, lower cost, ease of deployment and low maintenance. PONs are usually associated with star components (see page 50) since other configurations are difficult without active components (see page 50).

Wireless access. Several technologies are available or under development to provide mobile personal communications services (PCS), broadband wireless local loop access (WLL) and wireless LAN access. Potential systems are described in the section following.

2.2.2 Broadband Wireless Technologies

Cellular wireless networks are the ubiquitous access method for today's mobile communications services, currently providing cellular telephony and paging. These services are already heavily subscribed, enjoying exponential growth, with projections for this to continue into the foreseeable future [6]. A second generation
of cellular mobile radio systems, which includes GSM, is now being widely deployed on a global scale, providing a mobile digital service. Whilst still primarily voice-oriented, second generation systems now support a range of personal communications services (PCS), expanding to narrowband data, including fax, email and www access [7]. Items in bold are representative of the wireless access systems investigated in this project.

These seminal mobile communication systems are now set to be superseded by broadband technologies. A ‘third generation’ of mobile systems will support a wide range of services and teletraffic types, with user access rates of several Mb/s. Table 2.1 shows current and projected technological ‘generations’ of wireless communications and their associated services.

In addition to their capacity to provide mobile communication, wireless networks offer the primary advantage over wireline alternatives in the speed and ease of their deployment, presenting lower costs in terms of capital equipment, construction and maintenance. In many contexts it will be uneconomic to deploy new wireline technology to provide broadband access. Situations dictating wireless local loop (WLL) access might include rural deployment and areas where cabling is at particular risk of damage or prohibitively expensive to deploy. When network deployment is mature, costs remain low compared with wired access due to smaller maintenance requirements and the lower cost of adding subscribers [8]. As the deregulatory move away from monopoly providers progresses, wireless systems are likely to prove particularly attractive to emerging access providers who face large capital investments on market entry.

Commercial (as opposed to previously mentioned geographical) environments in which WLL is likely to feature include:

- established monopoly or near-monopoly operators with underdeveloped infrastructure and a need to satisfy regulatory requirements for penetration, waiting lists and network development,
- emerging operators who face large capital investments on entry to recently deregulated but highly developed markets,
CHAPTER 2. OPTICAL NETWORKS AND HFR SYSTEMS

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<td>Fixed wireless loop</td>
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Table 2.1: Development of wireless communication systems, showing technological generations and their associated services

- existing operators facing new competition and needing to match the network development of other companies using WLL, or wishing to offer new services with an initial small subscriber base.

WLL access may also be deployed to support wireline infrastructure and allow services to be provided transparently over multiple access networks. Services are already being provided using such asymmetric connections since different applications have different traffic characteristics and require disparate BERs, data rates and network latency times. These parameters often differ widely between the upstream and downstream links, suggesting different access technologies for each
Tiered arrangements of multiple wireless networks with varying data rates are also emerging as an hierarchical approach able to provide seamless mobile access in many different environments. These ‘wireless overlay’ networks may consist of a wireless LAN to provide high-speed connection within a building, handing off to a cellular digital network for metropolitan area access and using satellite coverage for remote areas. Intermediate tiers are also possible depending on requirements.

Figure 2.2 maps the likely evolution of broadband wireless services, including both mobile and WLL access. Also shown are the different media, traffic types, cellular structure and access bandwidths associated with providing each example.

Figure 2.2: Data rates, access technologies (shaded boxes) and service support for wireless access networks.

Systems are now in development to extend wireless technology to provide WLL access to broadband network services. Wideband access is likely to be deployed using extensions of ‘cordless’ phone technology. These systems, deployed in a ‘low-tier’ cellular arrangement, are planned to support services up to ISDN basic rate [9, 10]. In common with DSL wideband access, these networks are intended...
as an intermediate solution between current narrow band systems and the deployment of true broadband access networks. Such broadband wireless systems are already under development with the aim of providing multimedia PCS and local-loop access to subscribers. Some of these technologies, including wireless LANs and MMDS systems have already found deployment. Research into broadband wireless access systems include:

**Wireless LANs** are already deployed as replacements or extensions to existing wired LANs, providing mobility within a building for portable computers [11]. Typically client / server data is carried, characterised by bursty traffic patterns. Future developments may include the transport of all communications, including packetised telephony, and the proliferation of PDAs in the LAN environment.

**Wireless ATM.** Extending ATM infrastructure using wireless access provides mobility, together with end-to-end ATM networking and its associated guaranteed service characteristics [12, 13]. Wireless ATM systems can connect to a single port of an ATM switch and provide access to a number of users, typically within a building or campus environment. This type of system could comprise or extend an ethernet system or intranet. Alternatively access could be provided over metropolitan areas, connecting mobile users to the core network.

**MMDS and LMDS.** MMDS, also known as ‘wireless cable’ is a well developed broadcast system designed to provide up to 33 analogue video channels where deployment of cable is impractical. Cell radii are typically 25 to 35 miles, with transmission between 2.1GHz and 2.7GHz. Digital video support is being developed using data compression techniques. Broadband access may be supported using high order modulation schemes such as 64-QAM and two-way access added by using a separate wired channel for upstream communication. MMDS has found increasingly widespread deployment, currently deployed in over 60 countries and reaching over 150 million premises. The growth of MMDS subscribers, particularly in recent years, illustrates the potential for wireless delivery in the local loop. Low
deployment costs are seen as a key economic driver for all MMDS network operators [14]. **LMDS** is a development of MMDS, transmitting at higher frequencies between 27GHz and 30GHz with cell radii of around 5 miles. It is designed to provide a two-way link supporting a range of broadband services, including VOD, broadcast video and telephony. Both systems require a static antenna for access, though a received signal may be distributed locally by cable.

### 2.2.3 Millimetre-wave Wireless Systems

In order to provide broadband services over wireless networks an increase in bandwidth is required over current RF carrier technology. This will necessarily require operating at higher frequencies, into the millimetre-wave region of the radio spectrum.

Commercial and military exploitation of the radio frequency spectrum for both fixed and mobile access is growing very rapidly. The microwave region (defined between 1 and 30GHz) is already heavily congested and tight regulations are imposed on operators transmitting in this area [15]. Heavy usage in this region and at lower frequencies has lead to increased pressure to develop systems working at as-yet unused higher frequencies. Millimetre-wave frequencies will be the next to be developed, extending from 30GHz to (nominally) 300GHz, encompassing wavelengths from 1cm down to 1mm. These higher frequencies will provide correspondingly high bandwidths and enable broadband traffic to be accommodated. The advantages of systems employing millimetre-wave carriers, though primarily associated with bandwidth, also encompass cellular network structure and efficiency of radio spectrum usage.

Millimetre-wave frequencies generate new network design parameters. Bands of high absorption due to oxygen molecule resonance characterise the atmospheric absorption spectrum, with intervening ‘windows’ of low attenuation. Propagation distances are further reduced by moisture in the air: both water vapour and droplets are strongly absorbing at mmw frequencies [16]. Building materials
have various absorption characteristics and network planning requires complex simulation of mmw propagation [17]. Such small propagation distances dictate a pico-cellular network structure, with cell diameters of the order of tens of metres. An advantage of small cell size is the high capacity for frequency re-use within the network, providing efficient use of the radio spectrum. The proximity of transmission frequency to absorption maxima ultimately determines cell size and consequently both network cost and efficient spectrum use [16]. Carrier signal frequency characteristics are thus a vital aspect of system design and networks are critically dependent on the generation of stable, narrow linewidth carriers to achieve efficient and cost effective operation within tight regulatory restraints.

The attenuation characteristics of mmw signals which make them so desirable for cellular network applications are very poor from the perspective of point-to-point and trunk connections where microwave and fibre links are likely to prove more economic.

### 2.3 Optical Transport Networks

Optical fibre has been developed from its conception in 1966 to its present position as the primary broadband telecommunications transmission media, with myriad applications from LAN to WAN communication. The qualities that have driven this research and development effort are well documented but most importantly include low attenuation, negligible external cross-talk, and potentially huge transmission bandwidth [18]. The continual deployment and development of communication networks together with burgeoning multimedia applications are driving demand for more bandwidth. Teletraffic content now comprises a range of information services, expanding from speech transmission to computer, static image and video data.

This shift to heterogeneous traffic content is driving intensive research efforts towards the development of novel optical network approaches. This research is intended to develop fibre systems from high capacity transmission links to a true
broadband networking technologies. Approaches include WDM systems, which increase available bandwidth and optically routed and passive optical networks which remove limiting electronic network components. These technologies are now maturing to the point where optical networks are becoming the economic choice over electronic equivalents to provide broadband communication services [19]. A brief overview of these systems and associated devices is presented in the context of the project.

2.3.1 WDM systems, Optical Routing and PONs

In order to better exploit the huge potential bandwidth available using optical fibre, (> 30THz in the low-loss region between 1.2μm and 1.6μm) the development of multiple-wavelength transmission has received much attention [20]. By attributing transmission channels to a series of frequency spaced optical signals, multiple channels may be carried in parallel on a single fibre. This technique is known as wavelength division multiplexing (WDM) and allows the capacity of existing fibre infrastructure to be increased many times without the need for additional cabling.

Early WDM systems used two transmission channels in the fibre low-loss bands centred at 1.3μm and 1.55μm. Contemporary research however configures channel spacings at less than 1 nm and these are referred to as high density frequency division multiplexed (HDWDM) schemes [21]. The minimum spacing between each channel is ultimately determined by the modulated signal characteristics. For direct modulation schemes, the laser linewidth is increased beyond the modulating signal bandwidth by induced chirp [22]. Channel spacing of the order of the modulated laser linewidth is termed frequency division multiplexing (FDM) [21].

Further innovation in the frequency domain is available in the form of routing signals through the network in the optical domain [23]. This relieves the transmission bottle-neck associated with the detection, decoding, switching and re-transmission used in electronic switching schemes. This allows the development
of high-speed all-optical networks which are resolved completely in the optical domain. Optically routed networks allocate destination addresses to signal channels via transmission wavelength, in addition to the encoded header address used in single channel packet-switched systems. The signal is either switched spatially or filtered spectrally: these are referred to as ‘wavelength routing’ (WRON) and ‘broadcast and select’ (BSON) schemes respectively. Broadcast operation relies on transmission of all wavelength channels to all possible destination nodes in the network, at which point spectral filtering takes place to obtain the desired signal. The routeing scheme uses diffractive or refractive elements to spatially couple wavelength channels into different fibres. The implementation of both schemes is critically dependent on the performance of several key optical components, notably frequency stabilised and tunable lasers, optical amplifiers and high resolution multiplexers. The broadcast scheme in particular relies on amplification of optical power levels in the system as the signal is repeatedly split for transmission to each network node. EDFAs provide the broadband, multichannel and non-regenerative amplification required. Passive optical networks (PONs) are a further development of all-optical configuration which remove amplification from the network entirely, producing a simple and highly reliable system by removing powered electronic components.

2.3.2 Key Network Devices

Innovation in fibre communication networks is achieved through development of optical devices and their novel configuration within the network. The performance of each type of fibre system is critically dependent upon the behaviour of certain key network components. This section gives descriptions of optical devices important to network performance in the context of this thesis.

2.3.2.1 DFB laser

DFB lasers are the standard source for optical telecommunications networks using single-mode fibre. DFBs offer line widths down to a few kHz and tens of mW
in optical output power. Modelling system performance in terms of key parameters such as SNR, eye-closure and BER is critically dependent on an accurate description of diode laser behaviour.

In addition, particular emphasis is placed on laser diode operation by some system types. As we have seen, tunable diode lasers constitute key components in WDM and optically routed systems. Several methods are available for tuning diode lasers including among others: thermal, mechanical, acoustooptic and injected current methods [20]. Frequency wander or induced chirp in these devices can broadly affect system performance over a much wider range of parameters than in a conventional fibre link. In WDM and optically routed networks, diode lasers can affect inter-channel crosstalk, routing behaviour and channel blocking, in addition to these system performance issues associated with conventional deployment. DFB laser modelling, including induced chirp and phase noise effects, is described in Section 4.3.

### 2.3.2.2 Erbium Doped Fibre Amplifier

Optical amplifiers have revolutionised optical communications and had a particular impact on the development of optical networking. WDM and optically routed systems in particular would be severely limited in scope without the capacity for optical amplification. Prior to their development, optical signals were boosted using electronic repeaters, which detected, electronically amplified and re-transmitted the signal. Repeaters are complex devices which pose a severe restriction on the bandwidth and transmission time through a network. Optical amplifiers allow a network to be transparent to the transmitting light source, providing broadband amplification with no electrical conversion of the optical signal required. In addition to configuration as non-regenerative repeaters, optical amplifiers may be used in pre-amplifier and power-amplifier configurations. Various types of optical amplifier are available including; semiconductor laser amplifiers (SLAs), Raman fibre amplifiers (RFAs) and Erbium doped fibre amplifiers (EDFAs) [24].
EDFAs are undoubtedly the most useful optical amplifier currently available in terms of signal gain, saturation levels and additive noise. Erbium ions (Er\textsuperscript{3+}) are incorporated into the core of single-mode optical fibre and this doped fibre is spliced directly into the fibre network. Er\textsuperscript{3+} amplifies in the 1.5\textmu m range. The amplifier is pumped by a semiconductor laser which is spliced onto the doped fibre as an integral part of the EDFA package. The pump laser may be configured in the direction of the signal known as a co-pumped configuration, or in the opposite direction known as counter-pumped. Pumping from both directions simultaneously is termed bi-directional pumping. The gain available to the optical signal is dependent on the population of excited ions in the doped fibre and this is dependent on the intensity of the pump light at that point. A dynamic gain distribution is set-up within the doped fibre as the pump light is absorbed through the fibre length and ions are depleted from their excited state through stimulated and spontaneous emission processes. Spontaneous emission is amplified in the doped fibre in the same way as stimulated emission. The accumulation of amplified spontaneous emission (ASE) in a chain of EDFAs constitutes a significant and sometimes limiting noise source in optical systems. Optical isolators may be employed to limit backwards-propagating ASE but forward ASE is transmitted through the network and amplified with the signal by each successive EDFA.

EDFA operation may be described through formulation of these gain and loss processes exhibited by the amplifier, leading to a rate equation approach to modelling device behaviour and ASE transmission through the network [24, 25].

2.3.2.3 Mach-Zehnder modulator

The Mach-Zehnder modulator uses interferometric conversion of an induced phase difference to produce intensity modulation of an optical signal. The device is based on the linear electro-optic effect (known as Pockel's effect) which is exhibited by certain materials, including semiconductors such as Gallium Arsenide (GaAs) and crystals such as Lithium Niobate (LiNbO\textsubscript{3}). Applying an electric field to these materials induces a localised refractive index change. The material is doped to produce a waveguiding structure which ensures that the optical field
2.3. OPTICAL TRANSPORT NETWORKS

optical signal in
$E \cos \omega_f \cdot t$

modulating voltage
$V_1(t) = V_n (1 + \varepsilon) + \alpha V \cos(\omega_m t)$

phase modulator

modulating voltage
$V_2(t) = -V_n (1 + \varepsilon) - \alpha V \cos(\omega_m t)$

optical signal out
$E \cos (\omega_f t + \omega_m t)$

optical coupler

$\omega_f$ = optical (laser) frequency
$2\omega_m$ = millimetre-wave frequency

Figure 2.3: Mach-Zehender modulator

propagates through this modulated region. The change in refractive index leads to a corresponding modulation of the phase of the propagating lightwave. The modulation rate of these devices is limited by the RC product of the drive circuit electrodes which are placed on either side of the waveguide [26]. Though there is a trade-off between frequency response and drive power, the upper modulation bandwidth is ultimately limited to a few GHz [27].

Mach-Zehnders use a configuration of two electro-optic phase modulators to produce an intensity modulation. A typical Mach-Zehnder modulator arrangement is shown in Figure 2.3. The modulating voltages shown are for the MODAL mmw generation system, described in Section 2.4.3. The incoming signal is split between the two arms of the Mach-Zehnder which each comprise an electro-optic phase modulator. In a balanced arrangement, as shown, the phases of the two optical signal components are modulated by an equal amount, in opposite directions. Recombination of the signal using an optical coupler (with associated 3dB loss) produces interference between the two waves. Intensity modulation of the final signal is thus achieved by controlling the relative phase difference between the two components. If the device is appropriately voltage biased, the original optical component is suppressed, leaving two modulation sidebands [28]. A residual component of the original signal will be observed if there is unequal splitting or phase modulation between the two arms, or reflections from the splitters and
Due to the upper bandwidth limit of Mach-Zehnder modulators they are not immediately applicable to millimetre-wave applications. Several novel techniques have been developed however which use Mach-Zehnders to impose a millimetre-wave carrier onto an optical signal with applied modulated subcarriers [27]. These are discussed in Section 2.4 following.

2.3.3 Optical Crosstalk in a Fibre Network

A source of degradation in a fibre system is generation of crosstalk due to imperfect component performance and spurious reflections in the network. Reflections can cause feedback to the transmitter or cause the signal to take alternate paths to the receiver. Figure 2.4 shows the effect of reflections and leaky routers in generating spurious network routes and optical feedback towards the transmitter. The signal from A to X is affected by crosstalk transmitted from terminal B and by reflections back to the transmitter. Channel B to Y is affected by spurious, delayed replicas of the intended signal caused by a reflected alternate path.

Reflections are inherently produced by fibre splices and imperfect component
performance, since actual transmissivities are always less than 100%. Though modern splicing techniques are designed to provide an almost seamless interface between fibre sections and components, it has been shown experimentally that misaligned splices and connectors can produce reflectivities of up to 0.24 [29].

The generation of spurious crosstalk components will depend on the type and size of fibre network. WDM networks are prone to crosstalk from other light sources. Passive optical networks are transparent from transmitter to receiver. PONs use spatial switches and optical filters to remove other wavelength channels and are thus dependent on the crosstalk isolation provided by these devices. Both WDM and single wavelength channel networks will produce delayed replicas of the original signal. These replicas arrive at the detector, or back at the transmitter, as delayed and attenuated versions of the original signal. Delay times are dependent on the additional propagation distance incurred via the spurious network route with respect to the intended network route. Larger and more complex networks will be at greater risk of generating crosstalk due to the larger numbers of components, splices and splitters they contain.

Crosstalk can have a range of effects on a fibre network, depending on the degree of isolation, direction, position in the network, frequency and phase characteristics. Though optical isolators can be employed to reduce backward propagating light, they only appear periodically in the network layout and may introduce reflections themselves. In addition to the linear crosstalk power that is coupled into the signal channel, photo-detection in the presence of crosstalk generates beat noise terms. Beating between signal and crosstalk components is termed interferometric noise, due to the phase to amplitude conversion which takes place. Linear feedback and IN can affect system performance at various points in the network, impacting the operation of transmitters, amplifiers, routers and receivers.

### 2.3.4 Interferometric Noise

Interferometric noise (IN) occurs in an optical network when the signal arrives at the receiver together with other unwanted lightwave components from the
network. IN is generated by beating between the signal and crosstalk components. The beat frequency is driven by phase disparity between the beating elements. Interferometric conversion of the phase difference leads to an intensity fluctuation, which appears on the detected signal.

Interferometric noise can be classified as coherent or incoherent depending on the degree of phase correlation between the signal and the interfering terms [30]:

**Coherent IN** occurs if the interfering crosstalk terms originate from the same laser and are delayed by less than the coherence time of the laser. A DFB laser with a linewidth of 10kHz has a coherence time of around $10^{-4}$ seconds, corresponding to a coherence length of 20km in standard fibre. In this case the phase terms are correlated and the amplitude of the signal drifts slowly, depending on the phase relationship between the components.

**Incoherent IN** occurs if the interfering crosstalk terms originate from different sources or from the same source with a delay greater than the coherence time of the laser. A laser which is directly modulated will produce chirp and will not behave as a single frequency source in this context, reducing the coherence time to the symbol period. In the incoherent case, IN appears as a beat signal on the photocurrent and we may describe its characteristics statistically by treating the phase of each optical component as an independent random variable.

We will not consider coherent crosstalk in this analysis since incoherent crosstalk is generally acknowledged as the most likely case to occur due to the tight constraints on delay time, and laser chirp. In order to characterise the case of incoherent IN we assume that the polarisation states of all the optical components are aligned. This is a worst case assumption which gives a useful bound on network design and has also been shown to be the most likely case to occur in practice [31]. If the transmitted optical signal electric field is $\sqrt{2P_s} \cos(\omega_0 t + \phi_0(t))$ we can write the total optical electric field arriving at the photodiode, signal + sum of
interferers, as

\[
E(t) = \sqrt{2P_s \cos[\omega_0 t + \phi_0(t)] + \sum_{i=1}^{N} \sqrt{2\epsilon_i P_s \cos[\omega_i t + \phi_i(t)]}} \tag{2.1}
\]

where \(E(t)\) is the electric field amplitude, \(P_s\) is the signal power, \(\omega\) is the angular frequency, \(\phi_i(t)\) is the phase, \(N\) is the number of interfering components and \(\epsilon_i\) is the power reflectivity of an interferer (= \(P_i/P_s\)). If we assume a normalised receiver responsivity, the photocurrent produced by the total optical field will be

\[
i(t) = E^2(t) = 2P_s \cos^2[\omega_0 t + \phi_0(t)] + \left[\sum_{i=1}^{N} \sqrt{2\epsilon_i P_s \cos[\omega_i t + \phi_i(t)]}\right]^2
+ 4P_s \sum_{i=1}^{N} \sqrt{\epsilon_i \cos[\omega_0 t + \phi_0(t)] \cos[\omega_i t + \phi_i(t)]} \tag{2.2}
\]

If we expand and neglect terms that appear at frequency \(2\omega\) we obtain

\[
i(t) = P_s \left[1 + \sum_{i=1}^{N} \epsilon_i + 2 \sum_{k=i+1}^{N} \sum_{i=1}^{N-1} \sqrt{\epsilon_i \sqrt{\epsilon_k} \cos[(\omega_i - \omega_k)t + \phi_i(t) - \phi_k(t)]}
+ 2 \sum_{i=1}^{N} \sqrt{\epsilon_i \cos[(\omega_0 - \omega_i)t + (\phi_0 - \phi_i)(t)]}\right] \tag{2.3}
\]

where

- \(P_s\) is the signal
- \(P_s \sum_{i=1}^{N} \epsilon_i\) is the total linear crosstalk power contribution
- \(2P_s \sum_{i=1}^{N} \sqrt{\epsilon_i \cos[(\omega_0 - \omega_i)t + (\phi_0 - \phi_i)(t)]}\) is the signal-crosstalk or primary beating
- \(2P_s \sum_{k=i+1}^{N} \sum_{i=1}^{N-1} \sqrt{\epsilon_i \sqrt{\epsilon_k} \cos[(\omega_i - \omega_k)t + \phi_i(t) - \phi_k(t)]}\) is the crosstalk-crosstalk or secondary beating

Therefore the total crosstalk, primary and secondary is given by

\[
IN = 2P_s \left[\sum_{i=1}^{N} \sqrt{\epsilon_i \cos[(\omega_0 - \omega_i)t + (\phi_0 - \phi_i)(t)]}
+ \sum_{k=i+1}^{N} \sum_{i=1}^{N-1} \sqrt{\epsilon_i \sqrt{\epsilon_k} \cos[(\omega_i - \omega_k)t + \phi_i(t) - \phi_k(t)]}\right] \tag{2.4}
\]
2.4 Optically Supported Millimetre-wave Networks

As we have seen, a system employing a millimetre-wave carrier will require a large number of closely spaced antenna units, due to the small propagation distances associated with millimetre-waves (Section 2.2.3). A potential problem therefore, is the distribution of the millimetre-wave signal to such a large number of remote sites. The use of copper cable to perform the delivery would prove too lossy at mmw frequencies, while local generation of the millimetre-wave would prove too expensive, since each site would require a high stability oscillator. Thus cheap, low-loss connection of antennas to a central network node, from where the millimetre-wave could be generated and controlled, is the optimal network solution. These requirements suggest the use of optical fibre to connect the central node and remote sites, reducing remote sites to electro-optic converters [35]. For micro- or picocell networks, remote sites might also be developed using passive technology, so that no amplification of mmw signals need be carried out [36].

Several methods have been proposed and developed for modulation of an optical source with RF or millimetre-wave subcarriers. These can be broadly divided into intensity modulation systems (using either direct or external modulation of an optical source) and those which employ dual optical frequency techniques, which
generate the mmw carrier by coherent mixing, similar to heterodyne communication systems [37].

**Direct modulation** Generally fibre / radio systems have used direct modulation or an external intensity modulator to introduce an RF subcarrier onto the output of a continuous wave laser source. Direct modulation is limited to an upper modulation frequency due to the resonance of semiconductor lasers, which is typically around 20GHz (see section 4.3), though mmw subcarrier modulation of up to 35GHz has been achieved experimentally [38, 39]. Additional problems associated with laser chirp and non-linearities compound this resonance limit.

**External modulation** External intensity modulators are already widely used in modulating laser outputs with RF signals. Currently, limited in available bandwidths of operation mean that these devices cannot be extended to transmit at the higher frequencies required by a millimetre-wave carrier.

**FM-IM conversion using fibre dispersion** A technique which has been practically demonstrated at millimetre-wave frequencies employs fibre dispersion to convert FM into amplitude modulation [40] and has been used to transmit analogue TV signals at 40GHz. The system requires the signal to be distributed over fixed lengths of fibre to remote units in order to obtain the correct dispersion. This is very restrictive in terms of network topology and the technique has not been adopted to date.

**Coherent Mixing** is a potentially simple solution, which generates the millimetre-wave by mixing two frequency displaced carriers at the detector. An analysis of coherent mixing is presented in the section following. Heterodyning techniques rely on the stable generation of two optical carriers, for which several methods are available. These include:

- **Optical frequency locked loops (OFLLs)** The most basic method available using OFLLs is to lock two lasers to the required offset frequency by means feedback from the output of the master laser to the drive circuit of the slave laser [28]. Because the two lasers lack
phase coherence, the electrical linewidth of the resulting millimetre-wave beat signal is typically twice that of the two optical sources. While this is not a problem at high data rates, lower rate applications would suffer. While very narrow linewidth lasers may be employed to improve performance, increased complexity and other associated limitations are incurred.

- **Optical phase locked loops (OPLLs)** are similar, though more complex than OFLLs, using a phase detector in the feedback loop between master and slave lasers. This leads to complete correlation of the phase noise between the two optical signals. This phase correlation produces a very narrow electrical linewidth after mixing on the photodiode: sub-Hz linewidths have been reported for microwave generating systems [41].

- **Mach-Zehnder modulator frequency-doubling method.** This method differs in that a single optical source is used to generate the two optical carriers. The two carriers are produced by passing a semiconductor laser output through a Mach-Zehnder modulator which has been appropriately biased. By using a single source the two carriers will be, by definition, completely phase correlated, producing a subcarrier with a very narrow electrical linewidth after mixing [27]. A description of carrier generation and modulation imposition using assemblies of Mach-Zehnder modulators is given in Section 2.4.3.

Other coherent mixing systems have been developed, using various technologies to generate the optical signal components [42, 43, 44, 45]. Though employing different optical signal generation methods, these systems use similar distribution networks and optical detection/mixing. System studies of signal distribution and detection might therefore be applied to a range of HFR networks, common in their utilisation of coherent mixing for mmw generation. Phase characterisation of the various systems will differ however, as will modulation schemes, though most make use of high order m-ary modulation in order to efficiently use the available RF spectrum.
2.4. **OPTICALLY SUPPORTED MILLIMETRE-WAVE NETWORKS**

2.4.1 **Generation of Millimetre-waves by Coherent Mixing**

An analytical description of subcarrier generation by coherent mixing is presented. The treatment is not dependent upon the method of generating the two optical carriers but provides a general description.

Two optical fields of frequencies \( \nu_1 \) and \( \nu_2 \) can be written

\[
E_1(t) = E_1 \cos(2\pi\nu_1 t) \\
E_2(t) = E_2 \cos(2\pi\nu_2 t) \tag{2.5} \]

If these fields are coupled into a fibre and incident onto a photodiode, the resulting photocurrent (assuming a normalised receiver responsivity) will be given by

\[
I_{ph}(t) = E_{total}(t) . E_{total}^*(t) \tag{2.7}
\]

\[
= (E_1 \cos(2\pi\nu_1 t) + E_2 \cos(2\pi\nu_2 t))^2 \tag{2.8}
\]

\[
= \frac{E_1^2}{2} + \frac{E_2^2}{2} + \frac{E_1^2}{2} \cos(2\pi(2\nu_1)t) + \frac{E_2^2}{2} \cos(2\pi(2\nu_2)t) \\
+ E_1 E_2 \cos(2\pi(\nu_1 - \nu_2)t) + E_1 E_2 \cos(2\pi(\nu_1 + \nu_2)t) \tag{2.9}
\]

Of these terms only one appears at the difference frequency \( \nu_1 - \nu_2 \)

\[
I_{ph}(t) \propto E_1 E_2 \cos(2\pi(\nu_1 - \nu_2)t) \tag{2.10}
\]

Thus by controlling \( \nu_1 \) and \( \nu_2 \) so that \( \nu_1 - \nu_2 = \nu_{\text{millimetre}} \) we can generate the millimetre-wave frequency simply by mixing the two optical carriers at the photodiode. If we modulate one of the optical carriers with an information signal, we find that the modulation is translated onto the millimetre subcarrier. Rewriting the two optical fields with applied modulation \( m(t) \) as

\[
E_1(t) = m(t) E_{01} \cos(2\pi\nu_1 t) \tag{2.11}
\]

\[
E_2(t) = E_{02} \cos(2\pi\nu_2 t) \tag{2.12}
\]

Mixing at the photodiode and neglecting terms not at \( \nu_1 - \nu_2 = \nu_{\text{millimetre}} \) as before, gives

\[
I_{ph}(t) \propto E_1 E_2 m(t) \cos(2\pi(\nu_1 - \nu_2)t) \tag{2.13}
\]
Various modulation schemes may be employed: $m(t)$ may be amplitude or phase related, or may consist of sets of modulated subcarriers, which appear as sidebands centred on the mmw frequency following detection. Thus coherent mixing provides a flexible, inexpensive and robust RF generation technique, using a minimum of components at the receiver.

### 2.4.2 System Applications, Services and Topologies

The demand for optically supported millimetre-wave systems is being driven by the desire for tetherless access to broadband services and the congestion of the electromagnetic spectrum at lower frequencies. A major application then, is likely to be tetherless provision of BISDN as offered by the broadband fixed network, including telephony, video-conferencing, high speed data and telepresence. These services require a full duplex broadband link.

Specific applications and services planned for the millimetre-wave region include some of the wireless access services outlined in Section 2.2.2. Deployment of wireless LAN services is seen as a likely development of optical millimetre-wave technology. Installation of such a system would be relatively simple, with a fibre network running through a building, connecting antenna sites which would serve specific rooms or working areas. The advantages over conventional LANs would include mobility and flexibility of connection with high bandwidth. Automation and control of industrial and manufacturing environments using a millimetre-wave wireless LAN derivative, has many advantages. Equipment and computers could be coordinated with a fibre-network building or site using radio links to provide flexible deployment and mobility [11]. Millimetre-wave systems are also likely to be used in meshed point-to-point networks over LAN areas, associated with the growth of EFTPOS (electronic funds transfer at point of sale) systems. These networks would be deployed within a store or group of shops to verify and transfer payment from customer accounts [46]. These systems may be integrated with sophisticated security checks such as visual recognition and signature verification systems, requiring high bandwidth connections [47]. Another likely
application area is that of vehicular information delivery, with regard to traffic monitoring and guidance. Such a system would provide information to drivers in their vehicles about accidents, traffic congestion and delays. The system would suggest alternative uncontested routes. A network would be deployed with antenna sites situated along major roads and motorways and connected to central distribution points using a fibre network. Radio spectrum at 63GHz to 64GHz and 76GHz to 77GHz has already been reserved for traffic information systems in Europe by the EU. ¹ A similar system might be deployed to provide broadband communication to train passengers, with antenna units placed along the track side.

Network layouts for these systems will very much depend on both the nature of the application and the technology deployed, though many topologies are available using fibre due to the low transmission losses available. Passive optical networks lend themselves most readily to star and tree topologies, shown in Figure 2.5(a), since other configurations are difficult to achieve without active components. Tree topologies are particularly associated with PONs as they closely match the geographical nature of access networks. Techno-economic drivers of network layout include distribution of the cost of head-end equipment since this represents a comparatively high proportion of network cost. Economic topologies would thus serve as many antenna units as possible from each central node. Topologies based on configurations such as the star or tree would achieve this, though many more variations and hybrids are possible. Further economy is available through the use of WDM technology, which would enable existing fibre infrastructure to be used, minimising the requirement for additional cable infrastructure.

Final transmission network layouts are conformed by geographical features, such as road patterns for traffic information systems, while cabling is limited to routes which are easily developed. In addition, buildings, terrain elevation and type will affect millimetre-wave propagation patterns and must also be captured in wireless network planning.

¹Spectrum allocated under EU directive number COM (92) 341 final SYN 441.
CHAPTER 2. OPTICAL NETWORKS AND HFR SYSTEMS

(a) Logical tree (left) and star (right) topologies

(b) Cell sites connected by a PON in tree configuration

Figure 2.5: Mapping a logical topology to a transmission network
An illustration of the mapping of a logical tree topology to a transmission network layout is shown in Figure 2.5(b), using a traffic information system as an example. The antenna sites are positioned to provide maximum coverage in terms of the road network, while fibre is run along the road-side for ease of deployment, using passive splitters to produce branching where appropriate.

2.4.3 RACE Project MODAL

The MODAL (Microwave Optical Duplex Antenna Link) project was funded under the European RACE program and is designed to provide mobile access to a broadband telecommunication network [28, 48]. It is configured to operate at 30GHz in a region of the electromagnetic spectrum which is currently almost unused. Coherent mixing is used for optical millimetre-wave generation, with the optical carriers produced from the same laser source using a Mach-Zehnder modulator. A schematic diagram of the millimetre-wave generation technique is shown in Figure 2.6.

2.4.3.1 Optical Millimetre-wave Generation and Modulation Imposition

The MODAL system architecture consists of two modulator assemblies. One is required to generate the two frequency-displaced optical components; the other is used to impose modulated subcarriers onto one of these components.

The first modulator is voltage biased at the $V_T$ point. This has the effect of suppressing the original optical component generated by the laser and generating two new spectral components, separated by twice the drive frequency of the modulator [27]. A schematic diagram of the MZ with appropriate biasing is shown in Figure 2.3 on page 41. As we have seen, coherent mixing of these two optical components on a photodiode generates a photocurrent at the difference frequency, which is twice the drive frequency of the modulator, requiring the modulator to be driven at only half the required millimetre-wave frequency. This frequency
doubling technique has been demonstrated to generate 36GHz microwave signals from a single DFB laser [48].

Following generation, the two frequency components are separated using a Mach-Zehnder interferometric filter. A second Mach-Zehnder modulator is then used to modulate one of the components with the information signal, consisting of several subcarriers each modulated with a TDM of digital data. The modulating subcarriers are at frequencies centred at 1.5GHz, producing modulation sidebands at frequencies of 28.5GHz and 31.5GHz as shown schematically in Figure 2.6. Mixing on a photodiode produces a mmw signal that includes the modulated subcarriers, allowing the second modulator to operate at low frequencies, broadly commensurate with the modulation bandwidth. The arrangement thus avoids the need for either modulator to operate at the millimetre-wave frequency.

The remote antenna unit is designed to be as inexpensive and simple as possible and consists of a photodiode, bandpass amplifier and antenna. By using optoelectronic integrated circuit technology a single chip could be mass produced, reducing the cost further. Several users would be supported by a single remote antenna unit. Each user would require only a receiving antenna and down converter unit, with the relatively high cost of the head-end equipment shared between many users.

MODAL was conceived and completed as a system demonstration of the Mach-Zehnder millimetre-wave generation technique. As a demonstration, the project was less concerned with applications and system deployment than with device and component development. The project was envisaged as an enabling technology which might be applied to both WLL and mobile networks by using MODAL-like systems as the transmission infrastructure for linking cell-sites. MODAL ideas and technology are being developed further under the ACTS programme in the form of FRANS.
Figure 2.6: MODAL system of optical millimetre wave generation and modulation imposition
2.4.4 ACTS Project FRANS

The RACE mobile communications programme has now ended but many projects are being extended into further developmental phases under ACTS. This new programme draws on RACE ideas and technology with particular emphasis on developing systems to the point of demonstrator trials [49]. FRANS is the successor to MODAL, using the HFR technology developed in that project and an existing ATM PON [50] which has demonstrated delivery of BISDN services. The integration of these systems is planned on the basis of using MODAL derivative technology as the access interface to the PON core, replacing the existing HFC interface with HFR technology. FRANS thus aims to provide wireless, bi-directional, broadband services transported end-to-end via ATM.

A schematic diagram of the FRANS system can be seen in Figure 2.7. This shows the optical line termination (OLT) interface between the core ATM PON and the HFR access network and constituent components of a central node and remote antenna unit. Each central node is designed to serve up to 64 antenna units over fibre lengths of up to 12km. Each antenna unit will broadcast at 30GHz to a group of subscriber sites over distances up to 1-2km. In common with MODAL, customer premises equipment will consist of a directional fixed antenna and interface unit similar to existing FTTH equipment. Though FRANS is designed as a fixed access system, development of a mobile derivative should be possible with resolution of issues such as cellular hand-off and radio channel fading. Migration of the system to other mmw carrier frequencies, particularly in the 60GHz range, is also likely.

Two field trials are planned for the system, exploring different combinations of up-link and down-link transmission technologies. The first uses a 622Mb/s ATM radio down-link with a 40Mb/s aggregate TDMA radio return link. The second uses a 155Mb/s ATM radio down-link with a multichannel CDMA return link supporting up to 5 users at 2Mb/s per user. The access rates on the downlinks are to be achieved through the use of high order modulation schemes, in particular 16-QAM. Forward error correction is also employed to cope with the demands of
QAM transmission over a radio medium.

The first field trial is planned to take place in Lannion, France and the second in Stuttgart. The aim is demonstrate the feasibility of the system and verify that a quality of service can be achieved that is comparable to a FTTH network.

### 2.5 Summary

The aim of this chapter has been to provide a review of aspects of telecommunication technology which are relevant to the project work. The advantages of optical
networks and broadband wireless networks have been discussed together with the portion of these technologies in the wider context of services, applications and alternative technologies. The effect of reflections in optical networks was discussed and interferometric noise singled out as potentially limiting in terms of system performance. Through a brief analysis IN was shown to have a greater impact on the received signal than linear crosstalk. Drawing from these ideas, the chapter finally focused on optically supported millimetre-wave wireless access networks as an important developing technology. The primary advantages of these networks were identified as high bandwidth, rapid deployment, flexibility of operation and low cost. The problems of optical millimetre-wave generation were discussed and coherent mixing was highlighted as a method producing potentially narrow linewidth signals and simplicity of operation. Two demonstration projects were described, MODAL and FRANS, together with a novel method of optical mmw generation which has been employed in both systems.
Chapter 3

Simulation Tools and Techniques

3.1 Introduction

This chapter provides a review of approaches to digital computer simulation of optical communication networks and the various software environments which support such work. Section 3.2 provides an overview of available techniques and discusses the advantages of software development within an existing commercial package. Some proprietary packages are then described before Section 3.3 describes the SPW platform in some detail, listing the features which led to its choice as the basis for simulator development and in particular its object-oriented network design facility. Methods for modelling optical signals are addressed in Section 3.3.1 and the facility for embedding custom code into SPW is reviewed in Section 3.3.2. Previous work using SPW in the optical network field is reviewed in Section 3.3.3 and the chapter concludes with a summary.

3.2 Simulation Platforms

Having identified a requirement for the simulation of optical communication networks to provide cost-effective and efficient performance assessment of potential systems, an appropriate methodology for computer simulation of optical net-
works at the physical level was now required. It was decided to use the Signal Processing Worksystem package as the development environment, not least because a considerable development effort had already established SPW with an optical networking capability. The package was originally chosen for a number of reasons, which this section will address.

Several development options were available for optical network simulation, and the most fundamental decision concerned the choice of either adapting of an existing simulation platform or writing software from scratch as a stand-alone environment. The obvious advantage of using a pre-developed and supported platform was that development time would be much shorter. A commercial package should offer a user-friendly interface and pre-developed libraries of simulation tools, ready to run in the environment. Provided a prospective package did not compromise the performance or flexibility required, it would be clearly advantageous to adapt an existing platform.

Initial simulation studies of communication systems using digital computers employed command line and interpreted language environments. These cumbersome tools contrast with contemporary packages which employ graphical user interfaces, exploiting modern windowed operating systems and powerful workstation and personal computer hardware. Several proprietary simulation packages were available for development, though none of these systems directly supported optical network modelling. Thus an additional selection criteria was the support of custom code, allowing additional functions to be purpose written and embedded into the package to enable optical systems modelling.

The main types of proprietary package available can be defined as either technical computing language environments or simulation environments. Simulation environments for telecommunications modelling further sub-divide to support network modelling at various architectural levels (such as those defined in the SNA and ISO architectures [51]).

MATLAB is an example of a technical language environment and has been used extensively in analytical work in this project [52]. It is available for both PCs and
workstations and uses a C-like expression language with matrix based representation and manipulation of data. Graphical output has facilities for 2D, 3D and animated figures. Various add-on tool-boxes are available to provide additional functionality. Examples of tool-boxes include SIMULINK which enables diagrammatical representation of time-sampled simulation code using a dynamically interactive windowed interface. The SIMULINK environment is very powerful and includes many of the features required for the proposed optical network simulator with a large library of pre-developed code, support for custom code and hierarchical programme construction.

Many dedicated simulation packages are available and typically support a specific architectural level of telecommunication system operation. Platforms for simulation of network and transport layers are commonly available and these model the flow of traffic through network topologies for given routing and control algorithms. Examples include OPNET modeller [53] and BONES, produced by the Alta Group [54]. BONES uses block diagram construction of network paths, switching and queueing elements and may be linked to lower physical level models produced in SPW, its sister package [55]. SPW is designed to be a general purpose signal processing platform, primarily designed for modelling electrical circuit systems at a physical behaviour level. SPW includes purpose written libraries specifically for the simulation of communications systems. A full description of SPW follows in Section 3.3.

Given these requirements, SPW has been selected by the research group as the most appropriate development environment for optical network modelling [20]. Thus the project adopted SPW together with extensive Alta library code and built on a platform of custom code already established by the group to support optical network modelling [20, 24].
3.3 Description of the SPW Environment

SPW is a graphically driven simulation package that is compiled to run under the UNIX OS. It is comprised of two sub-environments, the block diagram editor (BDE) and signal calculator. The BDE window is where systems are constructed from block icons used to represent signal processing functions. Blocks are, which are linked together to emulate signal flow through the network structure. Most obviously the BDE utility presents the system designer with an intuitive, mouse-driven, windows-based interface to the simulator. However, the design of the BDE and its combination with the other elements of SPW provide many other important features. These include:

- Graphical representation of network construction and signal flow, providing:
  - an intuitive method for system construction;
  - quick and easy "cut and paste" editing and reorganisation of network configurations;
  - easy and ordered editing of system / component parameters.

- A 'windowed' interface, providing:
  - a familiar environment for launching programmes, providing input, reading dialogue etc.;
  - several designs to be open for editing or simulation at once;
  - viewing / editing of hierarchical structure through sequentially nested windows;

- Provides a network editing utility distinct from data generation and analysis; these tools combine with the simulation engine to provide several simulation modes.

- An 'object-oriented' paradigm to network construction, achieved by:
  - allowing hierarchical network construction to an unlimited depth;
3.3. DESCRIPTION OF THE SPW ENVIRONMENT

- modular representation of code and system construction;
- encapsulating data;
- providing a clear representation of data flow through the system;
- classification of distinct data types and facility to cast others.

The inclusion of the OO features listed above gives the BDE a very powerful approach to network construction. Specifically the OO paradigm enables large networks to be constructed while minimising problems associated with data corruption and mis-direction. In addition, network complexity is managed, allowing large structures to be visualised and manipulated more easily. Traditional, procedural approaches would otherwise limit the size and complexity of systems. A screen-shot of sequentially nested hierarchical windows in the BDE is shown in Figure 3.1.

SPW also includes an extensive library of signal processing functions, comprising many standard modules, while optional add-on libraries extend these basic functions to encompass more specialised simulation areas. The basic libraries include: signal sinks and sources, for writing to and from data files; signal generators; numerous mathematical functions; vector handling blocks and vector transform blocks. The vector transforms include fast Fourier transform (FFT) and inverse FFT blocks, allowing signal representation to interchange between time and frequency domains at any point in the network. An example of an optional library is the communications module which includes specialist function blocks specifically designed for use in designing and simulating communications systems. The module covers areas such as: modulation and demodulation; encoding and decoding schemes; specialist filters; estimators and random generators. Library blocks may be combined together using the hierarchical facility, to produce extended functionality attributed to a single block symbol representation, broadening the scope pre-developed code.

The generation and analysis of signals is handled by the Signal Calculator utility which provides data manipulation, pre- and post-simulation. Signals can be represented in both serial and vector form. A vector signal passes the data in a
parallel format allowing many signal values to be processed at a single data port during the same sample period, providing the capability for multi-dimensional vector calculations. SPW also reserves a specially formatted two-dimensional array for representation of complex signals. A built in analytical sub-environment allows signal waveforms to be viewed as eye-diagrams, distribution histograms or transformed using FFTs and IFFTs.
3.3. DESCRIPTION OF THE SPW ENVIRONMENT

As mentioned in Section 3.2, SPW can be linked to the BONES simulator [54]. This provides a facility for modelling at higher network levels while providing the facility to embed detailed physical-level description in the network structure. SPW has a further facility for dynamically linking MATLAB programmes into SPW designs, allowing co-simulation and sharing of data files between the environments. MATLAB may thus be used to specify algorithms in the design that are suited to the textual format it uses.

3.3.1 Modelling Optical Signals

In considering the sampling requirements for an optical transmission system, it is clear that simulation of the optical carrier electric field (at around $10^{14}$ Hz) would require prohibitively long simulation times using contemporary workstations. Resolution of typical modulation signals however, requires a disparately smaller sampling rate. To quote an example by BT researchers [56]:

Consider, as an example, a 1 1550nm DFB source modulated at 300Mbits/s. If there is no laser chirp, the highest information frequency is 150MHz which leads to a Nyquist sampling rate of 300MHz. To simulate 128 bits will therefore require 128 time samples. If the source chirps by 0.2nm, the highest information frequency rises to 25GHz giving a Nyquist rate of 50GHz. To simulate 128 bits will now require $\approx 21500$ time samples. If a wavelength multiplexed network and/or a system transmitting amplifier noise covering a bandwidth of 60nm is now considered, the Nyquist rate becomes 7.5THz. To simulate the 128 bits will now require at least 3.2 million time samples!

It is clear then that modelling a base band signal, either a sub-carrier frequency or the information modulation itself, relative to the optical carrier frequency and with measured chirp, is required. This base band signal is then passed through the system as a scalar power signal and/or complex electric field, depending on the behaviour to be captured by the model.
The problem of modelling multiple-wavelength signals in WDM systems and FDM systems has been addressed by the group and other researchers [20, 24, 56]. The approach adopted for the project models signals in the time domain with the addition of an additional vector element to each signal for frequency identification. The resulting vector components are then modelled independently through the system. The frequency response of devices is modelled by using FFT conversion between the time and frequency domains. This was found to be the most effective approach since other methods for direct combination of vectors in the frequency domain was too restrictive in terms of relative bandwidths and ease-of-implementation.

Other methods have been developed for modelling optical systems in SPW. These are similar to the adopted approach in the respect that all methods model optical signal components independently through the network in order to reduce the required number of samples. So similarly, components are only combined when propagation through non-linear or other signal-dependent components is required, while frequency dependent elements are handled using vector transforms. Actual signal representation differs between methods however. An alternative approach developed at University of Wales, Bangor appends a sampling frequency value and centre wavelength value to each complex spectral signal component. Thus formed, the vectors from each wavelength channel are then multiplexed together in the time domain to form a 'WDM vector' [20]. Contribution from ASE produced by optical amplifiers in the network is appended to this concatenated signal to form a final vector which is transmitted through the network. Since sampling and noise information is contained within each transmitted vector, the signal may be reconstructed at any point in the network.

As we have seen, British Telecom researchers also acknowledge the problems associated with physical-level modelling of broadband optical systems using a signal-based environment such as SPW [56]. The BT approach again models optical signal components independently through the network, treating signal elements such as ASE, polarisation, electric field and individual WDM signal channels entirely separately [56]. Libraries of custom code, representing various
optical network elements, have been developed at BT using their own signal representation.

3.3.2 Custom Coded Blocks

Since SPW does not directly support optical network modelling, purpose written custom code must be written to provide the functionality of various optical network components which cannot easily be represented by library functions. A facility is provided within SPW to allow custom code, written in ‘C’, to be embedded into the BDE utility for use in system designs.

In order to embed the code into the BDE graphical environment, a block symbol for representation in the editor must be designed. This symbol includes all the inputs and outputs to the system hierarchical level and these I/O ports are defined by the data type they will handle. Editable parameters are listed in a separate screen and linked to the block symbol. This enables the user to define parameter values by double-clicking the block symbol in the system design and editing the instances that appear in the nested child parameter window. The source code itself is written to a specified structure and inserted into a template created by a compiler-like facility that considers the constituents of the symbol and parameter screens. Macro commands are defined within SPW for handling complex signals and functions, while other macros provide standardised functions to ensure custom code compatibility with future SPW releases. The source code is then compiled to a binary image using GCC and the block dynamically linked to the SPW libraries for testing and subsequent use in system design.

The source code structure required of CCBs by SPW did impose some constraints on the development of code in comparison to producing stand-alone models from scratch. This was expected, and acceptable, since development flexibility was not impaired so as to significantly encumber the environment. The code structure imposed by SPW delineated the programme into three sections; the initialisation function, the run output function and the termination function. The initialisation and termination sections are run once only for each simulation, as the first and
last executions respectively. The run output function is executed once for each sample in the simulation. Specially defined variables handle data sharing between the functions and also to the system environment via the input and output ports of the module. This structure is relatively easy to achieve if the program is written this way from inception. Porting programs to the environment which were written as stand-alone executables however, proved difficult.

Additional problems were encountered during the course of the project with regards to interaction between CCBs and pre-developed library code. Since library modules are not supplied with source code, their precise behaviour is undetermined. Problems occurred with aspects of library block behaviour which were not guaranteed by Alta Group. These problems are preemptively checked in each module and the user alerted. Details of error conditions and warnings can be found in the software documentation in Appendix B on page 145.

3.3.3 Previous Optical Network Modelling Using SPW

To date, several groups have used SPW to model optical communication systems [56, 57]. Within the UCL telecommunications group (and previously within the Bangor Electronic Engineering Department) there has been a great deal of optical network modelling development built onto the SPW platform. These projects have produced a library of optical components embedded into the SPW environment as CCBs linked to the standard libraries, and were available for use at the project outset. Optical components that have already been modelled within the group include:

- Optical sources, including lasers, white light, WDM array sources;
- Optical amplifiers, including EDFAs, SOAs;
- Network elements, including optical multiplexers, and demultiplexers, fibre, filters;
• Specialised elements, such as a Mach-Zehnder modulator for the FRANS project.

Specifically, work has been centred around the development of a basic simulation environment [20, 24] and its subsequent application to issues of network performance. Examples of network design studies that have been addressed using SPW are: the effects of homodyne beat noise on WDM network performance [57, 58]; fibre-radio hybrid systems [59, 60] and the modelling of fibre-radio CDMA spread-spectrum systems [61].

3.4 Summary

In this chapter we have discussed our requirements for an optical network simulator, reviewed the software development options available and commented on the advantages and drawbacks of using a commercial package as a development platform. Some of the available package types were reviewed with examples of textual environments and network-level simulators given. We then determined the suitability of of the SPW simulation package by Alta Group as an appropriate development platform in the context of some of the available features, particularly the object-oriented BDE utility and provision for including custom code. Some basic ideas for the modelling of optical signals was then presented, commenting upon previous work in using SPW in this way, particularly with regard to the problem of WDM signal representation. The procedure for embedding CCBs into the package was described with a view to presenting the following chapters. Finally previous work in the area of optical network modelling using SPW was briefly reviewed and types of library CCB developed during the course of these studies was summarised. The next chapter describes the method used for modelling linear crosstalk effects, by the development of CCBs.
Chapter 4

Simulation of Optical Feedback

4.1 Introduction

This chapter describes the modelling and simulation work completed with respect to capturing the effects of linear optical feedback into network devices. The cases chosen for study were those of the semiconductor laser and Erbium-doped fibre amplifier (EDFA). Both these devices were implemented as CCBs (see Section 3.3.2) in SPW. The semiconductor laser block was written from scratch, while the EDFA model was modified from existing modelling work undertaken by S. L. Zhang [24, 25]. Section 4.2 is an overview of the general techniques employed in the modelling of feedback issues, in particular the neglection of beat terms. Sections 4.3 and 4.6 give overviews of the laser and EDFA models respectively. Some example simulation results are presented in Section 4.4 and the chapter is concluded with a summary.

Full documentation covering design and use of the optical networking modules described can be found in Appendix B.
4.2 Optical Feedback Modelling Considerations

From Equation 2.2 we saw that the effect of the addition and detection of two optical fields was the addition of the power of the two lightwaves together with an additional term which appears as a beat signal, driven by the phase disparity between the lightwaves. As mentioned in Section 2.3.4, if the two waves are incoherent with respect to each other the additional component appears as a beat signal driven by the frequency disparity between the fields. When considering optical feedback however, the beat term can contribute negligibly, since many optical components will reject the signal, depending on its frequency.

In the case of the semiconductor laser, a resonant frequency for modulation of the output intensity is determined by the lifetime of electrical carriers and photons within the active region of the device. A typical DFB laser will have an intensity modulation resonance of around 20GHz, corresponding to a carrier lifetime of around $5 \times 10^{-11}$s. If the optical feedback is displaced with respect to the laser emission frequency by more than this resonant frequency, the laser will be negligibly affected. This corresponds to a difference of about 1.6nm in wavelength at an operating wavelength of $1.55\mu m$. If optical feedback originates from a separate source it must be propagating at a wavelength displaced by less than this value from the laser wavelength. If the laser is experiencing feedback from its own output, reflected by imperfect network components, then this will also be displaced in frequency by some amount from the current emission wavelength.

In the case of direct modulation, the laser will be undergoing carrier-induced frequency excursions at periods determined by the modulation rate. If the modulation is digital and has a period less than the round-trip time to the reflective network element, the returning lightwave will be at a different frequency to the current emission wavelength. Induced chirp limits feedback round-trip times to the data rate for RZ codes. In the case of NRZ codes, feedback round-trip times are limited by the run-length of consecutive identical symbols, though this is also likely to be limited by the application of techniques such as line coding [62]. For
an EDFA, the lifetimes of photons in the doped fibre length is of the order of $10^6 \text{s}$, almost completely negating any beating effects.

In general, effects caused by coherent feedback are ignored in the modeling here. Any light reflected back into the laser cavity and maintaining coherence with the laser output will produce non-linear coherent interference which will potentially dwarf the effects captured in the simple treatment offered here. Though feedback is modeled on a simple linear power-coupling basis, limited to incoherence between the reflected signal and the laser, it provides a computationally efficient modeling methodology, which is simple to implement in system design and gives can useful results in the conditions previously stated.

### 4.2.1 Feedback Generation CCBs

Several new CCBs were created to simulate optical feedback from the network. Each of the network devices, which were modeled under the influence of optical feedback, were written with a system-level input configured for photon injection. Blocks were created to simulate groups of reflections or backscatter from the network. The optical power output from these blocks can simply be summed and applied to the feedback input of the device models. The CCBs were written in vector format in the frequency domain to allow many wavelength channels to be handled simultaneously and broad optical spectra to be modeled. A BDE diagram of the DFB laser model with feedback from a multiple reflection generator and Rayleigh backscatter generator is shown in Figure 4.1. The function of the feedback generation CCBs shown is as follows

**Vect_fibredly** simulates multiple reflections of a broad spectrum optical signal from up to 6 reflections in the network. The delay time associated with the each of the returning signals is determined by the value of 6 block inputs. These allow the distances to each reflection to be determined dynamically during the initialisation of a system. A disconnected fibre length input results in a default delay, set in the parameter screen. A global input for sampling frequency is included together with refractive index and reflectivity parameters.

**Rayleigh_scatt** generates the backscattered optical power produced by Rayleigh scattering from the network fibre. It has inputs for fibre length (allowing
dynamic initialisation), sampling frequency, average input power (in vector form) and wavelength comb, for frequency definition of the input power vector. The comb is of the same format described in Section 3.3.1. The block calculates the backscatter at each wavelength sample of the input vector, assuming a constant average power distribution along the fibre length. The parameter screen includes a 'yes / no' string for initialising the fibre to its full length. The simulation otherwise models the time required for the light to propagate through the fibre and the backscattered power to return to the fibre end, providing quick estimations of backscattering.

**Vect.blockmean** calculates the average of an input vector, using a simple 'block mean' method. The code also checks for an over large sample count and 'cancellation' due to restricted storage of data types. In the event of these effects occurring, the block is held in its current output state and a warning message printed in the SPW system view-port.

**Vect.incatten** calculates the integral of the attenuation along the fibre axis. Since the scattered light propagates from every point along the fibre length, the total scattered power from a fibre experiences the integral of the fibre attenuation as it propagates back along the fibre. To provide compatibility with Rayleigh.scatt the module may similarly be used incrementally (using the propagation distance of the light in the fibre as the upper integration limit) or initialised to the full fibre length.

**Rayleigh.dist** block is a second method developed for generation of Rayleigh backscattered optical power, which uses a dynamic representation of the signal propagating within the fibre. The module is described in more detail below.

Other CCBs were also developed, including some general vector manipulation blocks not specific to optical network modelling.

The backscatter calculation shown in Figure 4.1 comprises three of the above CCBs: The mean of the optical power is calculated and applied to the scattering calculation block, which produces backscattered power at each optical frequency
4.2. OPTICAL FEEDBACK MODELLING CONSIDERATIONS

Figure 4.1: BDE diagram showing application of optical feedback to DFB laser model

spacing to which an integrated attenuation is applied. This method takes no account of the distribution of optical power within the fibre.

The Rayleigh_dist module uses a more accurate method of Rayleigh backscatter generation which models the fibre as a series of concatenated sections, capturing the dynamic power distribution propagating its length. The optical power launched into the fibre is stored in a two dimensional circular-buffer which spatially represents the round-trip time to the end of the scattering fibre and back and spectrally represents the signal bandwidth. At each sample every second element in the buffer is summed, starting with the second element, with attenuation proportional to the round-trip to the centre of each fibre section. This produces a total backscattered power at the end of the fibre with delay and attenuation applied dynamically.
This method of backscatter calculation can prove very computationally intensive depending on the spatial resolution required. The spatial resolution of the calculation (i.e. the length of each fibre section) is set in the parameter screen, allowing a user-defined trade-off between accuracy and computation time.

4.3 Semiconductor Laser Model

The semiconductor laser was modelled through solution of rate equations for the carrier and photon populations in the active region of the device. The treatment assumes a single longitudinal mode and a single spatial mode of operation and applies the rate equation solutions to the entire laser cavity [63]. This provides a simple yet effective model which has been widely applied [20, 64, 65].

To capture the effect of linear crosstalk an additional term was added to the photon population rate equation. This represented, in the first instance of the model, the effect of a reflection from a splice at the end of the fibre pigtail included with the diode laser package. With the addition of this term the photon population rate equation for the active region becomes

\[
\frac{dP(t)}{dt} = G \cdot P(t) - \frac{P(t)}{T_P} + R_{sp}(t) + S(t)
\]

where \( P(t) \) is the photon population, \( G \) is the device gain, \( T_P \) is the average photon lifetime and \( R_{sp}(t) \) is the number of photons produced by spontaneous emission. \( S(t) \) is a feedback injection term generated by a single reflection travelling back into the cavity and is given by

\[
S(t) = P(t - \tau) R_{splice} \frac{L V_g}{4} \ln \left( \frac{1}{R_1 R_2} \right)
\]

where \( \tau \) is the time delay associated with a round trip of laser light from the cavity to the splice and back, \( R_{splice} \) is the reflectivity of the splice, \( R_{1,2} \) are the reflectivities of the cavity ends, \( V_g \) is the group velocity and \( L \) is the cavity length. The feedback was handled as linear crosstalk stored in a circular buffer, dynamically sized at run-time to temporally represent the round-trip to the splice in the pigtail fibre.
The rate equation for the carriers in the active region is given by

\[ \frac{dN(t)}{dt} = \frac{I(t)}{q} - \frac{N(t)}{T_n} - GP(t) \]  

(4.3)

where \( N(t) \) is the carrier population, \( I(t) \) is the injection current to the active region, \( q \) is the charge on an electron and \( T_n \) is the lifetime of an electron in the active region. The device gain \( G \) is given by

\[ G = \Gamma V_g a \left( \frac{N}{V_a} - N_0 \right) \left( 1 - \epsilon \Gamma \frac{P}{V_a} \right) \]

(4.4)

and the spontaneous emission rate \( R_{sp} \) is given by

\[ R_{sp} = \frac{\beta_{sp} \Gamma N}{T_n} \]

(4.5)

where \( \Gamma \) is the optical confinement factor, \( V_g \) is the group velocity, \( a \) is the differential gain coefficient, \( N_0 \) the carrier density at transparency and \( \beta_{sp} \) the fraction of spontaneous emission coupled into the lasing mode.

By considering these expressions term-by-term, we can intuitively relate them to the physical processes involved in laser operation. In terms of the photon population given by Equation 4.1, we can see that the first term gives the photons produced by spontaneous emission, the second the loss of photons moving through the cavity in terms of average photon lifetime, the third term photons produced by stimulated emission and the final term the additional photon injection due to optical feedback from the network. In terms of the carrier population, given by Equation 4.3, we can attribute the first term to the number of carriers injected into the junction, the second gives the carrier loss through spontaneous photon emission and the last term gives the carrier loss due to stimulated photon emission.

The coupled rate equations are solved simultaneously, using numerical methods to give the instantaneous photon and carrier populations in the active region. The optical power output for each facet, \( P_f \), of the laser cavity can be related to the photon population by

\[ P_f = \frac{P(t) h c V_g \alpha_m}{2\lambda_0} \]

(4.6)
where $c$ is the speed of light (in vacuum), $h$ is Planck's constant, $\lambda_0$ is the wavelength of operation and $\alpha_m$ is the loss from the cavity end facets, given by

$$\alpha_m = \frac{1}{2L} \left( \frac{1}{R_1 R_2} \right)$$  \hspace{1cm} (4.7)

where $R_1$ and $R_2$ are the reflectivities of the end facets, defining a cavity of length $L$. This is simply the loss from a Fabry-Perot cavity due to imperfect reflectivity at the end facets. Equation 4.6 is again intuitively obvious if we note that $V g \alpha_m$ is the rate at which photons are incident upon, and thus transmitted through, the end facet mirrors. Thus Equation 4.6 provides the power output of the laser model. Though optical power output is obviously a necessity for our laser model, in order to capture phase related effects we must consider the propagation of the light as an electric field.

### 4.3.1 Phase Characterisation

To generate an electric field output from our laser model we must calculate the electric field amplitude produced and also characterise the phase behaviour of the laser. The output power from a facet $P_f$, given by Equation 4.6, can be related to the propagating electric field $\bar{E}$ by

$$\bar{E} = \sqrt{P(t)} \exp[j\phi(t)]$$  \hspace{1cm} (4.8)

where the instantaneous angular phase $\phi(t)$ is given by

$$\phi(t) = 2\pi \int_0^t \Delta \nu(t')dt'$$

$$\phi(t) = 2\pi \int_0^t \Delta \nu(t')dt'$$  \hspace{1cm} (4.9)

where

$$\Delta \nu(t) = \frac{\alpha}{4\pi} \Gamma V_g a \left( \frac{N}{V_a} - \frac{\bar{N}}{V_a} \right) \left( 1 - \Gamma \varepsilon \frac{P(t)}{V_a} \right)$$  \hspace{1cm} (4.10)

where $\alpha$ is the linewidth enhancement factor of the laser and $\bar{N}$ is the steady state value of the carrier density in the active region at a defined reference wavelength of operation.

An inherent characteristic of any semiconductor laser is that direct intensity modulation will cause the laser to 'chirp'; that is a corresponding modulation of
the frequency of the laser output. Chirp produced by the model was calculated relative to the reference frequency defined for \( \bar{N} \) in Equation 4.10. The chirp behaviour of the laser is discussed in more detail in Section 4.4.

### 4.3.1.1 Phase Noise

The instantaneous phase of the laser, as given by Equation 4.9, assumes a perfectly monochromatic output with no linewidth. In an actual semiconductor laser however, linewidth broadening is always observed. The salient contribution to laser linewidth is the presence of phase noise. In some cases the phase noise can become a limiting impairment and so we need to capture its characteristics in the model.

Phase noise is driven by random fluctuations in the phase of the output light caused by the spontaneous emission process in the cavity, inherent to laser operation. Each spontaneous emission event causes a small jump of random magnitude and sign in the phase of the electric field output by the device. Viewed over time, the effect of these spontaneous events is that the phase executes a random 'walk' away from its initial value; the mean-squared deviation is approximately linear with time. Figure 4.2 shows the simulated phase noise excursion of a laser with a linewidth of 16MHz over 0.05\( \mu \)s: the plot also provides an indication of the coherence time of the laser.

We can describe the phase noise of a laser as a Wiener-Lévy random process, the instantaneous phase of which is again given by Equation 4.9 [66]

\[
\phi_{pn}(t) = 2\pi \int_0^t \Delta \nu(t') dt'
\]  

(4.11)

where the instantaneous frequency \( \phi_{pn}(t) = 2\pi \int_0^t \Delta \nu(t') dt' \) is determined by a Gaussian distribution [67]. We can introduce the phase noise \( \phi_{pn} \) to the laser model by simply adding \( \phi_{pn} \) to the phase component of the output electric field and in SPW the random phase fluctuation was produced using a Gaussian random generator function. The variance of the phase distribution is determined by the linewidth of the laser [68]

\[
\sigma^2_{\phi_{pn}}(t) = 2\pi \Delta \nu_{laser} t
\]  

(4.12)
This is intuitively obtained by visualising the random walk executed by the phase, comprising random spontaneous emission events. Using Equation 4.12 we can see that for a laser of linewidth 16MHz and sampling frequency of $10^{12}$Hz, we obtain a variance for the Gaussian generator of around $10^{-4}$.

Figure 4.3 shows a BDE diagram of the laser CCB with phase noise added to the electric field output using library function blocks.

### 4.3.2 Implementation of the Model

Now that we have established the coupled rate equations and expressions to generate outputs of optical power, chirp and electric field, we must concern ourselves with implementing these expressions into a programme to produce numerical solutions. Since the rate equations 4.1 and 4.3 are to be solved numerically, initial values for the variables must be found. Due to the non-linear behaviour of the
coupled equations, accurate starting values must be calculated to obtain consistent results from the model. A first estimate of the initial conditions can be found by solving the rate equations at \( t = 0 \). Using this starting solution, steady state values may then be calculated for a given drive current. If we put \( t = 0 \) into equations 4.1 and 4.3 (and assume there is no carrier diffusion) we can rearrange them to give

\[
N^2 \frac{\Gamma V_g a}{T_e V_a} (1 - \beta_{sp} \Gamma) + \frac{I}{q} \left( \frac{1}{T_p} + \Gamma V_g a N_0 \right)
+ N \left( \frac{\Gamma^2 V_g a N_0 \beta_{sp}}{T_e} - \frac{1}{T_e/T_p} - \frac{\Gamma V_g a N_0}{T_e} - \frac{I \Gamma V_g a}{q/V_a} \right) = 0
\]

which we can solve for \( N \) and hence obtain \( P \)

\[
P = \frac{\beta_{sp} \Gamma^2 N}{T_e} \frac{1}{T_p - \Gamma V_g a \left( \frac{N}{V_a} - N_0 \right)}
\]

Using these starting values we can iteratively solve the rate equations until a steady state value is reached. This is done at the reference wavelength, using the ‘current at reference wavelength’ parameter to obtain \( \bar{N} \). The process is then repeated for the specified bias current to obtain starting values for \( N \) and \( P \). These values are substituted into the rate equations and solutions are now found under application of the drive current. At each sample period the drive

Figure 4.3: BDE diagram showing application of phase noise to laser electric field output
current input is read and the rate equations iterated at the specified rate (see below). The resulting values calculated for each output are written to the system environment at the end of the sample period.

The equations are coded in the 'C' language and embedded into SPW using the CCB creation procedure described in Section 3.3.2. The CCB of the laser model has inputs of drive current and sampling frequency, which allows the sampling time of the system to be determined dynamically. The drive current input is offset by the bias current specified in the parameter screen. A low-pass filter can be connected to the drive input to modify the frequency response of the modulating signal. An optical input port allows feedback from the network to be generated at the system level using the generation CCBs described previously. Outputs are power and chirp (which are scalar values) and electric field (a complex vector). The CCB symbol representation as it appears in the BDE is shown in Figure 4.4(a). Parameters were made user definable according to those deemed most relevant in the specification of a semiconductor laser. The parameter screen, which is accessible from the BDE by double clicking the CCB symbol, is shown in Figure 4.4(b). Parameters pertaining to laser specification (such as active volume, linewidth enhancement factor and wavelength of operation) are grouped together under the heading 'LASER'. This includes a 'solution frequency' which defines the frequency at which the rate equations are iterated. Though user-definable this is ultimately determined by the average carrier and photon lifetimes in the active region and is separate from the sampling rate at which the top-level system design is simulated. Twelve reflections in the network can be described in terms of reflectivity and propagation distance. These are grouped together under the heading 'SYSTEM REFLECTIONS'. Parameters for calculating backscatter are grouped with values describing system and network attributes under the heading 'FIBRE AND SYSTEM'. Two yes/no options are also included to disable the feedback calculations in the code. This enables the block to be used without feedback modelling or for the optical input port to be used to model feedback effects.

Now that we are considering multiple reflections and backscattering the photon injection term $S$ in Equation 4.1 becomes a sum of all the photons travelling back
4.3. SEMICONDUCTOR LASER MODEL

(a) Laser symbol

(b) Parameter screen, showing typical values for a DFB implementation

Figure 4.4: BDE symbol and parameter screen for laser CCB
into the cavity:

\[ S_{\text{total}}(t) = S_{\text{scatt}} + \sum_{i=1}^{N} S_i(t + \tau_i) \quad (4.15) \]

where \( S_i \) are back-reflections produced in the network. \( S_{\text{scatt}} \) is the total Rayleigh-backscattered photon number reaching the cavity from the network, and is calculated in the CCB by the summing the contributions from \( N \) small sections of fibre

\[ P_{\text{scatt}} = \sum_{n=0}^{\infty} P(t - \tau_n) \left( 1 - e^{-\gamma_R T_{\text{sam}} c/n_2} \right) \left( 1 - \cos(\sin^{-1} \frac{n_1}{n_2}) / 2 \right) / 10^2 \alpha x_n / 10 \quad (4.16) \]

where \( \gamma_R \) is the Rayleigh scattering factor for the fibre, \( T_{\text{sam}} \) is the internal (laser) sampling period, \( n_1 \) and \( n_2 \) are the indices of the fibre core and cladding respectively, \( \alpha \) is the attenuation of the fibre (in dB/km) and \( x_n \) is the distance from the laser cavity to the fibre section. By treating the backscatter in this way the distribution of optical power in the fibre is captured, providing a dynamic scattering calculation along the fibre length, as described in Section 4.2.1.

The optical feedback input port appears in the rate equations as a further photon injection term and allows the feedback generation CCBs described earlier to be used as an alternative, or in addition to, the internal feedback calculations.

### 4.4 Simulation of Laser Behaviour

As we have seen from Equation 4.10 direct intensity modulation of a diode laser, also produces a frequency modulation. This is an inherent aspect of semiconductor laser operation that is caused the dependence of the refractive index of the active region on the electrical carrier density. The drive current can therefore be used to produce both FSK and ASK of the laser output, with the dominant form determined by the characteristics of the drive current modulation. This effect also requires the system sampling frequency to be carefully set when directly modulating the laser, since induced chirp is likely to be much greater than the modulation bandwith. The maximum chirp excursion is dependent on the modulation depth of the drive current though it is typically around 20GHz.
Simulations were carried out to assess the impact of light reflected back into the laser cavity using various combinations of reflectors. Initial studies assumed a single reflection occurring at the splice between the fibre pigtail from the laser package and the network fibre. Direct digital modulation was applied to the model, with a period less than that of the propagation time to the pigtail splice and back. Various reflectivities at the splice were simulated.

Two reflection induced effects were observed in the laser output, as expected; intensity modulation and induced frequency chirp. For OOK, direct modulation, the reflected feedback is proportional to either a zero or unity symbol power level. This leads to four possible combinations of transmission state in the presence of a single reflector: A transmitted one, with crosstalk from either a one or zero; and a transmitted zero, again with crosstalk from either a one or zero. Of these four states, eye closure will be produced by the bottom of the eye being raised, corresponding to a transmitted zero and reflected unity symbol.

Eye-closure due to crosstalk on a transmitted zero is dependent on laser operating...
Figure 4.6: Eye diagrams of laser output power

(a) No applied optical feedback

(b) Fresnel reflection from pigtail splice
conditions. Figure 4.5 shows the amount of eye-closure as a function of laser bias current (added to the modulation signal), for a crosstalk isolation of 0.04. Eye-closure is at a maximum when the laser is operating close to threshold, but becomes constant as drive current is increased. Figures 4.6(a) and 4.6(b) show eye diagrams to illustrate optically-induced eye closure from a single Fresnel reflection compared with a fully isolated device. The transmitted eye represents filtered data and a laser biased at $\approx 10\%$ above the threshold drive current.

The effect of a single Fresnel reflection from the splice at the end of the fibre pigtail on the laser output is shown in Figures 4.7(a) and 4.7(b). Laser behaviour with no feedback applied is shown as dotted lines. Comparison of the two figures shows that intensity modulation driven by optical injection behaves similarly to current injection into the active region. Feedback induced chirp however, is driven in the opposite direction to current induced chirp. This occurs because the injected photons deplete the carrier population in the active region, lowering the refractive index. Induced chirp increases in range to $\approx 12\text{GHz}$ in the presence of a Fresnel reflection, corresponding to a wavelength excursion of approximately 0.1nm at 1.55\text{$\mu$m}. The laser is also driven to a different central operating frequency. These effects are particularly significant in systems employing HDWDM or FDM where inter-channel crosstalk may be induced. Multiple reflections increase chirp linearly with respect to the total returning power, independent of the number of crosstalk components present.

Change in the range of induced chirp and mean centre frequency with increasing crosstalk is shown in Figure 4.8. Also shown is the induced chirp on zeros, which represents the maximum frequency excursion driven by the optical feedback. Chirp is shown relative to operation at a reference frequency of 1550nm at which the laser operates under bias. Crosstalk isolation is shown as a proportion of transmitted signal power. The frequency offset associated with a HDWDM channel spacing of 0.1nm is plotted to give an indication of potential system impact. Plots for 1, 4 and 8 reflectors are defined by total crosstalk isolation.

Differences in the range of chirp and maximum frequency excursion can be seen between a single reflector and multiple reflectors, while mean frequency excursion
Figure 4.7: Laser behaviour with a Fresnel back-reflection
remains almost independent of the number of reflectors. The difference in range is due to the distribution of the reflected optical power reaching the laser cavity. For the case of a single reflector, only 4 combinations of transmitted and reflected symbols are possible, giving the largest range of induced chirp for a given crosstalk isolation for half the time (assuming equal numbers of ones and zeros).

### 4.5 Comparison of Backscatter Generation Methods

As described in Section 4.2.1, two methods were developed to calculate Rayleigh backscattered light from a fibre: a computationally intensive method which modelled the dynamic distribution of power propagating in the fibre and an average power technique, which assumes a uniform power distribution. A comparison of these two generation methods was performed to determine the importance of the
simulated average power in the fibre on the calculation.

Figure 4.9 shows total backscatter generated by the Rayleigh.dist CCB for various bit rate signals. The simulations modelled a normalised power signal at 1.55\mu m propagating through a 5km length of silica fibre with 3dB/km attenuation using a spatial resolution of 2.0m. Modulation rates of 400kb/s, 2.85Mb/s and 20Mb/s (all with full extinction) are shown together with a plot assuming a uniform, average power distribution. From the results it is clear that high modulation rates, with pulse lengths much smaller than the fibre length, are most validly approximated to a uniform power distribution, with only small variation from the uniformly distributed approximation. Low modulation rates however, with pulse lengths comparable to the fibre length, produce large power fluctuations within the fibre and a high variance around the mean scattering value. This is as expected: if modulated pulse lengths are of the order of the scattering fibre length, we would expect the variance of the total back-scattered power to approach 0.25.
From these results we can see that the applicability of each method depends on the modulation employed and scattering fibre characteristics. Clearly for a low modulation rate or short length of scattering fibre (which might be defined by the position of an optical isolator) the dynamic module will produce the most useful results. The method incurs much longer simulation times and is limited in its accuracy by the spatial resolution or sampling frequency of the system, while. The averaging module is more computationally efficient but provides a poor approximation in the presence of large dynamic power variations.

4.6 EDFA Model

The EDFA model was modified from a CCB created by S. Zhang [24, 25, 69]. The model treats the doped fibre as a group of concatenated sections through which the optical signal is modelled sequentially with respect to photon and excited ion population. This allows a dynamic gain distribution along the length of the fibre to be captured by the model.

Output from the EDFA for a single sample is achieved by iteratively solving rate equations, initialised by calculating the conditions at the extreme ends of the doped fibre. These are then used to achieve solutions for the two end sections which are in turn used as the initial conditions for the two adjacent sections. This sequential process continues from one end of the fibre to the other, through each section, in forward and backward directions. Initial conditions are derived by modelling the pump light through the fibre and calculating the resulting gain for each section. The signal, forward and backward ASE are then calculated, modelling propagation end-to-end and starting from each fibre end simultaneously.

An important aspect of EDFA operation is behaviour under conditions of gain saturation. Saturation conditions occur if the pump light intensity is insufficient to produce enough excited Erbium ions to amplify all the propagating signal photons. This is caused by a combination of reduced pump intensity or increased signal intensity. Signal photons that exceed the number of excited Erbium ions
available to produce stimulated emission will experience absorption in the fibre, leading to sharply reduced gain and high signal to noise ratios. These conditions first occur in the ends of the doped fibre, since this is where signal intensities are highest and in co-pumped systems, where pump levels are lowest. Such dynamic population modelling along the fibre length means that saturation conditions can only properly be modelled using a concatenated sections approach.

Modification to the EDFA was in the form of two additional vector inputs for optical feedback. Since the EDFA is a bi-directional device, inputs were required for feedback propagating forwards and backwards through the doped fibre. These were labelled pre- and post-device feedback inputs respectively, relating to the position of feedback generation relative to the amplifier. The inputs are of vector format to allow broadband spectra to be modelled [25]. Other inputs to the EDFA are also of vector format, including the absorption and emission spectra of the fibre. The feedback inputs use the same frequency reference comb as these other vector inputs. The CCB symbol representation for the EDFA showing system-level inputs, including the pre- and post-feedback ports, is shown in Figure 4.6. The feedback ports are used as initial conditions for the optical power at each end facet of the doped fibre: Figure 4.11 shows the configuration of EDFA sections and injection of optical feedback.

Figure 4.10: EDFA CCB symbol
4.6. EDFA MODEL

optical power is modelled through the fibre in opposing directions, each section initialising its neighbour.

- doped fibre
- direction of signal propagation
- end facet of section N initialised with backwards propagating feedback
- end facet of section 1 initialised with forward propagating feedback

Figure 4.11: EDFA model comprising $N$ concatenated sections

4.6.1 EDFA Behaviour with Feedback

An experiment with a non-isolated EDFA was conducted for qualitative comparison with simulated results. Some initial simulations were carried out to aid in the construction of an EDFA at GEC-Marconi Research Centre, using data sheet information. The model predicted an optimised fibre length of 17.0m to produce maximum gain at 1.55 $\mu$m, and this was confirmed during construction of the amplifier by measurements made as the doped fibre length was cut back.

Figure 4.12: EDFA experimental ASE measurement setup

The effect of a Fresnel reflection on the ASE output of a second non-isolated amplifier was measured experimentally using the set-up shown in Figure 4.12. Angled splices at the doped fibre ends minimised connector reflections. A Fresnel back-reflection was introduced to the amplifier by connecting a length of SMF to each end of the doped fibre in turn. The fibre had an angled splice for connection
to the EDFA and was cleaved normally to the fibre axis at the other end. ASE output was measured from each end of the EDFA using an optical spectrum analyser. Results were obtained in co- and counter-pumped configurations for forward and reverse propagating ASE [70].

Figure 4.13: Experimentally measured ASE from an EDFA with a Fresnel back-reflection

The measured ASE spectra can be seen in Figure 4.13. The figure shows the ASE produced by the EDFA at the opposite end to the Fresnel splice. Results are shown for a co-pumped configuration, though no appreciable differences were observed for counter-pumping.

The simulated ASE output from the EDFA model is shown in Figure 4.14. This also shows the ASE output from the fibre end opposite to the reflector. Both figures show the reflected feedback increasing ASE across the spectrum and by \( \approx 2\text{dBm} \) in the region of highest gain. ASE in the reverse direction was fractionally reduced \((< 0.1\text{dB})\) in the reverse direction for both experimental and
simulated cases. This was attributed to gain re-distribution in the amplifier. Direct simulation of the experimental case was not possible due to a lack of data on the EDFA, which was an old design. Differences in the shape of the spectra might be attributable to the length of the experimental EDFA being longer than expected, causing greater pump absorption at the gain peak between 1.53\textmu m and 1.54\textmu m. This is typical for older designs of EDFA which employed longer doped fibres and lower gains compared with contemporary equivalents. Qualitative comparison of the two figures does illustrate a general agreement on the effect of a back-reflection on the ASE produced, and shows the model producing gain in the forward and backward directions which is consistent with experimental observation.
4.7 Summary

The development of ideas and software to capture effects of back-reflections in a fibre network have been described in this chapter.

Following the introduction, the rationale for considering only linear crosstalk power was explained and the main software modules written for the purpose of crosstalk generation were described. Two device models were then presented: first, in Section 4.3, the DFB laser CCB was described, together with a method for phase noise characterisation in SPW using library blocks. Some results showing behaviour of the model were presented in Section 4.4, particularly with respect to operation under the influence of a Fresnel back-reflection from the fibre pigtail end. Eye-closure was found to be highly dependent on the bias current applied to the laser with worst case closure occurring at threshold levels. Frequency chirp showed dependence on the number of reflecting terms present and crosstalk isolation levels required to limit chirp for WDM applications were commented upon. Section 4.5 compared the methods of Rayleigh backscatter calculation, with regard to determining the the applicability of each. The dynamic calculation was shown to be more accurate, especially for low modulation rates or where the length of scattering fibre was short. The averaging method yielded good results for high modulation rates and was potentially orders of magnitude more computationally efficient. Finally, in Section 4.6, the modified EDFA model was presented, with inputs added to the CCB for the application of bi-directional feedback. Comparison of results from the model was made with experimental data, which provided good agreement in the design of an amplifier and general agreement on the effect of a Fresnel reflection at one end of the doped fibre.
Chapter 5

Interferometric Noise in HFR Networks

5.1 Introduction

This chapter is involved with analysis and simulation of interferometric noise in fibre networks carrying optically generated millimetre-wave signals. The study is concerned with systems which use heterodyne mixing of two optical carriers at the receiver to produce a millimetre-wave carrier signal. Analysis of IN impact on OOK modulation schemes is presented in Section 5.2. Network simulations to replicate and verify the analytical predictions are described in Section 5.3 and results compared. Simulations to determine the impact of IN on phase-related modulation schemes are presented in Section 5.3.2. The chapter concludes with a summary.

5.2 Analysis of IN with OOK Modulation

Interferometric noise is produced in an optical network when the wanted signal is received together with delayed and attenuated replicas of itself. IN is com-
prised of combinations of beat signals which occur between the signal and the spurious replica components, as described for single wavelength channel systems in Section 2.3.4. This chapter includes an analysis of IN generation in a system employing coherent mixing of two optical carriers to produce an intermediate carrier frequency, presented for the first time [71]. This coherent mixing method has been presented in Section 2.4.1 together with associated methods of modulation imposition. Figure 5.1 is a schematic diagram of the FRANS system labelling the optical frequency components at intermediate stages in the system together with spurious signal replicas from the network.

As we have seen, the millimetre-wave subcarrier is produced by mixing two optical signals displaced in frequency by $\omega_m$ (the millimetre-wave frequency), at the receiver. The information modulation $m(t)$ is imposed onto one of the optical carrier signals. We assume the two optical components arrive at the detector together with one or more attenuated, delayed and consequently phase-shifted replicas, caused by propagation along alternate, spurious light-paths in the network. The polarisation states of all interfering component are assumed to be aligned with those of the carrier signals, which is a worst case assumption shown to be the most likely to occur in practice [72].
5.2. ANALYSIS OF IN WITH OOK MODULATION

From these assumptions, the component optical electric fields impinging on the photo-detector in a system employing coherent mixing at the receiver, are given by

\[ E_{\text{total}}(t) = E_1 e^{j(\Omega - \frac{\pi}{2})t + j\phi(t)} + m(t)E_2 e^{j(\Omega + \frac{\pi}{2})t + j\phi(t)} \]

\[ + E_1 \sum_{k=1}^{N} \sqrt{\varepsilon_k} e^{j(\Omega - \frac{\pi}{2})(t - \tau_k) + j\phi(t - \tau_k)} \]

\[ + E_2 \sum_{k=1}^{N} \sqrt{\varepsilon_k} m(t - \tau_k) e^{j(\Omega + \frac{\pi}{2})(t - \tau_k) + j\phi(t - \tau_k)} \]  

(5.1)

where \( m(t) \) is the modulating signal, which for OOK is given by

\[ m(t) = \sum_{l=-\infty}^{\infty} a_l p(t - lT) \]  

(5.2)

where \( E \) is the electrical optical signal amplitude, \( \omega \) is the millimetre wave angular frequency, \( a_l \) is the random binary data; \( a_l \in \{a, 0\} \) with both events having equal probability of occurrence. \( p(t) \) is the digital data pulse shape, \( \Omega \) is the angular frequency of the laser diode, \( \phi(t) \) is the laser phase noise, \( \tau_k \) represent the time delays of the interferers, \( N \) is the number of interferers, \( \varepsilon_k \) is the crosstalk isolation as a ratio between the power of the \( k \)th interferer and the power of the wanted signal.

Referring to Equation 5.1: the first term is the first, unmodulated optical carrier; the second term is the second, modulated optical carrier; the third term comprises the sum of crosstalk components arising from the first optical carrier; the fourth term is the crosstalk arising from the second, modulated carrier. Assuming all components impinge on a receiver of normalised responsivity, the resulting photocurrent will be the square of the sum of these signals \( i(t) = |E_{\text{total}}(t)|^2 \). Expanding this expression gives

\[ i(t) = |E|^2 (1 + m(t)) \]

\[ + [1 + m(t)] |E|^2 \sum_{i=1}^{N} \sqrt{\varepsilon_k} \]

\[ + E^2 m(t) \cos(\omega_m t) \]

\[ + 2|E|^2 \sum_{k=1}^{N} \sqrt{\varepsilon_k} \text{RE}[m(t - \tau_k)] \cos \left[ \omega_m t + (\omega_o - \frac{\omega_m}{2}) \tau_k + \phi(t) - \phi(t - \tau_k) \right] \]
\[ + 2|E|^2 \mathbb{E}[m(t)] \sum_{k=1}^{N} \sqrt{\varepsilon_k} \cos \left[ \omega_m t - \left( \omega_0 - \frac{\omega_m}{2} \right) \tau_k - \phi(t) + \phi(t - \tau_k) \right] \]
\[ + 2|E|^2 \sum_{k=1}^{N} \sqrt{\varepsilon_k} m(t - \tau_k) \cos \left[ \omega_m (t - \tau_k) \right] \]
\[ + 2|E|^2 \sum_{k=1}^{N} \sum_{j=k+1}^{N} \sqrt{\varepsilon_k \varepsilon_j} m(t - \tau_j) \cos \left[ \omega_m t + \omega_0 (\tau_j - \tau_k) - \frac{\omega_m}{2} + \phi(t - \tau_k) - \phi(t - \tau_j) \right] \] (5.3)

This represents the total photocurrent before bandpass filtering: The first term is the received signal power from the wanted signal components; the second term is the sum of the linear crosstalk power from all interferers; the third term is the primary beating between the signal components, which constitutes the millimetre-wave carrier signal; the fourth and fifth terms comprise secondary beating between the signal and interfering components, which forms the main salient contribution to IN on the received signal; the fifth and sixth terms are tertiary beating between the crosstalk components themselves, which also contribute to IN. The contribution of tertiary beating is small compared to the secondary beating terms however, since \( \sqrt{\varepsilon_k \varepsilon_j} \ll \sqrt{\varepsilon_k} \) and is thus neglected in this analysis. Also since the passband of the receiver accepts only signals centred on \( \omega_m \), the millimetre-wave frequency, we can disregard terms not at this frequency. This gives

\[
i(t) = E^2 m(t) \left\{ \cos [\omega t] + \sum_{k=1}^{N} \sqrt{\varepsilon_k} \cos [\omega t + \Phi_k] \right\}
\[ + E^2 \sum_{m=1}^{N} \sqrt{\varepsilon_m} m(t - \tau_m) \cos [\omega t + \Phi_m] \] (5.4)

If we assume incoherence between the signal and interfering components then \( \Phi_k = \left( \frac{\omega}{2} + \Omega \right) \tau_k + \phi(t) - \phi(t - \tau_k) \) and \( \Phi_m = - \left( \frac{\omega}{2} + \Omega \right) \tau_m - \phi(t) + \phi(t - \tau_m) \) are independent identically distributed random variables described by a uniform distribution. Since direct modulation is not employed, this limits the delay time \( \tau_k \) associated with each replica signal to greater than the coherence time of the laser. This is a worst case assumption, ensuring maximum IN beating.

Referring to Equation 5.4: the first term is the desired, primary beat signal produced by coherent mixing; the second is an IN term resulting from beating between the modulated carrier and the replicas of the unmodulated carrier; the third term is an IN term resulting from beating between the unmodulated carrier
and replicas of the modulated carrier. Equation 5.4 illustrates the interferometric phase to amplitude conversion which takes place, so that IN is observed on the received signal.

Thus the signal input to the decision circuit is given by

\[
i_0(t_s) = E^2a_0 + IN + n(t)
\]

\[
IN = E^2a_0 \sum_{k=1}^{N} \sqrt{\varepsilon_k} \cos(\Phi_k)
\]

\[
+ E^2 \sum_{k=1}^{N} \sqrt{\varepsilon_k}a_k \cos(\Phi_k)
\]

(5.5)

where \(t_s\) is the sampling time, \(a_0\) is the binary random variable of the wanted signal and \(a_k, \ k \in \{1, 2, ..., N\}\) are the binary random variables of the noise term arising from the beating between the reflections of the modulated carrier with the non-modulated carrier. \(n(t)\) is the additive Gaussian receiver noise.

In order to assess the impact of such noise on system performance, we need to capture the statistics of the random variable which describes the IN. Here we adopt the moment generating function, MGF, approach. It can be shown that the MGFs for the signal plus IN, for a transmitted ‘1’ and a transmitted ‘0’, can be expressed as follows:

\[
M_0(s) = \prod_{k=1}^{N} \left[ \frac{1}{2} I_0 \left( E^2 \sqrt{\varepsilon_k} a_s \right) + \frac{1}{2} \right]
\]

(5.6)

\[
M_1(s) = M_0(s) \prod_{k=1}^{N} I_0 \left( E^2 \sqrt{\varepsilon_k} a_s \right) e^{E^2 a_s}
\]

(5.7)

\(I_0(.)\) is the modified Bessel function of the first kind of order zero. The modified Chernoff bound, MCB, can then be used to provide a tight upper bound on the BER [73]

\[
\text{BER} \leq \text{MCB}(s) = \frac{M_n(s)}{2\sqrt{2\pi}\sigma_n} \left[ M_0(s) e^{-sD} + M_1(-s) e^{sD} \right] \quad ; \quad s > 0
\]

(5.8)

where \(D\) is the decision threshold and \(M_n(s)\) is the MGF of the additive Gaussian receiver noise [73]. \(\sigma_n^2\) represents the power of the Gaussian receiver noise.
An alternative, simplified representation treats each IN r.v. as Gaussian, adding the IN variance to that of the receiver noise to provide a Gaussian Approximation (GA) to the BER:

\[
\text{BER} = \frac{1}{2} Q \left( \frac{E^2 a - D}{\sqrt{\sigma_n^2 + \sigma_1^2}} \right) + \frac{1}{2} Q \left( \frac{D}{\sqrt{\sigma_n^2 + \sigma_0^2}} \right)
\]

(5.9)

where

\[
\sigma_1^2 = \frac{3 a^2 E^4}{4} \sum_{k=1}^{N} \xi_k \quad \text{and} \quad \sigma_0^2 = \frac{a^2 E^4}{4} \sum_{k=1}^{N} \xi_k
\]

(5.10)

and \( \sigma_1^2 \) and \( \sigma_0^2 \) are the power of the IN on ‘1’s and ‘0’s, respectively. \( Q(.) \) represents the error function. Note that the GA uses only minimal information concerning IN statistics (just the variance): we can, by considering the central limit theorem, reasonably expect this to provide a good approximation for \( N \) sufficiently large, but for a small number of interfering terms this model is obviously inadequate since the sinusoidal PDF is bounded whereas the Gaussian is not.

### 5.2.1 Analytical Results

Gaussian approximation of the noise and the modified Chernoff bound were used to estimate the system performance implications of IN on systems employing coherent mixing: as described in the previous section. The BER was determined, using both techniques, as a function of the network crosstalk isolation, when distributed equally between varying numbers of interferers. The analysis assumes a receiver which would yield a BER of \( 10^{-9} \) in the absence of IN. Results, shown in Figure 5.2, demonstrate that for small \( N \) the GA overestimates the BER and the crosstalk isolation tolerance. This is as expected, since the IN distribution is tightly bounded, as demonstrated in the previous section and schematically shown in Figure 5.9. For a single interferer \( (N = 1) \) we can see that Gaussian approximation of the IN leads to an overestimation of the crosstalk isolation tolerance of approximately 2dB. As the number of interferers increases the interference tends to a Gaussian process (by the central limit theorem) and the GA improves in accuracy, ultimately underestimating the MCB. In actual network deployment however, crosstalk at the receiver is usually dominated by a small number (most
5.2. ANALYSIS OF IN WITH OOK MODULATION

Figure 5.2: BER versus total crosstalk isolation for OOK with IN generated by a varying number of interferers.

commonly 1 or 2) of interfering terms, suggesting the GA as an inappropriate method for estimating the performance of real systems.

Figure 5.3 shows the power penalty imposed on a system as a function of crosstalk isolation, where we define power penalty as the additional signal power required to maintain a BER of $10^{-9}$ despite the presence of IN. Again both GA and MCB approaches were used for the calculation, which provides further illustration of the overestimation of IN impact when using the GA. For a power penalty of 2dB, the GA overestimates the crosstalk isolation tolerance by 4dB for a single interferer ($N = 1$) and 2dB for $N = 2$. The inaccuracy is again is due to the unbounded nature of the Gaussian distribution, which also leads to erroneous error floor predictions. Figure 5.4 shows the GA predicting an error floor for a single interferer, for which the MCB shows no evidence.
Figure 5.3: Imposed power penalty versus total crosstalk isolation for OOK with IN generated by a varying number of interferers

Figure 5.4: BER versus signal power for OOK with IN generated by a single interferer
5.3 OPTICAL FEEDBACK MODELLING CONSIDERATIONS

The exponential response of PP to varying crosstalk isolation shown is driven by the Gaussian nature of the noise distribution on the detected signal, and is clearly most pronounced on the GA and also for both analytical methods when considering many (>8) interferes. For the MCB, with a small number of interferes (<8), the bounded distribution produces a gentler degradation in performance, as measured in this case as PP. This is driven by the weight of the Gaussian distribution moving across the decision threshold. The error probability, and hence imposed power penalty rises exponentially as the variance of the noise distribution approaches the decision threshold.

5.3 Simulation of IN

Simulation of the coherent mixing process was designed to ensure that the system was as close as possible to the analytical assumptions [74]. Two separate complex electric fields were carriers were generated as the output from the Mach-Zehnder modulator. Phase noise was generated using the method described in Section 4.3.1.1 and was added to both signals. Modulation was applied to one of these carriers, which was initially OOK but was extended to include other modulation schemes using the same carrier and crosstalk generation methods. The detector comprised a square-law amplitude to intensity conversion and passband filter. Alternate paths were included by sending the combined carriers through buffer delay lines with attenuative scaling. Multiple reflectors were simulated using multiples of these delay lines. The spurious components were added to the combined signal fields prior to detection.

5.3.1 ASK Modulation

Initial results were concerned with replicating the values produced by the analysis of OOK. Figure 5.5 shows simulated and analytical values for standard deviation of the IN noise power as a function of number of interferers, each of crosstalk isolation of -26dB. The detector passband was set to include the entire IN bandwidth, though this also included tertiary beating contributions. A demodulated OOK bit sequence with IN from a single interferer of -20dB crosstalk power is shown in Figure 5.7. The simulated distribution of the IN was also analysed and compared with analytical expectation. Figure 5.7 shows the bounded and Gaussian distributions associated with 1 and 8 interferers respectively, for a total crosstalk power isolation of -14dB in both cases. The PDFs are normalised to the probability of a received zero without IN. The noise on each symbol is composed of two super-
Figure 5.5: Simulated and analytical results for IN standard deviation

Figure 5.6: Demodulated OOK bit sequence with IN from a single interferer
5.3. SIMULATION OF IN

Figure 5.7: PDFs for an OOK signal with IN from (a) 1 and (b) 8 interferers

Figure 5.8: PDFs for OOK with IN generated from an increasing number of interferers
imposed PDFs: this was predicted by the analysis, and is shown schematically in Figure 5.9. Figure 5.8 shows the change in shape of the IN PDF as the number of interferers is increased. The total crosstalk power isolation for all the simulations was equal at -20dB. This total crosstalk power was split equally between the interferers in each simulation. The log of the total number of events simulated was plotted against amplitude samples of a normalised signal. The events represent the entire eye to produce a large, consistent sample count. The results show the trend from a tightly bounded distribution for 1 interferer as the system tends to a Gaussian as the number of interferers increases.

For a single interferer, the distributions on both one and zero symbols each comprise an addition of two independent PDFs. All four PDFs have equal weighting assuming equal numbers of zero and unity bits are transmitted. Figure 5.9 is a schematic illustration of the IN PDFs and the signal-crosstalk beating combinations from which they originate. OOK modulation with full extinction is assumed.

A zero symbol signal distribution consists of a delta function A and arcsine distribution B. These are attributable to specific beating combinations. Distribution A occurs on a transmitted zero with a reflected zero present as the interfering crosstalk. Thus no interferometric beating occurs since there are no two frequency displaced components, leading to a perfectly defined signal level. Distribution B occurs when a zero symbol is received with crosstalk from a unity bit. Beating occurs between the modulated crosstalk component and the unmodulated signal component. In the same way, a transmitted unity bit produces an arcsine distribution C when received with crosstalk from a zero symbol due to beating between the modulated signal and unmodulated crosstalk components. Two beating combinations are produced if both signal and crosstalk are unity bits, between the unmodulated signal and modulated crosstalk and between the modulated signal and unmodulated crosstalk components. This produces a distribution which is convolved from two PDFs, shown as D. The simulations also include the effects of tertiary (crosstalk-crosstalk) beating, which is neglected in the analysis for simplicity as a negligible IN contribution.
5.3. SIMULATION OF IN

Bit error rate (BER) calculations were performed on the simulation data to determine the impact of changes in the IN distribution on system performance. As with the distribution analysis, the total crosstalk power isolation was constant across the simulations at $-14\,dB$. BER calculations were based on a statistical $Q$ function analysis of the mid-eye point, since it was not possible in terms of computation time to simulate individual error events [75]. The receiver was defined in terms of thermal noise power, and the $Q$ function analysis performed on the electrical power SNR. Results from the BER analysis are shown in Figure 5.10 as a function of number of interferers. The IN variance is also plotted and shows constant behaviour as the number of interferers increases. The results illustrate an increase in BER which is attributable to a change in the shape of the IN distribution and is independent of IN variance. Saturation of the BER occurs as the distribution tends to a Gaussian.

Figure 5.9: Schematic of OOK signal PDF with IN generated by a single interferer, relating distribution to signal-crosstalk beating
5.3.2 FSK Modulation

Since interferometric noise is a phase driven phenomenon, which is observed as an amplitude perturbation, the impact on angle modulation schemes might be expected to be less than that for ASK. To this end, simulations were conducted using FSK modulation to assess the impact of IN on the received signal.

The receiver comprised a square-law, bandpass detector with amplitude limiting applied to the signal following detection and optimised for each simulated crosstalk isolation. An FM discriminator with an ideal linear response over the modulation frequency excursion was used to recover the base-band signal.

In common with OOK, IN was found to produce a bounded distribution for a small number of interferers and tend to Gaussian as the interfering terms were increased. For a given signal power however, IN variance was much reduced compared with OOK if appropriate clipping was applied. A $Q$ value was calcu-
lated to give a measure of the relative IN degradation, where \( Q = \sigma_{IN_i}/P_{sig} \) and where \( \sigma_{IN_i}^2 \) is the variance of the IN on transmitted unity symbols and \( P_{sig} \) is the received mean signal power. Figure 5.11 shows the optimisation of \( Q \) through clipping of a received signal with IN generated by a single interferer at -13dB. The optimised value of \( Q = 56.5 \) compares with \( Q = 23.1 \) for OOK under the same conditions of IN generation.

5.4 Summary

This chapter has presented an analysis of interferometric noise in optical systems which employ coherent mixing of two carriers for generation of an RF carrier frequency. Simulation results were also presented in support of the analytical predictions.

Following the introduction, an analysis of IN for OOK modulation was presented and contributions from signal-signal, signal-crosstalk and crosstalk-crosstalk beat-
ing were identified and termed primary, secondary and tertiary beating respectively. Secondary beating was determined the salient IN contribution and the MGF of the beat terms was derived by characterisation of the phases of the interfering terms. Two methods were employed for estimation of the system impact of IN using the derived MGFs: a Gaussian approximation (GA) of the IN noise distribution and the modified Chernoff bound. These methods were used to estimate the BER and power penalty resulting from IN as a function of network crosstalk isolation and number of interferers. The GA was found to overestimate the impact of IN for small numbers of interferers, due to the unbounded nature of the distribution. The MCB approach gave a very tight upper bound on the effects of IN and agreed well with the GA for large numbers of interferers. Simulation results for the IN variance for OOK agreed well with analytical predictions. The distribution of IN was confirmed by simulation which also revealed the predicted dependence of BER on the number of interferers. Further simulation of FSK modulation showed a greater resilience to IN than ASK.
Chapter 6

Illustrative HFR System Studies

6.1 Introduction

The purpose of this chapter is to illustrate the capability of the developed software and techniques in representing complete fibre systems, with HFR networks used as examples. The chapter seeks to show the flexibility and ease of system simulation available using SPW and the developed blocks while presenting some illustrative results which comment on the relative merits of each network design. Systems are assessed in terms of received signal waveforms, eye-diagrams, signal distributions and bit error rates.

A system employing direct modulation of a laser diode is the first study, illustrating the implementation of physical component models as a system and showing the type of immediate signal analysis that is available in SPW.

Sections 6.3 and 6.4 look at the MODAL and FRANS systems respectively. These are simulated to determine the effects of IN generation on the BER of both systems. MODAL is simulated for both ASK and FSK modulation, and FSK is shown to be much more resilient to IN. Simulation of the FRANS system shows the models capability to handle multiple frequency channels simultaneously and results showing the general impact of IN, inter-channel crosstalk and effect
of increasing the power budget are shown. The Chapter is concluded with a summary.

6.2 Direct Modulation HFR System Simulation

In order to illustrate the implementation of modules combined as a system study, a simple HFR network was simulated. Physical models of components were included to show the impact of various device parameters on the system performance.

The developed laser model was used in the transmitter, with phase noise added using the method described in Section 4.3.1.1. This allowed variation of parameters such as laser linewidth, drive current bias, modulation depth and optical feedback isolation. The laser was modulated directly with an RF subcarrier at 10GHz which carried an OOK signal at 1Gb/s. A set of alternate propagation paths was also included to capture forward propagating crosstalk and the impact of interferometric noise generation on the system. Direct detection was employed using a photodiode and coherent demodulation carried out prior to analysis, which was handled using hierarchical blocks for mid-eye sampling and BER estimation.

Error rates were estimated by Q function analysis, as described in Section 5.3.1. The calculation is performed on a single mid-eye sample point from each received symbol, ensuring optimal error rates by alleviating symbol transition anomalies and filtering effects. This method was also used in the other systems described later in this chapter. Hierarchical blocks developed to perform mid-eye sampling and BER estimation can be seen Figure 6.1 which shows a screen-shot of the BDE top-level system design, and an hierarchical window, displaying lower-level detail of the laser, phase noise generation and associated parameters. Salient parameters in each block structure are elevated through the hierarchy to assist variation and optimisation in each simulation.

Results illustrating the effect of multiple propagation paths and optical feedback on the system can be see in Figure 6.2 which shows eye-diagrams of the de-
Figure 6.1: BDE windows showing direct modulation HFR system and laser block detail
Figure 6.2: Received eye-diagrams and base-band signal for direct modulation HFR system, showing IN and laser feedback effects.
modulated base-band signal. These show eye-closure induced by the presence of: (a) no interference; (b) a single -15dB forward-propagating interferer; and (c) 4 x -15dB forward-propagating interferers. Plot (d) is a section of the demodulated bit sequence used to produce (b), clearly showing the beat signal as a perturbation on half the received unity bits, with linear crosstalk also visible. The IN is characterised by the derivation in Section 2.3.4 since only a single optical carrier is involved, though its impact on the system is reduced through filtering. Beating is observed on both ones and zeros as a result of non-zero extinction, since the laser is biased above threshold to avoid signal overshoot and to maximise eye-opening.

The impact of feedback from the pigtail splice into the laser on the system can be seen in (e), which shows the effect of a -15dB reflector situated 0.5m from the laser diode; and (f), which shows the same pigtail reflection with additional crosstalk from a -15dB interfering replica arriving at the detector, with IN centred on the linear crosstalk levels produced by the laser feedback.

Figure 6.2 illustrates the capability of SPW with the addition of the developed software and techniques to simulate a complete optical system and provide immediate analysis within the environment. This analysis can extend beyond the signal waveforms and eye-diagrams shown to parameters which confer impact on the system as a whole; such as BERs, error floors and power penalties. These are used to illustrate system design in the sections following.

6.3 MODAL Simulation

The MODAL system was described in Section 2.4.3, which detailed the method of mmw generation and modulation imposition. Several alternate modulation arrangements might be employed in MODAL, including ASK, FSK or QAM, depending on system requirements. A proposed development of the MODAL system was the use a frequency division multiplex (FDM) to provide additional bandwidth from a remote antenna unit. Generation of multiple frequency signals
was to be achieved by transmission of multiple modulated carrier signals, each
displaced in frequency from a single un-modulated carrier by increasing amounts.
Figure 6.3 schematically shows the optical signal components used to generate the
proposed MODAL multiplex. The systems developed under RACE programme
however, used only a single modulated carrier to achieve single frequency trans-
mision, as do the systems simulated in this section.

The simulated MODAL system generates a mmw carrier frequency of 31.5GHz
by optical heterodyning and delivers data at 1.55Mb/s using either ASK at FSK
modulation. The signal is distributed to 8 antenna sites over 12km of fibre from
the central node. Both ASK and FSK schemes were simulated: the BDE top-level
view of the MODAL system employing ASK is shown in Figure 6.4.

### 6.3.1 ASK Modulation

Simulation of the splitting ratio of signal amplitudes between the two optical car-
rriers had no effect on the IN noise contribution in terms of percentage eye-closure
or distribution, as expected. In terms of IN generation and the resulting BER,
the parameters with most impact were found to be the distribution of crosstalk
power between interferers and the extinction ratio of the MODAL source. Fig-
Figure 6.4: BDE diagram of ASK modulated MODAL

Figure 6.5 shows the effect of these two parameters in terms of the system BER. Four interferers were present in the simulation, with one dominating to a varying degree over the other three equal terms: the total crosstalk isolation was constant at -20 dB over all simulations. Signal bias level is measured relative to eye opening at the receiver. The detector sensitivity was optimised to yield a BER of 10⁻⁹ in the absence of interference, and the decision threshold adjusted for each bias level.

The plot shows increasing BER as the ratio of interferer dominance decreases. This is expected, driven by the change in shape of the IN distribution, and shown previously in Figure 5.10. The plot was produced in MATLAB from results produced by the BER test set hierarchical block. Dynamic linking between MATLAB and SPW is available through library blocks, as described in Chapter 3. The effect of non-zero extinction, in this case in the form of a signal bias, is also shown to be detrimental to the BER. This occurs as the variance and shape of IN spreads due to beating on all combinations of received symbols and replicas, including zeros.
CHAPTER 6. ILLUSTRATIVE HFR SYSTEM STUDIES

The change in signal distribution driving the error rate degradation can be seen in Figure 6.6 which shows: (a) full extinction and one interferer dominating by 17dB; (b) bias of 25% and one interferer dominating by 17dB; (c) full extinction with all 4 interferers equal; (d) bias of 25% and all interferers equal. These distributions represent the four corner points of the surface shown in Figure 5.10. All distributions were plotted by extracting a single mid-eye sample from each symbol. Histogram analysis of signal distributions is immediately available in the SPW signal calculator environment, requiring a minimum of effort in the windowed environment.

6.3.2 FSK Modulation

FSK subcarrier modulation was also simulated for the MODAL system, since this was found to be more resilient to IN in Section 5.3.2. The FSK MODAL system used a frequency excursion of 50GHz between symbol levels and employed a linear FM discriminator at the receiver. The BDE system representation is shown in Figure 6.7: comparison with Figure 6.4 illustrates the simple re-configuration of systems using SPW, with only the modulation source and detector block differing between the designs. The FSK receiver consists of further hierarchical blocks performing coherent demodulation blocks, clipping, linear FM discrimination and lowpass filtering.

The effect of IN on the BER of the system was simulated as a function of interferer dominance and total crosstalk isolation. The system was simulated with a receiver sensitivity of -30dBm yielding a BER of $10^{-9}$. As in the ASK specification, the PON uses an 8-way splitter to distribute the signal to 8 base stations over 12km runs of fibre from the central node, requiring the transmitter to launch $\approx 0.4$ mW of power into the network to achieve a BER of $10^{-9}$ in the absence of crosstalk or clipping. Crosstalk was generated from 4 interferers, with one dominant term, in the same way as the ASK system described previously.

The effect of IN on the system BER can be seen in Figure 6.8 which also gives a comparison of resilience to IN with ASK modulation. The plots shows the FSK
6.3. MODAL SIMULATION

Figure 6.5: BER versus interferer dominance and laser bias for ASK MODAL

Figure 6.6: Received signal distributions with IN for ASK MODAL
system largely unaffected by the distribution of interfering power between replica terms, with the equally distributed case (0dB dominance) producing the lowest error rates. The ASK system used for comparison is identical to that used for the FSK results, with the input power increased to 3.3mW to give the same nominal BER in the absence of interference using the same detector specification. All interfering terms were equal for the ASK simulations and full extinction modelled to produce OOK, for which analytical results were shown in Figure 5.2 on page 103. The FSK system is clearly much more resilient to IN with $10^3$ difference in the error rate at -30dB crosstalk isolation. FSK remains almost unaffected by IN until crosstalk levels approach -25dB, whereas in contrast ASK is affected by even the lowest levels of crosstalk, showing degraded error rates at -45dB. These BERs are likely to degrade further in the absence of full extinction, as was shown in the previous section. In general, Figure 6.8 shows that for a given BER, FSK shows $>15$dB tolerance to forward propagating crosstalk compared with ASK.

The effect of increasing the clipping ratio is shown in Figure 6.9. The ratio was varied by maintaining the clipping parameters and increasing the power launched into the PON. Ratios are shown on the graph for each input power and were
calculated as unclipped signal amplitude to clipping cutoff amplitude. Crosstalk was generated from 4 interfering terms of equal power. It is clear from the 0.4mW trace that clipping provides resilience to IN with the un-clipped system underperforming the others by several orders of magnitude. The Figure also shows that while error rates improve with increasing signal power at low crosstalk levels, high levels of crosstalk produce BERs that show little improvement with increasing optical power, suggesting the existence of error floor at high crosstalk levels.

The results suggest that in designing resilience to IN into MODAL or similar heterodyning system using FSK, we can power budget for a clipping ratio of between 2 and 4 (in this simulated case achieved using 0.9 to 1.6mW transmitter power) and have confidence that the system will perform optimally without requiring an additional margin for possible IN generation in the system. In contrast, a system employing ASK should focus system design on reducing crosstalk generation in the network since such a system would be impacted by even low crosstalk lev-
6.4 FRANS Simulation

The FRANS system was introduced in Section 2.4.4, which described its support for ATM, allowing multiple users and data types to be multiplexed onto a single data channel. FRANS transmits the ATM multiplex using one of several possible modulation schemes on a single mmw frequency channel from each base station, requiring a single optical subcarrier pair for generation. Figure 6.10 schematically shows the optical signal components used to generate the FRANS multiplex. Results from such a system using either ASK or FSK modulation would be similar to those obtained for the MODAL system in the previous section, depending on
transmission and detection parameters and the distribution network.

A proposal for future development of FRANS uses a hybrid multiplexing system which combines elements from MODAL and the original specification for FRANS. The proposed format involves transmitting a frequency division multiplex, as proposed for the MODAL system, with each frequency channel carrying an ATM multiplex. Figure 6.11 schematically shows the optical signal components used to generate the hybrid multiplex. The system would provide access to multiple users on each of the multiple frequency channels transmitted from the base station.

Figure 6.10: Optical signal components for FRANS

Figure 6.11: Optical signal components for multi-frequency FRANS
Figure 6.12 shows the BDE diagram of the hybrid system, using 5 individual frequency channels. Each channel is modulated with a filtered ASK time-division multiplex at 100Mb/s comprising a total of 5 separate random bit sequences of on-off keying. The channels are spaced in frequency by 500MHz, which includes a buffer to avoid inter-channel crosstalk interference, with the centre channel at 31.5GHz. Each channel is filtered out following detection and Q analysis performed to obtain the BER. The distribution network is again configured as a PON with an 8-way splitter feeding 12km lengths of fibre connecting each remote antenna unit. The deployed version of FRANS is likely to use more and closer
spaced channels than the simulated system, with each carrying a lower bit rate signal. The system specification was necessarily modified for the purposes of simulating error rates while maintaining the correct carrier frequencies.

Interferometric noise was introduced to the system, generated by a single interferer (which might also represent a saliently dominant interferer). The BER response to the IN was calculated for channel 3, since this should be most affected by interference from the other channels by virtue of its central frequency location at 31.5GHz. Simulations were carried out to assess the impact of IN and to determine the effects of beating and crosstalk generated by replica signals and imperfect filtering from other channels.

![Figure 6.13: BER versus crosstalk isolation for multi-frequency FRANS](image)

IN generated by signal combinations from other channel carriers should fall outside the bandpass of the filters used to extract the signal for each channel. This was confirmed by the simulation results in Figure 6.13, which show only a small
degradation in the BER. The error rate degradation is constant with respect to crosstalk isolation, suggesting the presence of inter-channel interference rather than IN beating from other channels. In this way the simulations demonstrate the facility to isolate parameters and effects in analysis of system performance that would be difficult in other forms of development and testing.

![Graph showing BER versus signal power for FRANS with IN generated by a single interferer.](image)

**Figure 6.14:** BER versus signal power for FRANS with IN generated by a single interferer

The effect of increasing the system power budget is shown in Figure 6.14 which shows the BER response to an increase in transmitter power. A single -20dB interferer was used to generate IN in the system as the input power was increased and threshold optimised accordingly. The plot reveals the absence of an error floor, as predicted by the analysis in Figure 5.4 on page 104. The result suggests, in contrast to that found for the MODAL FSK system, that in amplitude modulated systems an increase in the system power budget will lead to consistently lower error rates and resilience to the effects of IN generation.
6.5 Summary

The purpose of this Chapter has been to demonstrate the capability of the developed tools and techniques in representing complete telecommunication networks. This was done by building a number of HFR systems and using simulation to determine the impact of certain device and network parameters on the system as a whole.

The first section illustrated the implementation of physical component models as a basic HFR system using direct modulation of the laser diode to produce the mmw carrier. The impact of reflections in the form of feedback into the laser and as IN generating replicas at the receiver was determined. Received signal waveforms and eye patterns were presented for combinations of IN generating terms and laser feedback to show the types of results and analysis immediately available using the SPW sub-environments such as Signal Calculator and Signal Analyser.

In sections 6.3 and 6.4 two specific HFR systems were modelled, namely the MODAL and FRANS projects which were described in Chapter 2. MODAL was modelled using ASK and FSK modulation schemes and the impact of IN on the system error rate determined for both. For ASK, laser extinction ratio and interferer dominance were determined in terms of BER, while for FSK the general resilience of the modulation scheme to IN degradation was demonstrated, with > 15dB tolerance to IN when compared with ASK. The effect of increasing the power budget was also examined and optimisation of the system budget with respect to IN was commented upon. The FRANS system was shown to have good inter-channel isolation and the absence of an error floor in the presence of IN was shown, as predicted by analysis. The results from the MODAL and FRANS simulations illustrated the differences both in terms of resilience and in system design between ASK and FSK modulation when a system is subject to interferometric noise. In addition, the power of the modelling platform was demonstrated and range and immediacy of tools available for system analysis within the SPW environment were shown.
Chapter 7

Conclusions

7.1 Research Results

The aim of the project was to investigate the effect of spurious signals which may occur in optical network and the resulting research may be categorised into two areas:

- The development of software modules and techniques to model the effect of linear optical feedback in an existing fibre network simulation environment.

- The analysis and simulation of interferometric noise impact on optical heterodyning hybrid fibre/radio systems, presented for the first time.

7.1.1 Optical Feedback SPW Modules

As has been shown in this Thesis, spurious signals in fibre communication networks can pose a limiting problem in the design and maintenance of systems. While modelling of optical networks has become established as an essential part of the design and development process, the capturing of spurious signal effects is often not addressed by platform developers.
A capability for modelling optical feedback propagating backwards through a fibre network has been added to the SPW optical network simulator. The work draws on established ideas and embeds them into an established network simulation environment, providing a graphical interface and pre-developed libraries for simple system design and optimisation. All network elements are represented by icons and systems are formed by block-diagram constructions using these symbols. Modules were written for feedback generation, including blocks representing reflective elements and Rayleigh backscatter from runs of fibre. Models for a semiconductor laser and EDFA were also created, with input ports for simple connection to the feedback generating blocks. Full documentation for the modules was written and is provided in Appendix B.

7.1.2 Interferometric Noise in Heterodyning HFR Systems

Interferometric noise has gained increasing attention in the study of optical networks, in particular its effect on WDM networks and digital systems. IN occurs in a system due to reflections and multiple paths occurring in the network which cause the signal to arrive at the detector together with spurious replicas which interfere with the signal to cause an amplitude fluctuation on the resulting photocurrent. The effects of IN are often much more severe than for linear crosstalk since the perturbation generated is proportional to $\sqrt{\epsilon}$, where $\epsilon$ is the power attenuation of the generating interferer, and is $< 1$.

Analysis and simulation of IN was extended to address the impact on heterodyning HFR systems in which two optical carriers of displaced frequency are used to generate an intermediate RF carrier. The analytical study predicted IN signal distributions for these systems by consideration of the beating combinations between the two signal components and the crosstalk replicas for OOK modulation. These distributions were confirmed by simulation. The analysis also predicted the impact of IN in terms of BER and power penalty, based on formulation and bounding of the moment generating functions for the IN distributions. Simula-
tion results confirmed the predicted error rates and showed that FSK modulation was much more resilient to IN than ASK.

7.2 Summary of Thesis

Introductory material was presented in Chapter 2 with the aim of providing context to the project and presenting the current state of development of the technologies studied. Alternative access networks were presented to provide context for wireless access; millimetre-wave systems were then identified as a likely future wireless technology. Relevant aspects of optical network technology were also covered, with a view to introducing hybrid fibre radio (HFR) systems, in which millimetre-wave networks might be supported by optical transport infrastructure. The current state of development of HFR systems was critically reviewed in the final section which concluded with a presentation of two technology demonstration projects.

Chapter 3 provided a brief introduction to the various simulation and modelling techniques available for optical network analysis. Several software options were reviewed, culminating in the choice of Signal Processing Worksystem (SPW) as the preferred development environment. A description of SPW followed, focusing in particular on the object-oriented design interface and methods developed for handling broadband optical signals.

Modelling feedback in fibre networks was tackled in Chapter 4, with several custom-coded modules for SPW presented. These included feedback generation blocks and models for a semiconductor laser and EDFA. Results obtained from the models were presented, indicating the effects of optical feedback on the laser in terms of eye-closure and induced chirp, and for the EDFA in terms of ASE emission and gain.

The effect of forward propagating crosstalk was investigated in Chapter 5, which studies the impact of interferometric noise on heterodyning HFR systems. A general mathematical description of IN was developed for OOK modulation, treating
crosstalk as a sum of attenuated and time-delayed replicas of the carrier signals, with no phase correlation. The moment generating functions for the IN distributions were found and the system impact determined using the modified Chernoff bound and by Gaussian approximation. Analytical results were confirmed by simulation, which also predicted the impact of IN on phase-related modulation schemes would be much reduced.

Demonstration of the developed custom-coded modules, hierarchical blocks and techniques as complete systems was provided in Chapter 6, which used HFR networks as examples. The first system used direct modulation of the semiconductor laser module to produce the carrier frequency and output signals from the system were presented in terms of eye-diagrams and signal waveforms. MODAL and FRANS were then studied for susceptibility to IN generation, as specific examples of current heterodyning HFR technology. The MODAL study showed that FSK offers much greater tolerance to IN, confirming predictions from Chapter 5. Crosstalk isolation requirement to achieve given BERs were presented for both ASK and FSK schemes and were found to be consistent with analytical predictions. A multi-frequency FRANS system was also described and simulations used to determine IN impact and show the absence of an error floor as predicted by the analysis.

The work presented further builds on an established optical network modelling environment and contributes additional tools and techniques to its further development. A modelling capability for capturing forward and backward propagating spurious signals has been addressed for application to several specific and some general optical network types. The impact of IN in heterodyning HFR systems has been evaluated and confirmed by simulation, including end-to-end system studies which demonstrate the capability of the developed tools to capture effects and aid design of complete optical networks.
7.3 Suggested Further Work

There is scope for further work in the general areas covered in this Thesis. The developed software and techniques build on the already well developed optical modelling capability in SPW and further system studies should be easy to undertake, with little or no additional development required. Some suggestions for further work follow:

- Extension of feedback modelling to include other network devices such as the semiconductor optical amplifier and wavelength translation devices.

- Development of the studies systems (particularly FRANS in view of its experimental development and deployment) to include higher order modulation schemes. FRANS is likely to employ 16-QAM in its field trials.

- Verification of the effects of IN in optical heterodyning HFR systems by experiment. These might use the basic MODAL carrier generators developed under the RACE programme or take place as part of the development and field trials of FRANS under the ACTS programme.

- Extend the range of optimum parameters for optical heterodyning HFR systems to provide resilience to IN. These might include optimum receiver decision thresholds and various modulation parameters (such as extinction ratio, modulation depth, type of scheme) including further work on the use of angle modulation schemes, including phase modulation.

- Produce a study of typical spurious signal generation in fibre networks, characterising both reflected and scattered feedback and alternate path generation through different types of network. Statistical data might be obtained by considering the structure of various network types, including the positioning of potentially reflective elements such as splices, cross-connects, splitters and couplers and by determining the effect of any isolators employed. These statistics might then be emulated in the simulator to give more realistic indicators of the potential impact on systems.
Appendix A

Derivation of Interferometric Noise MGF

This Appendix includes the full analysis of interferometric noise in HFR systems presented in Chapter 5, where reference was made to the results and intermediate workings presented here. A full description of coherent mixing can be found in Section 2.4.1 and IN generation in Section 2.3.4. The FRANS system and generated optical signal components are described in Chapter 5.

The appendix concludes with a derivation of the modified Chernoff bound (MCB) which is used to assess system performance implications.
The component optical electric fields impinging on the photodetector in a system employing coherent mixing at the receiver, are given by

\[ A = E_1 e^{j(\omega_0+\frac{\omega_m}{2})t+j\phi(t)} \]

\[ B = m(t)E_2 e^{j(\omega_0-\frac{\omega_m}{2})t+j\phi(t)} \]

\[ a = E_1 \sum_{k=1}^{N} \sqrt{\varepsilon_k} e^{j(\omega_0+\frac{\omega_m}{2})(t-\tau_k)+j\phi(t-\tau_k)} \]

\[ b = E_2 \sum_{k=1}^{N} \sqrt{\varepsilon_k} m(t-\tau_k) e^{j(\omega_0-\frac{\omega_m}{2})(t-\tau_k)+j\phi(t-\tau_k)} \]

(A.1)

where \( m(t) \) is the modulating signal, which for OOK is given by

\[ m(t) = \sum_{l=-\infty}^{\infty} a_l p(t-lT) \]  

(A.2)

where \( E \) is the optical signal electrical field amplitude, \( \omega_m \) is the millimetre wave angular frequency, \( a_l \) is the random binary data; \( a_l \in \{1,0\} \) with both events having equal probability of occurrence. \( p(t) \) is the digital data pulse shape, \( \omega_o \) is the angular frequency of the laser diode, \( \phi(t) \) is the laser phase noise, \( \tau_k \) represent the time delays of the interferers, \( N \) is the number of interferers, \( \varepsilon_k \) is the crosstalk isolation as a ratio between the power of the \( k^{\text{th}} \) interferer and the power of the wanted signal.

Referring to Equation A.1: \( A \) is the first, unmodulated optical carrier; \( B \) is the second, modulated optical carrier; \( a \) represents the sum of crosstalk components arising from the first optical carrier; \( b \) represents the crosstalk arising from the second, modulated carrier. Assuming all components impinge on a receiver of normalised responsivity, the resulting photocurrent will be the square of the sum of these signals

\[ i(t) = |A + B + a + b|^2 = |A + B + a + b||A + B + a + b|^* = |A|^2 + |B|^2 + |a|^2 + |b|^2 + 2\Re(AB^*) + 2\Re(Aa^*) + 2\Re(Ab^*) + 2\Re(Ba^*) + 2\Re(Bb^*) + 2\Re(ab^*) \]  

(A.3)

Resolving these terms yields: the linear signal power \( |A|^2 + |B|^2 \), given by

\[ |A|^2 = E_1^2 \]

\[ |B|^2 = E_2^2 |m(t)|^2 \]  

(A.4)
the linear crosstalk power $|a|^2 + |b|^2$, given by

$$ |a|^2 = E_1^2 \sum_{k=1}^{N} \sum_{l=1}^{N} \sqrt{\varepsilon_k \varepsilon_l} e^{j(\omega_0 + \omega_m \tau_k)(t - \tau_k)} e^{-j(\omega_0 + \omega_m \tau_l)(t - \tau_l)} $$

$$ = E_1^2 \sum_{k=1}^{N} \sum_{l=1}^{N} \sqrt{\varepsilon_k \varepsilon_l} e^{j(\omega_0 + \omega_m \tau_k)(t - \tau_k) + j\phi(t - \tau_k) - j\phi(t - \tau_l)} $$

$$ = E_1^2 \sum_{k=1}^{N} \varepsilon_k + E_1^2 \sum_{k=1}^{N} \sum_{l=1}^{N} \sqrt{\varepsilon_k \varepsilon_l} e^{j(\omega_0 + \omega_m \tau_k)(t - \tau_k) + j\phi(t - \tau_k) - j\phi(t - \tau_l)} $$

$$ = E_1^2 \sum_{k=1}^{N} \varepsilon_k + E_1^2 \sum_{k=1}^{N} \sum_{l=1}^{N} \sqrt{\varepsilon_k \varepsilon_l} \cos \left( \frac{\omega_m}{2} (\tau_j - \tau_k) + \phi(t - \tau_k) - \phi(t - \tau_j) \right) \quad (A.5) $$

following the same procedure for $|a|^2$ gives for $|b|^2$

$$ |b|^2 = E_2^2 \sum_{k=1}^{N} \varepsilon_k |m(t - \tau_k)|^2 + E_2^2 \sum_{k=1}^{N} \sum_{j=1}^{N} \sqrt{\varepsilon_k \varepsilon_j} m(t - \tau_k) m^*(t - \tau_j) + \cos \left( \frac{\omega_m}{2} (\tau_j - \tau_k) + \phi(t - \tau_k) - \phi(t - \tau_j) \right) \quad (A.6) $$

the primary (signal-signal) beating, which forms the ‘wanted’ modulated millimetre-wave carrier signal, given by

$$ 2\Re(AB^*) = 2E_1E_2 \Re \left[ e^{j(\omega_0 + \omega \tau_k) t + j\phi(t)} e^{-j(\omega_0 + \omega \tau_l) t - j\phi(t)} m^*(t) \right] $$

$$ = 2E_1E_2 \Re [m(t)] \cos(\omega_m t) \quad (A.7) $$

the secondary (signal-crosstalk) beating $2\Re [(Aa^*) + (Ab^*) + (Ba^*) + (Bb^*)]$, which will form the salient interferometric noise contribution, given by

$$ 2\Re(Aa^*) = 2E_1^2 \Re \left[ e^{j(\omega_0 + \omega \tau_k) t + j\phi(t)} \sum_{k=1}^{N} \sqrt{\varepsilon_k} e^{-j(\omega_0 + \omega \tau_l) t - j\phi(t)} \right] $$

$$ = \Re \left[ 2E_1^2 \sum_{k=1}^{N} \sqrt{\varepsilon_k} e^{j(\omega_0 + \omega \tau_k) t + j\phi(t)} \right] \quad (A.8) $$

$$ 2\Re(AB^*) = 2 \Re \left[ E_1 e^{j(\omega_0 + \omega \tau_k) t + j\phi(t)} E_2 \sum_{k=1}^{N} \sqrt{\varepsilon_k} m^*(t - \tau_k) e^{-j(\omega_0 + \omega \tau_l)(t - \tau_k) - j\phi(t - \tau_k)} \right] $$
APPENDIX A. DERIVATION OF INTERFEROMETRIC NOISE MGF

\[ 2E_1E_2 \Re \left[ \sum_{k=1}^{N} \sqrt{\varepsilon_k} m^*(t - \tau_k) e^{j(\omega_0 \tau_k + \omega_m t - \frac{\omega_m \varepsilon_k}{2})} e^{j\phi(t) - j\phi(t - \tau_k)} \right] \]

\[ = 2E_1E_2 \sum_{k=1}^{N} \sqrt{\varepsilon_k} m^*(t - \tau_k) \cos \left[ \omega_m t + \left( \omega_0 - \frac{\omega_m}{2} \right) \tau_k \right. \]

\[ + \phi(t) - \phi(t - \tau_k) \]  \hspace{1cm} (A.9)

\[ 2\Re(Ba^*) = 2 \Re \left[ m(t) e^{j\left(\omega_0 - \frac{\omega_m}{2}\right)t + j\phi(t)} E_1E_2 \sum_{k=1}^{N} \sqrt{\varepsilon_k} e^{-j\left(\omega_0 - \frac{\omega_m}{2}\right)(t - \tau_k) - j\phi(t - \tau_k)} \right] \]

\[ = 2E_1E_2 \Re[m(t)] \Re \left[ \sum_{k=1}^{N} \sqrt{\varepsilon_k} e^{-j\omega_m t + j\left(\omega_0 - \frac{\omega_m}{2}\right)\tau_k} e^{j\phi(t) - j\phi(t - \tau_k)} \right] \]

\[ = 2E_1E_2 \Re[m(t)] \sum_{k=1}^{N} \sqrt{\varepsilon_k} \cos \left[ \omega_m t - \left( \omega_0 - \frac{\omega_m}{2} \right) \tau_k - \phi(t) \right] \]

\[ + \phi(t - \tau_k) \]  \hspace{1cm} (A.10)

\[ 2\Re(Bb^*) = 2 \Re \left[ m(t) e^{j\left(\omega_0 - \frac{\omega_m}{2}\right)t + j\phi(t)} E_2^2 \sum_{k=1}^{N} \sqrt{\varepsilon_k} m^*(t - \tau_k) \right. \]

\[ \times e^{-j\left(\omega_0 - \frac{\omega_m}{2}\right)(t - \tau_k) - j\phi(t - \tau_k)} \right] \]

\[ = 2E_2^2 \sum_{k=1}^{N} \sqrt{\varepsilon_k} \Re[m(t)] \Re[m(t - \tau_k)] \Re \left[ e^{j\left(\omega_0 - \frac{\omega_m}{2}\right)\tau_k} e^{j\phi(t) - j\phi(t - \tau_k)} \right] \]

\[ = 2E_2^2 \sum_{k=1}^{N} \sqrt{\varepsilon_k} \Re[m(t)] \Re[m(t - \tau_k)] \cos \left[ \left( \omega_0 - \frac{\omega_m}{2} \right) \tau_k \right. \]

\[ + \phi(t) - \phi(t - \tau_k) \]  \hspace{1cm} (A.11)

and the tertiary (crosstalk-crosstalk) beating, given by

\[ 2\Re(ab^*) = 2E_1E_2 \Re \left[ \sum_{k=1}^{N} \sqrt{\varepsilon_k} e^{j\left(\omega_0 + \frac{\omega_m}{2}\right)(t + \tau_k) + j\phi(t - \tau_k)} \right. \]

\[ \times \sum_{j=1}^{N} \sqrt{\varepsilon_j} m^*(t - \tau_j) e^{-j\left(\omega_0 - \frac{\omega_m}{2}\right)(t - \tau_j) - j\phi(t - \tau_j)} \right] \]

\[ = 2E_1E_2 \sum_{k=1}^{N} \varepsilon_k \Re[m(t - \tau_k)] \cos[\omega_m(t - \tau_k)] \]

\[ + 2E_1E_2 \sum_{k=1}^{N-1} \sum_{j=1 \neq k}^{N} \sqrt{\varepsilon_k \varepsilon_j} \Re[m(t - \tau_j)] \cos \left[ \omega_m t + \omega_0 (\tau_j - \tau_k) \right. \]

\[ - \frac{\omega_m}{2} (\tau_k - \tau_j) + \phi(t - \tau_k) - \phi(t - \tau_j) \]  \hspace{1cm} (A.12)
We assume the total electric field is incident upon a detector of normalised responsivity and with a bandpass filter centred on $\omega_m$. Thus we neglect terms not centred around the millimetre-wave frequency, giving a post-detection photocurrent of

$$i(t) = 2E_1E_2 \Re \{m(t)\} \cos(\omega_m t)$$

$$+ 2E_1E_2 \sum_{k=1}^{N} \sqrt{\varepsilon_k} m^*(t - \tau_k) \cos \left[ \omega_m t + \left( \omega_o - \frac{\omega_m}{2} \right) \tau_k \right]$$

$$+ \phi(t) - \phi(t - \tau_k)$$

$$+ 2E_1E_2 \Re \{m(t)\} \sum_{k=1}^{N} \sqrt{\varepsilon_k} \cos \left[ \omega_m t - \left( \omega_o - \frac{\omega_m}{2} \right) \tau_k - \phi(t) \right]$$

$$+ \phi(t - \tau_k)$$

$$+ 2E_1E_2 \sum_{k=1}^{N} \varepsilon_k \Re \{m(t - \tau_k)\} \cos[\omega_m(t - \tau_k)]$$

$$+ 2E_1E_2 \sum_{k=1}^{N} \sum_{j=1 \neq k}^{N} \sqrt{\varepsilon_k \varepsilon_j} \Re \{m(t - \tau_j)\} \cos \left[ \omega_m t + \omega_o(\tau_j - \tau_k) - \frac{\omega_m}{2}(\tau_k - \tau_j) + \phi(t - \tau_k) - \phi(t - \tau_j) \right]$$

(A.13)

The last two terms represent tertiary beating between the modulated and unmodulated components of the replica signals, appearing at $\sqrt{\varepsilon_k \varepsilon_j}$. Since the secondary crosstalk appears at $\sqrt{\varepsilon_k \varepsilon_j} \gg \sqrt{\varepsilon_k \varepsilon_j}$, we neglect the tertiary beating as a negligible contribution to the total interferometric noise.

Applying coherent demodulation - by multiplying the detected signal by a local oscillator (LO) signal $K \cos(\omega_m t)$ and filtering out multiple carrier frequencies - gives

$$i_0(t) = KE_1E_2 p(t)$$

$$+ KE_1E_2 \sum_{k=1}^{N} \sqrt{\varepsilon_k} p(t - \tau_k) \cos(\phi_k)$$

$$+ KE_1E_2 \sum_{k=1}^{N} \sqrt{\varepsilon_k} \cos(\phi_k) p(t)$$

$$+ n(t)$$

(A.14)

where

$$n(t) = \text{Gaussian receiver noise}$$
and where
\[ \phi_k = \left( \omega_0 - \frac{\omega_m}{2} \right) \tau_k + \phi(t) - \phi(t - \tau_k) \]
and modulation function \( p(t) \) is an OOK bit stream, is given by
\[ p(t) = \sum_k a_k \text{rect} \left( \frac{t - kT}{T} \right) \]
The constant \( K \) results from multiplication of the detected signal by the LO signal to produce the intermediate frequency (IF) and subsequent multiplication to produce the base-band signal.

If we consider the case of a single interfering component, i.e. \( N = 1 \), Equation A.14 becomes
\[ i_0(t) = E_1 E_2 a_0 + E_1 E_2 \sqrt{\varepsilon_1} a_1 \cos \phi + E_1 E_2 \sqrt{\varepsilon_1} a_0 \cos \phi \quad \text{(A.15)} \]
where \( a_0 \) and \( a_1 \) are the levels associated with bit transmissions at \( t \) and \( t - \tau \) respectively.

If we now consider the case of a zero bit transmission, with full extinction (i.e. \( a_0 = 0 \)), we obtain, for a sample interval \( t_s \)
\[ i_0(t_s) = E_1 E_2 \sqrt{\varepsilon_1} \cos \phi a_1 \quad \text{(A.16)} \]
From this we can derive the moment generating function for a zero transmission:
\[ M_{i_0}(s) = \frac{1}{2} + I_0 \left( s E_1 E_2 a \sqrt{\varepsilon_1} \right) \quad \text{(A.17)} \]
where \( a \) is an arbitrary bit level.

Similarly, considering the case of a unity bit transmission gives
\[ i_0(t_s) = E_1 E_2 a \]
\[ + E_1 E_2 \sqrt{\varepsilon_1} a_1 \cos \phi a_1 \]
\[ + E_1 E_2 \sqrt{\varepsilon_1} a_0 \cos \phi a_1 \]
this yields an MGF
\[ M_{i_0}(s) = e^{s E_1 E_2 a} \]
\[ + \frac{1}{2} \left( 1 + I_0(s \sqrt{\varepsilon} a E_1 E_2) \right) \]
\[ + I_0(s \sqrt{\varepsilon} a E_1 E_2) \quad \text{(A.19)} \]
So for the general case of a sum of interferers, the MGFs for a zero bit transmissions becomes

\[
M_{IN_0}(s) = \prod_{k=1}^{N} \left[ \frac{1}{2} I_0 \left( s \sqrt{\varepsilon_k} a E_1 E_2 \right) + \frac{1}{2} \right] \quad (A.20)
\]

and for a unity bit transmission

\[
M_{IN_1}(s) = M_{IN_0}(s) \prod_{k=1}^{N} I_0 \left( s \sqrt{\varepsilon_k} a E_1 E_2 \right) \quad (A.21)
\]

From these expressions we can derive expressions for the variance on the IN

\[
\sigma_0^2 = \left. \frac{d^2 M_0(s)}{ds^2} \right|_{s=0} = \frac{a^2 (E_1 E_2)^2}{4} \sum_{k=1}^{N} \varepsilon_k \quad (A.22)
\]

\[
\sigma_1^2 = \left. \frac{d^2 M_1(s)}{ds^2} \right|_{s=0} = \frac{3a^2 (E_1 E_2)^2}{4} \sum_{k=1}^{N} \varepsilon_k \quad (A.23)
\]
Appendix B

Software Documentation

This Appendix includes software documentation for some of the CCBs developed during the project. The manuals are written in the style used by GEC-Marconi Research Centre for documenting CCBs, hierarchical blocks and top-level systems in SPW. The documents are self-contained and cited references are listed at the end of each module specification. The hardware and software specification listed in the first section is common to all documented modules and is thus omitted thereafter.
B.1 DFBv9

Semiconductor Laser Diode Model

Module name: DFBv9
Library name: mattlib
Issue status: 1
Date: 16 / 6 / 97
Author: Matthew Darby

Abstract

This module simulates the behaviour of a semiconductor diode laser through solution of coupled rate equations for the photon and carrier populations in the device. Direct modulation can be applied using a drive current input port and output is in terms of instantaneous power or as a complex exponential. Various editable parameters relating to laser specification are provided together with additional parameters relating to linear feedback from the network into the device. Laser behaviour is modelled. The block may be held.

Description

This block represents a semiconductor diode laser model with additional parameters allowing linear optical feedback into the device to be modelled. Provision is made for up to 12 reflections and Rayleigh backscattering to be simulated. The feedback is characterised as linear power crosstalk which affects the photon and carrier population densities in the active region of the device. These are given by

\[
\frac{dP(t)}{dt} = G \cdot P(t) - \frac{P(t)}{T_p} + R_{sp}(t) + S(t)
\]

\[
\frac{dN(t)}{dt} = \frac{I(t)}{q} - \frac{N(t)}{T_n} - G \cdot P(t)
\]

where \( P \) is the photon number, \( N \) is the carrier number, \( T_p \) is the average photon lifetime in the active region, \( T_n \) is the carrier lifetime, \( G \) is the device gain, \( R_{sp}(t) \)
is the spontaneous emission factor, $N(t)$ is the carrier population, $I(t)$ is the drive current injection into the active region and $q$ is the charge on an electron. Derivations of these equations can be found in [3].

$S(t)$ is a feedback injection term which is a sum of contributions from up to 12 reflectors and backscattering which have been specified in the parameter screen. The scattering is calculated using the same method as the Rayleigh_dist block, taking in to account the optical power distribution within the fibre [1]. The method differs in that the DFBv9 block does not model frequency sampled signals, using only a monochromatic wavelength channel with scalar signal representation. Similarly reflections are handled as described in the Vect_fibredly block [2], again using scalar signal representation. Inputs to the block are for sampling frequency, drive current and optical feedback. The drive current is in terms of amps and the modulating signal should be offset by a bias, which may be added directly to the signal or specified in the parameter screen for summing to each sample of the modulating input signal. The optical feedback input allows external modules to be connected and provide linear crosstalk power injection instead of, or in addition to, the blocks internal feedback calculations. Output from the block is in the form of scalar quantities of power and chirp and also as a complex electric field. The scalar power output may be inserted into a broadband optical signal vector at the appropriate frequency sample bin for passing to a broadband optical component model [1]. Chirp is relative to a reference frequency, which is defined in the parameter screen.

**Input and Output Vectors**

Table B.1 details the input and output ports to the block.

**Parameters**

Table B.2 details all editable parameters which appear in the parameter screen, with default values for a DFB laser implementation.
## Error Conditions

The most likely causes of a floating point exception error in the module is the setting of parameters or inputs to ‘unrealistic’ values. Typical values for a DFB implementation are set in the original model instance. Parameters most likely cause to a floating point exception are

- Incorrect scaling of the drive current. Typically only a few tens of mA are required.

- Incorrect of the bias current level. Again, typically only a few tens of mA are required. Note that the parameter bias should be set to zero if already included in the drive signal file.

- Too much optical feedback has been applied.

- ‘Internal’ sampling frequency parameter is too large. This needs to be set at a value smaller than the photon and carrier average lifetimes, typically around $10^{12}\text{Hz}$.

- Sampling frequency input is disconnected.

Aliasing problems in the output, particularly the complex electric field at high induced chirp rates, will result from an insufficiently high sampling frequency.
<table>
<thead>
<tr>
<th>Name</th>
<th>Type</th>
<th>Default Value</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>IB</td>
<td>double</td>
<td>20e-3</td>
<td>bias value for drive current input</td>
</tr>
<tr>
<td>IR</td>
<td>double</td>
<td>35e-3</td>
<td>drive current bias required for operation at the reference wavelength</td>
</tr>
<tr>
<td>Sam</td>
<td>double</td>
<td>1e12</td>
<td>'internal' sampling frequency at which the rate equations are solved. This should be an exact multiple of the system sampling frequency and less than the photon and electron average lifetimes</td>
</tr>
<tr>
<td>W</td>
<td>double</td>
<td>1550e-9</td>
<td>wavelength of operation (m)</td>
</tr>
<tr>
<td>L</td>
<td>double</td>
<td>5.0</td>
<td>linewidth enhancement factor</td>
</tr>
<tr>
<td>V</td>
<td>double</td>
<td>5.6e-17</td>
<td>volume of the active region</td>
</tr>
<tr>
<td>GC</td>
<td>double</td>
<td>4.5e-23</td>
<td>gain compression factor</td>
</tr>
<tr>
<td>O</td>
<td>double</td>
<td>0.4</td>
<td>optical confinement factor</td>
</tr>
<tr>
<td>NO</td>
<td>double</td>
<td>1e-24</td>
<td>carrier density at transparency</td>
</tr>
<tr>
<td>DG</td>
<td>double</td>
<td>2.5e-20</td>
<td>gain coefficient</td>
</tr>
<tr>
<td>B</td>
<td>double</td>
<td>3.9e-4</td>
<td>spontaneous emission factor</td>
</tr>
<tr>
<td>CT</td>
<td>double</td>
<td>1-9</td>
<td>average carrier lifetime in the active region</td>
</tr>
<tr>
<td>PT</td>
<td>double</td>
<td>1.4e-12</td>
<td>average photon lifetime in the active region</td>
</tr>
<tr>
<td>CG</td>
<td>double</td>
<td>74948114.5</td>
<td>group velocity (m/s)</td>
</tr>
<tr>
<td>A</td>
<td>double</td>
<td>45e2</td>
<td>alpha m</td>
</tr>
<tr>
<td>yesscat</td>
<td>string</td>
<td>'y'</td>
<td>determines whether scattering is to be simulated: Only a 'y' response will initiate the scattering calculation</td>
</tr>
<tr>
<td>scatt coeff</td>
<td>double</td>
<td>3.283e-5</td>
<td>Rayleigh scattering coefficient for the fibre in terms of $\lambda^{-4}$ [1]</td>
</tr>
<tr>
<td>raylength</td>
<td>double</td>
<td>100.0</td>
<td>fibre length to be simulated for Rayleigh scattering (m) factor which reduces the computation time and spatial accuracy of the scattering calculation by integer multiples</td>
</tr>
<tr>
<td>undersam</td>
<td>integer</td>
<td>1</td>
<td></td>
</tr>
<tr>
<td>n</td>
<td>double</td>
<td>1.50</td>
<td>refractive index of the fibre core</td>
</tr>
<tr>
<td>n2</td>
<td>double</td>
<td>1.47</td>
<td>refractive index of the fibre cladding</td>
</tr>
<tr>
<td>aten</td>
<td>double</td>
<td></td>
<td>Fibre attenuation (dB / km)</td>
</tr>
<tr>
<td>Pig</td>
<td>double</td>
<td>300.0</td>
<td>Length of the fibre pigtail (mm)</td>
</tr>
<tr>
<td>R</td>
<td>double</td>
<td>0.0</td>
<td>Reflectivity at the pigtail connector</td>
</tr>
<tr>
<td>yesrefl</td>
<td>string</td>
<td>'y'</td>
<td>Determines whether reflections are to be simulated (including the pigtail connector): Only a 'y'responsescalingse will initiate the reflections calculations</td>
</tr>
<tr>
<td>r1 to r12</td>
<td>double</td>
<td>0.0</td>
<td>Reflectivities for reflectors 1 to 12 (normalised power isolation)</td>
</tr>
<tr>
<td>x1 to x12</td>
<td>double</td>
<td>1.0</td>
<td>Propagation distances to reflectors 1 to 12 (m)</td>
</tr>
</tbody>
</table>

Table B.2: Editable parameters for DFBv9 CCB
The required sampling frequency rate may be reduced by changing the reference operation frequency $W$ and associated drive current $I_R$.

**Computer Configuration**

This specification is common to all modules documented for ‘mattlib’.

<table>
<thead>
<tr>
<th>Hardware</th>
<th>Sun SPARCstation 5</th>
</tr>
</thead>
<tbody>
<tr>
<td>Operating system</td>
<td>SunOS 4.1.3 / BSD 4.3</td>
</tr>
<tr>
<td>SPW version</td>
<td>3.1</td>
</tr>
<tr>
<td>Compiler version</td>
<td>GCC version 2.5.8</td>
</tr>
</tbody>
</table>

**References**


(a) Laser symbol

(b) Parameter screen, showing typical values for a DFB implementation

Figure B.1: BDE symbol and parameter screen for laser CCB
B.2 mulEDFA_fb

Multichannel EDFA Model

Module name: MULEDFA_FB
Library name: mattlib
Issue status: 1
Date: 16 / 6 / 97
Author: S. L. Zhang modified by M. Darby

Abstract

This module simulates the behaviour of an Erbium-doped fibre amplifier (EDFA). Output is in terms of optical signal power, amplified spontaneous emission (ASE) and amplifier gain. Broadband optical signals and multi-wavelength channel transmission are supported. The device is modelled through iterative solution of rate equations and treats the doped fibre as a series of concatenated sections, allowing dynamic gain distribution within the fibre to be captured. The block may be held.

Description

The behaviour of an Erbium-doped fibre amplifier (EDFA) is simulated by this module. Broadband signal representation is achieved through frequency sampling of input and output power signals. Each power signal spectrum is defined by a wavelength reference vector, which is the same length as the signal vector and contains the central wavelength value for each sample bin [1]. The doped fibre is modelled as a series of concatenated sections which allows dynamic gain and signal variation along the fibre length to be captured. Rate equations for the densities of the photons and excited Erbium ions within each section are solved iteratively and steady state solutions found. These solutions are passed to the two adjacent sections, in the direction of propagation, and used as initialisation
values for solution for these sections. In this way signals are modelled through the
doped fibre in forward and backward directions simultaneously. Detail regarding
the iterative solution of the rate equations and method of concatenated sections
representation can be found in [2].

Two additional ports were added as a modification of a previous model [2] for
this issue: Vector input ports named pre.fb and post.fb. These ports allow
optical feedback to be entered into the model. Optical feedback power generation
modules should be connected directly to these ports: Post.fb handles feedback
travelling backwards from the network into the EDFA and pre.fb handles reflected
feedback travelling in the signal direction and originating in the network before
the EDFA.

The modelling of complex electric fields is not handled directly by the module.
Optical power signals should be applied to the model as described, in order to
obtain the gain developed by the device, which is defined in terms of the sigwave
wavelength reference comb. A wavelength dependent value can be extracted [3]
from the gain vector and applied to the electric field signal by way of a complex
multiplier block [4].

**Input and Output Vectors**

Table B.3 details the input and output ports to the block. All vectors are of
length DEFAULT_VECLEN (see Table B.4). Data for emispec, abspec fasepow,
basepow etc. which is less than this value must be followed by null data in
order to comprise a vector of DEFAULT_VECLEN length for input to the block.
Vectors which are a concatenation of multiple signals (such as sigoutp) must be
decomposed using extraction blocks set to the length of the constituent vectors
in order to manipulate the separate signals.

Signal source blocks used as vector data input may be held in their output state
for the duration of the simulation. It is recommended that these vector inputs
are included at a lower-level with signal and feedback outputs only connected
### Table B.3: Input and output port descriptions for mulEDFA fb CCB

<table>
<thead>
<tr>
<th>Name</th>
<th>Input Type</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>emispec</td>
<td>vector double</td>
<td>doped fibre emission spectra</td>
</tr>
<tr>
<td>absspec</td>
<td>vector double</td>
<td>doped fibre absorption spectra</td>
</tr>
<tr>
<td>specwave</td>
<td>vector double</td>
<td>wavelength definition comb for emispec and absspec</td>
</tr>
<tr>
<td>sigpower</td>
<td>vector double</td>
<td>input signal power</td>
</tr>
<tr>
<td>sigwave</td>
<td>vector double</td>
<td>wavelength definition comb for sigpower</td>
</tr>
<tr>
<td>sponwave</td>
<td>vector double</td>
<td>wavelength definition comb for ASE output</td>
</tr>
<tr>
<td>spdeltan</td>
<td>vector double</td>
<td>wavelength spacing between each consecutive pair of sponwave definition</td>
</tr>
<tr>
<td>gammas</td>
<td>vector double</td>
<td>fibre containment factor at the wavelengths defined in the sponwave comb</td>
</tr>
<tr>
<td>gamman</td>
<td>vector double</td>
<td>fibre containment factor at the wavelengths defined in the sigwave comb</td>
</tr>
<tr>
<td>alphas</td>
<td>vector double</td>
<td>fibre loss at the wavelengths defined in the sigwave comb</td>
</tr>
<tr>
<td>alphan</td>
<td>vector double</td>
<td>fibre loss at the wavelengths defined in the sponwave comb</td>
</tr>
<tr>
<td>pre_fb</td>
<td>vector double</td>
<td>forwards propagating optical feedback power (generated in the network before the amplifier)</td>
</tr>
<tr>
<td>post_fb</td>
<td>vector double</td>
<td>backwards propagating optical feedback power (generated in the network after the amplifier)</td>
</tr>
<tr>
<td>hold</td>
<td>double</td>
<td>standard SPW hold input (held if positive)</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Name</th>
<th>Output Type</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>basepow</td>
<td>vector double</td>
<td>optical output power (W)</td>
</tr>
<tr>
<td>fasepow</td>
<td>vector double</td>
<td>forward propagating ASE power (produced locally in this amplifier)</td>
</tr>
<tr>
<td>asewave</td>
<td>vector double</td>
<td>backwards propagating ASE power (produced locally in this amplifier)</td>
</tr>
<tr>
<td>asedelta</td>
<td>vector double</td>
<td>induced frequency chirp (Hz)</td>
</tr>
<tr>
<td>sigoutp</td>
<td>vector double</td>
<td>signal output power and forward ASE power in a vector splice; the vector is thus of length sigoutw = asewave + sigwave</td>
</tr>
<tr>
<td>sigoutw</td>
<td>vector double</td>
<td>wavelength definition comb for sigoutp, comprising two concatenated vectors: first vector defines the signal, second defines the ASE</td>
</tr>
<tr>
<td>gainsig</td>
<td>vector double</td>
<td>amplifier gain produced</td>
</tr>
</tbody>
</table>

Hierarchically to the system-level design.

### Parameters

Table B.4 details the editable parameters of the module, which all appear in the parameter screen.
B.2. MULEDFA.FB \hspace{1cm} MULTICHANNEL EDFA MODEL

<table>
<thead>
<tr>
<th>Name</th>
<th>Type</th>
<th>Default Value</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>DEFAULT_VECLEN</td>
<td>integer</td>
<td>200</td>
<td>length for all input and output vector ports</td>
</tr>
<tr>
<td>aspec</td>
<td>integer</td>
<td>38</td>
<td>number of points in emispec and abspec spectral data</td>
</tr>
<tr>
<td>am</td>
<td>integer</td>
<td>10</td>
<td>number of WDM channels</td>
</tr>
<tr>
<td>an</td>
<td>integer</td>
<td>100</td>
<td>number of spectra points in ASE spectra</td>
</tr>
<tr>
<td>ak</td>
<td>integer</td>
<td>20</td>
<td>number of concatenated sections constituting doped fibre length</td>
</tr>
<tr>
<td>tao21</td>
<td>double</td>
<td>0.01</td>
<td>average photon lifetime of excited state Erbium ions</td>
</tr>
<tr>
<td>ar</td>
<td>double</td>
<td>2.0e-6</td>
<td>radius of the doped fibre core (m)</td>
</tr>
<tr>
<td>Ital</td>
<td>double</td>
<td>20.0</td>
<td>length of the doped fibre (m)</td>
</tr>
<tr>
<td>sigmapa</td>
<td>double</td>
<td>1.85e-25</td>
<td>stimulated absorption cross-section for the doped fibre at the pump wavelength</td>
</tr>
<tr>
<td>sigmape</td>
<td>double</td>
<td>0.7e-25</td>
<td>stimulated emission cross-section for the doped fibre at the pump wavelength</td>
</tr>
<tr>
<td>wavelp</td>
<td>double</td>
<td>1.48</td>
<td>wavelength of the pump light (μm)</td>
</tr>
<tr>
<td>alphap</td>
<td>double</td>
<td>0.0002</td>
<td>fibre loss at the pump wavelength (dB / km)</td>
</tr>
<tr>
<td>gammap</td>
<td>double</td>
<td>0.3</td>
<td>fibre containment factor at the pump wavelength</td>
</tr>
<tr>
<td>Nt</td>
<td>double</td>
<td>5.0e25</td>
<td>density of Erbium ions in the doped fibre (m³)</td>
</tr>
<tr>
<td>iPPump</td>
<td>double</td>
<td>0.1</td>
<td>optical power of the pump source (W)</td>
</tr>
</tbody>
</table>

Table B.4: Editable parameters for mulEDFA_fb CCB

Error Conditions

No warnings are printed. The most likely causes of a floating point exception are the setting of parameters to physically unrealistic values. Values representing an in-line booster configuration are set in the original model instance of the parameter screen, though these will require vector inputs for implementation, which are typically entered from data sheets. Likely causes of problems encountered when running the block include

**Indefinite run-time** can occur if high gain saturation has occurred, causing the model to oscillate. This can be a combination of a doped fibre length which is too long, pump power which is too small or input signal power which is too large.

**Floating point exception** will occur if the optical signal input, sigpower, is zero (through gain calculation). Signal values should be inserted, at the appropriate defined wavelength points, into a negligible valued but non-
zero vector covering the full vector length represented by sigwave. Floating point exceptions are also likely to be caused by incorrect definitions or disconnection of wavelength reference comb vector inputs, which should be specified in metres.

References


B.2. MULEDFA_FB  MULTICHANNEL EDFA MODEL

(a) EDFA symbol

(b) Parameter screen, showing default values

Figure B.2: BDE symbol and parameter screen for EDFA CCB
B.3 Rayleigh_dist

Dynamically Distributed Rayleigh Scattering

Module name: Rayleigh_dist
Library name: mattlib
Issue status: 1
Date: 16/6/97
Author: Matthew Darby

Abstract

This CCB calculates the backscattered optical power from a length of fibre. The block uses a dynamic representation of the optical power distribution within the fibre. Vector representation of the input signal, allows broadband frequency sampled spectra to be modelled. The block may be held.

Description

This block implements a two-dimensional circular buffer to calculate the amount of backscattered optical power generated by a broadband optical signal propagating through a given length of fibre. The optical signal input and output spectra are represented in a broadband format which is defined in terms of a wavelength reference vector [1].

Each input vector is stored to a circular buffer address, representing in two dimensions the distribution of power within the fibre both spatially and spectrally. The fibre is treated as a series of concatenated sections, allowing dynamic power distributions to be captured by the model. The spatial resolution of the calculation (i.e. the length of each fibre section) is set in the parameter screen, allowing a user-defined trade-off between accuracy and computation time. The round trip
from the fibre end to the mid-point of each fibre section is calculated for each section as the light propagates along the fibre, and used to apply attenuation to each section individually. Scattering is calculated in the same way as for the Rayleigh_scatt block [2] using the Rayleigh scattering coefficient, in terms of $\lambda^{-4}$, which defines the material dependent variables as a single parameter

$$\gamma_{coeff} = \frac{8}{3} \pi^3 n^8 \rho^2 \beta \epsilon K T_F$$

where $\rho$ is the average photoelastic coefficient, $\beta$ is the isothermal compressibility at a fictive temperature $T_F$ is the (related to the anneal temperature), and $K$ is Boltzmann’s constant. Resultant scattering is then subject to the angle of acceptance and attenuation of the fibre. The default value for $\gamma_{coeff}$ is $1.895e-28 \lambda^{-4}$, which is typical for silica glass.

Input and Output Vectors

Table B.5 details the input and output ports to the block.

Parameters

Table B.6 details the editable parameter of the module which all appear in the parameter screen.
APPENDIX B. SOFTWARE DOCUMENTATION

<table>
<thead>
<tr>
<th>Name</th>
<th>Type</th>
<th>Default Value</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>DEFAULT_VECLEN</td>
<td>integer</td>
<td>20</td>
<td>length for input and output vector ports</td>
</tr>
<tr>
<td>scattcoeff</td>
<td>double</td>
<td>1.895e-28</td>
<td>Rayleigh scattering proportionality $(\lambda^{-4})$, defined in Section B.3, with a default value typical for silica glass</td>
</tr>
<tr>
<td>spat_freq</td>
<td>double</td>
<td>250.0</td>
<td>default spatial sampling length if S_freq input is disconnected</td>
</tr>
<tr>
<td>def_length</td>
<td>double</td>
<td>1e8</td>
<td>default fibre length if length input is disconnected</td>
</tr>
<tr>
<td>n_core</td>
<td>double</td>
<td>1.5</td>
<td>refractive index for the fibre core</td>
</tr>
<tr>
<td>n_cladding</td>
<td>double</td>
<td>1.47</td>
<td>refractive index for the fibre cladding</td>
</tr>
<tr>
<td>atten</td>
<td>double</td>
<td>3.0</td>
<td>fibre attenuation (dB / km)</td>
</tr>
</tbody>
</table>

Table B.6: Editable parameters for Rayleigh.dist CCB

Error Conditions

A warning is printed if the either fibre length or sampling frequency are initialised to zero. This occurs if either input is disconnected or zero-valued during system initialisation and if the associated default parameters are also set to zero. The simulation will terminate and either of the following messages are displayed in the system view-port: Rayleigh.dist block has one or more length parameters initialised to zero; Rayleigh.dist block has sampling frequency initialised to zero.

Very long run-times may result if the fibre length is long and the spatial resolution is set to a small value. If this is the case the spatial resolution should be increased accordingly.

References


RAYLEIGH_DIST  DYNAMIC RAYLEIGH SCATTERING

(a) Rayleigh.dist symbol

(b) Parameter screen showing default values

Figure B.3: BDE symbol and parameter screen for dynamic Rayleigh scattering CCB
B.4 Rayleigh_scatt

Uniformly Distributed Rayleigh Scattering

Abstract

This CCB calculates the average backscattered optical power from a length of fibre. The block has inputs for average optical power of a broadband signal in vector format, defined by a reference wavelength vector \([1]\). Output is unattenuated. The module is designed for use with the Vect_blockmean and Vect_distatten CCBs \([2,3]\). The block may be held.

Description

This block calculates the backscattered optical power from a length of fibre as a broadband optical signal propagates through it, using an average power approach, which assumes a uniform distribution of optical power along the fibre axis. The signal input accepts a broadband frequency sampled power signal and associated wavelength reference vector \([1]\). The input should be the average optical power in the fibre at each frequency sample, as calculated by Vect_blockmean \([2]\). The scattered power is calculated by

\[
\gamma_{\text{coeff}} = \frac{8}{3} \pi^3 n^8 \rho^2 \beta_c K T_F
\]

where \(\rho\) is the average photoelastic coefficient, \(\beta\) is the isothermal compressibility at a fictive temperature \(T_F\) is the (related to the anneal temperature), and \(K\)
is Boltzmann’s constant. Resultant scattering is then subject to the angle of acceptance and attenuation of the fibre. The default value for $\gamma_{\text{coeff}}$ is 1.895e-28 $\lambda^{-4}$, which is typical for silica glass.

The block also contains a string parameter which allows the block to model the full length of fibre from the first sample of a system following initialisation. This gives quick estimations of scattering from the entire fibre rather than simulating the time taken for the light to propagate the length of the fibre and back.

### Input and Output Vectors

Table B.7 details the input and output ports to the block.

### Parameters

Table B.8 details the editable parameters of the module, which all appear in the parameter screen.
APPENDIX B. SOFTWARE DOCUMENTATION

<table>
<thead>
<tr>
<th>Name</th>
<th>Type</th>
<th>Default Value</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>DEFAULT_VECLEN</td>
<td>integer</td>
<td>20</td>
<td>length for input and output vector ports</td>
</tr>
<tr>
<td>scattcoeff</td>
<td>double</td>
<td>1.895e-28</td>
<td>Rayleigh scattering proportionality (λ^-4), defined by Equation B.1</td>
</tr>
<tr>
<td>def_length</td>
<td>double</td>
<td>1e8</td>
<td>default fibre length if length input is disconnected</td>
</tr>
<tr>
<td>n_core</td>
<td>double</td>
<td>1.5</td>
<td>refractive index for the fibre core</td>
</tr>
<tr>
<td>n_cladding</td>
<td>double</td>
<td>1.47</td>
<td>refractive index for the fibre cladding</td>
</tr>
<tr>
<td>fullinit</td>
<td>string</td>
<td>'no'</td>
<td>option to initialise block to simulate the full fibre length</td>
</tr>
</tbody>
</table>

Table B.8: Editable parameters for Rayleigh_scatt CCB

Error Conditions

A confirmation is printed if the block has parameter fullinit set to 'yes': Rayleigh_scatt block initialised to full length. Continuing...

A warning is printed if the either fibre length or sampling frequency are initialised to zero. This occurs if either input is disconnected or zero-valued during system initialisation and if the associated default parameters are also set to zero. The simulation will terminate and either of the following messages are displayed in the system view-port: Rayleigh_scatt block has one or more length parameters initialised to zero; Rayleigh_scatt block has sampling frequency initialised to zero.

References

RAYLEIGH.Scatt symbol

(a) Rayleigh_scatt symbol

RAYLEIGH SCATTERING BLOCK PARAMETERS

<table>
<thead>
<tr>
<th>MAIN PARAMETERS:</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rayleigh scattering proportionality</td>
<td>1.695e-28</td>
</tr>
<tr>
<td>Default vector length</td>
<td>20</td>
</tr>
<tr>
<td>Default fibre length</td>
<td>250.0</td>
</tr>
<tr>
<td>Refractive index, core</td>
<td>1.5</td>
</tr>
<tr>
<td>Refractive index, cladding</td>
<td>1.17</td>
</tr>
<tr>
<td>Initialise to full length? (‘yes’)</td>
<td>no</td>
</tr>
</tbody>
</table>

(b) Parameter screen, showing typical values for silica glass

Figure B.4: BDE symbol and parameter screen for uniform Rayleigh scattering

CCB
B.5 Vect_blockmean

Vector Mean Estimator

Module name: Vect_blockmean
Library name: mattlib
Issue status: 1
Date: 16/6/97
Authors: Matthew Darby

Abstract

This block performs a mean average calculation of a vector input. The block checks for cancellation and an over-large sample count. The block may be held.

Description

The block takes an input vector of user-defined length and performs a mean average calculation on each element in the same way as the SPW library block 'mean estimator' used in its 'block' mode [1]. The block also checks for cancellation, prevents out of memory errors and may be held.

Input and Output Vectors

Table B.9 details the input and output ports to the block.

Parameters

The only editable parameter, DEFAULT_VECLEN, appears on the block and is detailed in Table B.10.
### Error Conditions

The following error conditions may occur during the execution of a simulation including the module. Any error report is preceded by the line “*****Warning***** Vect_blockmean held on sample no. #” where # is the sample number on which the warning event occurred.

- **due to CANCELLATION** The accuracy of the storage class has been exceeded and the block is held in its current state.
- **due to LARGE SAMPLE COUNT** The maximum sample count allowed before an ‘out of memory’ error occurs has been exceeded, based on the size of the storage class for the system specified in B.1. The block is held in its current output state.

### References

Figure B.5: BDE symbol and parameter screen for Vect.blockmean CCB
B.6 Vect_incatten

Incremental Integrated Attenuation

Abstract

This module applies the integrated attenuation of a length of fibre to an input signal. It is designed to be used with the Rayleigh_scatt block to produce an average backscatter calculation [1]. The block may be incremented with each sample or implemented at a constant length. The block may be held.

Description

This module is designed to be used in conjunction with the Rayleigh_scatt block which calculates the total Rayleigh backscatter from a length of fibre, assuming a uniformly distributed power signal [1]. Backscatter is generated by Rayleigh_scatt from every point in the fibre core and thus experiences an integral of the fibre attenuation during propagation back along the fibre length. This integrated attenuation is calculated by Vect_incatten and applied to an input signal. Vector representation is used for input and output signals, allowing broadband spectra to be modelled [2].

The total attenuation applied to the signal is calculated from

\[ P_{scatt} = \frac{P}{L} \int_{0}^{N} \alpha^2 x \, dx \]

where \( N \) is the length of the fibre, \( P \) is the total scattered power calculated by Rayleigh_scatt and \( \alpha \) is the fibre attenuation, which is specified in dB/km in the
Table B.11: Input and output port descriptions for Vect_incatten CCB

parameter screen. The integration is calculated symbolically for each sample of the simulation, using half the current propagation distance for limit $L$, which defaults to the user-specified fibre length (using an input port or default in the parameter screen) once this value has been reached.

In order to provide compatibility with the Rayleigh_scatt module the parameter screen contains a string parameter which allows the block to model the full length of fibre from the first sample of a system following initialisation. This gives quick estimations of scattering from the entire fibre rather than simulating the time taken for the light to propagate the length of the fibre and back.

**Input and Output Vectors**

Table B.11 details the input and output ports to the block.

**Parameters**

Table B.12 details the editable parameters of the module, which all appear in the parameter screen.
### Table B.12: Editable parameters for Vect_incatten CCB

<table>
<thead>
<tr>
<th>Name</th>
<th>Type</th>
<th>Default Value</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>DEFAULT_VECLEN</td>
<td>integer</td>
<td>20</td>
<td>length for input and output vector ports</td>
</tr>
<tr>
<td>def_length</td>
<td>double</td>
<td>250.0</td>
<td>default fibre length if length input is disconnected</td>
</tr>
<tr>
<td>atten</td>
<td>double</td>
<td>3.0</td>
<td>fibre attenuation (dB / km)</td>
</tr>
<tr>
<td>n_core</td>
<td>double</td>
<td>1.5</td>
<td>refractive index for the fibre core</td>
</tr>
<tr>
<td>fullinit</td>
<td>string</td>
<td>'no'</td>
<td>option to initialise block to simulate the full fibre length</td>
</tr>
</tbody>
</table>

### Error Conditions

A warning is printed if the fibre length is initialised to zero. This occurs if the length input is disconnected or zero-valued during system initialisation and if the associated default parameter is also set to zero. The simulation will terminate and the following message is displayed in the system view-port: Vect_incatten block has length parameters initialised to zero.

Similarly, if the signal frequency input is zero during initialisation the system terminates with: Vect_fibredly block has signal frequency parameters initialised to zero.

### References


INCREMENTAL FIBRE ATTENUATION BLOCK PARAMETERS

<table>
<thead>
<tr>
<th>MAIN PARAMETERS:</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Vector length</td>
<td>20</td>
</tr>
<tr>
<td>Default fibre length</td>
<td>250.0</td>
</tr>
<tr>
<td>Fibre attenuation (dB/km)</td>
<td>3.0</td>
</tr>
<tr>
<td>Refractive index, core</td>
<td>1.5</td>
</tr>
<tr>
<td>Initialise to full length? (‘yes’): no</td>
<td></td>
</tr>
</tbody>
</table>

(b) Parameter screen, showing default values

Figure B.6: BDE symbol and parameter screen for Vect_incatten CCB
B.7 Vect.fibredly

Multiple Back-reflections

Abstract

This CCB simulates the effect of reflected optical feedback from a fibre network. Up to 6 reflective elements may be modelled at different points in the network, the distance to which can be dynamically allocated for each reflector. Vector representation of the input signal allows broadband frequency sampled spectra to be modelled. The block may be held.

Description

This block implements a series of circular buffers with attenuative scaling to simulate the effect of a round trip to various spurious reflections in an optical network. The optical power input is in vector format, allowing a sampled frequency spectrum to be modelled [1]. Other inputs include propagation length definitions for each buffer and for sampling frequency, allowing dynamic allocation of these values at run-time. Parameters in the linked window allow fibre attenuation, refractive index and input/output vector length to be specified. Default propagation distances for each reflector are present and used in the case of the inputs being disconnected or zero-valued.
Input and Output Vectors

Table B.13 details the input and output ports to the block.

Parameters

Table B.14 details the editable parameters of the module, which all appear in the parameter screen.

Error Conditions

To avoid divide-by-zero and memory allocation errors, a warning is printed if the fibre length is initialised to zero. This occurs if any of the length inputs are disconnected or zero-valued during system initialisation and if the associated default
B.7. VECT.FIBREDLY  MULTIPL e BACK- REFLECTIONS

![Diagram of Vect_Fibredly symbol]

(a) Vect_Fibredly symbol

<table>
<thead>
<tr>
<th>VECTOR FIBRE-REFLECTIONS DELAY BLOCK PARAMETERS</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>MAIN PARAMETERS</strong></td>
</tr>
<tr>
<td>Reflectivity</td>
</tr>
<tr>
<td>Default fibre length u (m) 1000.0</td>
</tr>
<tr>
<td>Default fibre length v 1000.0</td>
</tr>
<tr>
<td>Default fibre length w 1000.0</td>
</tr>
<tr>
<td>Default fibre length x 1000.0</td>
</tr>
<tr>
<td>Default fibre length y 1000.0</td>
</tr>
<tr>
<td>Default fibre length z 1000.0</td>
</tr>
<tr>
<td>Vector length 20</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>MISCELLANEOUS PARAMETERS</th>
</tr>
</thead>
<tbody>
<tr>
<td>Initial value 0.0</td>
</tr>
<tr>
<td>Fibre refractive index 1.5</td>
</tr>
<tr>
<td>Fibre Attenuation (dB / km) 3.0</td>
</tr>
</tbody>
</table>

(b) Parameter screen, showing default values

Figure B.7: BDE symbol and parameter screen for Vect_Fibredly CCB

Parameters are also set to zero. The simulation will terminate and the following message is displayed in the system view-port: Vect_Fibredly block has one or more length parameters initialised to zero. Similarly, if the signal frequency input is zero during initialisation, as is the value in the parameter screen, the system terminates with: Vect_Fibredly block has signal frequency parameters initialised to zero.

B.8 Vector Signal Representation

Module name: Vect_fibredly
Library name: mattlib
Issue status: 1
Date: 16/6/97
Authors: Matthew Darby

Abstract

This documentation describes the vector signal format used in several of the optical network modelling blocks found in library mattlib and in other modelling work completed at UCL. The vector format allows broadband optical power signals to be modelled through a network.

Description

Several of the blocks in mattlib, and in other UCL optical network modelling projects, use a common vector signal representation, allowing broadband optical power signals to be modelled through the network. Blocks in mattlib which use this format include mulEDFA_fb [1], Rayleigh.dist [2], Rayleigh_scatt [3] and Vect_fibredly [4].

Signals are stored in vector format, allowing an array of elements to be passed to a single port address during a single system sample period. Each signal stored in this way has two components: a signal vector array containing the amplitude values at each frequency sample; and a frequency reference vector array of the same length as the signal vector, containing the wavelength values at each sample bin of the signal vector array. The vector components of a signal spectrum are shown schematically in Figure B.8. Reading or writing of any element within the power signal spectrum by a block is referenced to the wavelength value stored at the same element position within the reference vector. The power spectrum
signal elements should be expressed in watts and wavelength reference values in metres.

Wavelength reference values are typically saved in a file and input to the system by a signal source which is held in its output state for the duration of the simulation. Wavelength reference combs may be non-linear, as in the case of EDFA input spectra \([2]\), though this must be consistent with the signal type.

### Error Conditions

Problems, and possibly floating point exceptions, will occur if the wavelength reference vector is zero-valued at any element. Spectra which have less samples than the default reference vector for the system should be padded with zero signal vector values at the extra sample points (unless the signal uniquely defined by a separate wavelength reference vector).

### References

[1] M. Darby “mulEDFA_fb, Multichannel EDFA Model” Page 152


Figure B.8: Vector representation of broadband signals in SPW
Bibliography


