TRANSVERSAL FILTER MMIC
DESIGN FOR MULTI-Gbit/s
OPTICAL CDMA SYSTEMS

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Statement of originality

Unless otherwise stated in the text, the work presented in this thesis was carried out by the candidate. It has not been presented previously for any degree, nor is at present under consideration by any other degree awarding body.

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Abstract

In this thesis, the approach of the distributed-amplifier based transversal filter for multi-Gbit/s Optical CDMA systems is addressed. Of particular interest is the research into circuits that enable handling high rate sequences for high-speed system applications. Different distributed transversal filter structures were considered, in particular those that allow extending the range of filtering functions by including positive and negative tap gain weight control. A novel transversal filter topology that enables operations at multi-Gbit/s application is designed and its behaviour is studied. The work demonstrates that pulse generation and correlation functions in the electrical domain can be accomplished using the designed versatile and reconfigurable transversal filters.

The newly developed filter is dubbed the triple-line transversal filter in reference to the filter cell design that distributes the parasitic capacitances of the devices along three artificial transmission lines; a common-input line and two output drain lines. It is demonstrated that this topology can be designed for chip rates exceeding 40 Gbit/s. The practicalities of the triple-line transversal filter were assessed using a 0.2 μm-gate length HEMTs and following the design rules of a commercially available MMIC foundry process.

A new tap gain weight control technique was specifically designed for the triple-line filter topology. A special bias level calculation technique was developed and used so that the filter can be modelled and operated with constant distributed characteristics. Wideband delay transmission lines were modelled and designed for filter implementations. An assessment of the MMIC transversal filter is provided via computer simulations in the time and frequency domains. A novel framework based on differential-mode scattering parameters was derived to investigate the various frequency responses of the triple-line structure. The filter satisfies the first Nyquist criterion and is suitable for decoders and encoders of CDMA systems.

Overall, this thesis addresses design issues that provide an insight into the practicality of receivers and transmitters for multi-Gbit/s CDMA systems.
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List of Symbols

OTDM  All-Optical TDM systems
AM-VSB  Amplitude modulation-vestigial band
AWG  Arrayed-waveguide grating
ATL  Artificial transmission line
BER  Bit error rate
CDM  Code division multiplexing
CTD  Charge Transfer Devices
DPSK  Differential phase shift keying
DA  Distributed amplifier
EDFA  Erbium-doped fibre amplifiers
FDM  Frequency division multiplexing
FSK  Frequency shift keying
GaAs  Gallium Arsenide
HBT  Heterojunction bipolar transistor
HEMT  High electron mobility transistor
IM-DD  Intensity modulation and direct detection
ISI  Intersymbolic interference
LED  Light emitting diode
LAN  Local Area Network
MAC  Media Access Control
MESFET  Metallic Schottky barrier FET
MAN  Metropolitan Area Network
MIC  Monolithic integrated circuit
MMIC  Monolithic microwave integrated circuit
MUI  Multiple user interference
NRZ  Non-return to Zero
OCDMA  Optical Code Division Multiplex Access
<table>
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<th>Acronym</th>
<th>Description</th>
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<tr>
<td>PSK</td>
<td>Phase shift keying</td>
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<td>P-HEMT</td>
<td>Pseudomorphic-HEMT</td>
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<tr>
<td>RTL</td>
<td>Real transmission lines</td>
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<tr>
<td>SIK</td>
<td>Sequence inversion keying</td>
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<td>SiGe</td>
<td>Silicon germanium</td>
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<td>SCM</td>
<td>Subcarrier multiplexing</td>
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<tr>
<td>SAW</td>
<td>Surface Acoustic Wave</td>
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<tr>
<td>SDH</td>
<td>Synchronous Digital Hierarchy</td>
</tr>
<tr>
<td>SONET</td>
<td>Synchronous Optical Networks</td>
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<tr>
<td>S-OCDMA</td>
<td>Synchronous-CDMA</td>
</tr>
<tr>
<td>TDM</td>
<td>Time division multiplexing</td>
</tr>
<tr>
<td>TWA</td>
<td>Travelling-wave amplifier</td>
</tr>
<tr>
<td>VSWR</td>
<td>Voltage standing wave ration</td>
</tr>
<tr>
<td>WDM</td>
<td>Wavelength division multiplexing</td>
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<tr>
<td>WAN</td>
<td>Wide-Area Networks</td>
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CHAPTER 1

Introduction

The very large bandwidth potential of optical fibres has enabled many innovations in the field of telecommunication networks. The advent of the single-mode optical fibre led to the development of networks with transmission capacities of several tens of Gbit/s and distances over several kilometres [1]. A further advance in the transmission over optical fibre was attained by the seamless integration of erbium-doped fibre amplifiers (EDFAs) into the optical links [2] that allows compensation for distance attenuation as well as losses of optical devices. These developments made it feasible to have networks in which the path between network nodes remains entirely optical. In addition to the transport functions, all-optical networks encompass different functions to multiplex data, mainly relying on optical devices so as to avoid critical bandwidth limitations associated with electrical devices. Due to its propagation and delay properties, researchers recognised that the optical fibre line is appropriate for optical signal-processing [3,4]. Parallel multiplexers based on different fibre optic delay lines have been implemented for all-optical multiple access networks [4,5]. Code sequence generators, convolvers and frequency filtering can be implemented using optical delay lines (fibres) as a matter of course. Initial work in the field of Optical Code Division Multiplex Access (OCDMA), was focused on optical delay-line signal processing to correlate and encode functions. Conversely, electrical components, such as microstrips, were dismissed for such implementation due to the high level of propagation losses and bandwidth limitations [3]. With the advances in high-speed active device technologies and the sub-millimetre control of transmission lines, signal processing using monolithic microwave integrated circuit (MMIC) is currently a viable alternative for high-speed optical communication systems. Accordingly, in this thesis newly developed MMIC circuits and design strategies are considered for achieving code generation and convolution in the electrical domain for systems applications working at several Gbit/s.
1.1 Optical and electrical processing for OCDMA systems

OCDMA was proposed in the mid of 1980s to provide optical fibre networks with high-speed connectivity, random asynchronous operation and network control simplification. These attractive features are the driving forces in the research of code division multiplexing for fibre networks. In code division multiplexing, user data is spread resulting in a higher bandwidth when compared to the information bandwidth. A practical multiplexing technique attempts to access the vast bandwidth of the optical fibre with the aim of including multiple-access capability. Some advantages include security as the probability of intercepting a message is reduced and lower network complexity.

Although significant advances have been made by using more efficient and practical multiplexing techniques, the hardware to develop encoders and decoders for high-speed optical networks is still in its initial stages. Indeed, recent investigations are being conducted into the design of lightwave integrated circuits for encoders and decoders of OCDMA systems. For instance, planar lightwave integrated circuits based on arrayed-waveguide grating (AWG) have been reported in [6]. Planar lightwave circuit design based on AWG has key advantages such as high reproducibility, lower power requirements and scalability. Based on their simulation results, the authors foresee the possibility of integration and high scalability (up to 200 frequency bins with 25 GHz channel spacing). Those attributes could out-perform other traditional methods for encoding reported in the literature and its practical implementation could deliver an advanced solution for high-speed OCDMA networks. Nonetheless, at the present, there are no reports on practical results of such encoders.

The initial method to encode data was time-domain encoding using non-coherent sources. Technologies considered for pulse generation and correlation in the electrical domain were Control Transfer Devices (CTDs) [7] and Surface Acoustic Wave (SAW) devices [8]. The first are capable of providing long delays (approximately 1s) and typically operate at frequencies of about 10 MHz [9]. Conversely, SAW devices can operate typically at frequencies as high as several hundred megahertz and delays of some microseconds [10]. Both devices offer the possibility of filter implementations with thousands of taps which can be switched to different values or incorporated in the transducer itself. The number of channels that can be created with this technology and with time-domain encoding can support some tens of users, lagging behind the number of channels that can be created with other multiplexing techniques.
Chapter 1 Introduction

The use of the optical fibre as a delay medium for signal-processing has been proposed for encoders / decoders for OCDMA systems [5] so as to avoid the bandwidth limitation of electrical devices. In this approach the encoding function is performed by the optical fibre and photodiodes connected at the end detect the envelope of the optical carrier. Optical fibres maintain a constant propagation loss in a wide range of frequencies. In a single-mode optical fibre, the bandwidth is mainly limited by dispersion, resulting in a bandwidth higher than 100 GHz-km [11]. Another favourable characteristic of the optical fibre delay lines for filter implementations is that the intervals between taps can be made with good precision because the group delay in fibre is proportional to the permittivity of the fibre material, requiring lengths of optical fibre within a millimetre to achieve a delay of some picoseconds [3]. It was in the mid 1980s that researchers aimed to develop improved encoders and decoders based on optical fibre technology [3]. However, in spite of performing the encoding and decoding functions in the optical domain, the electrical devices of the post detection stages of a receiver have to work at high speed rates (chip rates), which makes the implementation more difficult. This affirmation will be analysed in Chapter 3 in the context of OCDMA systems.

This thesis concentrates on the design, modelling and optimisation of high-speed electrical encoders and decoders based on distributed transversal filters and suitable for time-domain encoding applications. The utility of such electrical encoders and decoders rests in their ability to attain high-speed. High rate sequences with pulse widths falling in the picosecond region can be handled and filtered electrically. For high rate OCDMA systems, this is illustrated in the block diagram shown in Figure 1.1 below. Figure 1.2 shows a more detailed diagram of the transversal structure of the encoder and decoder transversal filters discussed in this thesis.

Accordingly, this thesis provides studies of the filters design and performance, identifying the main factors that limit the range of filtering functions and the potential application in lightwave systems.

![Figure 1.1 Transversal filters for high speed OCDMA encoding and decoding](image-url)
Chapter 1 Introduction

1.2 Thesis Organisation

Following this introductory chapter, multiplexing techniques for fibre networks are presented in
Chapter 2. The second chapter commences by describing the main multiplexing techniques for
fibre networks, with a context of technologies proposed for multiplexing data. A description of
the functional characteristics that entails such multiplexing techniques is outlined. In addition,
an account of the prospects to develop all-optical networks is provided. Chapter two ends with
an introduction of two hybrid multiplexing techniques that has been proposed to reap the
benefits of the multiplexing techniques.

Chapter 3 provides an overview of code division multiplexing technique for fibre networks. It
commences with a general description of OCDMA systems outlining some aspects related to the
current limitations to their deployment. Following this, some basic definitions and concepts of
coherent and non-coherent OCDMA schemes are provided. Non-coherent time domain
encoding, which is the scheme used for analysing distributed structures, is then introduced. A
more recently proposed encoding scheme for OCDMA systems such as amplitude spectral
encoding is introduced. In particular, this scheme is more suited to various encoding systems
since it imposes less stringent requirements on the optical processors based on the currently
available technology. The fundamentals of the spectral encoding technique will be described in
the context of non-coherent OCDMA systems. This chapter ends with a brief examination of
different technologies that have been utilised for encoding and decoding user data and reported
in the open literature.

Chapter 4 is concerned with distributed structures used for high-speed lightwave systems. It
commences with a detailed description of the distributed amplifier, active devices and
techniques which serves as a reference for the analysis and design of circuit designs. Some concepts such as the gain mechanism, synchronisation and design uniformity are highlighted. These basic concepts allow us to establish the foundations of distributed amplifier design. Other structures, which utilise different gain and delay partitioning techniques from those of the conventional distributed amplifier, are introduced. The first transversal filter concept based on distributed principles is described. Such precursory filter was designed as a monolithic integrated circuit (MIC) based on GaAs MESFET and provides the foundations of a new class of distributed circuits. The design criterion used for those tunable filters is based on the maintenance of pulse shape integrity. Following this, other subsystems for pulse-shaping applications designed as MMICs are described. The proposed schemes include tunable post-detection filters, shaping filters for high-speed optical soliton receivers and adaptive equalisers for lightwave systems, among others. This chapter ends with an account of the electrical characteristics of the topologies used in the above developments and the factors that limit their performance.

Chapter 5 deals with distributed structures with positive and negative tap gain weight control. The development of a novel transversal filter that maintains impedance matching over a wide range of frequencies is proposed. The filter design is aimed for the generation / reception of high-rate sequences. The chapter commences by introducing distributed structures with bipolar capacity to serve as a reference to analyse key drawbacks that have to be taken into account in the design of filter for receiver and transmitter implementations. Those limitations are overcome by the developed novel transversal filter, which is dubbed the **triple-line transversal filter**. The potentialities and limitations of the filter approach are investigated throughout this chapter from different aspects, such as electrical characteristics, tap gain weight control and MMIC implementation. Impedance loading uniformity and a novel tap gain weight control especially designed for such structure are investigated. For the first time, a description of the performance based on differential propagation of signals is suggested. Following this, a capacitively coupled active cell design is proposed. This technique is investigated using the electrical characteristics of a HEMT process, showing that such structure can be designed for very high-speed applications. At the end of the chapter small signal analysis is performed to gain an insight into the circuit elements and parasitics that limit the response.

Chapter 6 discusses the design of a distributed transversal filter for the generation / reception of high rate sequences using the triple-line topology. The chapter commences with an intuitive model that provides the guidelines for the design of high-rate transversal filter. Following, delay circuits suitable for on chip implementation are proposed and modelled. Such elements are connected between consecutive cells so as to set the filter tap delay and preserve pulse shape
Chapter 1 Introduction

integrity. A methodology is then developed for the triple-line transversal filter by which a wideband impedance matching between filter cell sections and delay lines can be achieved. Following this, an implementation of the transversal filter is carried out using a popular MMIC process. A MMIC 7-tap transversal filter was designed using HEMT and microstrip technology for 40 Gbit/s systems applications. The chapter ends with the description of a layout of the transversal filter.

Chapter 7 presents simulation based assessment of the performance of the MMIC filter. Time domain and frequency domain analyses are carried out by tuning the response to different code sequences. An appropriately modified methodology of differential scattering parameters is developed and then used for circuit frequency domain analysis. In addition, stability and noise analyses of the MMIC transversal filter are performed using appropriate HEMT models. The cascade of a pair of transversal filter, acting as a transmitter and receiver, is analysed. Reciprocal sequences, a concept associated with inverse filtering, are utilised for testing the transmitter-receiver pair. The analyses are aimed to assess the capacities of the filter structure to achieve pulse generation and correlation functions for multi-Gbit/s applications.

Finally, Chapter 8 concludes the thesis, summarising the main developments and identifying areas where further research may be appropriate.

The thesis includes three appendices. Appendix A has a short description of the MMIC process used. The Appendix also includes the small signal parameters extracted through an optimisation process based on the manufacturer’s device models and a small signal equivalent circuit. Appendix B details the mathematical derivation of the transfer function of the distributed amplifier-transversal filter structure when each of its lines is terminated by its image impedance. The formulae derived are generic and were used in the analysis of the designed filter given in Chapter 5. Appendix C describes the mixed mode s-parameters definitions and analytical derivation developed for the triple line filter and used in the treatment of the filter in Chapter 7.

1.3 Contributions to Research Field

The thesis reports novel circuit designs and design techniques that allow the generation and detection of different multi Gbit/s CDMA signals using a simple, yet versatile, distributed amplifier structure. A novel structure is developed to effect transversal filtering functions that can be modified by changing the bias of the active devices employed. The main contributions of this thesis are in the design of the new circuits and in the development of systematic
techniques to study the circuits' behaviour in the time and frequency domains. These contributions may be listed as follows:

- Different transversal filter structures were studied, in particular those that allow extending the range of filtering functions by including positive and negative tap gain weight control. The design allows versatility as the filters can be tuned to a predetermined filtering function by adjusting external bias voltages.

- A novel distributed transversal filter topology, which is named the triple-line transversal filter, is developed for transmitter and receiver implementations. Due to its high frequency capabilities, the filter is appropriate for shaping / generation of multi-Gbit/s sequences.

- A new approach for the design of distributed amplifier delay lines is proposed. A topology based on input line coupling capacitors was developed to achieve wideband impedance matching between filter cell sections and delay lines. The appropriate conditions for capacitor value choice were derived and such design was verified analytically and by simulation.

- A new tap gain weight control technique was specifically designed for the triple-line filter topology. The utilisation of such technique allows maintaining uniformity of the artificial transmission lines for continuous gain control.

- A MMIC 7-tap transversal filter was designed using HEMT and microstrip technology for 40 Gbit/s systems applications. The MMIC transversal filter is based on the triple-line topology. Layout of the filter is presented and simulation results indicate the efficacy of the design techniques proposed in the thesis. The design proves to be stable at the design operation conditions.

- A generic transfer function for the distributed amplifier, with ATLs matched by their image impedances, is derived. This is done by extending the work of Chen in [74,69] so that it can be applied accurately to structures where the counterpropagating waves (on the ATLs) are taken into account.

- An intuitive model of the distributed transversal filter that allows gaining an insight into the filtering of short pulses is established. The description of the model, which is in full agreement with simulation results, sustains that intrinsic limitations of the transversal filter make its optimisation in the time and frequency domain two different processes.
Chapter 1 Introduction

• An incremental model of the filter cell was derived, from which the responses of the filter can be analysed by assuming differential-mode and common-mode wave propagation on artificial transmission lines. A set of differential scattering parameters were derived so as to analyse the responses of the triple-line structure. This work is based on the extension of the 1995 treatment of Bockelman and Eisenstadt [105].

• Time domain simulations were carried out for theoretical and practical filter structures so as to analyse the effect of different components in the response. It is proved that the MMIC transversal filter design can maintain pulse shape integrity and low inter pulse interference. The filter is used for reconfigurable transmitters and receivers.

• The cascading of a transversal filter pair, acting as a transmitter and receiver, was analysed via reciprocal sequences. The output of the filter pair in cascade, which is effectively the convolution between two waveforms, maintains the in-phase amplitude of the convolution function for both periodic and aperiodic functions. It is proved that the filter proposal satisfies the first Nyquist criterion.

The contributions made during the course of this research have led to the following publications:


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6. Aguilar-Torrentera, J. and Darwazeh, I, 'High-speed Electrical Encoder/Decoder for Fibre Optic CDMA systems', International Symposium on Telecommunications, IST 2003 Proc., Organised by the University of Iran, Isfahan, Iran, August 2003, pp 484-8


CHAPTER 2

Introduction to Multiplexing Methods
for Fibre Networks

The transmission of information over guided media has had outstanding impact on telecommunication networks, first with low-speed transmission paths based entirely on copper cable and subsequently by upgrading transmission links using optical fibre. The low attenuation of optical fibre, vast bandwidth and the feasibility of amplification without the need to convert data into the electrical domain have enabled new network concepts to emerge such as the all-optical networks [17] in which the transmission remains entirely in the optical domain. Such optical fibre communication systems can meet the requirements of data transmission of many nodes' transceivers spanning over distances of several kilometres.

The advances in optical networking are associated with the capacity to exploit the available bandwidth of optical fibres. Multiplexing techniques can create an aggregate network capacity that satisfies the transmission requirements of many transceivers interconnected in the network and each working at speeds of several hundreds of megahertz (speed of electronics). Optical processing becomes the natural method for multiplexing signals of such networks. For the time being, there are some prevalent limitations of the optical processing that pose difficulties to accomplish certain functions. Despite the present technological barriers, in the past few years, there have been substantial advances in the improvement of optical devices and systems that have enabled new developments in the multiplexing techniques and networks [28]. This chapter introduces the principal multiplexing techniques used for optical networks and serves as a support for Chapter 3 which is focused primarily on the code division multiplexing (CDM) technique for optical networks.
Chapter 2 Introduction to Multiplexing Methods for Fibre Networks

2.1 Wavelength division multiplexing

Wavelength division multiplexing (WDM) technique enables the transmission of the information by accessing a channel defined in the wavelength domain, in which the wavelength of an optical carrier is modulated with the user data. Each optical carrier occupies a distinct wavelength space in the available optical spectrum. The operation of an optical fibre can be either in the window at 1.3 \( \mu \text{m} \) with a bandwidth of 18 THz or in the window at 1.5 \( \mu \text{m} \) with a bandwidth of 12.5 THz. Given the vast bandwidth of the optical fibre, each waveband that supports a network channel can exceed the transmission speeds of the terminals. Rather, the channel speed is limited by the capacity of accessing the optical bandwidth using optical devices. In principle, several hundreds of transmitter nodes can access the available optical spectrum.

WDM networks have adopted different topologies, which can be divided into WDM link, broadcast and select network and wavelength routing network [15]. Figure 2.1 shows the common topologies of WDM optical networks.

![Figure 2.1 Three forms of WDM: (a) WDM link, (b) broadcast and select network (c) wavelength routing network](image)

The WDM link scheme basically describes point-to-point links and does not represent properly a network. In long-haul links, some channels can be amplified in the optical path between the transmitter and receiver by the same erbium-doped fibre amplifier (EDFA). The wavelength multiplexer consists of laser diodes with wavelengths tuned to the central frequency of the optical filters at the demultiplexer followed by photodetectors to recover the user data.

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1. The attenuation at this window is in the order of 0.2 dB/km in a dispersion-shifted monomode optical fibre [11]

2. In WDM systems and in other multi-access optical networks, intensity modulation direct detection (IMDD) schemes are often implemented by virtue that the additional gain of 3 or 6 dB obtained in coherent detection are frequently consumed by losses of a more complex receiver implementations [27].
Conversely, broadcast-and-select network are most useful for LANs and MANs, in which multiple nodes are connected to the network. All channels operate at distinct wavelengths and are combined in a passive network that distributes the signals to all receiver nodes. A frequently used network topology is the star one, which is based on passive couplers as depicted in Figure 2.2. The star network topology is preferred over others configurations such as bus or ring because of its lower loss (which increases logarithmically with the number of taps) and more uniform power distribution [15].

![Figure 2.2 (a) Passive 8×8 star coupler formed by interconnecting 3-dB coupler](image)

(b) N×N star topology for multi-access optical-fibre network (after Mestdagh, D., [15])

In broadcast-and-select networks, transmitters and receivers are equipped with tunable filters and lasers to provide different network connections. In such application, the Media Access Control (MAC) sets the frequency of the tuned receiver or transmitter before the transmission of packets. Such tunable transceivers require time for stabilisation before the transmission takes place and that depends on the specific technology used. When the receiver nodes are fixed to distinct frequencies, the connection is established by tuning the transmitter to receiver’s frequency. Another possibility in this network is the tuning of the receiver while keeping the frequencies of the transmitters unchanged. Although both schemes enable data to flow within the network, there are some important considerations to be taken into account in the choice of the topology such as the tuning range of the optical filters, tuning speed (submicrosecond tuning times for packet-switched applications [15]) and number of resolvable channels.

The third category is the wavelength routing network. This network topology enables several thousands of users to transmit their information making it appropriate for Wide-Area Networks (WANs). A connection is established by determining the wavelength of the

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1 Due to its suitable characteristics, the star network will be considered for other multiplexing techniques such as in the optical code division (CDM).
transmitted signal and the node through which the signal is routed. This is accomplished by exchanging control of wavelengths at different frequencies [15,16].

In conventional WDM systems, the distance between channels can be of the order of several nanometres in wavelength, much in excess of the bit rate of channels. With advances of optical tunable filters and lasers as well as in modulation techniques, the space between different wavelengths channels can be reduced to make more efficient the spectrum usage. In particular, the networks that reduce the distance between wavelengths lower or equal to 1 nm, or equivalently to 130 GHz at 1.5 μm centre frequency [14], are termed dense-WDM networks [15].

The need for improving the spectrum usage is not because of a scarceness of the optical fibre spectrum, but because most widely-tunable semiconductor diode lasers tune only over a 10 nm range [30], a reduced fraction of the available optical spectrum of 200 nm. With the possibility of transmitting many channels in WDM networks, crosstalk is the most significant impairment of the network and constitutes the main limitation in reducing the frequency spacing between channels. In this regard, it has been established in [29] that for different kinds of modulations, coherent or non-coherent, a channel spacing of approximately six times the bit rate allows minimising channel cross talk. For terminals running at several mega-Hertz, the channel spacing that ensures low crosstalk is lower than 1 nm. Therefore, the number of channels that can be created depends on the effectiveness in rejecting interference from unwanted adjacent channels. Reported experiments in computer network [30] show that Fabry-Perot filter can satisfy the selectivity in dense-WDM network supporting several thousands of terminals.

At the present, the WDM technique is most prominent choice in the construction of networks and the front-end research of high-speed optical networks.

2.2 Time Division Multiplexing

Time division multiplexing (TDM) technique attempts to make use of the vast bandwidth of optical fibre by combining the information of multiple terminals in a high-speed optical stream. The transmission of information of many terminals using a single optical link allows reducing costs as the implementation of the network requires only a high-speed transmitter-receiver pair. This scheme has been adopted by transmitting pulses onto the optical fibre at rates of gigahertz while accomplishing multiplexing functions in the electrical domain [14,16].

In time division multiplexing each terminal can transmit data in packets or bits in an assigned time-slot. The assignment of a slot can be fixed, in which each terminal node transmits
its information in a pre-assigned TDM frame, or terminals are allowed to transmit the whole data packet when the access is granted (packet-based TDM). Users transmitting information are sequentially polled by the multiplexer allowing making use of the optical channel during a TDM frame. In a packet-based TDM the address of the intentional receiver can be included in the header so that the redistribution of the packets can be achieved at the receiver (downstream) in the same frame interval. By this means there are no contention problems at the output of the multiplexer. In accordance with the multiplexing schemes, the synchronisation of the transmitter needs to be done prior to the transmission of the packet or bit. At high-speed operation, packet-multiplexing is preferred over bit multiplexing as a dead time (guard time) is included in the frame, which relaxes the need of complex and expensive circuits that synchronise all nodes at the bit rate [16].

Another feature of TDM is the possibility of allocating TDMA / TDM transmission indistinctly in the same frame [15]. In the dynamic allocation of frames, the stream of packets in higher digital hierarchies can be sent to the receiver end with the same header address so that users can transmit or receive information at higher bit rates. By disregarding the headers and additional control bits in each slot, data is transmitted at speeds approximately equal to the number of terminals connected to the network (N) times the data rate of each node (BRch). This transmission efficiency requires that the information of node transmitters be stored only during a frame time at the time the multiplexer polls a specific transmitter.

The technology for commercially available TDM systems is currently located at 2.488 Gbit/ [12]. Recent advances in circuit technologies based on III-V compound semiconductor and silicon transistors shows the feasibility of circuit for TDM systems at 10 Gbit/s and beyond [13]. Current technologies based on SiGe BJT or GaAs FETs enable analog and digital circuit applications at higher bandwidths with improved performance. The eventual upgrading of TDM systems depends on the maturity of a specific technology that shows a sustainable improvement over others. Nonetheless, this scene is still incomplete given the current innovations in the field of high-speed circuits [12].

A schematic of a 10 Gbit/s TDM system based on SiGe bipolar ICs [12] is depicted in Figure 2.3. The design meets the specifications of SDH-64 TDM link and the high-speed data transmission is assigned to a specified wavelength. On the transmitter side, a multiplexor combines 4-low speed channels to form a 10 Gbit/s data stream. Prior to launch into the optical fibre, the modulated optical carrier is modulated with the data. An external modulator can be constructed as a Mach-Zehnder interferometer [18] or an electroabsorption type modulator [19]. Electroabsorption modulators are often preferred when the driver design is limited by the output power since lower swing voltages are needed when compared with those of drivers for Mach-Zehnder modulators. On the receiver side, the small signal generated by the photodetector is amplified by a low noise amplifier and a main amplifier with automatic gain control capability.
Chapter 2 Introduction to Multiplexing Methods for Fibre Networks

The optical wavelength demultiplexer separates the data streams. For such aim a retiming circuit and clock extraction are required. Such functions can be achieved with high-speed electronics either in SiGe bipolar transistors or InP heterojunction technologies for operations at several tens of GHz [13].

![Diagram of a 10 Gbit/s electrical-TMD link](image)

**Figure 2.3 Scheme of a 10 Gbit/s electrical-TMD link [12]**

For the time being, the TDM technique encounters some difficulties to allow higher speed operation, not only because of the lack of commercially available integrated sub-systems [12] but also because data transmission of TDM links at higher speeds are prone to optical fibre dispersion [14]. Multiplexing is a crucial function that electronics cannot accomplish at speeds of hundreds of gigahertz. For such reason multiplexing and demultiplexing functions for high-speed networks are currently envisaged in the optical domain. Initial attempts to construct ultrafast networks considered using optical delay lines and electro-optical switches. The functions based on such technologies were accomplished assuming each transmitter node is assigned to a fixed time slot, which restricts the configurations of other slot assignments for higher speeds.

Emerging concepts such as photonic packet switching (all Optical TDM systems, OTDM) at multiplexing speeds at 100 GHz have been proposed, however; as a long term solution for high-speed networks given the present limitations of optical processing [17]. The system requirements cannot be satisfied with present technologies. For example, buffering functions are required to process the header of the packet at times as large as a TDM frame. The optical technology is still unable to provide logic and memory functions required for such multiplexing functions. At the present, buffering and header recognition are functions best accomplished in the electrical domain [17] while optical functions are efficient in transmission of high-speed pulses. Given the anticipated difficulties in the implementation of OTDM

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*At the present, the rates for TDM systems are 9.95 Gbit/s OC-192 SONET or SDH-64 links [12]*

*Similar optical processors based on optical fibres were proposed for multiplexing OCDMA signals [4]*
networks, other schemes that allow increasing the speed operation of optical networks have to be considered. For instance, hybrid systems, where WDM channels with specific wavelengths of separation serving as carriers of TDM links, can provide a transport capacity of the order of several Tbit/s [16].

2.3 Subcarrier Multiplexing

In the subcarrier multiplexing (SCM) technique a number of channels are created by electrically modulating dedicated subcarriers with the data. The subcarriers are subsequently allocated to an optical spectrum by modulating the intensity of an optical carrier. That entails the use of microwave technology that effects the frequency multiplexing, while optical functions are only utilised for transmission over optical fibre, optical generation and photodetection. Since the nature of the information can be analog, SCM networks is a compatible with schemes of video transmission of CATV signals using subcarriers modulated in amplitude and vestigial-sidebands (AM-VSB) [20]. Other schemes of modulation, such as FSK and PSK have also been proposed for subcarrier transmission [21]. Those schemes show the flexibility of the SCM technique to transmit analog or digital data in a variety of modulation formats.

There are basically two schemes in SCM networks. When a single modulated carrier drives the optical transmitter, the channel assignment is termed single-channel SCM. On the other hand, when a number of frequently-separated RF subcarriers are combined to modulate a common optical source at a single wavelength, the channel assignment is termed multi-channel SCM. The last scheme is preferred due to the advantage of transmitting simultaneously multiple channels on a single laser thereby improving the spectrum usage.

Figure 2.4 shows a multi-channel SCM system. The fibre optic transport consists essentially of a star coupler used to interconnect permanently all transmitters and receivers. SCM multiaccess networks provide independence among channels since the nature of the user information can be analog or digital. As shown in Figure 2.4, there is neither a need for synchronisation between each channel nor a high-speed master network clock [21]. Prior to the transmission of information, the local oscillator is stabilised at the selected frequency of the receiver. At the receiver end, optical subcarriers are down converted into the RF domain and filtered at each receiver node. Electrical filters tuned to the receiver assigned frequency possess the required selectivity to ensure minimal crosstalk among channels.

Given the vast bandwidth of optical fibre (terahertz) and the megahertz bandwidth of the transmitted data, it is apparent that the practical limitation in the generation of SCM channels lies in the subcarrier generation using microwave techniques and the laser modulation bandwidth. In spite of the fact that laser modulation bandwidths can be on the order of
gigahertz, SCM system can create a large number of channels given the capacity of tightening the space between transmitted subcarriers (comparable with the bandwidth of the baseband signals). The number of SCM channels, $N$ assuming optical double side-band amplitude modulation can be approximated by [15,16]:

$$N \equiv \frac{B_e}{2B_{ch}}$$

where $B_e$ is the electrical bandwidth of the system and $B_{ch}$ is the information bandwidth of each channel. The electrical bandwidth depends on the modulation bandwidth of the optical modulator. With the current technology a laser can transmit 20 - 50 channels depending on the modulation format and the data streams rates.

**Figure 2.4 Subcarrier multiplexing system [15]**

There are two practical limitations that thwart the capacity to create more subcarriers and therefore the number of channels. First, the modulation bandwidth of electro-optical modulators cannot exceed some gigahertz, otherwise the analog signal is greatly distorted and secondly, the peak amplitude of the composed signal can be of such a level that the non-linear behaviour of the laser can induce a significant level of intermodulation distortion. For such reason, the transmitter needs to run at lower power levels with the consequent inability to compensate for splitting power losses associated with node transmitters added in the network.

High-speed SCM networks have been recently reported in [22] showing the feasibility of the technique for optical networks working at several Gbit/s. Experimental results in [22] show that SCM technique is less impaired by chromatic fibre dispersion than an equivalent
Chapter 2 Introduction to Multiplexing Methods for Fibre Networks

TDM link as the dispersion of the channel depends on the speed per channel. The operation of a 4-channel SCM at data speeds of 2.5 Gbit/s is less deteriorated than that of a 10 Gbit/s TDM link [22]. Two factors contribute to the lower degradation as linked to fibre dispersion. First, the low spectral distance between subcarriers transmitting data streams at lower speed per channel and secondly, the utilisation of optical single band modulation that reduces the transmission bandwidth. In the reported experiment in [22], the sensitivity of the TDM receiver is severely degraded as a function of the distance while the SCM is less sensitive to this variable. Those results show that SCM system has a good potential as a multiplexing technique for long-haul high-speed optical networks.

2.4 Hybrid multiplexing proposals: WDM-SCM and FDM-CDM

In a SCM system, the optical losses and the sensitivity of the receiver determine ultimately the number of SCM channels that can be used in the optical network. A method to increase the number of channels may be based on the transmission of data on different SCM channels using different lasers at different wavelengths, resulting in a WDM-SCM system. At the receiver end, optical filters demultiplex wavelengths of transmitters. Each transceiver in the network was assigned a unique wavelength of a specific cluster and frequency of the subcarrier. The subcarrier transmission could be of different terminals using conventional techniques of WDM described above. In a experimental system based on this hybrid scheme [16] four separated wavelength lasers transmit 50 AM-modulated SCM channels given a total capacity of distribution of 200 channels. The transmission of optical carriers at different wavelengths each modulated with different subcarriers can make more efficient use of the transport capacity of the optical fibre than in the case of pure SCM networks.

A more elaborate hybrid system has been proposed by which the transmission power of the laser can be increased hence the number of active users in the optical network. In this network, the intermodulation noise associated with the laser modulation characteristic can be suppressed or diminished to such a level that the interference associated to additional users in the network does not impair significantly the reception. The hybrid system proposed for optical networks is a frequency division multiplexing-code division multiplex (FDM-CDM) system [8]. In this scheme, the reception of coded signals is more robust against non-linear interference at the cost of spreading data over a large (electrical) bandwidth.

This scheme shares similar characteristics with those of WDM-SCM in terms of holding the transport functions in the optical domain while the subcarrier generation is effected in the electrical domain. In this scheme, the transmission on a stable subcarrier allows the transmission
Chapter 2 Introduction to Multiplexing Methods for Fibre Networks

of user information asynchronously. Here, different optical sources are modulated by composite signals constituting the subcarriers modulated with encoded data. Figure 2.5 shows a cluster comprising $M$ users in a FDM-CDM network. Each receiver has a unique combination of subcarrier frequency and code, thus each transmitter has to adjust the frequency of its local oscillator to the frequency $(f_1, f_2, \ldots, f_M)$ and set the code sequence that corresponds to the receiver address. The number of channels that can be created in the network is equal to the number of subcarriers per cluster, $M$ times the number of the pseudo-orthogonal codes, $N$.

At the receiver end, each receiver node has a SAW matched filter which has a band-pass characteristic close to a “sinc-function”. Every receiver is uniquely defined by a pseudo-orthogonal code and a centre frequency. That frequency can be directly adjusted in the SAW filter as shown in Figure 2.5. The delay and bandpass characteristics of SAW devices enable programmable transversal filters with lengths as large as hundreds of taps and a centre frequency from tens to hundreds of megahertz [10]. When the SAW filter is used as a convolver, its quality factor is adjusted to the data rate and nulls on each side of the main lobe by one data rate away from the central frequency $f_k$. By this means, the nulls of the filter can be set at the adjacent frequency subcarriers, $f_{k-1}$ and $f_{k+1}$ so as to increase the frequency selectivity of the receiver and efficiency of the spectrum usage. When the sequence of the receiver matches with the transmitter sequence, the SAW filter responds with an autocorrelation peak. If the sequence does not match, the response is a cross-correlation side lobes.

![Figure 2.5 A cluster of a FDM-CDM system with SAW convolvers tuned to the transmitted carriers at frequencies $(f_1, f_2, \ldots, f_M)$](image)

The practical implementation of the FDM-CDM system uses DPSK modulation to eliminate the need for synchronous carrier recovery at the receiver. The DPSK demodulator
comprises a delay line, with time delay equal to the bit rate, a multiplier and low pass filter [8,16]. The signal at the input of matched filters corresponds effectively to composite signals of all the transmitters in the network with the encoded data. A practical implementation of such system is described in [16].

At high peak signal modulation, the laser generates harmonics and intermodulation products which depend on the instantaneous amplitude of the composite signal. The analysis of the interference associated with the non-linear modulation characteristics of the laser takes into account the intermodulation products that results in cochannel interference at the input of the receiver (with the SAW filter programmed with the matched sequence). Such interference can be modelled as waveforms that hold the orthogonal characteristics as in the case of the multiple user interference (MUI), but weighted by corresponding coefficients associated with the harmonic and intermodulation products of the laser [8]. From this consideration, the effect of non-linear interference can be reduced at the output of the filter in relation to processing gain of the SAW filter. Hence, there is no need to "back-off" the laser power to reduce the level of interference associated with the non-linear behaviour of the laser as in the case of SCM techniques.

SAW-based transversal filters can be used for spread spectrum applications at chip rates on the order of several hundreds of megachip/s. Another electrical device that has been used for CDMA applications is Charge-Transfer Devices (CTDs), which exhibit speed operations of some megachip/s [7] and can be implemented with some hundreds of stages. In the following chapter, an account of the methods of modulation and devices implemented using such technologies is provided. To the best of the author’s knowledge, those developments are the only two proposals in which electrical devices are used to perform the correlation functions for optical CDMA fibre networks.

In this thesis, newly developed circuits designed for handling high-rate sequences are proposed for multi-Gbit/s system applications [23]. Previous developments based on this methodology are adaptive equalisers for high-speed applications [25] and distributed amplifiers with embedded pulse shape capability [24], aimed at different applications in high-speed systems. Inspired by such developments, this thesis explores distributed circuits as an alternative for high-speed OCDMA system implementations [23,26]. The intrinsic limitations of such filters will be investigated and their feasibility for OCDMA fibre networks will be discussed in final chapters of this thesis.
Summary

This chapter introduces the principal multiplexing techniques proposed for optical networks. Wavelength division multiplexing technique was first described. Such technique can be used for different networks such as local and metropolitan area networks. The use of such networks to LANs facilitates the deployment of networks with relative flexibility and simplicity. Device inadequacies such as low tolerances of the optical devices and frequency limit the node ability to select the desired optical channel. Dense-WDM networks have been demonstrated, in which the number of terminals working at several hundreds of Mbit/s is limited mainly by channel crosstalk.

Another multiplexing technique proposed for optical networks is TDM. This technique is appropriate for small networks basically due to the limited operating speed of optoelectronic transceivers. In this technique, in order to make use of the available optical bandwidth, pulses with widths falling in the femto-second region are required. Although such technology is currently available, its full use in TDM systems is still limited due to the high cost incurred in its deployment. For the time being, TDM has been adopted for the transmission of short pulses onto optical fibre while accomplishing multiplexing functions in the electrical domain. Several TDM links can be created in the optical network by modulating different optical carriers with TDM streams. TDM networks present difficulties for local area networks by virtue of the need for strict time management among terminals which reduces the flexibility to access to the channel.

Subcarrier multiplexing systems were also described. Such networks are flexible as various modulation schemes and information signals can be multiplexed and transmitted on the same optical carrier. The main problems of the application of SCM system are related to linearity and noise; thereby the inability to increase the power transmission to compensate for power splitting as high number of channels are added in the network. Recent efforts have been focussed so as to increase the speed of operation to some Gbit/s. Due to its inherent advantages; SCM can rival other multiplexing schemes such as TDM. They are also able to provide a useful contribution to spectrally efficient multiplexing systems. This advantage has been explored by combining this technique with other multiplexing techniques such as WDM and CDM.

The following chapter is focussed on the optical code division multiplexing technique. Some authors term CDM as the spread-spectrum version of TDM [15] due to its use of a high-speed stream of pulses spread by a pseudonoise sequence.
CHAPTER 3

Optical Code Division Multiple Access
for Fibre Networks

The Optical Code Division Multiple Access (OCDMA) technique has been proposed to improve the throughput and functionality of optical fibre networks. Since its first proposal in the mid-1980s [15,31] there have been significant advances in OCDMA components, systems and networks. Of particular interest has been the deployment of optical networks capable of working at speeds of multi-Gbit/s. High-speed OCDMA systems can exploit the vast bandwidth of optical fibre, relying for such aim on efficient ultra-fast optical components and technologies for the encoding and decoding functions. A distinct feature of the OCDMA systems is the variety of schemes proposed for optical networks which results from the methods of modulation and detection used in fibre optic communication systems.

This chapter presents an overview of OCDMA systems. It commences with an introduction to OCDMA systems highlighting their functional characteristics. Various coherent and non-coherent OCDMA schemes as well as the technological aspects related to their deployment will be described. The requirements of high-speed systems in the context of OCDMA applications such as the versatility and functionality of the receivers and transmitters will be highlighted. The physical limitation of the present technologies play an important role in the development of efficient high-speed OCDMA systems and those will be discussed alongside various CDMA systems. Non-coherent time domain encoding schemes, first proposed for OCDMA networks, will be described in some detail. More recently, OCDMA systems based on amplitude spectral encoding have been investigated. Practical implementations of spectral encoding systems impose less stringent requirements on the optical processors based on the currently available technology. The fundamentals of the spectral encoding technique will be
Chapter 3 Optical Code Division Multiple Access for Fibre Networks

described in the context of non-coherent OCDMA systems. This chapter ends with a brief examination of different technologies that have been utilised for encoding and decoding user data, from electrical devices employed in hybrid CDMA/FDMA systems to high-speed devices used in all-optical OCDMA networks.

3.1. General aspects of OCDMA networks

An important advantage of the OCDMA network lies in the possibility of transmitting the information asynchronously which facilitates sharing the optical channel among subscribers. This allows the information to be transmitted without delay as external timing or centralised control are not required. The network is thus greatly simplified provided that the transmitter and receiver include the necessary processing functions to encode and decode user data [15].

Synchronous-OCDMA (S-OCDMA) networks have also been proposed when users transmit the information in a scheduled pattern* [32,33]. Such networks require external synchronisation signals to attain synchronous transmission. Despite the inherent complexities in the distribution of synchronising signals to terminals, synchronisation increases the efficiency of the channel since for a given bit error rate (BER) more users can access the channel than those in the asynchronous scheme. Synchronous systems have the advantage that timing signals align data broadcast in the network and consequently the number of code sequences with good correlation properties is increased [32]. Although the requirements of time management are similar to those in TDMA networks, the utilisation of the OCDMA technique provides, as an intrinsic advantage, a degree of rejection of the interference associated with laser modulation characteristics and other non-ideal behaviour of optical devices† [33]. Synchronous-OCDMA schemes are of particular interest in certain applications and most of this chapter deals with asynchronous-class OCDMA systems.

Non-coherent OCDMA systems make use of intensity modulation and direct detection (IM-DD) schemes. User information is impressed onto a code sequence comprising unipolar pulses which in turn are directly detected (non-coherently) at the receiver. In the multi-user environment, users can occupy the same spectrum at the same time by transmitting the information using orthogonal spreading codes. An important consideration in OCDMA is the number of channels that can be created in the network. In the OCDMA technique, a large number of channels imply the utilisation of long sequences that occupy large bandwidth when compared to the information bandwidth. The generation of ultra-short pulses, in the femtosecond regime, has been proposed for high-speed systems, but the reception of CDMA signals is

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* This is the case of digitised video systems where transmission of output packets to the same destination is done in at regular intervals [33]
† Optical-CDMA systems employ very often high-power lasers to compensate for losses associated with both the passive optical network and the optical devices used in the receivers [8,16]
Chapter 3 Optical Code Division Multiple Access for Fibre Networks

severely degraded due to fibre dispersion and the non-ideal behaviour of optical devices [15,51]. Fibre dispersion is recognised as an important hindrance for developing efficient OCDMA networks [14,34,35] as the modulation bandwidth is much higher than the bandwidth of the user information.

OCDMA systems based on temporal encoding using conventional codes (such as Gold or Kasami) have been demonstrated practically at speeds of tens of Mbit/s [7,8]. At higher speeds, the performance of electrical-based correlators is limited either by the inability to construct encoding and decoding hardware with a large number of stages (effectively equal to the code length that can be used) or by the inherent bandwidth limitation of the devices. In order to circumvent the limitations of time domain encoding, amplitude spectral encoding of non-coherent optical pulses has been proposed [36]. In this approach, optical pulses are encoded in frequency patterns according to orthogonal codes. This method of coding is favoured since multiple user interference (MUI) can be completely eliminated; hence true orthogonal channels can be created. In addition, practical application of spectral encoding is feasible with “compact processors” and inexpensive non-coherent sources [36]. In practical implementations, the number of channels of such network is limited by the spectral resolution of the optical devices and the variation of the power spectrum of broadband sources [37].

3.2 Time-domain OCDMA

The majority of OCDMA schemes reported in the open literature make use of optical processing to perform the encoding and decoding functions. The usage of electrical devices [7,8] and electronic processing [38] to effect correlation functions has been proposed for OCDMA systems. In spite of the processing capacities of such devices, their speed is still much lower than the data rate at which the optical networks can work. Since the initial proposals of OCDMA systems [8,31], the encoding and decoding functions were shifted into the optical domain. The optical processing, however, still presents significant limitations in the capacity of processing OCDMA signals.

A fibre-optic CDMA system is depicted in Figure 3.1, where the nodes in the network are connected through a passive N×N star coupler [15]. The same power level of optical signals is detected in all terminals of the network. Transmission of user data includes the choosing of a unique code sequence that represents the address of the intended node receiver (also termed signature sequence). The encoding function is reconfigured according to the sequence. The optical source generates a pulse for each bit ‘1’ which is encoded in a stream of pulses and for each bit ‘0’ of the source information, the bit is encoded as an all-zero sequence. The encoded
signal is directly coupled into the input single-mode fibre and broadcast to all receivers as shown in Figure 3.1.

![Diagram of Fibre-Optic CDMA network](image)

**Figure 3.1 Fibre-Optic CDMA network**

At the intended receiver, the data is decoded by correlation. A fixed code is stored in a matched filter implemented either in the form of tapped delay lines [31] or actively, by storing the code in sequence generator and multiplying each code element with the incoming signal at the chip rate [39]. The intended receiver must be able to extract its address sequence and discriminate (reject) other transmitter addressees in the presence of multiple-user interference. The amplitude peak of the correlation is extracted by threshold devices and integrators. The model of a passive optical matched filter implementation is depicted in Figure 3.2 [31].

The correlation function is achieved optically. An optical filter capable of responding to amplitude modulated pulses (for instance On-Off keying modulation format) processes pulse sequences using tapped optical delay lines. In the non-coherent system, pulses at the input of the receiver are non-negative and must be narrow enough so as to not overlap and spread beyond the bit interval.

---

1 When the tuning speed of the encoder is not fast enough to change the corresponding address, a switched-encoding tapped line with all the receiver addresses could be implemented.
The overall response of a receiver, implemented as an $N$-tap transversal filter has the equation:

$$h_i(t) = \sum_{k=0}^{N-1} a_{k+1} \cdot P_{T_c}(t - kT_c)$$  \hspace{1cm} (3.1)$$

with $a_k$ corresponds to the coefficients of the signature sequence of the receiver and $P_{T_c}(t)$ is a rectangular function of width $T_c$ which starts at $t = 0$. In the optoelectronic implementation, the photodiode and electrical circuitry of the receiver needs to have integration times of the order on the chip period, $T_c$; otherwise the integration of interference at the output of the correlator could induce errors in the detection. Consequently, the electrical devices in the receiver must be able to operate at chip rates and bandwidth equal to $1/T_c$. Given the limited bandwidth of electrical devices, a passive implementation can be used for relatively low-speed systems [40]. In order to improve the operation speed in the optoelectronic implementation, i.e. negating the need of electrical devices working at the chip rate, an active correlator implementation can be considered [39,40]. Figure 3.3 shows such structure which can be used in a non-coherent (synchronous) OCDMA system [40].

The active correlator decodes user data by detecting the envelope of the encoded information. The structure performs the same functional operation as in the passive correlator receiver; however, its implementation needs electrical integrators and threshold electronics.
working at **bit-rate** speed, which reduces the bandwidth requirements of electronic devices. A crucial element of such receiver is the active multiplier; it can be implemented with acousto-optic modulators capable of working at several hundreds of MHz. Such structure thus can be used for high-speed systems [40].

As pointed out earlier, OCDMA depends on the methods of modulation and detection in which signal formats maintain their correlation properties. For instance, in the OCDMA system described above, pulses at the input of the receiver are detected non-coherently; as a consequence the correlator can only process unipolar pulses. Since initial experiences to apply the OCDMA technique to fibre networks, it was recognised that the processing of unipolar pulses poses an important problem in the efficient detection of OCDMA signals. For instance, it has been proved in [7] that by using unipolar version of sequences with good correlation properties (such as Gold sequences), the multiple user interference (MUI) at the output of the correlator becomes higher than in the case of processing bipolar pulses. Different schemes for time domain encoding of non-coherent systems were proposed to lessen the effects associated with the inability to transmit bipolar signals in the optical channel.

### 3.2.1 Conventional OCDMA receiver model

The response of the non-coherent optical receiver (taking the receiver 1 described by Equation 3.1 as reference) to a composite input signal, \( r(t) \) and a filter integration time, \( T \) is given by:

\[
O_i(T) = \int_{0}^{T} r(t) \cdot h_i(t) \, dt
\]  

(3.2)

where \( h_i(t) \) is the impulse response of the receiver. Equation 3.2 is the convolution integral assuming, without loss of generality, zero transmission time delay (\( \tau_1 = 0 \, \text{s} \)).

At the input of the correlator receiver, the composite signal \( r(t) \) of all the \( N \) users each transmitting their information on a base band signal, \( p_n(t) \) and time delay, \( \tau_n (< T) \) is given by:

\[
r(t) = \sum_{n=1}^{N} b_n(t - \tau_n) \cdot p_n(t - \tau_n)
\]  

(3.3)

with \( b_n(t) \) is the information-bearing signal of the \( n \)th user. For a continuos transmission, \( b_n(t) \) is given by:
where \( P_r(t) \) is a rectangular pulse of duration \( T \) which starts at \( t = 0 \) and \( b_{n}^{*} \) is the \( n \)th user data sequence that takes on 0 or 1 values. \( T \) is the bit rate and \( p_{n}(t) \) (Equation 3.3) corresponds the \( n \)th user transmitted sequence assuming On-Off keying modulation numerical period \( F = T/T_{c} \), which is equal to the sequence length. For continuous transmitted sequence, the baseband signal has can be described by:

\[
\begin{align*}
\sum_{j=-\infty}^{\infty} a_{j}^{*} \cdot P_{T_{r}}(t - jT_{c})
\end{align*}
\]

with \( a_{j}^{*} = a_{j+F}^{*} \) is the signature sequence. \( P_{T_{c}}(t) \) is a rectangular pulse of duration \( T_{c} \).

The impulse response of the optical correlator 1 matched to the signature sequence of the transmitter \( a_{j}^{1} \); \( j = \{1, \ldots, F\} \) is a time-limited response given by:

\[
\begin{align*}
h_{l}(t) = \sum_{k=0}^{F-1} a_{j+F-k}^{1} \cdot P_{T_{c}}(t - kT_{c})
\end{align*}
\]

where the signature sequence allocated to a correlator is time-inverted with respect to the code sequence set by the transmitter. By using this notation, the response of the receiver in Equation 3.2 can be described in terms of the correlation and autocorrelation functions.

In the asynchronous system, each user can transmit its information randomly at different time delays, \( \tau_{k} (< T) \) \[42\]. The decision variable at the output of the receiver, \( O_{i}(T) \) has the equation:

\[
\begin{align*}
O_{i}(T) = b_{0}^{1} \cdot \psi_{1,1}(T) + I_{N,1}
\end{align*}
\]

where \( b_{0}^{1} \cdot \psi_{1,1}(T) \) is the zeroth data of the first user multiplied by the autocorrelation function at time \( T \) (in-phase autocorrelation peak), \( I_{N,1} \) is the multiple user interference at the output of the receiver 1. Based on the above definitions and equations, it can be proved that \( I_{N,1} \) depends on the partial correlation functions of all the user sequences. The multiple-user interference is given by:

\[
\begin{align*}
I_{N,1} = \sum_{k=2}^{N} b_{k}^{1} \cdot \psi_{k,1}(\tau_{k}) + b_{0}^{1} \cdot \psi_{1,1}(\tau_{k})
\end{align*}
\]
where \( \psi_{k,1} \) and \( \psi'_{k,1} \) are the continuous-time partial cross-correlation functions given by:

\[
\psi_{k,1}(\tau) = \int_{0}^{\tau} p_k(t) \cdot h_1(\tau - t) dt
\]

\[
\psi'_{k,1}(\tau) = \int_{\tau}^{\infty} p_k(t) \cdot h_1(\tau - t) dt
\]  \hspace{1cm} (3.9)

The sequences employed in this receiver have to maximise the autocorrelation peak and minimise the multiple-user interference. The implementation of the passive structure, as depicted in Figure 3.2, based on optical-fibre delay line is constrained to use positive sequences (0,1). Since the intensities of input pulses are added, such optical receiver has an impulse response which is a positive function, i.e. \( h_1(t) > 0 \). Similarly, the cross-correlation functions are positive functions. As a consequence there is a significant increase in the interference for each user that transmits its information. Since conventional sequences become inadequate for this application [7], pseudo-orthogonal sequences (0,1) with sparse marks have been proposed to maintain the cross-correlation to a minimum [31,5]. Nonetheless, in order to generate a long sparsely marked sequence for high speed operation, the generation of ultra-short pulses and the implementation of lossy delay lines are necessary. Those requirements make this approach unattractive for system implementations.

### 3.2.2 Asynchronous sequence inversion keyed (SIK) receiver

Other methods of coding that circumvent the use of long sequences have been considered for non-coherent receiver structures. In particular, the implementation of a passive structure that allows bipolar codes (1,-1) to be used, thereby diminishing the underlying problems associated with a unipolar implementation such as the lack of orthogonality of the code sequences. Figure 3.4 shows a passive receiver implementation based on a unipolar-bipolar correlation [38].
Figure 3.4 Unipolar-bipolar correlator receiver, implemented code sequence (1,1,1,0,1,0,0,0)

In this system, each element of the sequence is defined by delayed pulses. The photodetectors in differential configuration provides the bipolar capacity required to add in anti-phase the photocurrent induced by the optical delay lines. The coupling of pulses into a photodetector determines the sign of the element which in turn are detected by a respective photodiode. A source of noise is added in the final stage which represents the thermal noise of the amplifiers, shot noise and dark noise of the photodetectors. In order to recover user data, a bipolar reference sequence may be correlated directly with the transmitter unipolar sequence.

For the unipolar-bipolar correlator the sequence is transmitted in a shift sequence inversion method. Here the code sequence of the $n$th user, $a^n_j$ has bipolar values (-1,1) and the transmitted baseband signal follows a continuous chip pattern, $p_n(t)$ consisting of unipolar pulses of duration $T_c$ (or chips) given by:

$$p_n(t) = \sum_{j=-\infty}^{\infty} a^n_j \cdot P_{p_n}(t - jT_c)$$  \hspace{1cm} (3.10)

where $\overline{a^n_j}$ is the code sequence, with values inverted according to the data sequence. For each transmitted data bit $b_n = 1$, the transmitted sequence is $\overline{a^n_j} = \overline{(1 + a^n_j)/2}$ and for each transmitted data bit $b = 0$, the sequence is inverted, $\overline{a^n_j} = \overline{(1 - a^n_j)/2}$.

The use of bipolar receiver improves the reception of user data since the cross-correlation functions are not limited to positive values; hence the multiple user interference $I_{N,1}$ may be reduced by using a receiver with bipolar capacity. It has been demonstrated in [38] that the unipolar-bipolar correlation functions of a SIK receiver are equivalent to those of a bipolar
receiver with data encoded onto a bipolar sequence. This assumption can be maintained if the sequence code used has the property $\Sigma_j a_j^* = 0$; i.e. the sequences are balanced [38]. The effects of imbalance in the receiver can be considered as multi-user interference which can be represented as a dc level, and that depends on the number of users transmitting their information in a bit period [38]. Unfortunately, many sequences with good correlation properties, such as Gold or Kasami, presents imbalance thus the SIK inverter is prone to significant multiple-user interference.

A modification of the unipolar-bipolar correlator receiver structure was introduced in [43] where the effect of unbalanced codes can be removed. The data transmission utilises the SIK technique. The receiver structure is depicted in Figure 3.5 in which an unbalanced 1:2 coupler splits the input power to both arms of the receiver; each arm has a Mach-Zehnder interferometer. The interferometers are driven by electrical signals corresponding to the unipolar sequences $A_i$ and $A_i^*$. Depending on the value of the applied voltage, the output power of the interferometer can vary from a maximum to zero. The elements of the sequences are complementary as required in the application. The optical pulses are converted into photocurrents and subtracted in the balanced receiver to form a bipolar electrical signal.

The active implementation of the receiver has important disadvantages to be taken into account. Mach-Zehnder interferometers need to be biased in both arms with a voltage level that corresponds to the signature sequence. The chip rate is therefore limited by the speed of generators. In addition, electrical generators have to be chip synchronised to the input sequence, which limits such system to low-speed operations [39]. Nonetheless, theoretical analysis in [43] could be used for a passive implementation, including for such aim an unbalanced 1:2 coupler and two complementary correlators based on optical fibre tapped delay lines.

![Figure 3.5 Optical unipolar-bipolar correlator receiver with unbalanced coupler [43]](image)
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The key concept of this technique is the introduction of imbalance in the power splitter; i.e. set $\alpha \neq 1$ so that the input signal is divided unequally between the upper and lower arms of the receiver. In this approach, a proper selection of the unbalance index, $\alpha$, allows offsetting the level due to code unbalance. It has been proved analytically, that the multiple user interference at the output of the unipolar-bipolar receiver, $I_{N,1}$, is equal to the interference of a correlator receiver assuming bipolar sequence reception. Khaleghi, et al. [43] demonstrated that this method can work for Gold sequences and large number of active users, whilst some inaccuracies in the detection process (such as offset effects) arise when the system has low number of simultaneous users.

The theory of unipolar-bipolar correlator receiver in [38,43] points out that the limitation associated with the inability to process bipolar signals can to some extent be counteracted either by modifying the receiver structure [43] or by using specific balanced codes [38]. There have been no reports (known to the author) on practical demonstration of unipolar-bipolar receivers using the SIK technique.

3.2.3 Amplitude spectral encoding

Non-coherent OCDMA systems employ signal formats which are compatible with intensity-modulation direct-detection (IM-DD) schemes. The detection of the signals in such systems is non-coherent; the photodetector converts the incident optical field into a photocurrent which is proportional to the intensity of the field. This consideration, however, does not reflect the fact that optical signals possess wave properties (amplitude and phase). The assumption that optical signals from all users add incoherently without interfering each other is inaccurate, since the field properties of optical signals give rise to random interference between optical signals in the photodetector. This source of noise becomes important in the detection of OCDMA signals [49]. The noise associated with the interactions of the optical fields in the photodetection process is termed beat interference noise and affects the performance of optical networks in which users occupy the same wavelength space simultaneously [15]. Beat noise is directly proportional to the total optical power falling in the photodetector hence depends on the number of active users in the network.

The above consideration is important for the non-coherent OCDMA network using spectral encoding technique, including the non-coherent time domain schemes mentioned above [52]. Spectral encoding is a technique well-suited for non-coherent CDMA systems in which true orthogonal channels can be created; hence zero interference arises at the output of the receiver. However, the performance of the receiver is impaired by receiver noise and beat noise. The last noise contribution becomes significant in networks supporting a large number of users.
and working at very high-speed since beat noise is proportional to optical signal in the photodetector and the equivalent noise bandwidth of the receiver [49].

In the spectral encoding technique, coding is achieved in the frequency domain by slicing the available power spectrum into frequency slots or bins [36]. Non-coherent sources with coherence times lower than the bit period, \( \tau_c \ll T \) (or effectively linewidth much higher than bit rate) can provide the bandwidth requirement for high-speed operations. In particular, sources that maintain a steady spectrum across the modulation bandwidth (such as LEDs) can be utilised [36]. A conceptual block diagram of a spectral encoding CDMA system is shown in Figure 3.6.

In this structure, the available bandwidth of the optical devices, \( \Delta \nu \) (given in nanometres) is divided into slices. The transmitter comprises a broadband source and an optical device with a defined amplitude function \( A(\nu) \), the receiver comprises an optical coupler, two (complementary) optical filters with amplitude functions, \( A(\nu) \) and \( \bar{A}(\nu) \) and photodetectors in balanced configuration. In the system, the transmitter sends a pulse with spectral distribution \( A(\nu) \) when the data bit is ‘1’ and nothing is sent when data bit is ‘0’. Coded pulses from all users are combined at the star coupler. At the receiver, the optical signal is split and decoded by two complementary decoders. A splitter is used to divide the received signal into two parts (for certain code application, the power division can be unequal [49]). The split signals are input into two decoders with complementary functions and the balanced photodetection affects the de-spread of user data. Multiple-user interference can be completely cancelled. Low-pass filter and threshold decision devices allow recovering user data, the requirements of the electrical stage is defined by the bit rate.

Figure 3.6 Diagram of the spectral encoding OCDMA system
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Each transmitter and receiver pair has a unique code assigned that is defined in the frequency domain. By assuming a uniform spectrum of the optical source, the corresponding frequency response of the \( j \)-th transmitter filter is \( C_j(f) \), where the amplitude of frequency slots depends on a binary code sequence \( C_k^j \), \( k = 1, \ldots, N \) and \( C_k^j \in \{0,1\} \). In spectral encoding, the optical bandwidth of encoded pulses is much larger than the bit rate; therefore the correlation properties of spectral encoding depend solely on wavelength and do not change with time. The output at the \( i \)-th receiver is determined by the data transmission of the \( j \)-th user, \( b_j^i \) at this fixed time and given by:

\[
O_i(0) = b_j^i \cdot \psi_{i,j}(0)
\]

with \( \psi_{i,j} \) is the autocorrelation function. In general, the correlation function is given by [41]:

\[
\psi_{j,i}(0) = \int_{-\infty}^{\infty} C_i(f) \cdot C_j(f) df
\]

Each frequency slot of the optical devices is bandwidth restricted so as to ensure that the different components of the transmitted pulse have the same time delay. By assuming this practical condition of the constituent optical devices (for instance fibre gratings [47]), the correlation function \( \psi_{j,i}(0) \) becomes a measurement of the energy of the optical signal that is coupled into the output of the receiver due to the \( j \)-th transmitter at time \( t = 0 \). By considering now balanced structures in which the filter functions are complementary; i.e. \( C_j^i(f) = 1 - C_j(f) \), the output of the transmitter has the equation:

\[
\psi_{j,i}(0) = \int_{-\infty}^{\infty} C_j(f) \cdot C_i(f) df - \int_{-\infty}^{\infty} C_j(f) \cdot C_i(f) df
\]

The bipolar receiver computes the difference of the power coupled in both arms. Kavehrad et al. [36] has shown that by using certain codes, the differential receiver can remove completely multiple-user interference. At the output at the receiver, such interference is described by the periodic cross-correlation of sequences \( X \) and \( Y \), given by:

\[
\theta_{XY}(k) - \theta_{XY}(k) = \sum_{i=0}^{N-1} x_i y_{i+k} - x_i y_{i+k}
\]
where $\overline{X}$ corresponds to the complementary sequence $X$; $\{x_i\}_{i=0,...,N-1} = \{1 - x_i\}_{i=0,...,N-1}$ with $x_i \in \{0,1\}$. The use of various sequences such as m-sequences, Hadamard or Gold codes result in zero output at the receiver [36]. Additionally, for the receiver proposed in [36], the number of codes is equal to the length of the sequences employed.

Practical implementations of spectral encoders based on lenses, two-dimensional mask and more recently fibre gratings have been proposed [36,37,48]. A fibre-grating implementation of a decoder is shown in Figure 3.7 [37]. The received spectrum of chips follows a pattern defined by the signature sequence. The generation of such signature sequence can come from a fibre-grating based encoder, in which the Bragg wavelength of each grating is adjusted to change effectively the signature pattern to a predetermined address code.

The decoder comprises a coupler, fibre gratings in the lower arm and balanced detectors. The correlation process can be achieved assuming complementarity in the signals coupled into two photodiodes (the reflected and transmitted fields complement each other in this arrangement). The balanced photodetection therefore computes the difference of signal power coupled to both encoder arms; thereby data recovery can be achieved according to the spectral encoding pattern.

![Figure 3.7 Fibre-grating decoder implementation for spectral encoding [37]](image)

Fibre-grating implementations facilitate versatile and functional encoders and decoders. Bit rates of the order on 10 Gbit/s could be feasible with fibre-grating implementations [49]. Nonetheless, there are some inherent limitations in this approach. Spectral encoding requires devices that are able to "slice" more thinly the available spectrum. In order to support more users in the system without spoiling the orthogonality among coded signals, higher spectral resolutions and steady broadband optical sources are needed. Such conditions, however, cannot be easily met in practical systems. Broadband sources such as luminiscence diodes (LEDs) have variations across the power spectrum and the optical resolution cannot be increased beyond the
3.3. Coherent-OCDMA systems

The wave nature of the optical fields in the optical fibre allows extending the data carrying capacity by manipulating the phase and amplitude of optical signals. The structures reported above for decoding data are limited to detect only the optical power of signals. In coherent-CDMA systems, the time coherence of the optical source is greater than the chip length. Therefore, in the optical correlator process, pulses are summed up coherently. The phases of individual pulses or chips influence the interference and amplitude of the pulses at the output. An important consideration for coherent networks is the need of synchronisation among nodes; otherwise the distinct phase of pulses cannot add coherently at the output of the correlator. By using such coherent processors, the encoded data can be recovered by obtaining at the output an autocorrelation pulse with high amplitude of the central peak and low sidelobes.

3.3.1 Coherent system using ladder networks

An important component of for all optical coherent CDMA system is the ladder network, which is depicted in Figure 3.8. The phase shifters allow reconfiguring the ladder function. A practical system demonstration of this ladder network is cited in [46]. The implementation comprises Mach-Zehnder interferometers to shift the phase of optical pulses and optical fibre delay lines to introduce delay for spreading [45].

![Figure 3.8 Two-phase ladder network of an encoder and decoder in cascade [46]](image)

In the two-stage ladder network, the transmitter and receiver are matched coherently, implying that the phases of the encoder \( \{ \phi_1^e, \phi_2^e \} \) and decoder \( \{ \phi_1^d, \phi_2^d \} \) allow pulses modulated by the interferometers to be added at the output. To obtain a coherent response at the output without pulse overlapping, the delay introduced by each fibre line has to satisfy the inequality:
\[
\sum_{i=1}^{K-1} T_i < T_K; \quad K < N
\]  

where \( T_i \) corresponds to the differential delay in the \( i \)-th stage. In this arrangement, a pulse introduced to the input port is spread as it passes through the optical ladder. At the encoder, the interferometers shift the coupled signal allowing the incoming pulses to take on four possible phases, two non-interfering pairs; \( \{0, \pi/2\} \) and \( \{\pi/4, 3\pi/4\} \). At the decoder, an inverse function is defined by setting the appropriate phases so that encoded pulses are despread. From a theoretical point of view, inverse decoding permits an ideal reconstruction of the input pulse (zero side-lobes in the auto-correlation function), providing that the decoder be synchronised with the incoming input pulse [45].

The pulse reconstruction can be performed by the encoder in the reverse order, combining two pulses (each modulated with a specific phase) and modifying its relative phase. In the frequency domain, the matching condition is traduced into a \( \pi \)-phase shift on the phase addresses of the encoder and decoder. The matching condition is given by:

\[
\{\phi_1^d, \phi_2^d\} = \{\phi_1^e, \phi_2^e\} + \{\pi, \pi\}
\]  

Four different combinations of phase patterns used in the two-ladder network are given in Table 3.1. In the ladder networks, the interference can be removed since the phase shifts can add in phase and antiphase to despread data.

<table>
<thead>
<tr>
<th>( \Phi_1 )</th>
<th>( \Phi_2 )</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>( \pi )</td>
</tr>
<tr>
<td>( \pi )</td>
<td>0</td>
</tr>
<tr>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>( \pi/2 )</td>
<td>( \pi/2 )</td>
</tr>
</tbody>
</table>

A practical system demonstration of this ladder network based on polarisation-preserving fibre is cited in [46]. In each stage of the ladder a phase control is added so as to permit the reconfiguration of the transmitters by switching the phase according to the signature sequence. In spite of the adequate correlation measurements in such demonstration, the performance of ladder networks is impaired by physical limitations of optical elements [51]. Several factors, such as fibre dispersion or optical source line-width, have an impact on the
propagation of ultra-short pulses. These limitations must be considered for practical implementations in optical networks. Accordingly, in order to facilitate the implementation of coherent systems, additional elements such as dispersion compensators and phase stabilisation loops are required. This makes the system more expensive and not competitive with other alternatives. Nonetheless, given that such complexity in the implementation is similar to that of filters used in WDMA, it is claimed in [45] that coherent CDMA is a viable technology to implement high-speed networks.

Other coherent networks have been proposed in the literature [52] which requires phase masks and optical threshold devices. The need of utilising expensive optical devices (in a non-integrated form) makes this approach inefficient and cumbersome.

### 3.4. Practical technologies for OCDMA systems

In reference to the encoding and decoding hardware, the technology for OCDMA systems is still immature and costly. In fact, the inherent deficiencies of this technology limit the development of efficient OCDMA networks [35]. A suitable technology should allow many channels to be created and for many encoding schemes proposed for OCDMA that require long processors. In particular, the number of channels that can be created in temporal addressing schemes depends on the time duration of the orthogonal signals and the number of chips in the signature sequence. Table 3.2 lists some existing technologies that have been employed for multiplexing and encoding CDMA signals, including the chip and bit rate and number of elements (chips or frequency slots) in parenthesis achieved in those processors. These practical demonstrations allow testing of the different schemes proposed in the literature. The viability of practical networks in the multi-Gbit/s regime is yet incomplete either due to the high cost incurred in its deployment (at a wide scale) or due the low reliability to perform the encoding functions.

### Summary

This chapter presents an overview of the main methods of encoding and decoding proposed for both coherent and non-coherent OCDMA systems. It commences with an introduction to OCDMA systems highlighting their functional characteristics. Various coherent and non-coherent OCDMA schemes as well as the technological aspects related to their development where described. Time domain encoding was the first proposal for high-speed OCDMA systems. A passive receiver structure based on optical fibre delay lines can be used for
achieving random and asynchronous access to the networks; such networks though require sparsely marked coded sequences to achieve appropriate bit-error rates which make this receiver difficult to realise with the available technology. Given the complexities associated with the use of unipolar receiver structures, new receiver schemes were proposed that circumvent the problems associated with unipolar receiver. The unipolar-bipolar receiver correlator scheme was proposed in the context of SIK systems. This scheme improves the performance of the receiver whilst using small length codes such as Gold or Kasami. Nonetheless, such receiver is prone to multiple user interference resulting from unbalanced effects associated with imbalance of the codes. However, a straightforward modification to the unipolar-bipolar receiver structure was proposed that allows removing the unbalanced effects at the output of the receiver, hence the performance of such receivers can be improved whilst using conventional codes such as Gold or Kasami sequences.

Table 3.2. Reported demonstration and technologies for OCDMA systems

<table>
<thead>
<tr>
<th>SYSTEM BASED TECHNOLOGY</th>
<th>DETECTION &amp; OPTICAL SOURCE</th>
<th>PRACTICAL IMPLEMENTATION &amp; (LENGTH OF SEQUENCES)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Charge Transfer Devices [7]</td>
<td>Non-coherent unipolar pulses, LED</td>
<td>20 Kbp/s, 2.54 Mchp/s (127)</td>
</tr>
<tr>
<td>Superficial Acoustic Wave Filters [8]</td>
<td>Heterodyne reception, GaAlAs LED at 0.875 µm</td>
<td>78 Kbp/s, 10 Mchp/s (127)</td>
</tr>
<tr>
<td>Coherent Optical Ladder Network [46]</td>
<td>Coherent matching using phase reference, coherent source</td>
<td>69 Mbps, 20 ps switched pulses (2 stages, 4 phase coded pulses)</td>
</tr>
<tr>
<td>Match-Zehnder &amp; Electronic PN-sequence generator [39]</td>
<td>Envelope Detection, CW laser</td>
<td>45 Mbps, 1.5 Gchp/s (30)</td>
</tr>
<tr>
<td>Fibre Gratings [47]</td>
<td>Correlation in time and frequency domains, non-coherent sources</td>
<td>500 Mbps, 6 Gchip/s (12 × freq. Hops)</td>
</tr>
<tr>
<td>Superstructured fibre gratings [48]</td>
<td>Coherent phase addressing, mode-locked soliton laser</td>
<td>2.5 Gbit/s, 160 Gchip/s (63)</td>
</tr>
</tbody>
</table>
Other philosophy for non-coherent OCDMA systems was described, which encode data in the frequency (wavelength) domain instead of the time domain. The structures proposed for such spectral encoders require inexpensive optical sources and direct detection receivers with bandwidth equal to the bit rate. Overall, this method of encoding allows true orthogonal channels to be created in which the signal to noise ratio at the receiver is limited by beat noise. Other aspects of the technology such as the inability to slice the spectrum more thinly to accommodate more channels have been discussed.

Coherent systems were described as a potential alternative for efficient OCDMA systems. In such systems, the optical carrier is modulated (according to the signature sequence) to encode user data. In this approach, pulses with coherence time larger than the bit interval are summed coherently in the correlation process. In those networks, the decoding of coherent pulses is dependent on the phase relationship of all encoders and decoders in the network. A "master network" that locks all decoders in the network must be added in the optical CDMA network. A proposal of encoders and correlators based on polarisation-preserving fibre in the form of a passive ladder networks was presented. The system demonstration is cited in Table 3.2. In spite of the adequate correlation properties of phase modulated pulses using compact processors, the performance of ladder networks is impaired by physical limitations of optical elements. Several factors, such as fibre dispersion or optical source line-width, were mentioned. Furthermore, in order to facilitate the implementation of coherent systems, additional elements such as dispersion compensators and phase stabilisation loops are required. This renders the system perhaps not competitive with other alternatives. Such aspects were addressed in order to describe the practical limitations in the deployment of coherent CDMA systems. Towards the end of the chapter, a table of practical demonstration is provided in order to account for the current limitation of technologies for OCDMA systems.
CHAPTER 4

Distributed amplification and transversal filter concepts in high-speed lightwave optical systems

4.1 Introduction to receiver and transmitter for lightwave communication systems

Optical communication systems have had an outstanding impact on data transmission and telecommunications in the last two decades. Synchronous Optical Networks, SONET and Synchronous Digital Hierarchy, SDH are currently working at data ranges of about 10 Gbit/s. This successful data transmission in high-speed TDM links can be achieved owing to the excellent capacities of high-speed circuits and the advances in semiconductor technologies to implement the digital and amplifying functions required. Transmitters and receivers have benefited from the capacity of integrating in a single chip a wide range of functions including multiplexing, retiming, reshaping and regeneration. Nowadays, research efforts are devoted to integrated communication circuits handling data rates at 20 Gbit/s or 40 Gbit/s communication systems [12]. It has been anticipated in [13] that many advanced integrated circuits working at 20 Gbit/s could be upgraded to high speed operations, 40 Gbit/s or even 60 Gbit/s by introducing technologies that solve satisfactorily the underlying limitation on the speed of integrated circuits; for instance, the reduction on interconnecting elements, transistor sizes and parasitic elements of integrated circuits.
For the optical transmitter, the modulator driver is a crucial circuit. An integrated circuit producing large output swings with extremely wide bandwidth and high speed is an essential constituent for intensity modulated systems. A suitable approach to modulator driver design is the distributed amplifier because of its exceptional bandwidth operations and its versatility to construct broadband sub-systems. The need of high power gains (20 – 30 dB) required to attain appropriate voltage levels to the electro-optical modulator and coupling low power signals from digital circuitry requires surmounting several limitations associated with the active device itself, in particular the transistor breakdown voltage and low output resistance. Distributed amplifiers have shown appropriate power handling capabilities and wideband performance compatible with the needs of external modulators [54,55].

For traditional optical receivers, the most important integrated circuit is the front-end amplifier because it determines the overall performance of the optical link. Performance criteria that must be satisfied are minimum signal distortion and high-sensitivity. Nonetheless, in communication links that employ an EDFA prior to the photo-detector to boost the optical signal, sensitivity is not an essential design requirement, meaning low-noise characteristics are not crucial for the amplifiers of post-detection electronic stages. Usually those amplifiers must have transimpedance gains larger than 50 dBΩ. For most optical links, sensitivity in excess of -20 dBm is adequate for attaining low bit error rates (10^{-12} – 10^{-9}) of systems working at tens of Gbit/s. The bandwidth requirement of standard SONET networks at 10 Gbit/s requires the amplifier to have a bandwidth of at least 7 GHz; 0.7 times the bit rate of the incoming signal of a NRZ bit stream [56].

Other important specification is the low cut-off frequency of the amplifier, which can be important when processing a continuous string comprising of recurring binary digits; a requirement for optical communication systems is that of amplifying signals with low frequency spectral content. Usually, the low frequency response must be extended to a frequency of 5 decades below than the bit rate. In order to avoid signal distortion, the gain ripple must be kept small and in most circumstances a deviation of ± 2 dB is usually tolerable [12]. Regarding the phase characteristic, the group delay variation across the passband must be maintained typically less than ±20 ps so as to reduce pulse dispersion that distorts the signal, this is translated into a high-linear phase characteristic and an amplifier gain / frequency characteristic with a smooth roll-off at bandpass edges [12].

The sensitivity of the front-end amplifier is determined by its input noise current density; which depends strongly on the gain of the amplifier achieved by a specific topology and the active devices employed. There are different preamplifier configurations appropriate for use in lightwave systems; the most widely used is the transimpedance amplifier topology since
it offers large dynamic range, wide bandwidth and high sensitivity when compared to low impedance or high impedance amplifier topologies [14]. Transimpedance amplifiers are well-suited for 10 Gbit/s SONET systems. At higher frequencies, conventional feedback amplifiers possess inherent bandwidth limitations to achieve such speed requirements even if such amplifiers employ devices with large figure-of-merits [57]. The intrinsic parasitic elements of the devices in turn reduce the stability margin thereby limiting the effective amplifier bandwidth that can be attained.

For emerging telecommunication systems operating at bit rates of 20 or even 40 Gbit/s, the distributed amplifier can fulfil the bandwidth requirements since its bandwidth can be comparable to the figure-of-merit of active devices. In addition, implementations based on active devices featuring low noise figures such as HEMTs can present equivalent input noise current densities as low as 10 pA/√Hz and bandwidth up to 40 GHz [58]. In this development, an optical amplifier was assembled with a high-speed photodiode on an alumina substrate and distributed amplifiers connected via bond wires.

Another strategy to improve the optical receiver’s performance and save costs is the full integration of the photodiode and travelling wave amplifiers. In this approach, active devices are integrated with a photodiode by using a pre-patterned substrate that serves as a growing stack for the photodiode and other for the active device. In the optoelectronic IC described in [59], a pin photodiode is coupled to strip optical waveguides for illumination. Essentially, the arrangement corresponds to an edge-illuminated structure where the optical path is perpendicular to the electrical field created in the absorbing (undoped) material of the pin photodiode. By this means, the absorbing layer provides a longer interaction length of light thereby improving the quantum efficiency [59]. In the fabrication process, a selective area regrowth produces separate device layer stacks. The HEMT buffer thickness is adjusted to the photodiode layer stack so as to obtain a planar surface of the wafer to define the HEMT gates by lithography. Once the stacks are defined to constitute all the structures based on the same InP material system, interconnections in the form of air-bridges are employed to interconnect coplanar waveguides. Four HEMTs with different areas were produced by this process so as to obtain a TWA with optimised amplitude-frequency characteristic [60]. Low input impedance for the TWA was considered to avoid the need of a matching resistor in series with the photodiode at the cost of this impedance level being a dominant source of noise. Similarly, the use of low input impedance reduces the overall amplifier gain to 45 dBΩ and sensitivity of -16.5 dBm at bit rates of 20 Gbit/s. Although integrated optical receivers have similar performance parameters to those encountered in some conventional structures, the optoelectronic receiver approach has inherent advantages, such as the reduction of parasitics, lower cost and higher reliability when compared to hybrid structures. As a result of that, optoelectronic receivers are
well-suited to attain the ultra-high speed operation required in advanced optical communication systems.

Given the importance of the distributed amplifier concepts in ultra-high speed optical communication system, this chapter outlines some variants of the distributed amplifier and the concepts that allow broadband transversal filtering with distributed amplifiers. Significant research work has been done in this field and will be described in the sections below.

### 4.2 Active device technologies

The performance of an amplifier is determined to a large extent by the active device technology employed for the implementation. Besides a low noise figure of active devices, which is an essential figure for front-ends implementations, the active devices must have other figure of merits that become relevant for high-speed optical systems. Usually, the current gain cut-off frequency, $f_T$, and the maximum frequency of oscillation $f_{\text{max}}$ are important figures for high-speed systems. Transistor based on modulation-doped structures or heterojunctions possess enhanced figures of merits. The high-electron mobility transistor, HEMT is a field-effect transistor which permits large concentrations of conduction electrons confined in the heterointerface formed between the doped layers (by utilising III-V compound semiconductors) and undoped GaAs layers. This electron confinement improves the mobility process because free electrons are physically separated from the parents' donor by diffusion mechanisms and the ionised donor scattering is drastically decreased in the active layer [61]. The mobility improvement results in active devices with larger transconductance gains. In addition, advances in lithographic processing techniques have allowed the manufacturing of HEMTs with gate lengths approaching a tenth of micron, which results in a reduction of the parasitic elements of the device.

The heterojunction bipolar transistor, HBT is other device based on heterostructures. The operation of such devices is similar to bipolar junction transistors, in which a small base current is used to control the much larger emitter current. HBT have better figures of merit than those of bipolar transistors because of the utilisation of the heterostructures in the junctions. HBT presents larger $f_T$ which is an indicative of its suitability for high speed applications. Nonetheless, an important disadvantage of bipolar transistors is their dependence on changes of temperature, which for certain applications can be critical.
Chapter 4 Distributed amplification and transversal filter concepts

A conventional structure of a pseudomorphic-HEMT is shown in Figure 4.1, it depicts a detail of the different layers utilised in the vertical structure. The term pseudomorphic arises from the fact that the active layer based on GaAlAs/GaInAs/GaAs system introduces a lattice constant different from that of GaAs semiconductor in the substrate, in contrast to GaAs/GaAlAs systems where the lattice constant is very close to that of the GaAs [62]. The n⁺-GaAs cap layer in Figure 4.1 present at the top of the device is used for creating Ohmic contacts at the source and drain terminals. This becomes essential for the low noise figure presented by HEMTs and high speed applications where parasitic elements play an important role in the maximum frequency of oscillation $f_{max}$ of the device.

![Figure 4.1 Cross section of a pseudomorphic-HEMT with gate triangular profile [63]](image)

InGaAs semiconductor systems provide larger conduction band discontinuities in the two-dimensional electron confinement thereby larger electronic concentration are obtained. Larger concentrations leads to larger device transconductances without modifying substantially the capacitance fixed by the thickness of the AlGaAs donor and the space layers. The concentration of electrons in the vertical two-dimensional layer can be controlled by varying the bias of the Schotky contact. In the modulation of the layer, free electrons are confined at the backside of the layer; which reduces the output conductance and provide less voltage irregularities near to pinch-off characteristics [61].

A variant of the pseudomorphic HEMT is the use of InP compounds in the substrate and other advanced material systems for channel region. Those materials have improved the high electron mobilities and peak drift velocities of HEMT [62]. In this thesis a pseudomorphic-HEMT, which scheme is depicted in Figure 4.1, has been characterised for its utilisation in distributed amplifiers.
4.3 The distributed amplifier concept

The distributed amplifier, also referred to as travelling-wave amplifier, TWA is essentially a structure with active devices connected in parallel and with propagating waves along its input and output ATLs. Distributed amplification is effectively an additive amplification because the overall gain is the sum of individual gains of the active devices, in contrast with multi-stage amplifiers connected in cascade where the overall gain corresponds to the multiplication of gains of individual stages. The first distributed amplifier developed as a monolithic integrated circuit (MMIC) was reported by Ayasli in 1981 based on MESFETs [64]. Figure 4.2 depicts the DA.

![Distributed Amplifier Diagram](image)

**Figure 4.2 MMIC distributed amplifier**

The amplifier in MMIC form has identical sections, consisting of drain and gate interconnecting transmission lines and identical active devices. The passive elements are predominantly short transmission lines designed to act as inductors which can be implemented either as microstrip transmission lines or coplanar waveguide. Newly developed DAs have improved the performance and bandwidth. For instance, it has been demonstrated in [66] that by using an appropriate design strategy, losses associated with active devices can be counteracted at such a level that losses of transmission lines become more significant in the design of such distributed amplifiers. Micromachined transmission lines used in [66] have proved suitable transmission characteristics for DAs with bandwidths of several tens of GHz.

In single-device distributed amplifier, passive coupling between drain and gate ATLs is maintained low by employing highly unilateral active devices. Passive couplings are mainly associated to “feedback elements” of active devices and their effects tend to increase with the frequency because of the dependence of the inverse gain parameter of practical transistors. If
parasitic elements can be neglected, the small-signal model of the device can be approximated by the unilateral one as depicted in Figure 4.3

![Figure 4.3 Small-signal HEMT model (left) and unilateral equivalent circuit (right)](image)

The use of highly unilateral devices improves the performance of distributed amplifiers since devices with high isolation characteristics present generally large figure-of-merits [62]. Likewise, low parasitic couplings ensure larger power gains and wide bandwidth [68,69]. By modelling the active device with the unilateral model, the distributed amplifier is represented by two transmission line circuits actively coupled by the device transconductance and those are referred to as the artificial transmission line (ATL). Figure 4.4 depicts a model of the distributed amplifier based on identical HEMTs and using identical interconnecting transmission lines.

![Figure 4.4 Small-signal equivalent circuit of HEMT-based distributed amplifier ATLs](image)
The gain mechanism of the distributed amplifier can be described in terms of propagating waves in ATLs. The uniformity of ATL sections ensures matching conditions in a wide range of frequencies thereby low reflections arise. In the distributed amplifier, little energy is absorbed by active device as waves propagate along ATLs. Signal applied at the input port travels down the gate ATL and then is absorbed by the terminal impedance. As the signal travels down the gate, each active device is in turn excited by the voltage wave. Voltages appearing across the input impedance of the device are amplified via the device transconductance. The output current of active devices are coupled to the drain line creating forward and reverse travelling waves. If the phase velocities of gate and drain lines are made identical; i.e. the distributed amplifier is phase synchronised, forward wave components coupled in drain-ATL are added in phase as they arrived at the output port. An amplified signal is obtained as a result of the additive contribution of all wave components. On the other hand, coupled waves travelling in the inverse or counterpropagation direction in drain lines are cancelled or absorbed by the resistance in the idle port, giving rise to a non constant gain characteristic over the amplifier bandwidth at that opposite port.

In practical implementations differences between drain and gate device capacitive impedance produce a group delay in gate-ATL greater than drain-ATL group delay. In order to equalise the phase velocity in ATLs, inductors in the form of short transmission lines $L_p$ are introduced in the drain terminal of all amplifier sections. Synchronisation can also be achieved by using thin film capacitors connected in parallel with drain terminals as a method for compensating the differences of capacitive load in drain and gate transmission lines [65].

In distributed amplifiers with identical sections and active devices there exists a limit on the number of active devices that would effectively increase the overall gain [70]. Beyond such limit the gain increase due to the fact that the addition of a device is not enough for compensating the attenuation of the entire amplifier. Moreover, dissipative elements in equivalent models of active device give rise to inhomogeneous coupled waves in both lines and non-constant gain characteristics. Different alternatives to improve the distributed amplifier performance have been proposed in the literature [69]. An alternative that retains the basic distributed amplifier structure was given by Niclas et. al. [71]. The gain characteristic of the distributed amplifier was improved by designing circuit elements with different parameters. In this approach, the uniformity of distributed amplifier is not preserved. It was demonstrated in [71] that by introducing distributed amplifier design with dissimilar sections, the principle of distributed amplification allows ultra-broadband operation whilst improving the gain uniformity across the bandwidth. The benefits of this approach will be later considered in other distributed amplifiers and circuit topologies in which circuit optimisation becomes an essential resource in the design process.
4.4 Fundamental circuit concepts

The distributed amplifier’s broadband capacity stems from the matching conditions met by the connection in cascade of identical sections comprising the artificial transmission lines (ATL). Artificial transmission lines differ from real transmission lines (RTL) in the fact that ATL is composed of lumped elements associated with the active device and interconnecting inductances (usually synthesised by short transmission lines). An ATL is basically a ladder of reciprocal sections. Its simplest structure is called a LC low pass filter section; an elemental low pass LC filter is depicted on Figure 4.5

![Figure 4.5 Low pass LC filter section (left) and T-low pass filter composed of low pass section (right)](image)

Connection of simple sections forms a ladder of symmetrical networks with identical input-output impedances. Two elemental sections connected in T, as depicted in Figure 4.5, form a symmetrical T-section. The purely reactive symmetrical two-port circuit is referred to as a constant-k filter since its series arm impedance and shunt arm impedance satisfy the following relation:

\[ \frac{Z}{\gamma} = k^2 \]  

(4.1)

If constant k-sections are connected in cascade as in distributed amplifiers and other passive networks, impedances seen looking into and away from each port are identical and termed image impedances [69,72]. The image impedance of a symmetrical T-section \( Z_{oT} \) is given by:

\[ Z_{oT} = Z_o \sqrt{1 - \frac{\omega^2}{\omega_c^2}} \]  

(4.2)
where $\omega$ is the angular frequency, $\omega_c$ is the angular frequency at cut-off and $Z_o$ is the characteristic impedance of the transmission line given, respectively, by:

$$\omega_c = \frac{2}{\sqrt{LC}}, \quad Z_o = \sqrt{\frac{L}{C}} \quad (4.3)$$

### 4.4.1 Filter sections and propagation parameters of ATLS

An approach to the analysis of the distributed amplifier involves a description of lumped-filtering sections using propagation functions. In this approach, it is assumed that the amplifier's response depends essentially on the loading effect of lumped elements periodically distributed along ATLS whilst dispersion and other physical characteristics of the transmission lines are not taken into account. Signals in distributed amplifiers can be treated as propagating waves due to the distributed nature of the interconnecting transmission lines. In this regard, a complete analogy between signals and waves suggests that the distributed amplifier can be analysed via electromagnetic theory in terms of distributed and lumped parameters [71]. Nonetheless, such task is not easy to undertake due to the inherent complexities of practical distributed circuits [73].

In a transistorised artificial transmission line\(^{\dagger}\), the nature of lumped elements distributed along the transmission line is complex when compared to the filter sections described above. In fact, the resulting transmission lines of practical distributed amplifier take a complex form when complete device models are included in artificial transmission lines [73, 74]. The image impedance concept can be extended to more complex structures by assuming that the distributed amplifier lines are terminated by their image impedances. Conventional parameters of linear circuits are used to represent each filter section which facilitates the analysis of the distributed amplifiers [72]. In order to determine propagation parameters of practical stages, the constituent T or II sections can be generalised as depicted in Figure 4.6.

Equations for generalised T and II section impedances are given respectively by [69]:

$$Z\omega_T = \sqrt{Z_1Z_2} \times \sqrt{1 + \frac{Z_1}{4Z_2}} \quad Z\omega_{\pi} = \sqrt{Z_1Z_2} \times \left( \sqrt{1 + \frac{Z_1}{4Z_2}} \right)^{-1} \quad (4.4)$$

\(^{\dagger}\) A terminology used commonly in first solid-state distributed amplifiers [74]
Chapter 4 Distributed amplification and transversal filter concepts

The propagation constant of reciprocal PI or T filters is defined by the current ratio:

\[ \gamma = \ln \left( \frac{I_1}{I_2} \right) \]  

(4.5)

where \( I_1 \) and \( I_2 \) are the input and output currents of the section and \( \ln(\cdot) \) is the natural logarithm function. The propagation constant per section is split into a real and imaginary part:

\[ \gamma = \alpha + j\beta \]  

(4.6)

where \( \alpha \) is defined as the attenuation of the electric wave and \( \beta \) is the phase constant, both are frequency dependent parameters. The inherent delay defined by the analogy with lumped sections can be obtained for each filter section from the derivative of the phase constant:

\[ \tau_d = \frac{d\beta}{d\omega} \]  

(4.7)

Phase and attenuation are fundamental parameters to determine the filter response in frequency domain. The current ratio is easy determined in sections matched on image basis. For symmetrical PI or T-sections depicted on Figure 4.6, the current ratio is given by [72]:

\[ \frac{I_1}{I_2} = 1 + \frac{1}{2} \frac{Z_1}{Z_2} + \sqrt{\frac{Z_1}{Z_2} + \frac{Z_1^2}{4Z_2^2}} \]  

(4.8)
Those fundamental equations allow obtaining propagation parameters of arbitrary filter sections. The relationship between the propagation constant and the propagation parameters of generalised structures is given by [72]:

\[ \gamma = \cosh^{-1}(1 + \frac{Z_1}{2Z_2}) \]  

(4.9)

The aforementioned concepts and relationships give valuable information to analyse the performance of wideband distributed structures using image propagation functions and virtually can be applied to obtain an approximated description of any complex structure with lumped sections terminated in image impedance [70,74].

### 4.4.2 Propagation parameters of lossless ATL

The lossless artificial transmission lines based on reciprocal LC-sections terminated with image impedance represent a first approximation to distributed circuits. This model, far from being utilised in detailed designs, permits outlining important characteristics of distributed circuits. The intrinsic parasitic elements of the active devices and the resistive elements utilised at input and output ports have a substantial impact on the behaviour of distributed amplifiers [20]. Similarly, the use of input and output resistances at input and output ports gives rise to impedance mismatch that in turn deteriorates the transmission of signals. These second order effects are not taken into account in lossless ATLs.

Lossless ATLs terminated in image impedance permits maintaining constant impedance in a limited range of frequency. This salient feature can be illustrated in the following lumped representation in which voltage and current waves are coupled to the artificial transmission line terminated in image impedance.

![Figure 4.7 Lossless artificial transmission line](image-url)
By using previous definitions and concepts, the impedance of the artificial transmission line looking away from the current source results to be:

$$|Z(j\omega)| = \begin{cases} \frac{L}{\sqrt{C}} & ; \omega \leq \omega_c \\ \frac{2}{C} \times \frac{1}{\omega + \sqrt{\omega^2 - \omega_c^2}} & ; \omega > \omega_c \end{cases}$$  \hspace{1cm} (4.10)$$

Where $\omega_c$ is the critical frequency defined by Equation 4.3.

Equation 4.10 shows constant impedance up to $\omega_c$ where travelling waves are coupled to the transmission lines without distortion. High attenuation will be introduced at frequencies beyond the critical frequency $\omega_c$. In practical distributed circuits, constant amplitude characteristic cannot be attained easily due to difficulties in obtaining appropriate image impedance values, that can be used for matching, with physically realisable circuits. Nevertheless, a large number of transmission line sections results in a good approximation to constant characteristics [69].

Another important property of the lossless ATL is the phase distortion introduced by cascaded sections. Since image impedance varies within the passband, the phase departs from linearity. By utilising the equation of transmission lines, attenuation and phase distortion of a single LC-section can be obtained as shown in Figure 4.8. It shows that the phase departs from linearity as frequency approaches the cut-off frequency. Since phase characteristic is not linear with the frequency, the group delay is not uniform within the pass band giving rise to pulse dispersion. In applications where the pulse shape characteristics must be preserved, both a linear phase and constant gain ensure amplification of bandlimited pulses with moderate distortion; however, this requirement imposes severe restriction on the practicability of the distributed amplifier.

![Figure 4.8 Attenuation and phase against normalised frequency (\(\omega/\omega_c\)) for constant k-filters](image-url)
Investigation into the transient characteristics of the distributed amplifiers is well documented in the literature [74, 75]. Those reports pertain to years when vacuum tubes were predominately employed as the active devices of distributed amplifiers. The thermionic emission of vacuum tubes does not involve dissipative effects; thus lossless transmission line as that described above are well-suited to describe distributed amplifier based on vacuum tubes. The transient response of such distributed amplifiers shows an identifiable rise time and a damped oscillation of significant amplitude for pulse amplification applications [75]. Conversely, in distributed amplifiers based on solid state devices, low oscillation response appears due to the presence of losses of the active devices [76]. Furthermore, losses associated with active devices improve the phase linearity of artificial transmission lines sections which reduces the amplitude of such oscillation components [75]. The oscillatory component of transient responses pertaining to distributed structures based on HEMTs will be analysed in Chapter 5.

4.5 Matrix amplifier

Theoretical analysis of conventional distributed amplifiers shows a limit on the number of active devices that effectively maximises the overall gain [70]. Beyond such limit, the gain of an additional device is not enough for compensating the increased attenuation factor of the entire amplifier which increases exponentially with the number of devices. The cascading of distributed amplifiers constitutes a viable solution to enhance the gain, resulting in a multiplication of gains of cascade stages. A modified distributed amplifier configuration, termed the matrix amplifier, was proposed by Niclas, et. al. [82] as a viable solution to build wideband high-gain amplifiers. The matrix amplifier is an alternative structure to build high-gain amplifiers combining the additive and multiplicative amplifying principles. The matrix amplifier consists of two rows of active devices sharing an artificial transmission line and is depicted in Figure 4.9.

In the analogy with two distributed amplifiers connected in cascade, the row of active devices stacked on the top corresponds to a single distributed amplifier. The difference between both approaches lies in that for the case of cascaded stages, amplified signals from the first distributed amplifier undergoes the same phase delay before signals are amplified by the following distributed amplifier, whereas in the matrix amplifier, signals amplified by devices of the first (input) row are amplified at different phase delays by active devices at the second row. For such aim, the ATL that concentrates drain and gate terminals of the active device enables the propagation of amplified signals from the low input row and also serves to feed at the same
time the second row of active devices using the stacked bias technique [65]. In this configuration, the multiplicative mechanism of the matrix amplifier gives, at mid frequencies, an overall gain equal to the squared gain of the distributed amplifier [83].

![Two-tier matrix amplifier diagram](image)

**Figure 4.9. Two-tier matrix amplifier [83]**

In the two-row matrix amplifier shown in Figure 4.9, the structure consists of three artificial transmission lines. The idle gate and drain ports are terminated with resistors that equal the characteristic impedances of their respective artificial transmission lines. Since the central artificial transmission line concentrates active devices of both rows, the capacitance per section on the central line corresponds to that of the parallel of the input capacitance of the upper transistor and the output capacitance of the transistor in the lower row. A detailed analysis given in [82] shows that device losses from both rows impair the gain uniformity and reduces the effective bandwidth. A method for improving the response is to increase the impedance of the central line by doubling the inductance whilst the effect of device losses is reduced. Niclas *et al.* [82] considered adjusting individual elements to enhance performance via optimisation. Without optimising each stage of the matrix amplifier, the gain and bandwidth performance is insufficient and lower than that of conventional amplifier connected in cascade.

The bandwidth of the matrix amplifier becomes somewhat smaller than the equivalent cascade of distributed amplifier when both designs use the same transistors. In favour of the matrix amplifier it is claimed that it shows a significantly better noise behaviour and higher input / output isolation [83]. For MMIC implementations the matrix amplifier is favoured; since
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The overall circuit area is reduced by approximately 65% in comparison with cascaded implementations [82]. This represents a valuable advantage to microwave designs since the area-per-chip of the prototype can decide its fabrication cost in monolithic integrated form.

4.6 Transversal filters and adaptive equalisers

The expansion of bit-rate × distance product of long-haul high-speed optical communications systems has imposed demanding performance requirements on optical transmitters and receivers. Several transmission impairments of the optical link degrade the bit error rate because of the increased intersymbolic interference, ISI. Among the impairments encountered in optical communication systems, chromatic dispersion is usually the dominant cause of ISI although non-ideal receiver responses and optical source fluctuations can become an important source of impairment in both coherent and non-coherent optical systems [85]. However, chromatic dispersion is a non-linear effect with conventional direct detection receivers [85, 86] and first order equalisers are unable to compensate efficiently for chromatic distortion since the phase of optical pulses is lost in postdetection receivers. For conventional IM-DD systems, linear equalisation based on tapped delay lines permits compensating optical fibre dispersion, polarisation dispersion and receiver bandwidth limitations [85] with a reduction of receiver power penalty by several decibels [88].

In pursuing such improvements, adaptive equalisers have also been considered for its use within the optical receivers. For instance, a transversal filter for direct detection receivers based on optical devices has been reported in [87], featuring tunable time delays with reconfiguration speeds at nanoseconds. Eventually such transversal filter could be included in adaptive equalisers for short-haul high-speed fibre optic communication systems. In spite of the fact that such transversal filter presents good performance characteristics, the cost incurred in the realisation of optical processors, their difficulty of integration and the need of bulky elements for reducing thermal drifts of optical devices makes this solution inappropriate for widespread utilisation in cost-sensitive optical communication systems.

Equalisation in the electrical domain can be attained at speeds of several Gbit/s using microwave transversal filters. There is a steadily growing research in the development of electrical equalisers due to the advantages that offers the integration of electronic subsystems with optical receivers. Microwave tapped-delay line filters have been used to compensate for linear and non-linear distortion in the electrical stages of optical receiver [88]; however, the capacity of adaptability to changing transmission characteristics and the capacity of reconfiguration can be important considerations to take into account in the selection of filter.
topology. Additionally, transversal filters must allow wide bandwidth operation in the microwave and millimetre-wave frequencies regime.

It has been recognised that distributed circuits possess good potentialities to fulfil those stringent requirements. Distributed-amplifier based transversal filters have excellent capabilities of monolithic integration; their transmission characteristics can be designed to obtain filters with appropriate phase linearity and bandwidths limited essentially by the figure-of-merit of transistors. Given the significance of wideband transversal filters in high-speed optical communication systems, different transversal filter concepts are outlined and their salient features described in this chapter.

### 4.7 Distributed-amplifier-based transversal filters

Conventional distributed amplifiers are designed to have drain and gate transmission lines with identical phase velocities; i.e. the complex phase shifts on both lines are essentially the same. Phase synchronised distributed amplifiers present a flat gain and almost linear phase characteristics in a wide range of frequencies. When the distributed amplifier is not phase synchronised, signals coupled to drain transmission line are out of phase at the output port. By relinquishing the need for uniformity in the artificial transmission line and allowing distributed amplifier to have different phase velocities and gains, input signals can be amplified and delayed in a similar way to transversal filters at bandwidths essentially limited by the figure-of-merit of the active device employed. The analogy between distributed amplifiers and transversal filter was established for the first time in [24] by showing that the transversal filter and distributed amplifier are functionally equivalent topologies, by which the gain and delay coefficients of the transversal filter can be selected to obtain a prescribed transfer function. A general topology of the microwave distributed amplifier circuit diagram featuring different delays and gains is shown in Figure 4.10 [24]

![Figure 4.10 (N+1)-tap microwave distributed amplifier circuit diagram](image)
Here, it is assumed that signals travelling along the gate delay line are sampled at each tap output. The sampled outputs are multiplied by their respective gain coefficients $G_1, G_2, \ldots, G_{N+1}$ and summed to form the final output. The transfer function of the topology described is given by [24]:

$$H_F(j\omega) = \sum_{k=1}^{N+1} G_k \exp \left(-j\omega \sum_{i=0}^{k-1} \tau_i \right)$$

(4.13)

where $G_k$ is the gain of taps and $\tau_i$ is the differential delay between adjacent taps.

A distinct feature of the distributed amplifier is the creation of two signal paths in the drain-ATL corresponding to the propagating and counter propagating coupled waves. By taking the output signal in the direction of propagating waves, the transfer function corresponds to the forward-gain mode of the distributed amplifier. The transfer function is given by Equation 4.13 with differential delay time given by:

$$\tau_k = \tau_{gk} - \tau_{dk}$$

(4.14)

with $\tau_{gk}$ and $\tau_{dk}$ are equal for the gate and drain of the $k$th-section of the microwave transversal filter. Figure 4.11(a) depicts a block diagram of two transversal filter topologies.

When the output signal is taken in the direction of the counter propagating wave, the transversal filter topology corresponds to that depicted in Figure 4.11(b). The transfer function corresponds to the reverse-mode gain function given by Equation 4.13 with differential delays corresponding to the sum of delays introduced by the gate and drain sections [24]:

$$\tau_k = \tau_{gk} + \tau_{dk}$$

(4.15)

From a practical point of view, a transversal filter based on the reverse-mode topology is favoured when larger differential delays and more compact designs are required. Given the requirements for large interstage delay in some filter approaches, the reverse-mode topology is used in MMIC transversal filter implementations. The flexibility in the selection of gain and delay elements implies that, when appropriate, a linear phase across the frequency range of interest is not a relevant design aim. It has been demonstrated that a judicious selection of filter interstage delays and tap gains [24, 91] allows obtaining specific distributed filter responses by abandoning the linear phase restriction across the bandwidth. This design flexibility has enabled new distributed transversal filters to be contrived for their use in high-speed lightwave systems. For instance, recent developments that utilises the analogy between the distributed amplifier and transversal include filter shaping filters for high-speed soliton filters [24, 91-93], the
construction of tunable post-detection filters [94,95]; and transversal adaptive equaliser for lightwave doubinary systems [25]. Given the significance of such developments in the design of sub-systems for high-speed lightwave systems, their salient features will be outlined in this chapter.

![Block diagrams of two distributed-based transversal-filter topologies](image)

**Figure 4.11 Block diagrams of two distributed-based transversal-filter topologies**

### 4.8 Microwave GaAs MESFET transversal filter concept

The first adaptive transversal filter operating at microwave bandwidths was proposed by Jutzi, W. in 1970 [95] showing for the first time that transversal filter based on distributed amplifier concepts is a feasible alternative to implement wideband tunable transversal filters. Historically, Jutzi had proposed this transversal filter concept ten years before Ayasli reported the first monolithic distributed amplifier based on MESFETs [95]. 1-GHz bandwidth operation was attained with early GaAs MESFETs in a monolithic integrated circuit (MIC) implementation. From this precursory implementation, it was concluded that the distributed transversal filter concept could work at high-speed operations by using transistors with improved figure-of-merits [95], clearly exceeding the bandwidth of other technologies.

The first transversal filter involves tapped delay lines with constant delay increment and active devices interconnected to achieve amplitude-weighting. The transversal filter is actively tuned to approximate the transfer function defined in discrete time basis.
Chapter 4 Distributed amplification and transversal filter concepts

\[ H_F(j\omega) = \left[ a_0 + \frac{1}{2} \sum_{n=1}^{N} a_n (\exp(j\omega NT) + \exp(-j\omega NT)) \right] \exp(-j\omega NT) \]  (4.16)

where \( a_n \) are gain coefficients and \( T \) is the uniform sampling time. Phase characteristics given in Equation 4.16 needs a filter implementation with \( 2N+1 \) taps and amplitude-weighting coefficients symmetrically distributed around the \( N+1 \) tap gain. The first implementation was a 15-tap transversal filter and the target filter function defined by eight \( a_n \) coefficients [95].

The realisation of a generic transfer function needs positive and negative filter coefficients, thus the filter must be designed with a method of adjusting tap gain weight and sign of individual filter tap. Jutzi came up with an equaliser scheme with capacity of changing the amplitude and sign of adaptive filter coefficients. A diagram of the distributed transversal filter is given in Figure 4.12.

The distributed structure consists of two rows of active devices interconnected via transmission lines. The filter topology is a triple-ATL structure and its implementation utilises a distributed amplifier by which the incident signal is split into two signals of opposite phase to feed two separate gate lines. A delay element in the second gate line compensates for the excess delay introduced by the distributed amplifier. The analogy between transversal filters and distributed amplifiers is evident by assuming that the structure can be viewed as two distributed amplifiers in the reverse-mode gain sharing a common drain-ATL.

Travelling-waves are coupled separately in both gate lines, which in turn excite the active device and induces a signal in the common drain-ATL via the device transconductance. The common drain-ATL linearly combines signals of both rows of active devices with essentially the same delay characteristics, thereby enabling the signal at the output port to be proportional to the difference of transconductances of the pair of active devices. The reverse-mode gain operation of the filter produces a differential tap delay which in this arrangement is equal to the sum of delays in the gate and drain sections of the transmission lines. Handling the counterpropagating wave component in common-drain ATL reduces the time delay required from elements interconnecting active devices that consume large amount of area hybrid or monolithic implementation.
From a practical point, active devices have losses that modify the transmission characteristics of the gate and drain ATLs. Waves travelling on different filter paths encounter different attenuation levels; consequently, the asymmetry of the filter taps around the centre tap gives rise to a reduced accuracy in the response. Losses of the distributed transversal filter have to be compensated using tap gain adjustment. Jutzi contemplated an active tuning technique based on the application of external gate bias voltage to control separately the transconductance of active devices. Device transconductance can be controlled from zero to the maximum device transconductance thereby compensating the accumulated attenuation and setting accurately filter coefficients. It is important to notice that the travelling wave in the common-drain line is prone to a high level of attenuation given the parallel connection of active devices in the drain line. Because of losses and other device limitations, the filter tap gains are lower than one and a\'s coefficients in Equation 4.16 are dubbed filter attenuation factors [95].

A complete analysis of the electrical characteristics of GaAs MESFET shows a large variation of input capacitance with the applied input voltage. It was found that the variation of the input capacitance in relation to the applied voltage level disrupts the uniformity of the gate lines, giving rise to reflections. Nonetheless, such dissimilarities could be reduced by increasing capacitive loading, adding periodically discrete capacitors $C_{\text{ext}}$ in gate lines as shown in Figure 4.12 [95].

The distributed transversal filter was implemented in hybrid form utilising Teflon coaxial cables having a 1 mm inner conductor diameter in all ATLs. Figure 4.13 depicts in
Chapter 4 Distributed amplification and transversal filter concepts

detail the hardware of equaliser taps in which the utilisation of discrete blocking capacitors, thin wires and bonded transistor chips is clearly associated with low microwave frequency implementations. In transversal filters implementations, the introduction of sufficient signal delay between adjacent taps is crucial for the filter performance. The delay between taps is the result of both the delay associated with the loading of transmission lines by parasitic (lumped) elements of active devices and the inherent delay of the electromagnetic wave in the transmission line. The differential delay in the hybrid implementation is given by:

\[
T = 2 \left[ \tau_T + \frac{\sqrt{\varepsilon_{rv}}}{c_o} d_v \right]
\]

where \(\tau_T\) is the delay of the tap-line sections, \(d_v\) is the length of delay lines, \(\varepsilon_{rv}\) is the relative dielectric constant (\(= 2\) for Teflon) and \(c_o\) is the velocity of light in a vacuum. Excessive parasitic elements interconnecting the gate and drain terminals, as shown in Figure 4.13, render a delay related to the loading, \(\tau_T\) only of 38 ps; 10% of the total tap differential delay obtained by long coaxial lines. The equaliser was constructed with a sampling frequency of 2.6 GHz; 380 ps differential delay. Considering the electrical parameters of transmission lines, the required length of coaxial lines was 32 mm [95].

The total equalisation span of the implementation is mainly determined by the filter response. Since the attenuation of coaxial transmission lines is negligible in the frequency range of interest, the main bandwidth limitation is associated to active devices employed. In bandwidth-limited distributed circuits, input pulses travelling through filter sections undergo pulse dispersion. In application where the pulse shape needs to be preserved, a criterion must be established to gauge the effect of pulse dispersion and other bandwidth limitations. The method consists in allowing the pulses to change their width and other characteristics as they travel through filter band-limited sections and then make comparisons with the transient characteristics of input pulses. If the final step response rise-time, \(T_R\) is allowed to be equal to (or lower than) a half of the pulse width, pulses can be distinguished with low inter symbol interference at the output of \(T\)-spaced transversal filters. For this equaliser scheme, the equalisation span was determined if the step rise time of each tap satisfies the inequality:

\[
T_R < 0.8 \cdot T/2
\]

This criterion limited the equaliser implementation to 15-taps at uniform sampling intervals of 380 ps. Besides the effects of impedance mismatch essentially associated to the distributed structure, the overall performance is affected by the physical limitations of the devices

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employed and the low reproducibility of the electrical characteristics of the artificial transmission lines. To the best of the author’s knowledge, apart from the hybrid implementation described above all microwave distributed transversal filters reported in the open literature have been implemented as MMICs.

![Diagram of an equaliser tap in hybrid implementation](image)

**Figure 4.13 Hardware scheme of an equaliser tap in hybrid implementation**  
*after Jutzi, W [95]*

### 4.9 Pulse-shaping distributed-based transversal filters

Distributed amplifiers have been designed to work as baseband front-end amplifiers in optical receivers followed by shaping passive filters previous to the decision circuit [92]. Using the analogy between distributed amplifier and transversal filters, distributed structures can be designed to achieve amplification and filtering / shaping functions, reducing by this means the number of interconnecting elements in the optical receiver. Essentially, two approaches have been considered for pulse-shaping transversal filter; those with amplitude-weighting coefficients symmetrically distributed around the centre of the structure and coefficients defined at constant
time intervals [91] and transversal filters designed with different stage delays selected to fit an asymmetric impulse response [90,92]. Those distributed amplifier topologies with embedded pulse shaping capability utilise the dissimilar phase characteristics of the drain and gate artificial transmission lines to obtain the interstage delay. Input signals are sampled at intervals shorter than the input pulse duration and in turn multiplied by tap gain coefficients.

Moreira et al [91] introduced a postdetection filter based on the forward-gain mode topology depicted in Figure 4.11(a) The transversal filter concept calls for the case where tap gains are different and input pulses are sampled at constant time intervals. The design is aimed to present a constant phase and symmetry around the centre of the structure. Tap gains were selected to satisfy a raised cosine spectral function to achieve minimal ISI at the optical receiver. Figure 4.14 depicts the distributed filter cell with the active device connected directly to drain and gain transmission line and additional sections of transmission lines in the gate line so as to increase the interstage delay. The subscript in the transistor size and transmission lines implies variability in the transistor areas (in correspondence with the filter coefficients) and additional transmission line sections to equalise interstage filter delays.

![Figure 4.14 Distributed transversal filter cell [91]](image)

This transversal filter was designed to shape high rate input pulses. In this filter approach the coefficients of the transversal filter are obtained by sampling pulses at regular intervals. GaAs MESFETs with three different areas where employed in a 5-tap implementation, presenting a good approximation to the transimpedance gain and group delay target responses [91]. In the implementation with MESFETs, the input capacitance $C_g$ depends on the transistor size that scales the stage gain or transconductance. Because of the variety of device gains, gate-ATL sections present a non constant capacitive loading, giving rise to different interstage delays. A method to compensate for the excess delay thereby achieving a uniform sampling is based on the introduction of additional inductors in the form of short transmission lines. In
MMIC implementations, thin transmission lines can be designed to increase the required inductance per section. Simulation results of a filter implementation based on a MMIC process, suitable for 10 Gbps optical receiver, was reported in [90].

A different approach to distributed transversal filter is based on the selection of different tap gains and delays to obtain an asymmetrical impulse response. A MMIC transversal filter topology featuring different interestage delays was introduced by Bojak, et. al [24] as a filter scheme to amplify and reshape high rate pulses. In a previous paper [92] it was suggested for the first time that transversal filter featuring unequal tap delays $\tau_k$ and equal / unequal filter tap gains $G_k$ could be implemented for use in optical receivers. Implementations based on the cases mentioned above enable pulse shaping functions to be performed at the same bit rate [92]. Simulation results suggest that the flexibility of the design process based on the case of unequal tap gains and unequal delays allows for compensating efficiently for losses and active device parasitics, resulting in a transversal filter with higher gains when compared to the design based on the case in which only delays are made unequal.

A methodology was proposed in [24] by which filter delay and gain elements are determined to construct the output pulse. The samples of the pulse designed for achieving minimal ISI, are non-linearly related with the time variable; a sharp input pulse is shifted by a fraction of the pulse duration; i.e. $\tau_k$ seconds, and multiplied by the correspondent tap gain $G_k$. By using this sampling and delay technique the filter responses can be tailored to satisfy specific time and frequency functions.

For the implementation with identical filter tap gains, the distribution of the capacitance in the artificial transmission lines is constant, as in the case of conventional distributed amplifiers. Different delays are set in correspondence to the inductance of the interconnecting transmission lines. By selecting appropriately filter delays, the range of needed inductances can be less restricted by the practical limitations of transmission lines. A 40 GHz-HEMT MMIC transversal filter based on the forward-gain mode topology was implemented using the distributed transversal filter analogy. The number of filter tap gains elements was chosen to be four utilising identical enhancement-mode HEMTs. In experimental results, the raised cosine target was compared with the forward gain parameter $s_{21}$ of the filter showing good agreement in a wide range of frequencies [24].
4.10 Adaptive transversal-filter proposals

The analogy between distributed amplifier and transversal filter was established as a potential method to share optimality between the amplifier front-end and post-detection signal shaping filter. Incoming pulses can be amplified and reshaped satisfying the first and second Nyquist criteria using the aforementioned transversal filter structure [24,90]. The functionality of optical receiver can be further enhanced by making the design tolerant to processes variations and capable of counteracting changes due to receiver non-ideal responses. Additionally, this approach allows the filter integration with other components of the optical receiver [90]. These design aims can be accomplished in the multi-Gbit/s regime by distributed transversal filters in which circuit parameters can be adjusted to obtain a wide range of filtering functions. These structures are termed adaptive transversal filters and are useful components for high-speed short-haul optical communication systems [90,93-95].

Electrically tuneable transversal filter permits implementing a wide range of filtering functions. The capacity of tuning the transversal filter to obtain a specific response is achieved by means of applying external bias voltages to adjust the gain element of distributed cells. In order to obtain a wide range of filtering functions, the desired transmission characteristics of the artificial transmission lines must be maintained without change; i.e. facilitating the adjustment of transfer function while keeping constant the distributed circuit characteristics. Active transversal filter structures featuring similar gain level and shaping pulse capability have been produced by two different research groups [90,93-94]. Those designs are aimed to compensate for distortion in high-speed lightwave systems. An adaptive transversal filter specifically designed to handle soliton pulses was proposed in [90]. This receiver implementation presents suitable functional characteristics and is based on initial work by Moreira et al [91], in which a symmetrical response is obtained by using an implementation with an odd number of filter taps. The filter cell consists of a single active device in common source configuration with transconductance gains varied by applying external voltage in gate terminal. Blocking capacitors are introduced in the gate line to adjust separately tap gains. Figure 4.15 depicts a diagram of topology suitable for MMIC implementations [90].

A MMIC transversal filter based on this configuration was optimised by taking into account losses and device parasitic elements of the active devices. In order to enhance the filter functionality by allowing cell gain variation, the filter was optimised involving transistors areas transmission lines [90]. Additionally, transmission line sections were also optimised to achieve various filtering functions; filter responses were obtained by electrically tuning filter taps separately in excellent agreement with simulation results. Experimental results using the post-
detection filter in an optical link show an improvement of the bit-error rate for a specific detection threshold level verifying the potential of the transversal filter concept [97].

![Distributed filter structure with variable tap gains](image)

**Figure 4.15 Distributed filter structure with variable tap gains [90]**

When a wide range of filtering functions is required, larger bias voltage levels need to be applied so as to change the transconductance of the active device. In practical FETs, when different external bias voltages are applied, the change in the modulation extent of the depletion region alters other electrical characteristics, prominently the input capacitance of FET. Since each device is biased separately, the irregular capacitive loading in the gate-ATL leads to large phase differences of signals amplified by different FETs; variations of the capacitive loading thwarts the constancy of distributed circuit characteristics. Such variations result in different time delays that cannot be considered in the optimisation process. Consequently, tap delay between adjacent stages varies with the applied voltage and only limited tunability can be obtained by using this structure.

In a separate development, Freundorfer, et al [93,94] proposed an adaptive transversal filter based on cascode cells as variable gain blocks. In this approach, a MMIC transversal filter was designed to enhance the range of obtainable filtering functions and to extend the functionality to preamplification, automatic gain control and group delay control. Figure 4.16 (a) depicts the adaptive transversal filter scheme based on cascode gain cells.

The concept underlying the adaptability of transversal filters refers to capacity of the filter to preserve pulse shape integrity in all ATL whilst filter tap gains are adjusted to obtain a predetermined response. When appropriate, the impedance loading of distributed cells must be maintained practically without modification; otherwise the bandwidth limitation introduces phase distortion on artificial transmission lines which cannot be easily compensated by equalisation itself.
The transversal filter based on cascode gain blocks is essentially a distributed amplifier with individual cell gain variation capability, where the gain of cells is set by applying external bias voltages to the common gate stages. For the transversal filter, the reverse-gain mode configuration was used in which tap gains are adjusted separately. The reverse-gain mode configuration allows increasing the tap delay per stage; thereby reducing the required additional time delay per stage. The use of cascode stages has inherent advantages; cascode cell presents a strong internal mismatching, that of the high output impedance of the common-source stage and the low input impedance of common-gate stage, which permits achieving amplification in a wide range of frequencies [81,98]. The high output resistance provides good matching conditions to coupling the drain ATL, whilst the low impedance of common-gate stage reduces the Miller capacitance and gives good output / input isolation. In the filter implementation, the gain of individual stages can be varied continuously from a maximum to a minimum value as a function of the transconductance of the common-gate stage [93,94] (although experimental results shows only binary full or zero tap gain weights). Bias control is applied to the common gate stage (VGc voltage as shown in Figure 4.16(b)). Experimental results of a full packaged MMIC transversal filter for 5-Gbit/s applications and based on cascode cells were described in [94].

In spite of the high isolation in the input / output ports and other desired characteristics presented by the cascode cell, the input / output impedances varies as a function of the external voltage VGc. Small-signal analysis of the cascode cell was carried out accordingly by performing simulation of cascode cell at different bias conditions to corroborate its electrical characteristics. Figure 4.17 shows our simulation results of cascode cell impedance, 50 Ω resistance loads and a 0.2 μm gate HEMT with pinchoff voltage of −0.7 V.
Chapter 4 Distributed amplification and transversal filter concepts

![Graphs showing magnitude and phase of output and input impedances for different values of VGS.](image)

**Figure 4.17** Magnitude and phase (degrees) of the output (left) and input (right) impedances of the cascode cells, plotted for different values of VGS; the gate to source voltage of the common-gate stage

Results in Figure 4.17 are obtained from simulations of the cascode stage to illustrate variations in the magnitude and phase of the output and input impedances as a function of the bias voltage of the common gate stage. The results show that the large output impedance of the cell is greatly reduced for lower values of VGS, corresponding to lower transconductance gains. To a lesser extent, the input impedance also varies with the VGS. These observations imply that a transversal filter based on cascode cells should feature limited tunability due to variations of its input and output impedances that would result in variations in the characteristic impedances of its gate and drain transmission lines. In a recent paper [94], details of the implementation based on this structure were outlined in which the use of additional capacitive loading in the gate and drain ATLs (not shown in the Figure 4.16(a)) allows reducing the variations of the cell impedance. Nonetheless, by adding external capacitors, the gate and drain transmission lines tend to reduce the effective filter bandwidth giving rise to an excessive phase distortion if input pulses contain significant frequency components.
In addition to bandwidth limitations, the transient characteristics of pulses change as they travel through transmission lines sections; the rise-time of pulses is increased whilst the interstage delay remains constant, leading to larger changes of pulses and limited filter tunability. Those effects must be considered carefully in the design of the artificial transmission lines since in MMIC implementations the interstage delay is essentially introduced by (lumped) elements of active devices periodically distributed along the ATLs.

In a fairly recent paper, Freundorfer, A, et al. [25] reported the design of a MMIC transversal filter for duobinary signals, which has positive and negative tap weights along with the capacity of setting tap gain weights continuously from a maximum to minimum value. Therefore this equaliser has increased the passband control. The relevance of this filter concept will be subsequently highlighted and its salient features discussed given its similarities with the new class of transversal filters developed in following chapters of this thesis.

Summary

This chapter has provided an introduction to distributed amplifiers and transversal filters for use in high-speed optical communication systems. The chapter commences by briefly outlining active device technologies and a comparison of the advantages of HEMT over GaAs MESFET, figures of merit, parasitic capacitance, source resistance and transconductance. HEMTs tend to present better figures of merit with larger $f_T$ and $f_{max}$ than those of GaAs MESFET, in addition to better noise performance. HEMT processes have become a mature technology for a wide range of microwave applications. The distributed amplification concept has been outlined regarding the transmission characteristics that enable broadband operations.

The chapter describes the gain mechanism of distributed amplifiers through the concept of phase synchronisation. A distributed amplifier can be represented by two artificial transmission lines actively coupled via device transconductances. When the phases of both transmission lines are identical, the distributed amplifier is phase synchronised and wave components are added in phase at the output port. An amplified signal is obtained as a result of the additive contribution of all wave components. An important variant of conventional distributed amplifier is the matrix distributed amplifier. The matrix amplifier is an alternative structure to build high-gain broadband amplifiers combining the additive and multiplicative amplifying principles. An advanced optical receiver based on the integration of TWA and pin photodiodes sharing the same InP substrate was outlined. Integrated optoelectronics are presently one of the most promising technologies to achieve optical receivers working at speeds
of several tens of Gbit/s. This technological advance is a firm prospect to enable IM-DD optical links working at speeds beyond 100 Gbit/s.

The versatility of the distributed amplifier has been extended to other distributed circuits for high-speed lightwave systems. The analogy between distributed amplifier and transversal filter topologies was highlighted as a methodology to create distributed structures with different capabilities. Transversal filters have been designed to achieve amplification and filtering / shaping functions embedded in a single structure. On the other hand, some adaptive transversal filters permit extending the functionality to preamplification, automatic gain control and group delay control, while other adaptive structures have been specifically designed to share optimality between amplification and filtering functions. Those structures allow the MMIC design to be tolerant to processes variations and adapt to different operating conditions.
CHAPTER 5

Transversal filter topology for generation / filtering of high bit-rate sequences

In the previous chapter, wideband transversal filters constructed for multi-Gbit/s lightwave system applications have been discussed. Based on the distributed amplifier analogy, such filters were designed effectively as fractionally-spaced transversal filters [24, 90]. Appropriate gain and delay partitioning techniques enable transversal filters to satisfy specific time and frequency functions required in postdetection stages of optical receivers.

Communication techniques involving the use of rate sequences have been considered for the improvement of data transmission in long-haul optical communication systems. Generally, transmit and receiver functions can be realised by a high-speed transversal filter in which tap gain weights are set according to a specific discrete sequence. For instance, multilevel signalling schemes have been introduced for lightwave systems to reduce bandwidth requirement of the optical channels thereby overcoming chromatic dispersion [114]. Multi-level signals provide a method for transmitting data on optical carriers and reducing the bandwidth around the carrier frequency.

Spread spectrum systems are another example of a system that involves the use of specific discrete sequences (pseudo-noise codes). In this case, each pulse is impressed on a high-speed stream of chips so that user data is spread over a larger bandwidth when compared to that of user data information.
Chapter 5 Transversal filter topology for generation/filtering of high bit-rate sequences

Advances of high-speed optical components and systems have increased notably the speed operation of optical communication systems. Congruent with this tendency, the utilisation of electrical filters for the transmission/reception of high speed sequences is currently a viable alternative. Figure 5.1 shows an intensity-modulation direct-detection (IM-DD) optical link in which adaptive filters have been attached to transmitter and receiver; a narrow input pulse is encoded in an electrical waveform. At the receiver, a version of the encoded waveform passes through a filter so that narrow pulses can be detected by performing a corresponding correlation function. In such system, the transmitter and receiver have to present bandwidths of operation compatible with chip rates at which data is encoded. Transversal filters based on distributed amplifier principles possess appropriate frequency and transient characteristics to achieve encoding and decoding functions in the electrical domain.

![Figure 5.1 High-speed photonic system for the reception and generation of binary / multilevel sequences](image)

In this chapter a novel transversal filter is presented; the topology provides the capacity of positive and negative tap gain weight control and is suitable for the generation/reception of high-rate sequences. The potentialities and limitations of the filter approach, as discussed in [100,23], are investigated. Physical limitations of the available MMIC process limit the number of filter taps; thereby limiting the number of obtainable functions. Theoretical analyses are carried out to investigate factors that affect the bandwidth and the performance of such a distributed transversal filter. This chapter also serves as an initial study of the novel structure to assess its feasibility for MMIC implementations.

5.1. An initial proposal with complementary device lines

Different communications systems involve the use of transmit and receive functions defined by bipolar or multilevel discrete sequences. Non-coherent Optical-CDMA systems employ differential receiver configurations and complementary optical filters to encode and
decode user data [41]. Receivers based on unipolar filter and single photodiode are restricted to decode unipolar codes with the disadvantages associated with unipolar encoding of CDMA.

For an electrical post-detection receiver, a balanced structure can be constructed by using two photodiodes in differential configuration followed by an electrical distributed transversal filter as shown in the scheme in Figure 5.2. Such structure, however, results in reduced bandwidth as the input receiver capacitance is doubled because of the antiparallel photodiodes.

![Figure 5.2 Receiver based on an electrical distributed transversal filter](image)

An electrical wideband topology that can provide bipolar capacity using two separate photodiodes is shown in Figure 5.3 [94]. The input signal is equally split to feed two separate detectors at the input of each transversal filter. The current of one detector is inverted with respect to the other by changing the polarity and the bias supplied to the photodetector. Two separated gate and drain artificial transmission lines provide the delay between taps. A microwave power combiner (for instance, a Wilkinson combiner) can be used to effect the addition of outputs of the transversal filters. The amplitude and polarity of the output pulses depend on the difference of the gain (i.e. transconductance) of active devices.

![Figure 5.3 Basic receiver with bipolar capacity using broadband power combiner](image)
A receiver based on this structure has key drawbacks that have to be taken into account. First, its implementation requires halving the input signal power whilst the equivalent input noise of such receiver is doubled when compared to the single-line based receiver, as the noise in both sections is uncorrelated. If such transversal filter is to be employed as a postdetection stage, the need of amplifiers in inverted and non-inverted configuration makes the receiver cumbersome to implement. In addition, variations associated with the limited reproducibility in its fabrication give rise to undesired filter behaviour associated with changes of the electrical characteristics of active devices. Therefore, other topologies best-suited for MMIC implementations have to be investigated.

5.2 The triple-line transversal filter approach

To overcome the limitations mentioned above a triple-line transversal filter with a single input line is proposed [100]. The transversal filter proposal parallels the precursory microwave filter concept of Jutzi, [95] in that both transversal filter designs employ two rows of active devices. However; Jutzi's design is constructed using a common drain-ATL, an external distributed amplifier for signal inversion and microwave splitter to feed separate input gate-ATLs. Such an elaborate topology presents inherent disadvantages for MMIC design and consequently is of limited functionality for optical communication systems implementations.

A novel structure was envisaged by which the sign and weight of the cell gain can be set without compromising the impedance uniformity on artificial lines. The design approach avoids the need of microwave devices in inverting and noninverting configurations with different reflection and delay characteristics. The structure consists of two rows of active devices sharing an input common gate-ATL and each of them with separate drain-ATL. The structure is depicted in Figure 5.4.

![Figure 5.4 Triple-line distributed transversal filter (DC circuitry omitted for simplicity)](image-url)

Figure 5.4 Triple-line distributed transversal filter (DC circuitry omitted for simplicity)
Chapter 5 Transversal filter topology for generation/filtering of high bit-rate sequences

In the filter approach dc voltages applied to gate terminals permit the control of active device transconductances. The half lines at the input allows coupling signals from the generator to the common-gate line. Input signals travelling along the common transmission line are coupled to external drain transmission lines using the amplifying properties of the active devices. Signal components in each drain-ATL travel towards terminal resistances. The structure uses the reverse gain mode topology, creating signals paths on both drain ATLs with essentially the same interstage time delay [24].

An active device in common source configuration effects signal inversion. As it is shown in following chapters, the direct coupling of the active inverter in the drain-ATL does not impair the transmission characteristics that encounter coupled travelling signals providing that the inverter present high level of isolation. Signal coupled by the inverter to the lower drain line is equally split in propagating and counterpropagating components travelling along the drain line. Such components are linearly combined with travelling-waves coupled by other devices on the lower drain line. Travelling-waves are combined with the same time delay at the summation point enabling positive and negative tap gain weight capacity. In the following, the capacity of control gain weight and sign of filter taps is also referred to as bipolar capacity.

5.3 Active cell description

The design discussed above is composed of active cells which consists of two active devices in common source configuration; each cell is capacitively coupled to the common gate ATL. Series capacitors inserted in the common gate-ATL block external bias voltages, thereby enabling each element to be biased separately. Figure 5.5 depicts a detailed diagram of the distributed cell. Thin overlay capacitors are utilised in MMIC implementations. Capacitive coupling in the common gate line plays an essential role on the design and performance of distributed structures such as the one proposed here and will be thoroughly addressed in section 5.7. Figure 5.5 also shows additional bond-wire inductances used for connecting the active cell with off-chip dc circuitry.

It is normally the case in MMIC implementations that components that consume large amount of area are usually placed off-chip whilst bond-wire are utilised to interconnect devices in both sides. However, large bond-wires have to be avoided so as to reduce parasitic resonance with decoupling capacitors, which usually are placed off-chip [65]. For the transversal filter cell, the effect of bondwires can produce undesired effects in the performance and its characteristics have to be considered carefully. Generally, high-speed active devices present reverse gain parameter that increases with the frequency. At high frequencies, the presence of “feedback
elements" of the device reduces the stability margin of the device and the maximum stable gain parameter sets the practical limitation of the active device. Nonetheless, the interconnection of devices through bondwire inductances to bias the device can produce instabilities at lower frequencies; specifically if a large reactive load associated with bondwires is connected in the gate terminal of the device. A series resistance \( R_G \) with values of hundreds of kilohms is included to avoid instabilities as the reflection scattering parameter can be closer to unity in submicron devices [102]. Analyses carried out in chapter 7 shows that a designed transversal filter based on this arrangement is a stable circuit.

Figure 5.5 Distributed active cell with external bias

At dc operation the transversal filter consists of two independent rows of active devices connected together on a drain line. A representation of the dc circuit and a frequently utilised biasing scheme for distributed amplifiers are shown in Figure 5.6. An external bias-T is connected to the filter drain outputs to provide the drain current for all HEMTs. The decoupling the drain line terminating resistor, using series connected capacitor(s), results in the reduction of dc current flow in the transmission.

Drain transmission lines invariably carry bias current to feed the active devices. As a consequence, losses associated with transmission lines reduce the voltage level applied to devices connected in the same row. In Figure 5.6, external resistors connected across the drain to source terminals represent the static resistance of the device at different polarisation points. The dc current fed to each active device varies according with the level of gate voltage bias applied to obtain a specific transconductance, the equivalent load varies hence the current flowing on transmission lines. The resistance load associated with all active devices has to be maintained large enough to avoid a large drop of voltage in drain terminals of active devices.
An advantage that the dual-drain topology offers over single drain-line filter topologies is that of requiring lower dc current flowing on transmission lines [101]; the impact of this outcome is important in the design of transmission lines since it is less restricted by the current handling capacity permitting a reliable design for higher bandwidth operation. In addition the topology has the advantage of making the drain-ATL characteristic impedance less complex.

**Figure 5.6 DC equivalent representation of transversal filter, where resistances are added to represent output static device resistances**

### 5.4 Impedance loading on artificial lines

The transmission and reception of high rate sequences require wideband transversal filters in which the maximum number of taps is determined by intrinsic limitations of the process. It is important to design the filter topology so that the impedance (on the output and input lines) is maintained constant as this results in low losses per stage therefore would allow a high number of stages. Unlike classical distributed amplifier designs, the filter design criterion is based on the maintenance of pulse shape integrity rather than an optimum number of stages to achieve certain gain level.

The triple-line filter should be designed to provide impedance matching when the filter is tuned to different tap gain values. In the process of tuning the filter response by changing the active device transconductance, other device electrical characteristics will vary and eventually give rise to changes in the impedance level of ATLs. By providing sufficient drain voltage to the active devices, the output impedance is maintained high in relation to the characteristic impedance of ATL. Such impedance conditions facilitate the design of drain-ATLs with almost constant transmission characteristics. Conversely, the input capacitance and device transconductance depend greatly on the depletion region extent modulated by the gate bias voltage. The gate to source capacitance of practical FETs, which corresponds approximately to the input capacitance of the device, can increase its value at higher voltages by a factor up to three times the pinch-off capacitance value [102]. This will affect the filter performance and
compromises the efficiency of signal transmission in the gate line. From the above discussion, it is important to develop a technique that avoids impedance mismatch and parasitic delays in the gate ATL. The gain control technique is based on the existing relationship between the transconductance and input capacitance of active devices at different bias points [94,100], low load capacitance variations are expected by using appropriate bias voltages. The technique is outlined below in order to establish its significance in the performance of the filter proposal.

At frequencies below the filter cut-off frequencies a simplified equivalent circuit (loss-free circuit) is shown in Figure 5.7.

When cell active devices are biased at two different bias voltages, the equivalent capacitance of both active devices connected in parallel is approximately equal to \( C_{g,1} + C_{g,2} \). For frequencies below the ATL cut-off frequency with inductance per section \( L_g \), both time delay and gate-ATL impedance section are given by

\[
\tau_d = \frac{1}{\omega L_g} \left( C_{g,1} + C_{g,2} \right)
\]

and

\[
Z_{og} = \frac{1}{\omega} \left[ \frac{L_g}{C_{g,1} + C_{g,2}} \right],
\]

respectively. Hence, the impedance and time delay will undergo variations as bias voltages are adjusted to different transconductances.

Appropriate matching conditions could be attained on the gate line if both constituent devices of all distributed cells were biased at two gate bias levels because nodes of the input line are always connected to the gate of a transistor biased for a maximum transconductance and that of a transistor biased for a minimum transconductance, resulting effectively in a transmission line with equal capacitance loading. This corresponds to the case of binary weight control, however; it is highly desirable to provide a technique that allows for other tap gain weight adjustments. It is shown below that the proposed gain weight control can provide suitable matching conditions for analogue weight adjustment.
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5.5 Novel tap gain weight control

A novel tap gain weight control that provides constant matching conditions and constant delay in transmission lines is proposed. In order to show the feasibility of the continuous weight adjustment physical processes relative to device transconductance and input capacitance have to be considered. An important relationship between those device elements establishes a transconductance to input capacitance ratio $g_m/C_g$ constant and approximately independent of the depletion region extent [103]. The ratio corresponds to the current cut-off frequency $\omega_T (= 2\pi f_T)$ when the steady-state carrier velocity becomes saturated at high channel potentials [102, 103]. In practical processes, $f_T$ varies with external bias voltages since the device’s transconductance and capacitance have different variations with bias as a result of the dependence of the depletion region dimensions on both drain and gate bias. Nonetheless, first order approximations of MESFET and HEMT processes show a transconductance directly and linearly proportional to the input capacitance of the active device under amplification conditions and a wide range of bias voltages [103]. Such linear approximation between both devices characteristic becomes invalid at low gate voltages, where the transconductance tends to zero and the capacitance drops to its pinch-off value and also in the onset of degraded transconductance performance at high gate voltages [102].

![Intrinsic transconductance against gate to source capacitance and linear interpolation, (normalised to HEMT gate width)](image)

Figure 5.8 Intrinsic transconductance against gate to source capacitance and linear interpolation, (normalised to HEMT gate width)

In order to verify the existing linear relationship a commercially available HEMT process, which is described in Appendix A, was characterised using technique of parameter extraction and microwave circuit optimisation at different bias conditions. Results of the
parameter extraction technique are in excellent agreement with device s-parameters for a wide range of frequencies. Such results are shown in Figure 5.8; the HEMT transconductance, $g_m$ as a function of the gate-to-source capacitance, $C_{gs}$ and the linear interpolation are displayed in the figure. The plot shows that the linear relationship is well approximated except below the pinch-off voltage and at high transconductance values.

Analogue weight adjustment implies that individual constituent devices will present any combination of capacitance and transconductance described by the relationship shown in Figure 5.8. When both constituent active devices are biased at the same voltage level, tap gain weight $|g_{k,1} - g_{k,2}|$ will be zero as both devices present the same transconductance whilst the capacitance loading is twice the capacitance of each device at such conditions. However, when the tap gain weight is set to a value different from zero, both devices are biased for different transconductances. In accordance with the functional relationship of Figure 5.8, the capacitance of one device will be reduced whilst the other increased; this capacitance variation suggests that the combination of capacitances could be maintained constant presupposing a linear relationship. In the proposed technique, it is fundamental to set a reference point at which an appropriate bias voltage will set the same transconductance in both active devices thereby obtaining zero tap gain. The reference may be set at the midpoint of the range of transconductance values, so that positive and negative cell gain are equally limited by the irregular performance of the active device in the pinch-off region and at high bias voltages, in other words gain weight control can produce the same range of positive and negative filter tap gains.

The technique is established as follows: cell gain is set by taking the difference between the constituent device transconductances, $g_{k,1}$ and $g_{k,2}$ equally separated from the reference transconductance; the linear interpolation predicts an equivalent cell capacitance $C_{g_k,1} + C_{g_k,2}$ approximately equal to $2C_{g,0}$ for any cell gain $|g_{k,1} - g_{k,2}|$. Accordingly, $C_{g,0}$ is defined here as an average capacitance and corresponds to the input capacitance of the device when a reference voltage is applied so as to obtain a zero tap gain at the output. Since the equivalent capacitance is constant for different bias conditions, tap gain control can be achieved without modifying the transmission characteristics of the gate line. Equally variation of transconductance from a reference transconductance, $g_{m,0}$ which ensures almost constant capacitance loading conditions, is accomplished by chosen transconductances with values fitted to the linear equation:

$$g_{k,1} + g_{k,2} = 2g_{m,0} \quad (5.1)$$
Unlike the linear relationship between the transconductance and capacitance, the transconductance of the HEMT is not linear with respect to the gate-to-source voltage as shown in Figure 5.9(a). The application of the fitting Equation 5.1 results in a nonlinear weighting function against applied voltages as shown in Figure 5.9(b). A reference transconductance of 311.0 mS/mm was chosen as the slope of the transconductance function at such point permits a more constant variation of both bias voltages to set a specific gain weight.

The continuous transconductance function of the cell shown in Figure 5.9(b) was obtained by curve fitting and interpolation from the knowledge of the device characteristics displayed in Figure 5.9(a) and assuming both active devices are biased at different Gate to Source voltages. The graph was obtained from parameters of a HEMT process with maximum intrinsic transconductance equal to 680 mS/mm and threshold voltage equal to –0.9 V. For a reference voltage of -0.5 V, the transconductance results to be equal to 341.0 mS/mm.

![Figure 5.9(a) HEMT transconductance against VGS, (b) cell gain weight $|g_{m1} - g_{m2}|$](image)

5.6 Generation of differential signals

Generation of differential signals is necessary to achieve positive and negative gain weight control in distributed cells. Differential signals are often associated with balanced circuits whose branches have elements with similar electrical characteristics. The differential pair is a prominent example of a balanced circuit; differential signals are amplified whilst the pair is not sensitive to the signal that is common to both input ports due to the symmetry of the balanced circuit. The Gilbert cell is a balanced circuit and consists of a direct interconnection of differential pairs, the transconductance gains of external stages are varied as a function of applied differential signals; hence a Gilbert cell can be considered as an amplifier whose gain and phase can be changed by applying external control signals (see Chapter 7).

In spite of the fact that the filter cell design proposal does not satisfy the requirements of a balanced circuit, the structure enables differential signals to be induced in drain line and with
the same functional characteristics as those of Gilbert cell-based transversal filters. The tap gain weight control as described in section 5.5 has important implications for the circuit design. In order to have an insight into the filter response, distributed cells are modelled by the small-signal equivalent circuit shown in Figure 5.10. In the schematic, induced currents $\Delta g_k V_{g,k}$ flow into both drain lines in opposite directions; such currents have equal magnitudes and are controlled by the same voltage appearing across the gate junction of both devices. Since in the gain weight control design the incremental element $\Delta g_k$ corresponds to the device transconductance variation from the reference $g_{m,0}$, a differential signal is induced by an equal variation of device transconductance. In this arrangement, differential signals are coupled to external drain lines with equal magnitudes and phase difference of 180 degrees.

The understanding of the processes involved in the generation of travelling-waves in transmission lines requires wave-based analysis which is carried out in Chapter 7. However, initial assumptions can be made at this stage so as to describe the bipolar capacity of the structure. By disregarding at this stage the active inverter at the output port, the distributed structure can be considered as a three port distributed circuit as shown in Figure 5.11 (output ports are placed at the right hand side for clarity), where active devices are connected to simple external structures with identical transmission characteristics. Voltage sources $V_{dd1}$ and $V_{dd2}$ represent different bias conditions associated with different dc loading conditions in both row of active devices.

![Figure 5.10 Small-signal equivalent circuit of the filter cell](image1)

![Figure 5.11 Bipolar waveform generation of the distributed structure](image2)
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If each pair of active devices capacitively coupled to a gate-ATL is represented by its equivalence of Figure 5.10, the response to a stimulus can be described in terms of the superposition of a reference voltage component and a differential waveform on both drain lines. Figure 5.11 shows a voltage that appears across the resistance in both drain terminals and consists of a reference voltage component and differential signals varying across the reference component, the time duration of voltages depends on the filter time span.

By considering that signals possess differential components referred to each other and to a ground plane, an inverter (for instance a wideband balanced-unbalanced component) coupled to both drain lines as depicted in Figure 5.11 enables the reception of bipolar waveforms whilst eliminating the induced reference component. A similar consideration arises when an active inverter is included to combine linearly signals on the lower drain line. Such method of operation maybe useful to bias both arms of an external modulator. Travelling waves of both drain lines provide differential signals whilst reference component corresponds to the switch voltage of the external modulator. Multilevel signalling schemes with electrical encoding / decoding could be deployed using this principle. Nonetheless, there are issues associated with the application of such filter to work as a filter/driver to be taken into consideration. For instance, parasitic couplings between transmission lines and bandwidth limitations introduce variations in the reference signals, which eventually contribute to chirping of modulated optical pulses [111]. Therefore; it is sensible to propose a transmitter scheme that utilises the transversal filter followed by a driver controlling appropriate swing voltages and accomplishing good CMRR.

The aforementioned concepts will be applied to filter analysis to assess its capacities to receive and transmit high rate sequences. Further analysis show that the filter can effectively mimic the output port of a differential circuit as propagating waves can be characterised by common and differential-mode components. Mixed-mode scattering parameters suitable for the analysis of differential circuits at microwave frequencies [104,105] are appropriate for analysing transversal filter responses and will be drawn upon in Chapter 7 for such aim.

5.7 Capacitively coupled active cells

The bandwidth is a key performance parameter of the transversal filter. The main restriction on extending the upper limit of distributed circuit bandwidth is associated with the input capacitance of the active devices [80]. In the proposed structure such limitation is increased due to the common-gate ATL topology. It is apparent that active devices featuring large figure-of-merits have been employed in the development of such filter [99]. Given that FET processes with large figure-of-merit are not always available to the designer, different
methods for enhancing bandwidth were investigated and considered for their application in the proposed topology.

A straightforward and efficient method to improve the bandwidth is the use of series capacitors for coupling active devices to the gate line. This technique has been successfully used to boost the frequency span of solid-state distributed amplifiers beyond 100 GHz [66] whilst reducing the effect of attenuation associated with losses of active devices. MMIC processes have readily made available small capacitors (with values from few femto-farads) in smart libraries which take into account parasitic distributed effects in the calculation of their dimensions [65,63].

The application of the gate-line capacitively coupled technique was first proposed by Ayasli, Y. [54] as a method to increase the power gain of the distributed amplifier. The capacitively coupled technique also referred to as “capacitive voltage division” is based on the insertion of small capacitors in series with active devices. The effect of this series connection is a reduction (scaling) of the capacitance load at the ATL. This allows selecting devices with high gate width to improve the power gain of the amplifier. The power handling capability of the travelling-wave amplifier is improved by tapering the capacitance division ratio in the gate ATL. Initial series capacitors have small values whilst the last capacitors have higher values so as to achieve the same swing applied to all FETs. An intrinsic disadvantage of this method is that the inhomogeneous capacitive loading along gate line leads to phase differences of signals passing through different FETs. The design is aimed to compensate for phase differences by introducing stubs of different electrical characteristics in the output line, resulting in a non-uniform cell design.

For the transversal filter, the use of small series capacitances increases the bandwidth as discussed in [100]. The maintenance of the group delay by using the gain weight control allows a uniform cell design. Time responses are basically determined by the bandwidth and uniform sampling as shown in a later stage. The basic concept of the capacitively coupled active devices can be depicted using Figure 5.13, which represent an active cell of a travelling wave amplifier.

![Figure 5.13](image_url)

**Figure 5.13. The unit cell and voltage division in the cell,** $C_{gs}$ = input device capacitance

$C_1$ = series capacitance and $r_g$ = device resistance (after Ayasli, et al [54])
This method of connection forms a capacitance voltage divider by which only a portion of the RF voltage in the input line will appear to the gate junction. Figure 5.13 depicts the series connection of the capacitance with a simplified RC circuit representing the input impedance of active device. Figure 5.13 also depicts a representation of the cell where input capacitance and transconductance are scaled by a factor $M$. Such factor corresponds to the ratio of the portion of the voltage, $e$ appearing across the input capacitance to the node voltage of the gate line $V$. The equation for the voltage divider is given by [54]:

$$M = \frac{e}{V} = \frac{C_1}{C_1 + C_{gs}}$$  \hspace{1cm} (5.2)

This capacitive division factor is well approximated if the voltage drop across the parasitic resistance is negligible; the voltage divider is independent of the frequency for $\omega C_{gs} r_g << 1$ [54]. Therefore, the bandwidth of a travelling-wave amplifier can be increased in proportion to the effective capacitance of the equivalent circuit. For a lossless gate-ATL of impedance $Z_{og}$, the cut-off frequency is given by the equation:

$$f_c = \frac{1}{\pi Z_{og} M C_{gs}}$$  \hspace{1cm} (5.3)

As a consequence, the insertion of the small capacitances in the gate line scales the gain and capacitance by the same $M$ factor [54].

### 5.7.1 Second order effects

For the transversal filter, the insertion of small capacitors to enhance the bandwidth has to be carefully considered given the complexity of the common gate-ATL. Since the functional characteristics described above undergo deviations as the active devices' capacitances change with applied bias voltages, the gate-line presents different transmission characteristics. At high-frequency this should result in a different cut-off frequency of the two devices in a single cell. However, in the filter bandwidth, changes of the active device cut-off frequency can be is considered here as second order effects as shown below. An equivalent small signal model of the active cell is depicted in Figure 5.14, where $C_{gs,1}$ and $C_{gs,2}$ are the input capacitance of the device and $R_g$ is the device input resistance. The node voltage $V$ is unequally divided in both
devices appearing as RF voltages \( V_{g1} \) and \( V_{g2} \) across the gate junction of each of the devices.

It is assumed that the voltage drop across the device resistance, \( R_g \) is negligible and the voltage divider is independent of the frequency.

\[
C_{\text{div}} = C_d S_{\text{rx}} R_{ds}
\]

Figure 5.14 Small-signal equivalent model of the cell

The relationships for the voltage division ratio for both constituent active devices are given by:

\[
M_1 = \frac{V_{g1}}{V} = \frac{C_{\text{div}}}{C_{\text{div}} + C_{g1}} \quad M_2 = \frac{V_{g2}}{V} = \frac{C_{\text{div}}}{C_{\text{div}} + C_{g2}}
\]

(5.4)

In particular, when both active devices are biased at the same reference voltage, the same portion of RF voltage appears in both devices. The capacitance division is defined as:

\[
M = \frac{C_{\text{div}}}{C_{\text{div}} + C_{g0}}
\]

(5.5)

where \( C_{g0} \) is the input capacitance of the FET when the reference voltage is applied to obtain zero tap gain. The capacitance load of the cell is equal to \( 2MC_{g0} \). However; when different bias voltages are applied to both constituent devices, the equivalent cell capacitance load depends on the capacitance \( C_{\text{div}} \) and the capacitance of constituent devices, the capacitance load of the cell is given by:

\[
C_T = M_1 C_{g1} + M_2 C_{g2}
\]

(5.6)

In order to show the capacitance load variation, the available HEMT (Appendix A) model under different bias voltages was utilised; for such aim the divisor capacitor \( C_{\text{div}} \) is set to be equal to the HEMT capacitance at the reference voltage \( \text{VGS} = -0.5V \); the \( M \) factor is equal to 0.5. Figure 5.15 depicts the capacitance load as a function of two external voltages.
Simulation results show an almost constant cell capacitance for different applied voltages. The capacitance load, $C_T$, results to be equal to $2MC_{g,0}$ with an error up to 7 percent. Another second order effect is related to the fact that the RF voltages ($V_{gs,1}$ and $V_{gs,2}$, see Figure 5.14) become unequal because of the voltage division effected by $C_{div}$. By considering the voltage dividers independent of the frequency, the actual transconductance of the cell and the approximation supposing equal RF voltages in both gate junctions of devices are:

$$G_m = \left| g_{m,1}M_1 - g_{m,2}M_2 \right| \approx M\left| g_{m,1} - g_{m,2} \right| \quad (5.7)$$

Figure 5.16 displays the variation of $M$ factors for both devices, showing that the approximation is maintained in the range of bias applied to both constituent elements; the error of the approximation is reduced for bias voltages close to the reference $VGS=-0.5V$. 

**Figure 5.15 Variation of capacitance loading of the cell, $M=0.5$**

![Figure 5.15 Variation of capacitance loading of the cell, $M=0.5$](image)

**Figure 5.16 Variation of voltage dividers, $M=0.5$**

(a) (b)
Figure 5.16 (a) shows the case when a constituent active device is biased to set a large transconductance. The input capacitance is increased resulting in a decreased amount of voltage coupled to the gate junction. Conversely, Figure 5.16(b) shows the case when the device is biased to set a low transconductance voltage. The input capacitance is reduced which increases the RF voltage coupled to the gate junction. Given the adequate approximation to a bias-independent voltage divider, the gain of the cell can be well approximated by Equation 5.7; the cell gain is equal to the difference between the transconductance (see Figure 5.9) scaled by the factor $M$.

5.7.2 Small-signal model of the cell

The variation of input capacitance with applied voltage constitutes the most important second order effect in the filter design. Regarding the gate line, parasitic elements such as input resistance undergoes low changes with applied voltage and can be considered as a bias-independent parasitic element. The small signal model of Figure 5.14 can be approximated by the model in Figure 5.17. The common-gate line section consists of a single branch of elements with a capacitance load, $2C_{g,0}$ and resistance $R_g/2$. This equivalence does not need further approximations than those mentioned above because in practical FETs the voltage drop across the parasitic resistance is negligible. In addition, the inequality $\omega C_g R_g << 1$ implies a voltage division independent of the frequency on both branches. This allows for approximating the equivalent resistance by a half of the input resistance of each device due to the parallel connection of active devices.

![Figure 5.17 An equivalence of the small-signal model of Figure 5.14](image-url)

Other effect not considered in the equivalence is the transconductance delay of active devices. Nonetheless, the intrinsic delay does not present large variations at different bias
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conditions (see Appendix A) and such constant delay can be neglected for practical conditions. The cut-off frequency of the gate transmission line depends on the capacitance per section of the transmission line, which can be considered approximately equal to $2MC_{g,0}$.

5.8 Small-signal analysis of the distributed filter design

Classical distributed amplifiers are designed for a constant gain amplitude over multi-octave bandwidths in which losses associated with the active devices limit the number of active devices and hence the level of gain. Practical restrictions of the circuit design establish that in order to satisfy a constant gain over the bandwidth a linear phase-frequency relationship cannot be satisfied simultaneously. As a direct result, pulse dispersion arises as a main form of distortion. Given the distributed nature of the transversal filter approach, the artificial lines of the transversal filter possess functional similarities with those of distributed amplifiers. Therefore, a design methodology has to be considered from which the amplitude and phase characteristics can be realised to maintain pulse shape integrity. The trade-off between amplitude and phase responses is analysed based upon the characteristics of the devices employed. Straightforward small-signal analysis permits gaining an insight into the effect of constituent elements (such as active devices, series capacitance and characteristic impedance) on the characteristics of the resulting distributed circuits.

A fundamental consideration must be made regarding the influence of device parasitic elements in the performance. For example, in HEMT or MESFET, the shunt feedback associated with the gate-to-drain capacitance, $C_{gd}$, the series feedback associated with the source terminal, $R_s$ and gate electrode inductance, $L_s$ introduce passive couplings that have a direct influence on the characteristic of the propagation modes in both drain and gate ATL [70, 71]. The effect of such “feedback elements” on the overall performance can be accurately accounted for by using the Cascaded Four-Port Matrix Method (CFPM), permitting a complete analysis of the propagation modes in both transmission lines of distributed amplifiers [71]. However, the utilisation of such accurate technique involves high complexity that potentiality makes the method difficult to apply for distributed circuits. Therefore, parasitic couplings between transmission lines are neglected in the analysis.

A manageable yet useful method adopted for small-signal analysis is the Image Transfer Function, ITF [18,70]. In the ITF technique, unilateral device models fitting the electrical characteristics of the device up to high frequencies have to be utilised for both accuracy and reasonable complexity. Specific unilateral models oriented to distributed amplifier designs such as those proposed in the open literature [106,107] can be used in the ITF technique so as to
obtain analytical close-form expressions. Initial analyses shown below utilise such unilateral models to get an insight into the filter bandwidth and transmission characteristics.

The transversal filter consists of common-gate and drain transmission lines having different transmission characteristics. Figure 5.18 shows the drain and common-gate line transmission lines, which are obtained by substituting the unilateral model in the triple-line transversal filter. The second drain transmission line is not depicted for simplicity. The coupling mechanism in the unilateral interaction of active coupling comes from the transconductances distributed along the drain line. The common gate has a capacitance load and parasitic resistance according with the equivalence discussed in Figure 5.17. In addition, a zero reflection at ports terminated in image impedance basis is assumed without reducing accuracy at low and mid band operation.

It is important to note in Figure 5.18 that the controlled current sources distributed along the drain line are subscripted in correspondence to wave paths formed in the image terminated transversal filter and not in sequential fashion in reference to the induced voltages on the gate line. The theory of propagating functions is applied to obtain the filter transfer function. Appendix B provides a detailed derivation of the transversal filter function using the image propagation factors of the filter stages and matched terminal impedances. By considering identical propagation coefficients in gate line sections and equal mid-shunt propagation factors in both lines, the resulting transfer function of the filter with N-identical stages is given by (Equation B.14):

\[
\frac{V_{out}}{V_{in}} = -\frac{1}{2} Z_{zd} \exp\left(\frac{\theta^g + \theta^d}{2}\right) \times \sum_{k=1}^{N} g_k \exp\left(-k\left[\theta^g + \theta^d\right]\right)
\]

(5.8)

with

\[
Z_{zd} = \sqrt{\frac{Z_{zd}^g}{Z_{zd}^d}} \sqrt{\frac{Z_{zd}^g Z_{zd}^d}{Z_{zd}^g Z_{zd}^d}}
\]

(5.9)

where the impedances and propagation constants are superscripted referred to the gate or drain line accordingly.

5.8.1. Propagation parameters of the common-gate line

The presence of resistive elements in both transmission lines as well as the series capacitance plays an important role in the response of the transversal filter. The transmission characteristics of the filter can be analysed by using conventional concepts of circuits matched in image impedance as described in section 4.4. By modelling device sections as shown in Figure 5.18, the equation of the propagation constant is given by [69]:

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\[ \theta_g = \frac{(\omega_c/\omega_g)(\omega/\omega_c)^2}{\sqrt{1 - (1 - (\omega_c/\omega_g)^2)(\omega/\omega_c)^2}} + j\cos^{-1}\left(1 - \frac{2(\omega/\omega_g)^2}{1 + (\omega/\omega_g)^2}\right) \] (5.10)

with the angular frequency normalised to the cut-off frequency \( \omega_c \) and the constant \( \omega_g \) given, respectively by:

\[ \omega_g = \frac{2}{R_g MC_g} \quad \omega_c = \frac{1}{Z_{0g} MC_c} \] (5.11)

Based on equation 5.10, the transmission characteristics can be evaluated. The real part of \( \theta_g \) describes the attenuation per section of the gate line periodically loaded with characteristic impedance \( Z_{0g} \) given by:

\[ Z_{0g} = \sqrt{\frac{L_g}{2MC_g}} \] (5.12)
At low frequencies, the real part of equation 5.10 can be used to obtain an approximation of the loss per section $\exp(-\alpha)$ introduced by the gate line. The approximation is:

$$\exp(-\alpha) \cong \exp(-4\pi^2 f^2 M^2 C_g^2 Z_{0,g} R_g / 2) \quad (5.13)$$

Equation 5.13 shows that the gate line can be considered as a lossy line with frequency-dependent attenuation. In addition, the frequency dependent attenuation can be reduced by using a small factor $M$ to such a level that the effect of losses of active devices can be neglected in the performance.

The imaginary part of the propagation constant given by equation 5.10 corresponds to the phase coefficient of the gate line. The variation of the phase response with the frequency is an indicative of pulse dispersion and becomes a critical effect to be considered in the filter design.

Figure 5.19 displays the group delay derived from the imaginary part of the propagation constant $\theta_g$, using the dissipative factor, $R_g / Z_{0,g}$ equal to 0.1 and two different M factors.

![Figure 5.19 Group delay derived from imaginary part of $\theta_g$, for $M = 1.0$ (left) and $M = 0.5$ (right)](image)

From Figure 5.19, it is clear that the use of small capacitance ratios reduces pulse distortion introduced by the line at such level that the group delay can be considered approximately constant, hence $t_d \times \omega_c \cong 1$. The phase constant of the gate line with these characteristics can be considered approximately independent of the frequency and constant group delay per section, $t_d$ along the bandwidth.
5.8.2. Propagation parameters of the drain line

The drain transmission line as depicted in Figure 5.18 is considered as a lossy line. Unlike the gate line, the resistive element $R_{ds}$ is connected in parallel with the capacitance and transconductive source having other transmission characteristics when compared with the gate line. The image propagation function of the drain transmission line is given by [69]:

$$\theta_d = \frac{\omega_d / \omega_c}{\sqrt{1 - (\omega / \omega_c)^2}} + j \cos^{-1} \left( 1 - 2(\omega / \omega_c)^2 \right)$$  \hspace{1cm} (5.14.1)

with $\omega_c$ equal to cut-off angular frequency and $\omega_d$ an angular frequency constant given by:

$$\omega_d = \frac{1}{R_{ds} C_{ds}}, \quad \omega_c = \frac{2}{\sqrt{L_d C_{ds}}}$$  \hspace{1cm} (5.14.2)

At low and medium frequencies, the drain phase function, $\text{Im}(\theta_d)$ possesses similar characteristics to the gate phase function if both line cut-off frequencies are made equal by using either additional capacitance in parallel or a series stub in the drain line. Therefore, the group delay and phase constant introduced by drain line stages can be considered similar to that of the gate line. In addition, the complex phase constant does not depend upon the shunt resistance, $R_{ds}$ given drain lines with low dispersion characteristics. On the other hand, the drain attenuation coefficient can be considered frequency-independent at frequencies below cut-off frequency. The attenuation per section has the approximation:

$$\exp(-\alpha) \equiv \exp\left( -Z_{od} / 2R_{ds} \right)$$  \hspace{1cm} (5.15)

Although the above equations related to the drain line provides an insight into the transmission characteristics, the description of the drain lines is actually more complex. In spite of the fact that the drain lines are terminated with image impedances, the drain impedance is not constant within the bandwidth of the transmission line. The existence of the shunt resistance $R_{ds}$ in the drain line introduces a reactive component that reduces the power coupled to the drain line at low frequencies. The impedance in the drain line is approximated by [69]:

$$Z_{od}^d \equiv \frac{L_d}{C_{ds}} \times \frac{1}{\sqrt{1 - j \omega_d / \omega}}$$  \hspace{1cm} (5.16)
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Such irregular low frequency behaviour is well known in solid state amplifiers and is an intrinsic limitation of the resulting drain transmission lines [69]. This is extended to the triple-line topology. At low frequencies the drain lines present strong impedance mismatch at the output ports and the performance is expected to be deteriorated. Nonetheless, as it is shown in a later stage many practical filter responses of interest can be obtained with low deviation at low frequencies.

The analysis of the transversal filter with embedded inverter is proposed in Chapter 7 and facilitates the description of the functional characteristics of such filter. At this stage, an expression for the filter transfer function can be obtained by assuming the electrical characteristics of the sections described above. Based on the modelling of filter sections terminated in image impedance basis, the analysis is extended for the dual drain line structure formed by two identical transversal filters coupled along the gate line and terminated with image impedances. By assuming that all transmission lines have identical cut-off frequency and delay per stage, a linear superposition of waves of both external drain transmission lines is obtained at the output port. Using Equation 5.8 for both terminal drain outputs, it is possible to arrive to the transfer function of the filter given by:

\[ H_f(j\omega) = \frac{Z_{sd} M}{2} \exp\left(\alpha(\omega) + j\omega 2t_d / 2\right) \times \sum_{k=1}^{N} (g_{k,1} - g_{k,2}) \exp\left(-k[\alpha(\omega) + j\omega 2t_d]\right) \]

with attenuation coefficient \( \alpha \) per filter stage equal to:

\[ \alpha(\omega) = \omega^2 M^2 C_{g,0} Z_{sd} R_g / 2 + Z_{sd} / 2 R_{ds} \]

and filter impedance with drain characteristic impedance \( Z_{sd} \) approximated by:

\[ Z_{sd} \approx \frac{Z_{sd}}{\sqrt{1 - (\omega/\omega_c)^2}} \]

The filter transfer function (Equation 5.17) describes distinct propagating characteristics that encounter travelling signals on different filter paths. From the above results, it is apparent that active devices featuring low dissipative effects can improve the response or when appropriate, frequency dependent attenuation can be reduced by using small \( M \) factors at such a level that pulse dispersion could be negligible. From such analysis, a filter may be designed whose performance is restricted by secondary loss mechanisms while attenuation can be counteracted by setting appropriate filter tap gains. Other issues such as the input pulse shape
Chapter 5 Transversal filter topology for generation/filtering of high bit-rate sequences

integrity on artificial lines determine significant aspects to be taken into account in the design and will be completely addressed in following chapters.

Summary

Transversal filter structures based on distributed transversal filters were proposed and discussed for their application in the context of optical communication systems. Two single-row transversal filters can be used for generation/reception of high rate sequences, but its implementation is costly and inefficient mainly due to the noise penalties intrinsic to the method of coupling optical power as well as by the level of noise at the output port. A novel transversal filter proposal was described as a viable option for the transmission and reception of high bit rate sequences. Distributed cells are coupled to a single gate line via small discrete capacitors. Impedance mismatch of filter cell and bandwidth operation were investigated. The maintenance of capacitance per section allows avoiding parasitic delays of signals while allowing a uniform filter design.

A novel active tuning technique especially designed for the transversal filter was proposed. The foundation of this technique lies on the linear approximation between the transconductance and input capacitance observed in the modelling of active devices under amplification conditions; hence the technique rests on solid physical bases and is proposed as a principle of operation of such filter. The gain weight control enables the propagation of reference signals and differential-mode signals associated with variations induced by changes in transconductance of active devices. Towards the end of the chapter, the filter transfer function of such filter was obtained using image transfer function technique, ITF. This function describes inhomogeneous propagating characteristics that encounter travelling waves at different filter paths. Analyses show that pulse dispersion can be controlled at the expense of employing small voltage capacitance divisions. This enables the design of transversal filter with flat group delay. In addition, frequency-dependent attenuation effects can be reduced by using small capacitive division ratios and losses on gate-lines.
CHAPTER 6

GaAs MMIC distributed transversal filter design

This chapter presents the design of a distributed transversal filter for the generation / reception of high rate sequences. The transversal filter is aimed to handle high rate pulses; the transient characteristics of input pulses applied to the transversal filter for sequence generation and filtering are compatible with pulse generation in the picosecond region available in high-speed electronic systems [115]. The filter presented in this chapter is designed to maintain pulse shape integrity. Short input pulses are delayed and weighted to the corresponding filter tap gain, pulse transient characteristics vary without compromising the receive and transmit functions. In the filter approach, the filter tap delay is made equal to the symbol rate.

Transversal filters based on distributed amplification allow pulse shape, gain control and other base band functions for very high speed applications. In particular, recently reported filters with bipolar capacity have been constructed and tested to achieve equalisation for high-speed lightwave systems. For instance Lee and Freundorfer [25] reported equalisation at bit rates of 2.5 Gbit/s using a 5-tap transversal filter; the tap delay was set to 40 ps which is shorter than the width of the input pulses. Therefore; the filter was constructed as a fractionally-spaced transversal filter. In a more recent development Wu, et. al. [116] demonstrated equalisation at chip rates approaching the filter cut-off frequency and beyond; a 7-tap fractionally-spaced transversal filter was utilised with a similar cell topology to that in [25]. Those examples show that it is feasible to use distributed transversal filters to achieve integrated equalisers for high-speed fibre-optic systems. This chapter is concerned with the practicality of a transversal filter featuring larger tap delay to perform transmit and receive functions at multi-Gbit/s. A 7-tap transversal filter for 40 Gbit/s applications is implemented following the rules of a
commercially available HEMT process. The chapter commences by establishing the requirements of filtering of short pulses and the functional capabilities of distributed transversal filters so as to allow the handling of high rate sequences. Following this, the triple-line filter topology, which was introduced in Chapter 5, is implemented in MMIC form using a commercially available process. The figure of merit and the electrical characteristics of the utilised Pseudomorphic-HEMT (PHEMT) process are obtained for a wide range of frequencies using the parameters of the process. The transistor geometry and the series capacitance are taken as the design variables, where the selection of such variables is accomplished following a trial-analysis and redesign process to achieve a symbol-rate transversal filter. The feasibility of constructing delay lines, suitable for on chip implementation, is described. Pulse shape integrity is accomplished by establishing the trade-off between amplitude tap gain and phase linearity. The associated limitations of the transversal filter topology based on the available process will be discussed. Towards the end of the chapter, the layout of the transversal filter following the rules of the GaAs MMIC process is presented.

The majority of this chapter is devoted to the design of the transversal filter, performance analyses from which the transversal filter concept can be assessed are given in the following chapter. Results described in this chapter have appeared in [99,100].

6.1. Filtering of short pulses; design considerations

MMIC transversal filters based on the analogy between distributed amplifier and transversal filter topologies consist essentially of active devices and transmission lines interconnected to perform a prescribed filtering function. Active device tuning allows a wide variety of filtering functions to be obtained. The propagation characteristics of artificial transmission lines are determined by the lumped parasitic elements of the active devices as well as by the impedance and electrical length of the transmission lines. In this chapter the design of the actively tuned transversal filter is carried out. Impedance matched delay circuits are introduced between consecutive active cells to obtain the necessary delay; the inverse-gain mode topology is utilised to increase the tap delay by designing the gate artificial transmission lines with similar delay and impedance characteristics [100].

In wideband filters aimed to preserve pulse shape integrity, a flat group delay is not the only condition to ensure pulse shape integrity. In addition, an appropriate amplitude gain function is required to ensure a specific transient response. An initial approach to the symbol-rate transversal filter encompasses a differentiation of components that determines the pulse shape characteristic from those associated with the generation of pulse pattern with amplitudes weighted to filter coefficients. Figure 6.1 depicts a block diagram that represents a symbol rate
transversal filter. Input pulses are sampled and in turn multiplied by positive coefficients corresponding to the gains of the active devices. Amplified signals are linearly combined in wideband external structures of both filter lines; the resulting signal of one line is summed in anti-phase with signals of the other line with the same delay characteristics. A low band pass filter with transfer function $H_p(j\omega)$ is included to represent the intrinsic frequency limitation of structure.

![Diagram of the 1/τ symbol-rate transversal filter](image)

**Figure 6.1 Block diagram of the 1/τ symbol-rate transversal filter**

The transfer function of the overall transversal filter represented in Figure 6.1 can be expressed as the product of the low pass function, $H_p(j\omega)$ and the overall shaping function as given below.

$$H_F(j\omega) = H_p(j\omega) \times \exp(j\omega \tau/2) \sum_{k=1}^{\infty} (g_{k,1} - g_{k,2}) \exp(-j\omega k \tau) \quad (6.1)$$

The composite transversal filter function of the block diagram assumes that all signals travelling in the different filter paths are exposed to the same bandwidth-limiting mechanism. This assumption is to some extent inaccurate since frequency analysis derived in the previous chapter shows that delayed pulses undergo different attenuation and dispersion characteristics. Hence the function $H_p(j\omega)$ approximates the bandwidth limitation of all transversal filter taps. The design of the symbol-rate transversal filter ought to consider the two aspects below:

1. The transversal filter is designed to maintain low levels of amplitude and phase distortion. The filter function $H_F(j\omega)$ is actively tuned to a target function defined by a finite impulse response function in the frequency domain. The discrepancies between the target function and the filter response are mainly associated with attenuation levels which can be counteracted by adjusting each filter tap gain.
2. The auxiliary filter function, \( H_p(j\omega) \) permits approximating the bandwidth limitation of all filter taps. Nonetheless, the dissimilarities in the transient characteristics of output pulses are mainly associated with the inherent bandwidth limitation of the transversal filter, in particular with the different dispersion and frequency-dependent attenuation levels in filter paths. In the implementation of a specific filtering function, tap gain weights are adjusted to obtain a pulse pattern whose amplitude correspond to the coefficients of the finite impulse response function. In this application, tap gain weight adjustment enables also compensating for the inter pulse interference associated with the transient oscillatory component.

The filter described in this chapter can be considered as symbol rate transversal filter since each delayed pulse is amplified by a single cell allowing a level of inter pulse interference for practical reasons [118]. The description of the filter and its time and frequency domain simulation are carried out in Chapter 7.

6.2. Description of the HEMT process

The main design goal is to achieve wideband operation with the additional requirements of handling ultrashort pulses. The utilisation of active devices with large transconductances and improved figures of merit is highly desirable for the symbol-rate transversal filter implementation since each delayed pulse is predominantly amplified by a single filter cell. A HEMT processes was chosen for the implementation since HEMT devices have larger transconductances and better figures of merit when compared to other device technologies. In addition, the utilisation of HEMT permits establishing suitable matching conditions to the distributed cell design [100,118]. A pseudomorphic-HEMT with a GaAlAs active layer was utilised for the implementation. Figure 6.2 shows the layer structure of the commercially available process. In general, PHEMT processes presents larger transconductances (1160 mS/mm as a typical figure in enhancement mode transistor [119]) than those of conventional HEMT processes and better noise figure due to the improvement of transconductance and reduced device input capacitance [119,65]

Heterojunction FETs are developed by introducing wide band gap materials beneath the gate so as to provide the donor and carrier concentration required for amplification [119], III-V compound semiconductors with high saturation velocities have made possible to fabricate submicron-gate-length devices featuring low intrinsic capacitances and large transconductance. In particular, PHEMT processes are implemented by utilising the GaAlAs/GaInAs/GaAs layers grown on a GaAs semi-insulated substrate. A pseudomorphic channel is created by introducing a very thin GaAlAs layer (30Å approximately) with different lattice number from that of the interfacing GaInAs layer. This results in a strained channel with a large conduction band
Chapter 6 GaAs MMIC distributed transversal filter design

discontinuity; larger bandgap differences between both material systems increases the higher concentration of two-dimensional GaAs and hence device transconductance. In addition, the lower bandgap of the GaInAs compound allows for a better carrier confinement at the back side of the interface, which results in a larger potential and a lower output conductance [119,65].

![Epitaxial layer structure of the pseudomorphic-HEMT process](image)

Figure 6.2 Epitaxial layer structure of the pseudomorphic-HEMT process

The commercially available GaAs MMIC process is developed by OMMIC, a manufacturing group of Philips. The HEMT process employed in this thesis is named ED02AH that stands for specific features of the device. Appendix A provides a detailed description of this process and its electrical characteristics. The ED02AH process was developed specifically for microwave applications up to the millimetre wave region; it is based on a 0.2 μm pseudomorphic-HEMT process either as depleted or as enhanced mode field-effect transistors. The equivalent small-signal model from OMMIC is depicted in Figure 6.3 [63].

![Small-signal equivalent model of the PHEMT process](image)

Figure 6.3 Small-signal equivalent model of the PHEMT process

The ED02AH process has been utilised for high-speed optical network interfacing. OMMIC-Philips has announced a 0.18 μm metamorphic-HEMT processes with $f_T$ of 160 GHz as a part of its production technologies for the near future. This process in not commercially available yet.
Chapter 6 GaAs MMIC distributed transversal filter design

Based on the technological parameter of the process, the author applied optimisation routines to obtain the small-signal parameters of the circuit model. Table 6.1 itemises the extracted intrinsic parameters of a HEMT with gate width of 4×15 μm biased at VDS=3.0 V and VGS=0 V. Table 6.2 itemises the extrinsic parameters of the process scaled to 60 μm gate width. Appendix A provides the details of the scaling and parameters extraction of the available HEMT process.

Table 6.1 Intrinsic parameters of the HEMT (4×15 μm, VDS=3V, VGS= 0V)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$G_m$</td>
<td>41.00 mS</td>
</tr>
<tr>
<td>$C_{gs}$</td>
<td>60.96 fF</td>
</tr>
<tr>
<td>$R_{gs}$</td>
<td>2.5 Ω</td>
</tr>
<tr>
<td>$C_{gd}$</td>
<td>9.88 fF</td>
</tr>
<tr>
<td>$T_d$</td>
<td>0.288 ps</td>
</tr>
<tr>
<td>$C_{ds}$</td>
<td>13.7 fF</td>
</tr>
<tr>
<td>$R_{ds}$</td>
<td>470.8 Ω</td>
</tr>
<tr>
<td>$R_{gd}$</td>
<td>2.2 Ω</td>
</tr>
</tbody>
</table>

Table 6.2 Extrinsic parameters

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$R_s$</td>
<td>11.25 Ω</td>
</tr>
<tr>
<td>$C_g$</td>
<td>6.0 fF</td>
</tr>
<tr>
<td>$L_s$</td>
<td>12.0 pH</td>
</tr>
<tr>
<td>$R_d$</td>
<td>11.25 Ω</td>
</tr>
<tr>
<td>$C_d$</td>
<td>0.67 fF</td>
</tr>
<tr>
<td>$L_g$</td>
<td>12.0 pH</td>
</tr>
<tr>
<td>$R_g$</td>
<td>0.66 Ω</td>
</tr>
<tr>
<td>$C_s$</td>
<td>1.01 fF</td>
</tr>
<tr>
<td>$L_d$</td>
<td>12.0 pH</td>
</tr>
</tbody>
</table>

Table 6.2 shows extrinsic parameters which are gate-width dependent, whereas the intrinsic parameters of Table 6.1 depend on both the gate width and bias voltages. It is worth noticing from Table 6.1, that the input capacitance to gate-drain feedback capacitance ratio $C_{gs} / C_{gd}$ results to be equal to 6.0. It is shown in [67] that in conventional structures, with gate and cross section profiles developed in traditional HEMT process, $C_{gs} / C_{gd}$ ratios can be of the order of 20. Therefore, regarding the shunt feedback capacitance, the available HEMT does not compare favourably with other popular processes. In addition, the factor $C_{gs} / C_{gd}$ alters noticeably the device performance at high frequencies and acquires more influence as the figure of merit of the device [61]. Other feature of the employed HEMT process is the source resistance $R_s$, scaled to the gate width, equal to 0.54 Ω•mm; this resistance becomes large when low gate width HEMT are implemented for specific applications.

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* Due to the need of counting upon accurate small signal models for analysis and design, the author tested small-signal analysis models given by the foundry. Some discrepancies were found and that led to the re-extraction of small-signals parameters by optimisation (Appendix A).
Chapter 6 GaAs MMIC distributed transversal filter design

The above electrical parameters are heavily associated with the recessed channel technology adopted in the development of the HEMT process. In fabrication, the active channel thickness is less restricted by tolerances of the device dimensions. Consequently the requirements for etching under the gate region and electro-beam lithography are reduced [63]. In terms of performance, a recessed HEMT possesses a large source resistance because the effective distance between source to gate is increased; in contrast with self-aligned processes in which the contact between both source to gate aligns itself with minimal separation. In addition, changes of the potential at the extreme edge of the drain terminal result in a large gate to drain capacitance $C_{gd}$ and a drop in the electrical potential near the drain terminal which is associated with changes in the field distribution particular to recessed channels [61]. The feasibility of active channel at lower potentials decreases the possibility of breakdown due to avalanching process which results in an increased breakdown voltage; the power handling of the PHEMT is consequently improved.

HEMT with triangular gate profile as that depicted in Figure 6.2, presents more losses when compared with other cross sections profiles [65,120]. For instance, the gate resistance of a HEMT process with a mushroom gate profile can be reduced by a factor of 0.15 when compared to that of a HEMT with triangular profile [62,120]. Similarly, with the advances of beam lithography, different profiles have been produced to improve the electrical characteristics of the process. This has a direct impact on the performance of conventional distributed amplifiers since the utilisation of such processes is translated into large gain, lower noise figure and return losses of distributed amplifiers when compared with implementations based on HEMT with triangular profile as reported in [120]. From the above considerations, the design of the triple-line transversal filter based on the available process is prone to different undesirable effects; such as low isolation and reduced drain impedance because of the inherent feedback effects at high frequencies.

6.3. Gain-Bandwidth product of the HEMT process

The trade-off between transconductance gain and the inherent bandwidth is determined by the figure of merit of the active device. For wideband applications, the device figure-of-merit corresponds to the gain-bandwidth product [102,121] or also termed the current gain cut-off frequency of the device, $f_T$. Active devices in common-source configuration distributed along transmission lines can be modelled as transconductive amplifiers as shown in Figure 6.4

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From Figure 6.4, the transconductance (extrinsic) of active device connected to a resistive load, $Z_{out}$ is defined by the ratio:

$$G_m = \frac{I_{out}}{V}$$

(6.2)

The input impedance of the active device $Z_{in}$ is predominantly capacitive (approximately equal to $C_{gs} + C_{gd}$); hence the device transconductance gain $G_m$ tends to increase with the frequency up to the limit given by the maximum current that the device can provide to the load. Here, the current gain with short-circuited load impedance is defined by the ratio:

$$H_{21} = \frac{I_{out}}{I_{in}} \bigg|_{Z_{out} = 0}$$

(6.3)

For an active device in common source configuration, the maximum current gain with short-circuited output impedance is given by:

$$G_I = \frac{g_m}{\omega(C_{gs} + C_{gd})}$$

(6.4)

where $g_m$ and $C_{gs} + C_{gd}$ are the intrinsic transconductance and the input capacitance of the active device, respectively. The gain current parameter $G_I$ is usually obtained at short-circuited output impedance assuming the input impedance is predominantly capacitive. $G_I$ is displayed in logarithm scale against the frequency and the frequency of transition at which the current gain parameter $|G_I| = 1$ is termed the current cut-off frequency. Using Equation 6.4, the current cut-off frequency is equal to [102]:

$$f_c = \frac{1}{2\pi\sqrt{g_m(C_{gs} + C_{gd})}}$$
Chapter 6 GaAs MMIC distributed transversal filter design

\[ f_T = \frac{g_m}{2\pi(C_{gs} + C_{gd})} \]  

(6.5)

In modern HEMTs and MESFETs, the cut-off frequency determined by Equation 6.5 can be up to 10% lower than the measured \( f_T \) [102,68,121]. The approximation above improves as the device under consideration has low parasitics [65]. In most cases the gain bandwidth product can be well-approximated by extrapolating the device current gain at low frequencies. However, it is shown that this is not accurate for the HEMT process considered. Using circuit simulation and the parameters of the HEMT process, the gain bandwidth product of the HEMT was analysed by extrapolating the values of the current gain at short-circuited output impedance, fitting appropriate transconductance and capacitance values. The maximum current gain parameter \( H_{21} \) against frequency is displayed in a logarithm graph. Figure 6.5 displays the maximum current gain parameter with a 6 dB/octave slope and the extrapolation of the maximum current gain at low frequencies, which indicates in an \( f_T \) equal to 66.9 GHz.

![Image of graph showing maximum current gain parameter](image.png)

**Figure 6.5 Maximum current gain \( H_{21} \) for a 4×15 μm HEMT**

Figure 6.5 displays a deviation of the maximum current parameter \( H_{21} \) at frequencies approaching \( f_T \). The large deviation from the linear interpolation stems from the large input resistance \( R_s \) and other parasitic elements in series with the input capacitance of the device. Because of the inaccuracy in obtaining \( f_T \), OMMIC has adopted other methods to estimate the frequency current cut-off frequency. Using on wafer measurements of parameters of the device, the HEMT \( f_T \) was obtained by linear regression of \( H_{21} \) (dB) against \( \log_{10} \) (frequency) between 1 and 40 GHz [63]; thereby obtaining a better estimate of the \( f_T \) when large deviations arise. Nonetheless, in the evaluation of the gain bandwidth consideration there are other effects to take into account. When effect of losses and other parasitic elements becomes significant at higher
frequencies, the model of the active device becomes more complex than that initially considered. A device model including parasitic elements gives a better estimate of \( f_T \) [65].

The available HEMT with 4×15 \( \mu \)m gate width has an intrinsic transconductance approximately equal to 40 mS. At first approximation, a distributed transversal filter using such HEMT and output impedance \( Z_{od} \) has a maximum gain per cell approximately equal to \( g_m Z_{od} / 2 \). For a 50 \( \Omega \) drain impedance, the voltage gain results to be approximately equal to one. Nonetheless, the transconductance of the device (or extrinsic transconductance) is lower than the intrinsic transconductance due to the feedback associated with the source resistance and the device transconductance can be approximated by:

\[
G_m \approx \frac{g_m}{1 + g_m R_s} 
\]  

(6.6)

The maximum gain per cell will be reduced by 70%. The gain bandwidth consideration of the HEMT process is also influenced by the presence of gate-drain feedback capacitance, \( C_{gd} \). In devices such as the HEMT, the shunt feedback gives rise to a substantial reduction in the output current. At high frequencies, the shunt feedback gives rise to an effective current gain lower than the maximum current parameter \( H_2 \) when a 50 \( \Omega \) impedance is connected at the output port, the differences between such parameters can be of the order of 3 dB. Such difference becomes negligible for processes with improved \( C_{gs} / C_{gd} \) ratios.

In order to increase the device transconductance, HEMTs with larger gate widths are implemented; device parameters such as \( f_T \) and \( g_m R_s \) product are device-area independent; thus in most cases an increase in the device dimensions results in a larger transconductances while the figure-of-merit remains almost constant to a first order approximation [102]. Nonetheless, the feedback associated with parasitic elements (\( C_{gd} \) and \( R_s \)) of the HEMT process gives rise to a significant reduction in the effective device \( f_T \) when the device dimensions are increased. The effect of the device feedback and limited gain bandwidth product will be assessed in the design of the transversal filter.

### 6.4 Delay circuits for transversal filter implementation

The design of delay circuits is of central importance for the symbol-rate transversal filter implementation. The microstrip transmission line is a suitable alternative for delay circuit implementations because of their low losses and broadband operations [65]. Microstrip
transmission lines are readily available in the MMIC process based on a GaAs substrate (relative permittivity of 12.9), such transmission line is surrounded by materials of different permittivities from which its electrical characteristics change with frequency. The non-homogenous behaviour in the frequency is characterised by microstrip permittivity which increases monotonically with the frequency [122].

For filter implementations, microstrip transmission lines introduce an intrinsic time delay as a function of the dielectric constant of the material that confines transversal electromagnetic waves. Short microstrip transmission lines are utilised to interconnect distributed cells constituting an artificial transmission line. Nonetheless, the delay of such artificial line is not sufficient in relation to the width of pulses to be filtered. Transmission lines become unsuitable for multi-Gchip/s MMIC filter implementations as the dielectric constant of practical material is not large enough to slow down transmitted waves; moreover, the length of transmission lines required to attain the necessary delays makes on chip implementations unfeasible. The physical length of transmission lines can be effectively reduced by connecting periodically discrete capacitors in shunt with high impedance transmission lines designed to act as distributed inductances; thus passive delay circuits are effectively constructed as low pass LC circuits. The design of microstrip transmission lines acting as inductances has the advantage of reducing the dimensions of transmission lines since the physical length of the lines is decreased in relation to the permittivity of the material in which the line is embedded. The variation of the permittivity with the frequency has a reduced effect in the characteristic that presents such transmission line when those are implemented as a high characteristic impedance transmission line (typically over 90 Ω) on thick substrates.

### 6.4.1. Implementation using the OMMIC process

Considering this practical approach for on chip implementations, the connection of discrete (overlay) capacitors along transmission line involves the use of an ac ground path, which is implemented using a via hole. Figure 6.6 shows a schematic of a delay line using the microstrip technology.

![Figure 6.6 MMIC implementation of a delay section using microstrip lines](image)
Delay stages are designed by connecting periodically discrete capacitances along lines and each ac grounded using a via hole. The MMIC process provides one or two-access via holes to achieve ground paths with high reproducibility [63]. Two-access via hole are used so as to avoid the parasitic associated with ground-bar schemes whilst allowing a reduction in MMIC area. Figure 6.7 shows two access via hole utilised in the implementation on a 100 μm-thick substrate and the equivalent circuit model. Via hole has a reverse side metallization which is continued into the via hole to make contact with the front-end metallization. A low resistance is included in the model to characterise the contact resistance.

### 6.4.2 Modelling of delay lines

The delay line approach described above constitutes a discrete section which electrical characteristics were designed for matching distributed cells of filter sections [100]. Practical design conditions such as the use of a low loss semi-insulated GaAs substrate and high-impedance microstrip lines on thick substrates enable a delay line section to be modelled as a discrete section with specific discrete capacitance and inductance per section. Figure 6.8 shows the delay line that consists of distributed capacitance and inductance per section ΔC and ΔL, respectively. The terminal capacitance of the microstrip transmission lines is added to the discrete capacitance of the overlay capacitor, \( C_D \). The cut-off frequency of circuit section and the characteristic impedance are given, respectively, by:

\[
f_c = \frac{1}{\pi \sqrt{\Delta L \cdot \Delta C}} \quad Z_o = \sqrt{\frac{\Delta L}{\Delta C}} \tag{6.7}
\]

The capacitance per section, \( \Delta C \) that includes the capacitance of the overlay capacitor is given by:
\[ \Delta C = C_D + \frac{\Delta L}{Z_{0,m}^2} \quad (6.8) \]

where \( Z_{0,m} \) is the characteristic impedance of the microstrip line used to form the inductance.

At high frequencies, passive devices can present parasitic behaviour that reduces accuracy. For instance, via hole has a high parasitic (about 27 pH as shown in Figure 6.7) which can be significant in relation to the inductance of the transmission lines; thereby rendering lower delay times.

For accuracy, it was considered to count upon suitable models for such delay lines. The grounded capacitors and distributed inductors, initially considered as low pass \( k \)-sections, now have been modelled as \( m \)-derived sections because of the considerable via hole inductance. A section of delay circuits and the equivalent \( m \)-derived section is depicted in Figure 6.8. Such equivalence permits modelling accurately the phase constant of the transmission line as the phase velocity of the artificial line has been modified by adding lumped capacitances along the transmission lines. The utilisation of \( m \)-derived sections to achieve gate and drain-line phase velocity equalisation is well-known in distributed amplifier implementations [65] and is utilised here to extend the time span of the transversal filter.

Based on the description of discrete sections, the cascade of \( m \)-derived sections forms a ladder circuit matched on image impedance basis. Using definitions given in section 4.3, the phase coefficient per section can be obtained as a function of the normalised frequency and is given by [126]:

\[ \cos(\beta) = 1 - \frac{2m^2(f/f_c)^2}{1-(1-m^2)(f/f_c)^2} \quad (6.9) \]

Equation 6.9 shows a phase constant that change with the frequency. As a consequence, each component of input pulses undergoes different phase velocity and phase delay \( t_s = (\beta/\omega) \) giving rise to pulse dispersion. Figure 6.9(a) shows the variation of the normalised delay versus...
the frequency for different $m$ factors. Figure 6.9(b) shows the standard deviation of $t_\phi$ in the normalised frequency interval $(0,0.5)$. It is apparent from such graphs that circuits featuring an $m$ factor approximately equal to 1.27 present an improved flat delay over a large range of frequencies (minimal standard deviation for $m$ equal to 1.27); whereas delay circuits featuring $m$ factor equal to 0.8 or 1.0 introduces larger dispersion levels.

![Image](image.png)

**Figure 6.9 (a)** Variation of the normalised delay versus the frequency
(b) standard deviation of normalised delay versus $m$-factor

The response of the wideband delay lines to an input pulse results in an asymmetric version of delayed pulses as pulse components undergo different phase velocities. The mean time of asymmetrical delayed pulses can be obtained from the definition of delay time given by Elmore [123]. For an $m$-derived section, the mean delay time, $t_d$ results to be equal to:

$$t_d = \frac{m}{\pi f_c}$$  \hspace{1cm} (6.10)

This group delay value is constant within the bandwidth where the phase shift of the single stage is linear. In circuits consisting of $n$-cascaded sections, the delay increases linearly as a function of the number of sections. From equations 6.7 and 6.8, the delay of $n$-cascaded sections is given by:

$$t_{d,n} = nt_d = n \times m \sqrt{\Delta L \cdot \Delta C}$$  \hspace{1cm} (6.11)

In order to model delay circuits using practical devices, two definitions of frequency arise: the ideal cut-off frequency and the 3dB cut-off frequency. The ideal cut-off frequency, also termed Bragg cut-off frequency in the context of travelling wave amplifiers [69,66], is the frequency at which the individual reflections from each periodically spaced shunt capacitances add in phase to maximise the reflection. Beyond such critical frequency the delay lines introduces large attenuation and represents the ultimate bandwidth limitation of the artificial
line. The Bragg cut-off frequency is associated with the group delay per section as given by Equation 6.10. The second definition relates to the fact that low dispersion levels imply increased phase linearity and amplitude falling smoothly with frequency. The frequency at which the amplitude response falls 3 dB is termed the 3 dB cut-off frequency, \( f_{c,3dB} \) and depends on the number of stages constituting the delay lines and on dissipative constants. Practical devices present parasitic elements that reduce the amplitude response at frequencies lower than the Bragg cut-off frequency.

For the basic \( m \)-derived sections described above (Figure 6.8) it is not possible to achieve \( m \)-factors larger than one to improve delay flatness. However, it is important to mention that there are configurations of circuits which allow circuit sections to have \( m \) coefficients larger than one and exhibit mutual inductance. Constant-\( R \) circuits present mutual inductance that can be designed as a bridged-T sections and improve the delay flatness [65,112]. Theoretically, constant-\( R \) circuits can be designed to have image impedance independent of the frequency [112]. Additionally, such circuits designed as bridged-T sections could increase the filter time delay whilst improving the bandwidth and impedance level. Delay flatness can be extended to higher frequencies; whilst the amplitude response varies with the frequency approximately as a Gaussian function [65]. Constant-\( R \) circuits are attractive for filter implementations given the need of increasing the interstage tap delay in MMICs with limited chip area. Coupling in densely packed inductors could be analysed using electromagnetic simulator, such implementations were not considered given the detrimental bandwidth limitation associated with the inductance of via hole.

### 6.4.3. Implementations with dissipative elements

Practical implementations of delay circuits based on microstrip transmission lines have an \( m \) factor lower than one. Using the GaAs MMIC process, the implementation of delay line sections has an \( m \) factor of 0.8 [100]. As a consequence, pulse dispersion arises as a main form of distortion of delayed pulses. The implementation of delay sections with dissipative elements improves the linearity of the phase function over a significant part of the pulse bandwidth giving rise to highly symmetrical delayed pulses. The amplitude response falling with the frequency reduces high-frequency oscillatory components. Dissipative lines used in this application include a small damping resistor, \( r_d \), in series with overlay capacitors.

Each delay section matched on image impedance basis has the propagation constant, \( \theta_d \), which depends on a dissipative constant, \( \alpha \). The real part corresponds to the attenuation per section \( \alpha_d \) and imaginary part to the phase function \( \beta_d \), is given by [75]:

\[
\theta_d = \sqrt{\frac{\alpha_d}{r_d}}
\]

\[
\alpha_d = \frac{1}{2} \ln \left( 1 - \frac{1}{r_d} \right)
\]

\[
\beta_d = \sqrt{\frac{\alpha_d}{r_d}}
\]
Chapter 6 GaAs MMIC distributed transversal filter design

\[ \exp(\theta_x) = \exp(\alpha + j\beta) = \frac{\sqrt{1-s^2} + as + ms}{\sqrt{1-s^2} + as - ms} \]
\[ s = j \frac{\omega}{\omega_c} \]  

(6.12.a)

where the dissipative element, \( a \) is normalised to the ratio:

\[ a = \frac{2mr_d}{Z_o} \]  

(6.12.b)

The choice of a dissipative constant different from zero is a compromise between delay, flatness, loss and reduction of the cut-off frequency. For example, Figure 6.10(a) shows the variation of the normalised phase delay for different dissipative constants, improving the delay flatness for large dissipative factors. Figure 6.10(b) shows the amplitude (image attenuation) response against the frequency. The 3 dB cut-off frequency is reduced as large dissipative constants are chosen.

![Figure 6.10 (a) Normalised phase delay versus normalised frequency](image)

![Figure 6.10 (b) image attenuation versus normalised frequency](image)

The temporal spread of delayed pulses associated with the bandwidth limitation can be measured via the step-response rise time, \( t_r \), following the conventional definition of time for delayed signal to rise from 10% to 90% percent of its final value. When designing a delay section it is important to maintain the image impedance matching condition. This is due to the distinct feature of such matching mechanism as it results in minimal bandwidth reduction when cascading several sections to increase delay. For instance, the distributed amplifier has better step-response rise times when compared with those of baseband amplifier based on the same device [76,69] in which large time constant arises mainly associated with the RC circuits and dissimilar input-output impedances.
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In a pioneering work on the characterisation of delay lines, Elmore and Sands [123] provided results of the measurements of the rise time of constant k-filters. In a range from 3 to 30 sections, the rise time of n-sections is represented very well by:

\[ t_{r,n} \approx \frac{1}{n} t_r \]  

(6.13)

This approximation is also valid for the case of cascaded m-derived sections [126]. This approximation compares favourably with multi-stage amplifiers in which rise times are degraded with a \( \sqrt{n} \) rule [125,127]; therefore superior transient responses can be obtained by transversal filters based on distributed principles.

Figure 6.11 depicts the effect of changing the number of sections of the transmission line matched in image impedance basis. The graph in Figure 6.11 is related to a normalised cut-off frequency. A degradation of the rise time arises as a result of a reduction in the higher frequency components.

![Figure 6.11 Rise time-cut-off frequency product as a function of delay sections](image)

Figure 6.11 Rise time-cut-off frequency product as a function of delay sections

When the delay circuit presents a linear phase in the frequency range of interest and as a consequence low overshoots in the transient response (lower than 5%), the bandwidth degradation \( f_{c,3\text{dB}} \) and the rise time of n-cascaded sections are given by:

\[ t_{r,n} \times f_{c,3\text{dB}} \approx 0.35 \]  

(6.14)

Based on the above definitions and parameters, the figure of merit of a delay line can be expressed by the delay time to rise time ratio assuming constant delay in the frequency band of interest [123]. Delay sections with large \( m \) factors and low dissipative coefficients such as that featuring an \( m \) factor equal to 1.27 presents better figure of merit than those circuits designed
with an $m = 0.8$ and dissipative constant different from zero. Equation 6.11 establishes that in order to achieve a specific delay and bandwidth operation, a delay stage consisting of $n$-cascaded circuits (or sections) with $m$-factor equal to 0.8 requires approximately an additional half of the number of sections when compared with those featuring an $m$-factor equal to 1.27.

Figure 6.12 shows simulation results of four identical delay stages based on the basic $m$-section of Figure 6.8, the details of the implementation are given in Table 6.3. The sections have an $m = 0.8$ to account for the via hole inductance.

![Figure 6.12 Voltage on delay stages for a sharp input pulse](image)

(a) (b)

Figure 6.12 Voltage on delay stages for a sharp input pulse (25 FWHH 7 ps fall and rise time), (a) lossless line (b) line with $a=0.16 \, r_d=5\Omega$

| Bgrass cut-off frequency | $f_c=63.3$ GHz | Characteristic impedance | $Z_{0,m}=100\Omega$
<table>
<thead>
<tr>
<th></th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>MIM capacitance</td>
<td>$C_D=0.71$ fF</td>
<td>Effective relative permittivity</td>
<td>7.52 at 40GHz</td>
</tr>
<tr>
<td>$\Delta C , , \Delta L$</td>
<td>$0.08 , pF, 198.4 , pH$</td>
<td>Attenuation</td>
<td>$0.08 , dB$ at $40 , GHz$</td>
</tr>
<tr>
<td>Transmission line parameters width, length</td>
<td>$w = 6.8 , \mu m$, $\ell=219.1 , \mu m$</td>
<td>Maximum current due to metal-migration</td>
<td>$40.7 , mA$</td>
</tr>
</tbody>
</table>
Figure 6.12 displays results of a dissipative line and those of a line without losses. The first has a transient response with little overshoot and symmetrical pulse while the amplitude is reduced at each stage as a result of the dissipative element. In addition, the rise time and fall time are increased as a result of the bandwidth limitation. For the case with a zero dissipative factor, the transient response shows less varying pulse amplitudes while the overshoots are more evident.

6.5. Considerations on the selection of the HEMT

An important consideration in the design of the transversal filter is the selection of the device geometry that satisfies certain performance requirements. Device elements such as the intrinsic transconductance, input capacitance and output resistance depend on the device area and play a fundamental role in the level of gain and bandwidth of the filter cell. Conversely, other device elements such as the gate-to-drain capacitance, $C_{gd}$ or source resistance, $R_s$ depend heavily on the channel technology adopted in the development of the process and those are completely determined when choosing the HEMT gate width so as to achieve a specific transconductance. For a distributed cell with the HEMT directly coupled to the drain artificial line, device feedback associated with $C_{gd}$ gives rise to reduced output impedance at high frequencies. In the selection of the transistor area, the output resistance is scaled inversely proportional with the gate width while the transconductance is scaled directly proportional to the device gate width. Those factors need a careful choice of the width geometry in accordance with the gain and bandwidth operation of the filter.

The selection of the gate width as a variable in the design is restrained by requirements and performance characteristics of the transversal filter. Transversal filter topologies, and in general distributed amplifiers with cells consisting in a single stage, are highly susceptible to bidirectional couplings between input and output transmission lines. Bidirectional coupling is associated with feedback elements of the active devices and such couplings could lead to high levels of loss which cannot be predicted by a unilateral model. Analysis of the distributed amplifier as a coupled structure was reported by Wong and Keli in [69] in order to get an insight into the mechanism of passive and active couplings when using bilateral models of active devices. However; the use of such analysis to optimise the performance based upon an individual selection of the device area is difficult to achieve given the complexities associated with using complete high-frequency equivalent circuits of the FETs. Moreover, the analysis based on the variation of the device equivalent circuit elements (one-by-one) to study the performance is fruitless since the device must be considered as an integral element in the MMIC
design. For this reason, the design based on a specific active device relies on the optimisation in which different elements are tuned to accomplish the performance.

In the artificial gate line, the attenuation depends on the input resistance of the device in common-source configuration. Such resistance is approximately equal to $R_s + R_G + R_i$ [106]. $R_i$ corresponds to the channel resistance which varies with external bias voltages. In practice, this resistance can be neglected because of its low value when compared with the other elements. $R_s$ and $R_G$ are the access resistance in the source and gate regions, respectively and both depend on the device gate width. The gate resistance $R_G$ is directly proportional to the device gate width and can be reduced by constructing the device as a parallel connection of the gate strips of identical widths [63]. Conversely, $R_s$ is scaled inversely with the device gate widths and can be large when low gate widths are chosen to maintain large output impedance levels. For the distributed filter design, and $R_s$ different from zero results in feedback effects*, which may lead to stability problems.

6.6. MMIC Transversal filter methodology

The design of microwave circuits starts with the selection of the proper circuit topology capable of providing the required performance and functionality. In the present implementation, the triple-line transversal topology was chosen mainly because of its low complexity as only two active devices are utilised to set the tap gain and sign and because of its good gain, bandwidth and noise performance [101]. Other designs, based on Gilbert cells as reported in [25,116], have also the bipolar capacity required to achieve the receiving function. Comparisons with distributed cell topologies reported in [25,116], in terms of performance as well as general aspects of those alternatives, were carried out in [101] and will be described in Chapter 7. The triple-line transversal filter design is based on identical distributed cells. As described earlier, the active matching technique provides uniformity on the common artificial gate line when different tap gain weights are set to a predetermined function; thus constant impedance and delay characteristics can be obtained without requiring phase compensation techniques as employed in capacitively coupled TWA [54] or in distributed amplifier with unequal characteristic impedance / electrical lengths [71,80]. Furthermore, the characteristic impedance and delay per section in the gate and drain artificial lines can be made approximately equal to increase the filter tap delay. Delay circuits described in section 6.4 can be designed to match the

* Because of the compromise arising from the selection of the device gate and the effect that the source resistance has on stability, an assessment of the filter performance has to include analysis of stability.
impedance of distributed cells and provide the transmission characteristics to handle picosecond pulses.

In order for design specification targets to be met, the design process is divided into different steps. These comprise a trial-analysis-and-redesign process in which two independent variables are involved; the capacitive division ratio $M$ and the transistor gate width which determines the input capacitance of the device. A redesign stage is then done if analysis results do not satisfy initial requirements; circuit simulation and optimisation are performed to compensate for nonidealities of the active device itself. The methodology of design is applied to the transversal filter by considering modelling each filter section using the small signal equivalent circuit of Figure 6.13.

![Figure 6.13 Small-signal equivalent circuit of the distributed cell](image)

The modelling of active cell by the equivalent circuit of Figure 6.13 was described in Chapter 5. Some simplifications were made in the small-signal cell model; for instance by neglecting the passive coupling within the device, in particular the effect of “feedback” elements of the HEMT. What the equivalent small-signal model of figure 6.13 can model accurately is the cut-off frequency of gate artificial lines, from which the delay per section can be predicted with good accuracy [99,100]. A basic equation that links the design variables of an ideal gate line cut-off frequency is given by:

$$f_c = \frac{1}{\pi Z_{o,g} (2MC_{in} + \bar{C}_g \ell_g)}$$

(6.15)

In equation 6.15, the device gate width is directly proportional to the device input capacitance $C_{in}$ and $M$ corresponds to the voltage division ratio effected by the external series capacitance with the input capacitance of the device, $Z_{o,g}$ is the characteristic impedance of the gate artificial line. $\bar{C}_g$ is the capacitance per unit length of the microstrip line and $\ell_g$ is the
Chapter 6 GaAs MMIC distributed transversal filter design

microstrip length designed for an inductance per section equal to $L_g$. The characteristic impedance and the approximate input capacitance are given, respectively by:

$$Z_{o,g} = \sqrt{\frac{L_g}{2MC_{inp} + C_g \ell_g}}$$

$$C_{inp} = C_{g,0} + (1 + g_m Z_{o,d}) C_{d,g,0}$$

(6.16)

with $L_g$ the inductance synthesised by short transmission lines. The approximation for the input capacitance $C_{inp}$ takes into account the Miller capacitance [106]; assuming the elements are determined at the reference voltage, $V_{GS} = -0.5$ V. $Z_{o,d}$ is the characteristic impedance of the drain line and is set to be equal to the gate characteristic impedance.

The design technique involves an appropriate selection of the ideal cut-off frequency which depends on the voltage division factor and gate widths so that the bandwidth limitation associated with cell design does not attenuate significantly components of the input pulses. Parasitic elements of active devices introduce frequency dependent attenuation on artificial lines. This lowers the 3 dB point of the transversal filter to a frequency below the ideal cut-off frequency. For instance; in distributed optical receiver preamplifiers, the required 3 dB point of the amplifier is set to a half of the gate cut-off frequency so as to minimise the group delay distortion [80,128]. Additionally, the relation maintained between the 3 dB frequency and the ideal cut-off frequency is also influenced by the reflection coefficient due to mismatch between the ATL characteristic impedances and terminal impedances; a return loss lower than 10 dB can be sustained up to frequencies close to $0.5f_c$ if the resistance of generator and terminal impedances are equal to the characteristic impedance of the artificial transmission line [130].

In order for pulse shape integrity to be maintained, the filter designer has to consider the link between the bandwidth and cut-off frequency. The capacitive division technique is used in the filter implementation to provide a degree of freedom in the design; the area of the device is chosen to accomplish certain gain and impedance level requirement whilst a proper selection of the $M$ factor permits setting the cut-off frequency, $f_c$ given by Equation 6.15 [100]. In the implementation, a readjustment of series capacitance can be achieved to accommodate the additional terminal capacitance of the interconnecting microstrips, $C_g \ell_g$; the series capacitance is reduced so as to obtain the required capacitance per section. The rest of the transmission line parameters are then chosen to synthesise the required inductance and complete the specification of each element.
6.7. Implementation using the ED02AH process

A design based on the triple-line filter topology was carried out using the trial-analysis and redesign approach and following the design methodology described in the previous section. A first implementation based on the HEMT process was aimed for an 8-tap transversal filter with a 3 dB cut-off frequency of 25 GHz. Initial design goals led to consider an implementation with a voltage division factor approximately equal to one and HEMTs with a 4×15μm gate width. This parameter selection allows carrying out filter performance analysis without making use of bandwidth improvements via small capacitance division ratios whilst the gate width could produce a gain per tap close to one. From the extracted parameters of the HEMT process (itemised in Tables 6.1 and 6.2) and Equations 6.15 and 6.16, the input capacitance of a device coupled to the gate line $C_{\text{in}}$, is equal to 75.6 fF. A series capacitance of 0.5 pF was chosen so as to set $M = 0.87$ for the filter implementation. The predicted loss-free cut-off frequency given by Equation 6.15 is equal to 49.8 GHz. In order to maintain pulse shape integrity; the filter bandwidth is set to 25 GHz which corresponds to input pulse width of 40 ps. Delay circuits were introduced between filter cells on drain and gate ATL to set a 40-ps tap delay. Other specifications were met in this initial trial. In this initial stage, the inverter was not included as simulation results were obtained by taking the difference between voltages appearing across drain terminal impedances.

![Figure 6.14 Time response to a sharp input pulse (FWHH equal to 40 ps and input amplitude of 15 mV), (a) nodes on gate line (b) output port](image)

Figure 6.14(a) shows time domain simulations of the filter when a 40 ps pulse is applied at the input port. The figure displays voltage nodes on the gate transmission line at different
Chapter 6 GaAs MMIC distributed transversal filter design
taps, showing a reduction in the amplitude in later filter stages. The first pulse presents ringing
and sags due to the reflections at the input at high frequencies. The low oscillatory component
of delayed pulses is a result of the phase linearity of gate line sections in which losses associated
with the input resistance of active devices plays an important role in determining the transient
response.

Figure 6.14(b) shows time domain simulations at the output port when the second and
eighth tap gains were set to maximum tap gain. Pulse amplitude is reduced mainly due to the
feedback effect of the device (3 dB reduction of the effective transconductance) and the $M$
voltage division. In addition, frequency dependent attenuation on drain ATL reduces high
frequency components of delayed pulses. The first pulse at the output shows a large undershoot
resulting from the effect of the reactive component at low frequencies. In spite of the fact that
the drain line attenuation can be considered almost constant in the frequency range of pulses as
discussed in Section 5.8, pulse distortion shown in graph of Figure 6.14(b), suggests that other
distortion mechanisms take place and become noticeable for pulse shape functions. In the light
of the above results, following trials were carried out by designing for a lower $M$ factor values.
A bandwidth limitation arises due to the feedback effect and the associated reduction of the
output impedance. High level of losses on the drain line reduces the amplitude of pulses. For
such reasons the HEMT was implemented with low areas so that the impedance load on the
drain transmission line is reduced.

A redesign process was carried out by reducing the small divisor capacitor to an $M$
factor equal to 0.5. Similarly, the HEMT gate width was reduced so as to enhance the
bandwidth. Design parameters (provided in Table 6.4) resulted in a cut-off frequency equal to
63.6 GHz, the 3dB point of the transversal filter should be approximately equal to 30 GHz
according to the criteria for low distortion adopted in the design of transversal filters. In
addition, such design characteristic matches the impedance and bandwidth of delay sections
implemented in Section 6.4. The specification of individual elements was achieved following
the design procedure described in the previous section. Figure 6.15 shows a schematic of the
transversal filter (the implementation was aimed to 7 taps) designed to the standard chip size of
the foundry process. The group delay per stage is approximately equal to 4 ps; therefore two
additional delay sections were required on gate and drain lines so as to set additive 25 ps tap
delay.

The inverter stage based on a HEMT in common-source configuration was designed by
optimising the HEMT area and capacitances. Its design method will be addressed in Chapter 7,
where the methodology of differential and common-mode signals is described, which provides a
foundation for the design. In the following the analysis of the transversal filter was done by
obtaining the forward gain parameter of the filter structure at the output of both drain terminals and taking the difference between the outputs.

![Transversal filter circuit topology, filter stage schematic and microstrip delay lines](image)

**Figure 6.15** Transversal filter circuit topology, filter stage schematic and microstrip delay lines

**Table 6.4 Parameters of the devices in gain stages**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Gate width</td>
<td>2 fingers×18 μm</td>
</tr>
<tr>
<td>Terminal capacitance of microstrip in drain lines</td>
<td>12.0 fF</td>
</tr>
<tr>
<td>Inductance of microstrips / Z₀,m (gate and drain)</td>
<td>99.2 pH / 100 Ω, 90 Ω</td>
</tr>
<tr>
<td>Additional drain capacitance to equalise delay in both lines</td>
<td>68.0 fF</td>
</tr>
<tr>
<td>M factor</td>
<td>0.5</td>
</tr>
<tr>
<td>Capacitance load per cell</td>
<td>88.0 fF</td>
</tr>
</tbody>
</table>
A transversal filter based on the novel topology was implemented as a MMIC following the rules of the ED02AH process from OMMIC-Philips. The AD02AH process is based on microstrip transmission lines. Figure 6.16 shows the layout of the MMIC 7-tap transversal filter. Single-metal layer transmission lines were employed to interconnect discrete elements in gate and drain ATLs. Multi-metal layer transmission lines, on the other hand, were not utilised given the layout restrictions with crossings which are employed to form the dc biasing path of each HEMT. The layout was generated using smart libraries provided by OMMIC and installed in Advanced Design System, ADS™. The layout was generated from an equivalent electrical schematic using the automated tool, resulting in slight modification of some device parameters or dimensions. That allows reducing pattern overlap and adjusting devices to be laid out on a grid of 0.5 μm. The chip size is 3.9 × 2.2 mm², within the range recommended by foundry rules.

In the centre of the MMIC pad frame, the gate line runs horizontally interconnecting overlay capacitors with 90°-bend transmission lines. Vertical lines correspond to thin NiCr resistances which interconnect device gate terminal to bond pads. Each resistance is split at some point so as to introduce crossing with the drain lines. Dry etched via holes were laid out along both sides of gate transmission lines. Overlay capacitors were employed to reduce the length of transmission lines and for blocking DC biases. Microstrip transmission lines in gate and drain lines are designed with different widths (12 μm in drain lines). This allows the biasing of devices up to a limit of ID=78 mA, which is appropriate to avoid transmission line metal-migration. Two-layer MIM capacitors (based on Silicon Nitride and Silicon dioxide (Si₃N₄+SiO₂) as insulator) were synthesised over 60 fF farads taking into account distributed parasitic elements.

Two-access via hole allows grounding discrete capacitors and HEMTs. The size of via hole is 120×120 μm² and minimum distance between the centres of adjacent via is restricted to 200 μm. The spacing between horizontal transmission lines and via holes is 70 μm, this proximity is sufficient to ensure low capacitive couplings between microstrip and layout.

Two capacitors of 20 pF (shown in the upper right hand side of the layout) grounded using a via hole and connected to bond pads were added for decoupling dc power supply. Bond pads were placed around the chip edge and at least 30μm from the street for dicing. In addition, RF probe pads were added to provide the ground-to-ground connections that are required by conventional probing instruments and aligned with pads according with assembly rules.

1 Similar layout considerations are analysed in the coupling between inductor and via hole in [131]
6.9 Frequency domain responses

The frequency response of the transversal is shown in Figure 6.17 and allows verifying that the MMIC transversal filter design can provide the required filter performance. Figure 6.17 (a) displays simulated results of the voltage gain of seven stages with tap gains weighted independently; i.e. a single filter tap gain was biased to a maximum gain whilst the rest were set for a zero tap gain. Normalised tap gain responses are consistent with the attenuation levels of filter taps as all filter taps were biased to obtain the same transconductance gain; the difference between the first and seventh gain is about 8 dB and mainly associated with the attenuation level.

To prevent signal distortion, the variation of the group delay over the frequency was kept to a minimum and stems from the small resistances connected in series with overlay capacitors and the associated losses of active devices, particularly due to large source resistance, $R_s$. At the upper cut-off frequency larger variations are evident. In addition, the filter gains decrease sharply at the cut-off frequency.

Time domain simulations are obtained in the following chapter where the performance of the filter is assessed.
Figure 6.17(a) Voltage gain normalised to first tap gain (b) group delay response

Summary

This chapter presents a methodology for designing transversal filter for the generation / reception of high rate sequences; the requirements of symbol-rate transversal filter has been outlined which allows establishing various practical issues related to the filter design. The practicality of such filter was assessed by using a commercially available HEMT process; a straightforward characterisation of transmission lines was proposed from which delay
sections based on microstrip transmission lines were designed. The transversal filter was effectively designed by interconnecting active cells with delay sections; a 7-tap transversal filter for 40 Gbps application was designed using a commercially available pseudomorphic-HEMT process. The number of the filter taps was chosen to be seven so as to satisfy the rules of the available GaAs MMIC process regarding the maximum chip size allowed by the foundry.

In the design process, the capacitive division technique and the transistor gate width were optimised in a redesign trial in order to satisfy the filter performance; it was shown that the distributed transversal filter topology based on the available HEMT process is highly prone to losses and distortion, mainly because of the low $C_{gs}/C_{gd}$ ratios and large source resistance. Such device characteristics are highly dependent on the recess channel technology of the available process.

Several practical issues were detailed; such as the methodology of design that takes into account the inherent bandwidth limitation of conventional distributed amplifiers and extended for the case of the filter design. Preliminary results given in this chapter lead to affirm that the performance can be further improved by using self-aligned processes featuring appropriate $f_T$. In addition, other transmission line technology such as coplanar waveguide could improve the cell design and delay characteristics. The utilisation of advanced processes will be advantageous to explore the limitation and, when appropriate, increase the number of filter taps for the application. The design considerations of the symbol rate transversal filter, as established at the beginning of this chapter, have been confirmed in this chapter extending the application of distributed transversal filter for the filtering / generation of high rate sequences.
Performance of the MMIC transversal filter for high-speed system applications

In the design of optical communication systems it is essential to choose filter topologies and active device technologies which satisfy the performance requirements of optical communication systems. The selection of the circuit topology becomes a fundamental part of the design. Noise, linearity and bandwidth are usually performance requirement to be considered although costs and practical constrains of MMIC design play a decisive role in the selection of the circuit topology [65]. In optically pre-amplified systems the sensitivity of the receiver is mainly determined by the optical amplifier rather than by the sensitivity of the electronic preamplifier [13]; therefore, the design methodology for such receivers is translated into less stringent noise and gain design requirements, making possible the use of low-gain topologies (such as distributed amplifier) in postdetection stages of the receiver. At frequencies higher than 10 GHz, the distributed amplifier is favoured since it provides the best compromise between bandwidth and sensitivity, whilst the parasitic elements of active devices becomes an important limitation on the gain and bandwidth achieved by other topologies [12].

Analyses carried out in this chapter show that the triple-line transversal filter provides an interesting alternative for CDMA filter implementation because of its functional characteristics and bandwidth. This chapter commences by introducing a small-signal equivalent model of the triple-line transversal filter with delay sections distributed along artificial transmission lines. Frequency domain simulations oriented to the analysis of differential circuit are used in filter analysis; this methodology allows obtaining important performance characteristics such as gain, cancellation of parasitic couplings and return losses.
that can affect the efficient transmission of differential signals. Following this, time domain analysis will be performed to show the versatility of the filter by tuning the response to different high rate sequences. Such time domain analysis confirms that, through the cascading of two transversal filters, the proposed designs can provide the bandwidth and time response to achieve the required encoding / decoding functions for transmitters and receivers of OCDMA. It is shown that operating conditions of the triple-line filter can provide the appropriate performance for the reception of high-rate sequences.

In the design of the symbol-rate transversal filter, the wideband operation is translated into low overall gains. Given the gain limitation that can be achieved (on the order of the transconductance of the active device), the utilisation of such filter for high bandwidth operations can lead to noise penalties that might compromise the overall sensitivity of the receiver. Filter noise will be analysed using complete models of the available HEMT process. It will be shown that the MMIC design is a stable circuit for the operating conditions. Towards the end of the chapter, other structures that provide bipolar capacity and similar functionality [25,116] will be discussed and compared with the triple-line filter. Results show that when the triple-line filter satisfies certain design criteria, it could be an appropriate alternative for Optical CDMA systems.

### 7.1 Model of the filter topology

The design of the transversal filter needs a stage to convert differential voltages to a single-ended response to allow an accurate measurement and further processing. Figure 7.1 shows the diagram of the transversal filter under consideration. The principle of operation is described as follows; input pulses are delayed and in turn amplified by each active cell. Signals coupled to both drain lines travel towards terminal impedances in both direction. In the inversion stage, the voltage wave of the upper drain line is coupled to the common-source HEMT via a series capacitor $C_b$ (see Figure 7.1). This capacitor blocks dc voltage in the upper drain line and divides the RF voltage appearing across the gate junction of HEMTs. The voltage $V_s$ at the input of the inverter stage is amplified to achieve signal inversion. The available output current of the HEMT is halved to form a forward wave which is combined linearly with signals propagating towards the output load. An additional delay line is included in the lower drain line to compensate for the inherent delay of the inverter. Results in this chapter show the efficacy of this technique.
A methodology to analyse the filter performance is developed from which the response can be conveniently characterised by differential signals propagating along drain transmission lines. Each filter cell satisfies the requirement of positive and negative tap gain weight using the voltage levels obtained by the gain weight control technique described in section 5.5. For analysis, the transconductance of the devices of a filter cell can be modelled as a linear combination of a fixed and incremental tranconductances \( g_{m,0} \) and \( \Delta g_k \), respectively. Cell transconductances are given by:

\[
\begin{align*}
g_{\text{max},k} &= g_{m,0} + \Delta g_k \\
g_{\text{min},k} &= g_{m,0} - \Delta g_k
\end{align*}
\]

Each active cell is modelled by equivalent small-signal circuit given in Figure 5.10 and repeated below, in Figure 7.2 for convenience.
Based on this circuit equivalence, it is possible to identify two different induced currents flowing simultaneously into drain lines. Differential signals are induced by incremental current sources with the same amplitude and opposite phase, whilst induced currents \( g_{m_0} V_{g,k} \) flow into both drain line having the same amplitude and phase. The induced currents associated with the fixed sources are independent of tap gain settings and may be taken as a common reference. Therefore, the filter response can be described in terms of its differential current sources. This is best represented by the incremental circuit model depicted in Figure 7.3.

**Figure 7.3** Incremental model of the filter with transconductances tuned for normalised coefficients \((A_{g_1}, A_{g_2}, A_{g_3}, \ldots, A_{g_N}) = (+1,0,0,\ldots)\)

Current sources distributed along drain lines have polarities set to the required filter function. For instance, in order to achieve the filtering function defined by the sequence \((+1,0,0,\ldots,0)\), the gain of the first cell is adjusted to the maximum transconductance, while the other sources are biased to a reference voltage to ensure zero signal level at the output port (i.e. \( \Delta g_k = 0; k = \{2,3,\ldots,N\} \) and \( \Delta g_1 = g_{max}/2 \)). In the inversion stage, the phase of the voltage phasor in the node of the upper drain line, \( V_{d2}(z) \) is shifted 180 degrees and combined linearly with the node wave of the lower drain, \( V_{d1}(z) \) with the same phase. By neglecting the delay associated with the inverter, the output signal corresponds to the difference between both phasors at \( z = N\ell_d \) given by:
where $\phi_g$ and $\phi_d$ corresponds to the drain-line and gate-line phase delay, respectively. The amplitude of the signal is proportional to input pulses, $V_{in}$ and is given by:

$$V_{max} = MZ_{zd}g_{max}V_{in}/2$$ (7.3)

If cascaded stages are adjusted to the code (0,-1,...0), the second tap gain is adjusted to maximum transconductance, $\Delta g_2 = -g_{max}/2$ and $\Delta g_k = 0$; for $k = \{1,3,...,N\}$. In this example, the polarity of the sources is inverted to obtain a pulse with positive amplitude on the upper drain line and negative on the lower drain line. The output signal corresponds to the summation of pulses of both drain lines in phase given by:

$$V_{out} = -V_{max} \exp(-j[3(\phi_g + \phi_d)/2 + 2\theta])$$ (7.4)

where $2\theta$ is the phase delay of sections in gate and drain lines which are made identical. Using superposition it is possible to arrive to the filter response for different sets of coefficients. The gain of an N-stage filter can be described by the equation:

$$\frac{V_{out}}{V_{in}} = \frac{1}{2}MZ_{zd}g_{max}\exp(j(\phi_g + \phi_d)/2)\times \sum_{k=1}^{N}c_k \exp(-j[k(\phi_g + \phi_d) + (k-1)2\theta])$$ (7.5)

where $c_k$ is defined as the fractional coefficient of a stage tap gain normalised to the gain of the stage with the highest gain coefficient. Note that in Equation 7.5, the maximum transconductance gain, $g_{max}$ has been factored out assuming the gain normalisation. The model described above may include device losses for accuracy. For such aim, the methodology presented in section 5.8 and the propagation constants of the equivalent small signal equivalent model could be used for the forward gain analysis.

Note that for linear phase, the phase constants can be expressed in terms of the equivalent delays

$\quad t_d = \omega \sqrt{C_dL_d}$ and $t_g = \omega \sqrt{C_gL_g}$; similar relationship can be used for the interestage delay lines.
7.2 Frequency domain analysis; scattering parameters

A feature of interest in the wideband filter cell design is that of the generation of differential signals to achieve a predetermined filtering function. For instance, Gilbert cell comprises a balanced input differential amplifier and source coupled-pairs to amplify input differential signals with a specific gain and phase in a wide range of frequencies (see section 7.6). Figures of merit of analogue circuits such as the fidelity of amplification of differential signals and rejection of common-mode signals are obtained using mixed-mode analysis [132]. With the need of differential circuits working at microwave frequencies, the analysis of conventional lumped circuits becomes inappropriate given the nature of signals to be processed and the inherent distributed-like behaviour of such circuits. Traditional scattering parameters enable the analysis of microwave circuits; however, the straightforward application of \( s \)-parameters cannot allow mixed-mode analysis. Recently, the theory of scattering parameters has been extended to differential microwave circuits in which the response can be described in terms of different modes of propagation [105, 133]. The application of mixed-mode scattering parameters to the transversal filter is useful for analysis given that the capacity of filtering / generation of bipolar signals are related to the propagation of differential signals on artificial transmission lines. Undesired responses such as parasitic couplings in drain transmission lines and impedance mismatching with terminal impedances need to be analysed carefully given the limitations associated with practical active devices.

7.2.1 Derivation of mixed-mode parameters

In order to assess the capacities of the filter proposal, the modes of propagation along circuit sections are analysed here using mixed s-parameter analysis. Propagation of differential signals in both drain lines are associated with node waves induced by periodically spaced cells. The filter structure under consideration has a single-ended input port and two-ended output port at the terminal nodes of both drain lines; \( i.e. \) the inverter stage is not included for the analysis. Figure 7.4 depicts the drain lines used to define the voltage and current waves in both drain lines. By assuming the bias technique in Chapter 5 applied to all distributed cells, signals propagation are described by two main modes of propagation; those are differential common-mode waves. At the output, both orthogonal modes of propagation can be characterised by taking the difference or sum of voltage phasors \( V_1(z) \) and \( V_2(z) \) in both drain lines, see Fig. 7.4). The differential mode of propagation is obtained by taking the difference:

\[
V_{\text{dif}}(z) = V_1 - V_2
\]  

(7.6)
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On the other hand, the fixed transconductance of the constituent devices $g_{m,0}$ is a common reference in the adjustment of transconductances in both rows of active devices. From Equations 7.1, the maximum and minimum transconductances of each active cell are related by relationship: $g_{\text{max},k} + g_{\text{min},k} = 2g_{m,0}$. Therefore; the corresponding voltage associated with the fixed transconductance that induces the common-mode of propagation is given by:

$$V_{cm}(z) = \frac{V_1(z) + V_2(z)}{2}$$  \hspace{1cm} (7.7)

The differential mode of propagation has associated mainly differential currents induced by current generators $\Delta g_k \cdot V_{g,k}$ flowing into both drains in opposite directions. In differential circuits, currents with equal magnitudes appear 180 degrees out of phase at the output port. Therefore; the differential mode current is defined as one-half the difference between currents flowing on both drain lines and it is given by:

$$I_{dm}(z) = \frac{I_1(z) - I_2(z)}{2}$$ \hspace{1cm} (7.8.a)

The current common-mode component is simply the total current induced by the transconductance of all distributed elements. The common-mode and differential-mode current is given by the current phasor [105]:

$$I_{cm}(z) = I_1(z) + I_2(z)$$ \hspace{1cm} (7.8.b)

The above definitions are used to characterise the simultaneous modes of propagation as a result of the level of gain of distributed cells and distributed characteristics. The description based on mixed mode analysis facilitates obtaining meaningful results from conventional s-
parameters. In order for mixed-mode scattering parameters to be applied, incident and reflected waves at both ports are defined by terminating the input and output impedances with an impedance identical to the impedance of the artificial line [105] in a similar way to conventional scattering parameters [135].

A distinct feature of mixed-mode analysis is the use of differential and common-mode waves directly applied to the output port. In the measurement of mixed-mode parameters, AC generators with the same amplitude and $0/180^\circ$ phase are applied to both drain terminals to launch common-mode and differential waves, respectively [105]. Incident and reflected waves at the nodes of the circuit are measured by the differential and common-mode voltages and current phasor defined by Equations 7.6 to 7.8. The methodology of measuring mixed-mode parameters is described in detail in Appendix C, where the linear equations to transform $s$-parameters into mixed-mode scattering parameters of the triple-line are derived. The following results of this methodology are focused on the interpretation of mixed-mode analysis applied to the triple-line filter structure (Appendix, Figure C.4). The frequency domain characterisation is itemised as gain parameters, input port reflection parameter, reverse transmission parameters, conversion mode parameters and output reflection parameters. The scattering parameters of the transversal filter are detailed as follows.

### 7.2.1.1 Gain parameters

Gain parameters $\{s_{DS21}, s_{CS21}\}$ correspond to the filter response when a stimulus is provided at the input port and when drain ports are matched to drain line. The difference between both parameters depends on the propagation modes on both drain transmission lines, either if common-mode components are taken at the output port (common-mode forward gain parameter, $s_{CS21}$) or by taking differential-mode components at the output port (differential-mode forward gain parameter, $s_{DS21}$). The last parameter corresponds to the desired response and depends essentially on the tap gain weight settings.

In order to show the filter tunability, the differential-mode gain parameter, $s_{DS21}$ was simulated for filters tuned to different discrete sequences. Appropriate tap gain weights were used for such aim; those correspond to the filter response to different pulse patterns as shown in time domain results in the section 7.5.2. Forward gain parameters are obtained by assuming terminal impedances matched to the drain artificial transmission lines. Figure 7.5 shows the amplitude of the forward gain parameter for two different sequences.
Figure 7.5 Forward gain parameter for (+1, +1,-1,+1,-1,-1,+1) (left) and for (+1,-1,+1,-1,-1,+1,+1) (right), the target finite impulse response is in grey line.

Figure 7.5 also shows, in grey lines, the target (ideal) filtering functions for comparison purposes. Gain components of the target filter function were obtained by the equation:

\[ H_T(j\omega) = \sum_{k=1}^{N} a_k \exp\left(-j2\pi k\omega / \omega_s\right) \]  

(7.9)

where \( a_k \) corresponds to the coefficients of the filter \( k = \{1,..,7\} \) and the maximum frequency displayed in the graph, at 40GHz corresponding to \( T_s = 25\,\text{ps} \), \( \omega_s \) is also inversely proportional to the sampling time \( T_s \). By comparing the calculated and simulated response in Figure 7.5, the bandwidth limitation of the filter at higher frequencies is evident. Conditions inherent to the transversal filter such as bandwidth limitation, inter pulse interference and accumulated attenuation make the optimisation in the time and frequency domain two different processes. This was pointed out in the design considerations for the symbol rate transversal filter in Section 6.1, in which practical limitations described in this chapter can be counteracted by adjusting filter tap gains to improve the response.

The common-mode gain parameter \( s_{C21} \) (orthogonal to the differential gain parameter) accounts for the propagation of common-mode signals; i.e. travelling signals that maintain the same amplitude and phase along drain lines and those are referred to a common ground. Unlike the differential-mode gain parameter, common-mode signals practically do not depend on tap gain weights. Figure 7.6 shows simulation results for the MMIC transversal filter biased for two different filter functions, the \( s_{C21} \) parameter for the filter sequence (+1, +1,-1,+1,-1,-1,+1) and the parameter for all tap gain weights equal to zero; i.e. all filter taps biased to the reference voltage. Changes in the device output impedances as a result of applied biases give rise to different attenuation levels on both lines, however; small discrepancies (\( \pm 1.5 \,\text{dB} \)) in the ripples at mid frequencies arise as a result of the variation of output impedances with bias voltages.
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Figure 7.6 Simulation results of the common-mode gain parameter for two different tap gain weights adjustments

7.2.1.2 Input port reflection parameter

The input port reflection parameter $s_{ss11}$ (which resulted to be equal to $s_{11}$ of conventional scattering parameters in the linear transformations in Appendix C) is determined by applying a source voltage at the input port and impedance match to drain lines. Figure 7.7 shows the simulation results of the reflection parameter and the voltage standing wave ratio.

Figure 7.7 Magnitude of the input reflection coefficient and input VSWR

A VSWR lower than 2:1 (or equivalently a $s_{ss11}$ lower than $-10$ dB) is sustained up to frequencies about 36 GHz. Practically identical broad-band VSWR was obtained for the transversal filter with different tap gains. Results suggest the transversal filter presents excellent matching characteristics with a preceding preamplifier or source. Similar input VSWR is obtained in distributed amplifiers [71]. A VSWR different from zero at low frequencies is associated with the use of transmission lines.
7.2.1.3 Reverse transmission parameters

The following parameters are obtained by assuming that the input port is terminated in a matched load. The determination of reverse transmission parameters \( \{s_{SD2}, s_{SC2}\} \) requires the application of pure differential-mode and common-mode signals at the output port, respectively. Simulation results displayed in Figure 7.8 clearly indicates that the filter has good input-output port isolation characteristics. Therefore, imbalance or mismatch with the impedance presented by the inverter or succeeding differential circuits has a negligible effect on the reflected power wave to the preceding amplifier. Practically the same results were obtained from simulations of transversal filter with different tap gains.

![Figure 7.8 Reverse transmission parameters for a filter with different tap gains](image)

7.2.1.4 Output port reflection parameters

The output port reflection parameters \( \{s_{DO2}, s_{CC2}\} \) are obtained by applying a differential or common-wave, while the input port is matched to the input impedance. Both parameters displayed in Figure 7.9 account for the power reflected to the output port caused by impedance mismatch. Reflection parameters better than \(-18\) dB are predicted in the frequency band of interest. Testing for different filtering functions practically gives the same results indicating that the output reflections are independent of tap settings.

![Figure 7.9 Output reflection scattering parameters](image)
7.2.1.5 Conversion mode parameters

When a common-mode wave is launched at the output port, part of the energy is reflected due to mismatching with the output impedance and part is coupled to the input port. In differential circuits, a remaining amount of energy is converted into differential-mode of propagation which is attributed to asymmetry or unbalance. Unbalance effects in the filter arise since parasitic couplings associated with “feedback elements” of active devices are not completely symmetrical; i.e. parasitic couplings are not completely cancelled by taking the difference between voltage phasors. Such unbalance stems from the non-reciprocity of the distributed circuit as well as the variations of shunt feedback capacitance $C_{gd}$ and output impedance of the active device with applied bias voltage. These effects are not noticed if unilateral devices are used.

Conversion mode parameter $s_{CD22}$ measures the conversion of differential-mode wave into common-mode wave whilst $s_{DC22}$ measures the conversion of common-mode wave into differential-mode wave. Those parameters depend on the filter tap gains. For example, when all filter taps are biased to the same voltage level, the identical electrical characteristics results in an infinite attenuation of the conversion mode parameters; thus a complete cancellation of parasitic couplings takes place. This is clearly not the case for a practical filter implementation. For example, Figure 7.10 displays parameters when the filter is designed for the sequence (+1, +1,-1,+1,-1,-1,+1). Hence, cancellation of parasitic couplings described by the parameter $s_{CD22}$ better than 30 dB can be achieved by assuming ideal inversion at the output port. A similar level of mode-conversion is obtained for other tap gains settings.

![Figure 7.10 Conversion mode parameters](image-url)
7.2.2 Comment on overall response

The performance of the transversal filter can be evaluated in terms of the parameters that affect the filter differential-mode response. The direct gain parameter $s_{DS21}$ determines the filter transfer function under perfect matching conditions; however, when output reflection and mode-conversion parameters are different from zero the filter function depends on the differential components of those parameters. Figure 7.11 allows comparing the parameters that affect differential-mode response. The figure displays a conversion mode parameter of approximately 30 dB lower than the direct forward parameter over a substantial part of the filter bandwidth. The reflection parameter $s_{DD22}$ presents ripples with amplitude increasing with the frequency given the unfeasibility of having perfectly matched impedances at both drain terminals. Figure 7.11 shows that the major parameter determining the response is the forward gain parameter $s_{DS21}$ and the minor parameters are the reflection parameter $s_{DD22}$ and conversion mode parameter $s_{DC22}$.

![Figure 7.11 Components of the differential-mode response for the MMIC filter tuned to two different sequences](image)

A comment on the performance at low frequencies response is fundamental for handling waveform patterns in Optical CDMA applications. When the filter response is not extended down near-dc frequencies, it may result in low frequency distortion such as dc wandering and pulse sag. Simulations in Figure 7.12 shows a 3dB cut-off at 23 MHz and with filters coefficients setting the first equal to +1 and the other six equal to 0. (Simulations for other gain tap settings indicate similar results). Despite of the limitation associated with low frequency response, such as the coupling capacitances and the reactive impedance associated with the output resistance of the devices $R_{ds}$, time domain simulations of the structure show that when using pseudonoise sequences, such low frequency operation does not restrict the application of the design to optical CDMA systems.
7.2.3 Single-ended filter frequency analysis

The addition of an inverter stage in the filter was described in the incremental model in Section 7.1 so that an inverter with unitary gain allows combining linearly signals in anti-phase at the output. The details of such an implementation are shown in Figure 7.13. The series capacitances $C_b$, $C_d$ are tuned and the HEMT gate width $W$ is chosen to achieve inversion. The optimisation commences by the selection of the series capacitance, $C_b$ and device geometry, $W$ so as to obtain an equivalent input capacitance of the inverter stage $M_{inv} \cdot C_{gs}$, ($M_{inv}$ is the capacitive division $C_{gs}/(C_{gs} + C_b)$ and $C_{gs}$ is the input capacitance of the device) approximately equal to overlay capacitors in delay lines. $C_d$ is a capacitance in series with the output resistance which can be adjusted to optimise the response at low frequencies. The implementation was achieved using the HEMT process though with some difficulties. The main drawback in the implementation was the high input resistance of the HEMT that introduces signal distortion in addition to the feedback capacitance which reduces isolation. In order to lessen the effects of the large input resistance, the gate width of the device was chosen to minimise such resistance to 5.7 $\Omega$ by setting $W = 33 \times 6 \mu m$. Following this design approach careful consideration has to be given to the increased dc current loading on the lower drain line, associated with the large area of the inverter HEMT. Figure 7.13 shows the details of the implementation and the forward gain parameter, $s_{21}$ of the single-ended filter. Those initial results suggest that the use of an improved HEMT process could make suitable an implementation with better characteristics.
Fig. 7.13(a) Schematic of the inverter, with $C_d = 0.8 \text{ pF}$, $C_b = 0.11 \text{ pF}$ and $R_g = 5 \text{M}\Omega$

(b) the corresponding simulated $s_{21}$ parameter (black line) and calculated (grey line)

7.3 Analysis of stability

Direct stability analysis was carried out because of the characteristics of the “feedback elements” of the HEMT employed as well as because of the parasitics of passive devices at high frequencies can contribute to a reduction to the resistance to oscillate. The single-ended output transversal filter was analysed using a 50 $\Omega$ impedance system in the filter ports. Stability circles were analysed using small-signal device models in the frequency range from 1 GHz to 50 GHz. A first approach was determining the output load conditions at which the input reflection coefficient, $\Gamma_{IN}$ is lower than one (|$\Gamma_{IN}$| < 1) and the source load conditions at which the output reflection coefficient, $\Gamma_{OUT}$ is lower than one (|$\Gamma_{OUT}$| < 1). Both stability conditions led to draw the stability circles in the load reflection $\Gamma_L$ plane for $|$\Gamma_{IN}$| = 1 and in the source reflection $\Gamma_s$ plane for $|$\Gamma_{OUT}$| = 1. A 50 $\Omega$ impedance load in both ports corresponds to the origin of the chart. Stability circles close to the origin are drawn in Figure 7.14, none of these circles encloses the centre of the chart as the radii of such circles are lower than their centres. The closest circles to the origin of the chart for the source and load have centres at 3.5 (at 33.9 GHz) and 1.7 (38.7 GHz), respectively.
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Figure 7.14 Stability circles in the Smith chart, source stability circle for $|\Gamma_{IN}| = 1$ (left) and load stability circle for $|\Gamma_{OUT}| = 1$ (right)

The above stability analysis, however, cannot indicate which area in the Smith chart (inside or outside of stability circles) provides stable impedance conditions at specific frequency; i.e. the load and source impedances for stability conditions [135].

In order to ensure that the filter is stable at the designed operating conditions, the Rollett stability factor, $K$ and measured stability factor $B_l$ were analysed. This set of parameters ensures stability if $K > 1$ and $B_l > 0$ simultaneously over a wide range of frequencies [135]. Simulation results based on the s-parameters of the transversal filter satisfy the conditions for stability.

Figure 7.15 Stability factor, $K$ and stability parameter, $B_l$

Simulation results in Figure 7.15 show the minimum value for the stability parameter $K$ at 37 GHz, near the frequencies at which the source and load circles are close to the origin of the Smith chart. From those results, it is concluded that for a 50 $\Omega$ impedance load the filter is a stable circuit.
7.4 Comment on noise performance

The HEMT process used is highly suited to high frequency-low noise design [58,63]. Nonetheless, the non-ideal behaviour of the HEMT at high frequencies led to the choice of devices with small HEMT gate areas and, consequently, the transconductance becomes low. In particular, the design of low noise distributed amplifier involves necessarily high level of gain. Nonetheless, it is worth mentioning that the low complexity of the filter may result in lower output noise current densities which will translate into low equivalent input noise, despite of the low gain [101,118].

Figure 7.16 shows the filter schematic to analyse the filter noise. The current sources \( i_{\text{upp},n} \) and \( i_{\text{down},n} \) represent the noise currents of the upper and lower rows of the filter, respectively and includes the thermal noise of terminal resistances. The interconnection of active devices sharing a single input line allows adding the noise current sources that model the gate noise of both active devices. As a result, the output noise currents in both arms present a level of correlation. By performing the inversion at the output port, assuming an ideal BALUN component that combines signals in anti-phase, the output noise density current, \( i_{\text{no}}^2 \) has the equation:

\[
\overline{i_{\text{no}}^2} = (i_{\text{down},n} + i_{\text{upp},n})^2 = i_{\text{down},n}^2 + i_{\text{upp},n}^2 + 2 \cdot \text{Re} \left( i_{\text{down},n} i_{\text{upp},n} \right)
\]  

(7.10)

where \( \text{Re} \) is the real part of the correlation between both noise current sources. Simulation results were obtained for two different sequences. Figure 7.17(a) and 7.17(b) display the output noise current density, (ONCSD) for a all-zero sequence and the sequence (+1,+1,-1,+1,-1,-1,+1), respectively.

![Figure 7.16 Circuit to characterise the output noise density current using a wideband BALUN transformer](image)
The computed noise includes the noise from both terminal drain impedances. It can be shown that in both cases, the noise is approximately equal and results from the effect of biasing cell elements for high and low transconductance gains. The average output current density is equal to 57 pA/Hz and 59.3 pA/Hz for the respective sequence as shown in Figure 7.17.

![Figure 7.17(a) output noise current density, (ONCSD) for an all-zero sequence, (b) the sequence (+1,+1,-1,+1,-1,-1,+1)](image)

The correlation between noise current sources resulted to be (in average) equal to $1.43 \times 10^{-21}$ pA/Hz and essentially associated with the noise elements contributions in the common-gate line.

### 7.5 Transient analysis of the transversal filter

The design of the distributed transversal filter ought to satisfy transient responses defined for its potential application for high-speed encoders/decoders for Optical CDMA. The transient response of distributed circuit is largely influenced by device parasitics and impedance mismatch effects. Unlike first models proposed to predict the transient response of loss-free vacuum-tube distributed amplifiers (for instance [74]), the response of distributed topologies depends heavily on losses of solid-state active devices as well as on the electrical characteristics of transmission lines. For such reason, it is essential to carry out time domain analysis based upon complete models of the devices otherwise the predicted response based on such analysis might lead to large inaccuracies. Theoretical analysis of solid-state distributed amplifiers by Wong and Han [69,134] shows that different sources of coupling-wave interactions, for instance passive parasitic couplings due to active devices, may influence the waveform propagation along artificial transmission lines. In a fairly recent article [76], Bianucci and Aitchison presented analysis of the transient response of solid-state distributed amplifier. Losses and parasitic elements of MESFET-based distributed amplifiers give rise to transient responses that
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differ from analytical models previously established [74]. Using computer simulations, the various components of the transient response were analysed leading to the conclusion that distributed circuits possess superior transient responses when compared to single stage amplifiers. Device losses and impedance mismatch are determining parameters that define the step response. Wong and Han also pointed out the lack of both theoretical and practical work on transient analysis of distributed amplification [69]. Studies on this topic could provide valuable knowledge on transient behaviour of distributed structures and therefore might be of interest for high-speed communication systems.

Analysis carried out in this chapter allows an assessment of the performance of the transversal filter design for high rate pulse applications. The following analysis is based on computer simulations using the fully-characterised models of the available MMIC process. The models of active and passive devices employed are of high complexity for good accuracy in the prediction and those are proprietary of OMMIC [63].

### 7.5.1 Transient response of the semi-ideal transversal filter

In the ideal symbol-rate transversal filter, the response is described by a sum of delayed pulses, each is weighted with a corresponding tap gain as depicted in the block diagram of Figure 7.18. Pulses at the output undergo the same bandwidth limitations associated with each filter tap derivation. In the diagram, the output block corresponds to a linear low-pass filter with transfer function $H_p(j\omega)$. The characteristics of output pulses are influenced by both low-pass effects and the shape of input pulses. In addition, the shaping function $H_p(j\omega)$ possesses a flat amplitude and linear phase functions in the frequency range $(-2f_c, 2f_c)$, where $f_c$ is equal to the filter cut-off frequency.

![Figure 7.18 Block diagram of symbol-rate transversal filter](image)
The low-pass filtering function introduces an infinite attenuation out of the frequency range set by $f_c$. The ideal low-pass filter that predetermines the response of such filter is thus given by the Fourier transform, $\mathcal{F}(\cdot)$ of the sinc function [124,138]:

$$H_p(j\omega) = \mathcal{F}(\text{sinc}(2\pi f_c t))$$  \hspace{1cm} (7.10)

Such filter satisfies the zero-ISI criterion if infinitely short pulses modulated by a discrete-time sequence are applied at the input [124]. A more realistic approach contemplates the use of pulses with finite energy and bandwidth of the order of the filter cut-off frequency, $f_c$. A basic knowledge on the filter behaviour may be obtained by calculating the response using rectangular pulses with pulse width of $\tau$ seconds. Pulse width is set to the reciprocal of the filter cut-off frequency, $\tau = 1/f_c$. The ideal low-pass filter can maintain the bulk of the energy of input pulses to provide suitable transient characteristics. Figure 7.19 shows the calculated response of the symbol-rate transversal filter, assuming the filter gain weights $(+1,+1,-1,+1,-1,-1,+1)$ and an output pulse with a finite width, $\tau$. The resulting peak amplitudes are not equal as in the case when infinitely short pulses applied at the input.

The calculated response shows the transient response as bandwidth limitation of the filter. Since the amplitude and shape of a single pulse is influenced by previous pulses as a result of the oscillatory components in the response, this results in inter chip interference. Figure 7.19 also shows an initial pulse of low amplitude associated with the non-causality of the ideal filter and low oscillatory components. Such results suggest that the peak amplitudes are influenced by low inter pulse interference effects when sharp pulses are applied to the input of the bandwidth limited filter.

Figure 7.19 Calculated response of the ideal symbol-rate transversal filter with normalised time delay $\tau = 100$ and for input rectangular pulses of width $\tau$
Given the complexities that arise in the design using such transversal filter topology at microwave frequencies [95], a semi-ideal filter was designed using distributed amplification principles. In the semi-ideal transversal filter, all the elements of the filter design were maintained as in the MMIC design without change but substituting transmission lines by ideal inductors; *inductor-based* delay lines were introduced between filter taps. It can be verified that the bandwidth limitation associated with the triple-line distributed structure using practical devices can give suitable transient responses for the application. Pulse generation and correlation functions were analysed for various transmit and receive functions using the distributed filter structure [100,118].

For example, filter gains were set to correspond to the (bipolar) maximal length sequence (+1,+1,-1,+1,-1,-1,+1). Figure 7.20 displays simulation results when *inductor-based* delay lines include losses (damping resistor of 5Ω or equivalently by dissipative factor, $a$ equal to 0.16) and without dissipative effects. The case of delay lines with losses is drawn in solid line whilst for delay lines without losses in dotted line. Both waveforms were obtained by using the same tap gain weighs and input pulse profile for comparison purposes.

![Figure 7.20 Filter responses to a single input pulse, response for a dissipative constant equal to 0.16 (black line) and zero losses (dotted line)](image)

The filter response displayed in Figure 7.20 shows pulse generation according to the discrete sequence. In these simulations, the gain weight control technique was applied to counteract losses in artificial transmission lines. The response for loss-free lines presents higher pulse peaks as a result of the different levels of attenuation in both filter implementations.
Simulation results of Figure 7.20 allow contrasting different transient characteristics. In general, the delay between pulses is kept constant following the design guidelines of the transversal filter given in Chapter 6. The first pulse in the response is wider than the input pulse as a result of the inherent bandwidth limitation of the filter. The oscillatory components stem from the application of short input pulses \( \text{i.e.} \) associated with the filter bandwidth-limitation and the multiple reflections. Figure 7.20 shows that the later effect is more pronounced for the case of loss free delay lines, whilst the use of lossy delay lines introduces larger attenuation at higher frequencies, thereby reducing the oscillatory components. In general, the addition of damping resistors in delay lines allows maintaining the pulse symmetry while improving the linearity of the filter function over a substantial part of the spectrum of input pulses. Consequently, lossy delay lines reduce the amplitude of the filter function with the frequency. The use of lossy delay lines reduces further the bandwidth in late pulses and degradation in the rise time becomes more noticeable than in the case of ideal filter.

7.5.2 Transient analysis of the MMIC transversal filter

Analysis of the semi-ideal filter is based on ideal inductances does not reflect the non-ideal behaviour of microstrips such as dispersion, skin effect or frequency dependent attenuation. Among the non-ideal effects of transmission lines, microstrip line losses acquire more relevance in the frequency range of operation, whilst dispersion effects are less noticeable when the inductances are synthesised using high impedance microstrip transmission lines on thick substrates [122]. The transient characteristics of pulses were analysed to obtain a pulse shape profile associated with different filter taps. This was done by weighting each tap independently, applying bias voltages of a single tap for a maximum gain whilst the other six were biased to the reference voltage so as to set a zero tap gain. Figure 7.21 shows the MMIC filter response of all taps settings for maximum tap gain (such results correspond effectively to the time domain version of the frequency characteristics in section 6.9, Figure 6.17). The input pulses have a rise and fall time of 5 ps, FWHM\(^{\dagger}\) of 20 ps and amplitude of 10 mV.

\(^{\dagger}\) The Full-Width-at-Half-Maximum (FWHM) parameter is also called Full-Width-at-Half-Height (FWHH) [85]
The delay between pulses in Figure 7.21 is approximately equal to 25 ps and in agreement with the group delay of the MMIC design. The transient responses of earlier taps shows lower rise time and less attenuation level compared to those of subsequent stages. Little overshoot in the response result from the linearity of the filter phase function. Simulation results in Figure 7.21 shows that the shape of pulses in late stages approaches to the form of a Gaussian pulse. This is consistent with measurements of the transient response of damped linear networks [123,129] showing output pulses approximated to pulses with Gaussian characteristics.

For linear phase filters, the rise time is usually considered to be equal to the pulse width of the filter impulse response. Different criteria have been established for defining the pulse width. For many practical systems, the pulse width can be defined either as the root mean square of the pulse [123,127] or as the full-width-at-half-height (FWHH) parameter [11]. The last criterion allows analysing pulse responses straightforwardly from measurements and is used here for convenience. Figure 7.21 clearly shows that late filter stages have larger pulse widths and larger step response as the high-frequency attenuation increases with the number of filter stages.

A basic criterion for the maximum number of filter stages can be established in which the inter pulse interference is mainly associated with the oscillatory component of the transient responses. For low inter pulse interference response, each stage ought to have a rise time lower than the differential tap delay; i.e. \( t_{r,n} \leq t_d \). Simulation results indicate that filter stages of the transversal filter have a FWHH lower than 25 ps in all filter sections. Hence, low inter-
pulse interference arises in the implementation of a specific filtering function. The above results lead to the consideration that in order to increase the number of filter stages, the filter should be designed with low rise time characteristics and large delay per stage. Nonetheless, the oscillatory components of filter taps must be taken into consideration when the number of filter stages is increased. The oscillatory components of all filter stages may be added in phase depending on the specific filtering function. High oscillation components in the coding pattern can give rise to significant inter symbolic interference (ISI) making this approach inefficient for Optical CDMA systems.

An improvement on the overall filter response can be achieved by increasing the amplitude of pulses. Two main mechanisms contribute to the reduction of the peak amplitude of pulses. These are; the attenuation introduced by artificial transmission lines (losses associated with active devices and transmission lines) and the fact that in damped linear filters the pulse-width amplitude product is constant [123,127], so that peak amplitudes are inversely proportional to pulse width and are reduced with the number of filter stages. Although delay lines with dissipative elements could reduce oscillatory components, it was decided to design the MMIC filter without including the small damping given the high attenuation of the microstrip transmission lines. Other methods to increase the delay whilst maintaining the phase linearity over a significant bandwidth (such as the use of constant R-sections) may be beneficial for increasing the number of filter sections.

The versatility of the transversal filter was tested by adjusting the response to different discrete sequences. As in the case of the inductor-based filter, the MMIC design can provide the required filtering functions although setting higher tap gains for latter taps was necessary so as to compensate for the attenuation introduced by transmission lines. Figures 7.22 and 7.23 show the transient responses when a 10 mV input pulse is applied. In both cases, normalised peak amplitudes correspond approximately to the filter sequence. The response in Figure 7.22 shows an undershot of 21%, whilst the response in Figure 7.23 has an undershot of 15%. Such differences depend on the oscillatory components which are added in phase or anti-phase according to the sequence used.

Results displayed above show low oscillatory components as a result of the inherent losses of microstrip lines. Table 7.1 shows the applied tap gain for time domain simulations displayed in Figure 7.22 and 7.23. Additionally, those tap gains were used for producing the scattering parameters in the section 7.2.
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Figure 7.22 Transient response of the filter to a single pulse
tuned to the sequence (+1,-1,+1,-1,-1,+1,+1)

Figure 7.23 Transient response of the filter tuned
to the sequence (+1,+1,-1,+1,-1,-1,+1)

Table 7.1 Sequences and tap gains for two different sequences (7-tap MMIC filter)

<table>
<thead>
<tr>
<th>Sequence 1</th>
<th>+1</th>
<th>-1</th>
<th>+1</th>
<th>-1</th>
<th>-1</th>
<th>+1</th>
<th>+1</th>
</tr>
</thead>
<tbody>
<tr>
<td>Tap gains</td>
<td>19.60</td>
<td>20.64</td>
<td>30.72</td>
<td>32.20</td>
<td>30.10</td>
<td>32.30</td>
<td>36.13</td>
</tr>
<tr>
<td>(mS)</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Sequence 2</td>
<td>+1</td>
<td>+1</td>
<td>-1</td>
<td>+1</td>
<td>-1</td>
<td>-1</td>
<td>+1</td>
</tr>
<tr>
<td>Tap gains</td>
<td>17.29</td>
<td>21.64</td>
<td>30.72</td>
<td>32.20</td>
<td>30.10</td>
<td>32.20</td>
<td>28.96</td>
</tr>
<tr>
<td>(mS)</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>
7.5.3 The transversal filter analysis as a matched filter

A fundamental consideration of a filter to be used in Optical CDMA systems is that of its ability to detect waveforms generated by sequences that present certain correlation properties. In our practical approach, waveforms consisting of short pulses are applied to the input port of the filter. The broadband operation of the receiver ensures the reception of high-rate sequences. For a matched filter in the receiver, the bandwidth sets a limit on the chip rate whilst the number of taps corresponds to the length of the sequence and therefore sets the bit rate.

In order to test the capability of performing the desired convolution functions, reciprocal sequences were used, following the methodology proposed in [136]. It has been proved that reciprocal sequences have the property of impulsive correlation functions; i.e. the response of a transversal filter provides an appertaining function with zero lateral lobes when the input sequence matches that of the filter sequence. This theoretical result allows for the application of reciprocal signals to test the cascade of transversal filters. Hereby testing the encode and decode functions together.

The construction of reciprocal sequences starts by choosing a maximal length sequence, for example; the binary $m$-sequence of length 7 (+1,+1,0,0,+1,0,+1) [137]. A straightforward algorithm is then applied to obtain the reciprocal partner. Time inversion of the sequence $(a_1,a_2,\ldots,a_N) \rightarrow (a_N,a_{N-1},\ldots,a_1)$ and the mapping (+1$\rightarrow$ +1 and 0$\rightarrow$ -1) of each element of the sequence yield the bipolar reciprocal sequence (+1,-1,+1,-1,-1,+1,+1), i.e.

$$ (+1,+1,0,0,+1,0,+1) \text{ time inversion } \rightarrow (+1,0,+1,0,0,+1,1) \text{ mapping } \rightarrow (+1,-1,+1,-1,-1,+1,1) $$

The above bipolar sequence corresponds to the filter sequence 1 in Table 7.1. The corresponding gain weights were used for analysis. A periodical signal based on the $m$-sequence and period of repetition of 5.71 GHz ($= (7 \times 25 \text{ ps})^{-1}$) is used as the input signal. The profile of chip pulses has the same characteristics as that of single pulses used to produce the results in Figure 7.22 and 7.23. The response of the filter, which is effectively the convolution of periodical signals, is depicted in Figure 7.24.
The filter response is in good agreement with an impulsive signal with low side lobes. The equal amplitude of the appertaining function each 175 ps over the time span confirms the capacity of reception of waveforms with zero ISI. The filter provides the functionality for encoding data at 5.7 Gbit/s or equivalently at 40 Gchip/s.

The aperiodic convolution function was analysed so as to corroborate that the filter can provide the suitable response. The aperiodic convolution between the $m$-sequence $(+1,+1,0,0,+1,0,+1)$ and the reciprocal $(+1,-1,+1,-1,-1,+1,+1)$ gives the pulse sequence:

$$
(+1,+1,0,0,+1,0,+1) \otimes (+1,-1,+1,-1,-1,+1,+1) = (1,0,0,0,-1,-1,4,-1,0,0,0,+1,+1)
$$  (7.11)

where $\otimes$ denotes the aperiodic convolution. Figure 7.25 shows the transient response of the filter to the aperiodic sequence.

Simulation results are in close agreement with the aperiodic sequence of Equation 7.11. It is noted though that negative pulses possess higher amplitude and distortion, such results
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correspond to non-optimised gain weights of the transversal filter (see Figure 7.22). The performance of this filter is analysed by verifying that the aperiodic convolution is a time-limited response. If the response of an aperiodic sequence in the time interval \( |t - t_0| \geq 2N T_{chip} \) is equal to zero (with \( t_0 \) equal the peak of the in phase auto-convolution function and \( N \) equal to the tap numbers) then the filter satisfies the first Nyquist criterion. This is the case for the response of the filter in Figure 7.25, in which the maximum pulse amplitude is equal to 1.65mV. This peak amplitude is approximately equal to the peak amplitudes of the 175-picosecond spaced pulses obtained from the periodic correlation function (see Figure 7.24). It is concluded that the filter introduces zero ISI and therefore is appropriate for Optical CDMA system applications.

7.5.4 The cascade of transversal filter

The cascade of transversal filter was tested with different sequences as well as filter implementations via simulations [100]. Bipolar and multilevel sequences can be generated / filtered using filters based on the triple-line topology. In particular, a double-ended transversal filter as a stage in the transmitter and a single-ended filter at the optical receiver could be used for coherent and non-coherent implementations. To prove the ability of the transmitter and receiver filters to operate together as sequence encoder and detector, the transmitter and receiver MMIC transversal filters were tested in cascade assuming for such aim a perfect signal coupling between both filters. The methodology of reciprocal sequences was used for such aim. The pulse response of the single-ended receiver is shown in Figure 7.26 showing the generation of the pulse pattern, although higher oscillation components when compared with results of the two-ended filter displayed in Figure 7.23.

![Response of the single-ended transversal filter](image)

Figure 7.26 Response of the single-ended transversal filter programmed with the sequence \((+1,+1,-1,+1,-1,-1,+1)\)
On the other hand, the reciprocal partner of the sequence in Figure 7.26 was programmed at the transmitter. The differential response of the filter to a single input pulse is shown in Figure 7.27. The response shows a large oscillatory component and stems from the addition in phase of the oscillatory components of filter stages which usually are cancelled when implementing a bipolar sequence.

![Figure 7.27 Response to a single pulse of the filter](image)

The two transversal filters, with tap gains biased to produce the responses shown in Figure 7.26 and 7.27, were cascaded, assuming perfect signal coupling. The response is shown in Figure 7.28 where the response shows an impulsive autocorrelation function and input periodic pulses (nearly 25 ps pulses and pulse repeating at 5.71 GHz)

![Figure 7.28 The response of cascade stages to a input pulses with repetition of 5.71 GHz](image)
Other simulation was carried out to analyse the effect of oscillatory components in the transmitter. In this case the transmitter was programmed with a bipolar sequence but adding a DC voltage level at the output (25 mV) to provide a unipolar sequence at the input of the receiver. The aim is lowering the oscillatory component as such components are added in anti-phase. The single-ended filter described above was used in the receiver. Figure 7.29 shows the transmitter’s response when a periodic sequence of pulses is provided at the input and the response of the cascaded stages. Figure 7.29 shows lower side lobes in the response when compared with results in Figure 7.27 and stems from the lower oscillation of the transmitter response.

7.6 Transversal filters with positive and negative gain weight control

Having designed and analysed the MMIC transversal filter for Optical CDMA applications, it is necessary to examine other filter topologies with similar functional capacities given that the selection of the filter topology is a fundamental part of the design process. Although the author is unable to present measured results of the transversal filter and then make some comparisons with measurements with other alternatives, part of the circuit design process
carried out in this work is to identify the potentialities and limitations of other filter cell topologies that eventually could be designed for Optical CDMA system applications.

Regarding the development of MMIC transversal filters suitable for the reception of bipolar signals, the filter proposed by Lee and Freundorfer [25] was the first MMIC design with the capacity of extending filtering functions with positive and negative gain weights; a five-tap transversal-filter based on Gilbert cells was developed and tested for duobinary partial-response receivers. Measurements of the fractionally-spaced filter reported in [25] show frequency and time domain responses suitable for the reception of duobinary signals. In a very recent paper (December 2003) Wu, et. al. [116] described a 7-tap 10 Gbit/s integrated transversal equaliser using a differential topology and based on a variation of the Gilbert cell design of [25]. To the best of the author’s knowledge, apart from the above developments, there are no reports of other transversal filters with bipolar capacity for high-speed lightwave systems.

### 7.6.1 Gilbert cell description

The Gilbert cell is a useful component in analog electronics given its capacity to control the level of amplification and phase inversion by using external voltages. The cell is a balanced circuit whose similar electrical characteristics on both branches allow amplifying differential signals and rejecting common-mode signals in function of the likeness of the electrical characteristics of active devices. Figure 7.30 shows the basic Gilbert cell where pure differential-mode signals are applied to the input differential port. The input pair stage feeds differential currents to the two upper source-coupled pairs. An external control voltage $V_{Gn}$ sets dc currents on the branches of the differential pair and controls the device transconductances. A differential current is obtained in external nodes whose amplitude depends on the difference between device transconductances [132].

Gilbert cells satisfy internal strong mismatching between the high output impedance of the input transistor and the low input impedance of the source-coupled stage ($Z_{in}$ see Figure 7.30), which gives rise to a low frequency dependence on the coupling of the cell stages. Due to the internal mismatching, the transconductance gain can be constant over multi-octave bandwidth [98]. In addition, the input impedance of the cell is maintained practically constant regardless the programmed gain weights. The utilisation of Gilbert cells in distributed circuits has two key advantages. First, differential pairs of the cells are connected directly without the need of interestage coupling capacitors. Its implementation involves a transistor-level design. As a direct consequence, a filter based on such cells can be designed to amplify signals at dc [25]. Second, the high level of isolation between the input and output ports relaxes the
requirement of the figure-of-merit of active devices whilst the high output impedance facilitates the design of drain artificial lines.

An important consideration in the MMIC filter design is the need of biasing the cell using low voltage levels. For such aim, active load impedances and current mirrors are implemented in a transistor level design. For low frequency applications, the tail current source can be modelled as an ideal current source in shunt with a tail resistance as depicted in Figure 7.30 [132]. When the differential pair is perfectly balanced and differential signals are applied at the input, a constant voltage appears across \( R_{\text{tail}} \). The superposition of currents flowing into \( R_{\text{tail}} \) results in current cancellation; therefore the resistance can be replaced by a short circuit in the small-signal model of the Gilbert cell. This results in a negligible dependence on the tail current source.

![Figure 7.30 Conventional Gilbert cell](image)

### 7.6.2 Transversal filter topologies

Two topologies based on Gilbert-cells have been implemented and tested as MMICs [25,116] and those are depicted in Figure 7.31. The fundamental difference between those filter implementations lies in the method of interconnecting distributed cells. The use of the differential filter topology [116], first suggested by Rauscher in [89], allows filtering of differential-mode signals and provides a level of rejection of common-mode noise appearing at input ports. Additionally, this configuration presents low dependence on the non-ideal behaviour of current sources since the cell works effectively as a balanced circuit. Given the balanced characteristics of the Gilbert cell, it is possible to switch the sign of tap gain without changing the impedance load on output artificial lines [116]. Conversely, the single input / output line topology permits more compact implementations. The differential topology offers a
higher gain per cell as the level of amplification of a single-ended cell is half of the gain of
differential cell assuming the use of identical active device and the same characteristic
impedances.

![Differential transversal filter topology](image1.png)

**Figure 7.31 Transversal filter topologies with bipolar tap gain weight control**

The schematic of Gilbert cells as implemented in the single-ended [25] and differential
topology [116] are shown in Figure 7.32. The implementation based on the single-ended
transmission line topology utilises Gilbert cell with an input terminal fixed to a reference
potential whilst an output node is connected to an internal resistance. The Gilbert cell in the
differential topology [116] (implemented in Bi-CMOS technology) is depicted in Figure 7.32.b.
This cell design has introduced other elements to improve gain control. Unlike the basic
structure, gain weight control is achieved by adjusting the tail current source, $I_w$ whilst the sign
is set by the difference relative to the potentials applied to the nodes $S_1$ and $S_2$. This bias scheme
has the advantage of increasing the linearity of the gain function against applied control voltage.
Nonetheless; the changes induced by the gain weight control modifies the bias condition of the
input pair giving rise to changes in the impedance loading on the input line and associated
problems. For this reason, additional input common-collector stages ($Q_{b1}, Q_{b2}, I_{b1}, I_{b2}$), not present
in the basic topology, are included in the differential structure so as to buffer input signals from
the differential inputs pairs ($Q_1, Q_2$). Input buffer stages allow for reducing the loading on input
artificial lines and maintain the impedance loading constant for different tap gain weights [116].

![Single-ended Gilbert cell-based transversal filter topology](image2.png)
In the filter implementation based on single-ended MMIC, a self-aligned MESFET process with an $f_T$ equal to 20 GHz was utilised. The delay per stage was set to 40 ps and the transmission lines cut-off frequency (Bragg cut-off frequency) was made equal to 25.5 GHz [25]. The testing speed of modified duobinary signals was 2.5 GHz [25] which is much lower than the filter cut-off frequency thus tested at rates lower than the chip rate. The differential topology as implemented in [116] was tested to equalise data at chip rates approximately equal to the filter 3 dB bandwidth, showing equalisation at chip rates of 10 GHz. In order to achieve such speed operation, the filter was designed with a BiCMOS process with an $f_T$ equal to 120 GHz and delay per stage of 50 ps. It is important to notice that the maximum chip rate is much lower than the figure-of-merit ($f_T$) of the active devices.

### 7.6.3 Discussion of filter proposals with bipolar capacity

It is apparent from the above schematics that the differential topology has the highest complexity and number of devices in input / output transmission lines. The use of such topology so as to retain the advantages of balanced circuits possesses important challenges to be undertaken. The implementation in [116] requires devices with high $f_T$ and additional buffers to reduce excessive load on base lines. As it is shown below, the gain per cell of the Gilbert cell is lower than the transconductance of the constituent active devices.
The topology in [25] does not require differential line configurations since the gate terminal of the input device is connected to a dc voltage source. The use of the cell as a gain block with a single-ended input port must be considered carefully. The main limitation to the performance is the unequal voltage applied to the devices of the input pair, which results from the finite conductance of the tail current source. This unequal impedance gives rise to different currents flowing into the resistance of the tail current $R_{\text{tail}}$ that do not maintain a 180-degrees relationship over the filter bandwidth. In addition, noise of the tail current source cannot be cancelled. Consequently, the overall Gilbert cell performance is deteriorated as its behaviour deviates from that of a balanced circuit. At this stage, it is difficult to determine the mechanism that sets 3 dB point of the filter since the frequency behaviour is influenced by the shunt impedance of the tail current source [132]. In [25] there is no analysis of the bandwidth limitation. Measured results show equalisation functions up to 8 GHz, lower than the filter cut-off frequency. A cell design with bandwidth much lower than the Bragg cut-off frequency of artificial transmission lines does not allow for equalisation at the chip rate.

An important advantage of Gilbert cell is that of the high isolation. By contrast, the triple-line filter of this thesis may be prone to losses associated with feedback effects of active devices. MMIC processes such as self-aligned ones best-suited for distributed amplifiers [120] (or a process that has a high $C_{gs}/C_{gd}$ ratio and low $R_s$) could be used to improve the performance of the design. Conversely, Gilbert cell-based designs are prone to asymmetries that result from the changes of the electrical characteristics of active devices and their limited repeatability. For instance, variations in doping or thickness of the channel region could influence changes in the pinch-off voltage, which may not be the same for FETs located in a different location of the MMIC. As a result, distributed cell can be exposed to imbalance and mismatch-induced effects in differential cells.

Two comparisons can be made between the Gilbert cell and the triple-line transversal filter implementations in the designs. First, Gilbert-cell implementations require large bias voltages that impose practical limitations on components of the MMIC. For instance; Gilbert cells are biased with a constant dc current regardless the programmed gain. If dc currents flow into microstrip lines, the current handling capacity of the transmission lines limits the number of stages. In the case of the triple-line transversal filter, the dc current is split into two transmission lines, therefore, their implementation is less restricted by the current handling capacity of transmission lines allowing a reliable design for higher bandwidth operation.

Regarding the single-ended topology, Gilbert cell results in a transconductances gain lower than that of its differential counterpart. For example, the normalised transconductance of the cell was plotted against the control voltage. This effectively is a linear gain-control with
applied voltage. Figure 7.33 depicts a maximum transconductance of about 200 mS/mm which is approximately a half of the maximum extrinsic transconductance of the HEMT (478 mS/mm). The reduction on the gain stems from the low internal impedance, $Z_{m}$ that is the load of the input stage. The gain of the Gilbert cell cannot be simply increased by choosing larger devices to implement the emitter-coupled pairs, since the associated increase of transconductance is offset by the reduction of the device output impedance ($Z_{in}$ in Figure 7.30).

![Graph of transconductance against input control voltage](image)

**Fig 7.33 Transconductance against input control voltage (using MMIC HEMTs)**

An important shortcoming of the single-line Gilbert cell implementation is that half of the output current is dissipated by internal impedance; thereby resulting in effective gain reduction. By contrast, in the triple-line transversal filter design, tap gain reduction arises when the voltage division capacitor is used; however, this allows enhancing the filter bandwidth. It has been shown that the capacitive division technique facilitates the filter design to a predetermined cut-off frequency of the gate transmission line.

Finally, the main advantage of the triple-line structure is in its simplicity and the low number of its active devices. Distributed structures are based on the concept of absorption of active device parasitics resulting in wideband structures, which are then translated into high-speed operations. By contrast, for the Gilbert cell design, additional capacitances in parallel with the gain blocks to increase the linearity of the filter sections (see Figure 7.31) suggest that the complex nature of the input impedance of the cells should reduce further the filter bandwidth. A filter cell with low complexity is more appropriate when pulse shape maintenance is a foremost design consideration.

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1 It is worthwhile mentioning that the Gilbert design proposed by Wu, et al. improves the linearity of the gain, which is important for the generation of coefficients in decision-feedback equalisers [116]. For optical-CDMA system, this requirement is not essential as the filtering function is previously defined by tap gain biasing.
Summary

In this chapter, analyses of MMIC distributed transversal filters have been presented. In particular, the capabilities and limitations of the transversal filter were studied through simulations. We demonstrated that the designed MMIC transversal filter allows handling at high rate sequences. The potential of such filters was assessed using frequency and time domain simulations and models of active and passive MMIC devices. The MMIC filter design is versatile and can be used for reconfigurable transmitters and receivers at speeds of multi-Gbit/s.

An approach to the transversal filter for Optical CDMA systems considers the analysis of an ideal filter assuming rectangular pulses at the input and weighted output pulses passing through an ideal filter. The calculated response of the “ideal sampler” presents pulses with low inter pulse interference as a result of the linear phase and the inherent bandwidth limitation. A more practical approach encompasses the analysis of distributed structures based on distributed amplifier principles, in particular the transversal filter based on the triple-line topology. The semi-ideal filter was designed based on MMIC devices, but using ideal inductors. Two cases were analysed, by interconnecting loss-free and lossy delay lines between active cells. The behaviour of both filters was compared showing that delay lines with dissipative elements can reduce the oscillatory components and increase the linearity of the filter function and improve the performance. However, frequency-dependent attenuation arises as a bandwidth limitation of the filter hence limiting the number of filter stages in the implementation. Inter pulse interference associated with the effects mentioned above can be compensated by tap gain weight adjustment so as to improve the quality of the response.

Time domain simulations for a single-ended and differential-output transversal filter were described. In particular, transient domain simulations of the MMIC transversal filter allow verifying that the filter can encode and decode data with negligible ISI. The response to a single sharp pulse showed low undershoots (15% and 21%) for two different filter sequence implementations. In addition, in order to compensate filter losses for those two sequences, the difference between the transconductance of the first and seventh tap gain can be of the order of 5 dB. For a 7-tap MMIC transversal filter, a criterion was established by which the response is mainly limited by undesired oscillations. When the step-response of rise time is lower than the time delay for all filter taps, the response for a specific sequence is limited by oscillatory components generated by filter taps. At those conditions, the encoding/decoding functions are achieved with practically zero-ISI.
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The methodology of differential scattering parameters for the analysis of the triple-line transversal filter was introduced. Derived scattering parameters for the transversal filter (which is detailed in Appendix C) allows wave-based analysis of the triple-line transversal filter. Frequency domain simulations based on such framework allow affirming that the MMIC transversal filter can provide the required performance and functionality as the filter function can be tuned to a predetermined sequence. The direct gain parameter of the filter determines the filter transfer function under perfect matching conditions, however, when output reflection and mode-conversion parameters are different from zero, the filter function depends on the differential components of those parameters. An input VSWR ratio of 1:1.5 was sustained over a substantial part of the bandwidth operation, the output reflection parameters (practically independent of the tap gains) are lower than -18 dB and the common-to-differential conversion mode parameter is approximately 30 dB lower than the gain parameter. The major parameter determining the response is the forward gain parameter and the minor parameters are the reflection and conversion mode parameter. The low 3dB cut-off for different gain weights resulted to be equal to 23 MHz and does not depend on tap gain settings. This effectively indicates broadband behaviour regardless the pseudonoise code used.

Noise analysis was carried out for the triple-drain line transversal filter. The average output noise current density of the transversal filter results to be equal to 57.9 pA/√Hz. Simulation results for different tap gain weight settings result in similar noise densities. The stability of the single-ended transversal filter was also analysed for a wide range of frequencies. The Rollet factor, $K$ was always greater than 100 and the stability parameter $B_1$ was greater than 0.8, satisfying the conditions for unconditional stability.

The cascading of two transversal filters was analysed using reciprocal sequences. To prove the ability of the transmitter and receiver filters to operate together as sequence encoder and detector, the circuits described above were tested in cascade, assuming perfect signal coupling between the two. Through the use of reciprocal sequences, the potential of such filters for the reception / generation of waveforms featuring pre-specified correlation properties were demonstrated. Given the significant level of oscillations in the MMIC transversal filter, the case of multi-level sequences was analysed using the semi-ideal approach.

Towards the end of the chapter, we carried out a comparison of the triple-transversal filter developed in this thesis and Gilbert cell structures. However, practical limitations such as the complexity in its implementation and the need of additional elements to reduce the load in gate transmission lines (such as capacitors or additional buffers) constitute an important impediment for using the Gilbert cell based transversal filter at high bit rates. We conclude that the filter presented here is an attractive option for OCDMA systems.
CHAPTER 8

Concluding remarks

This thesis has addressed MMIC circuit design for high-speed lightwave systems, the studies are focused on the development of systematic techniques to design and study the behaviour of multi-Gbit/s encoders and decoders. Based on distributed amplification concepts, a novel transversal filter topology has been proposed for generation / filtering of high-rate sequences. A gain and delay partitioning technique was specifically designed for symbol-rate transversal filters. Additionally, a gain weight control that maintains impedance and delay uniformity has been proposed and its application is based on physical relationships between device capacitance and transconductance of HEMTs and MESFETs. This novel design concept was established and its practical modelling issues were presented in detail. The design criterion is based on the maintenance of pulse shape integrity, which entails specific design variables not considered in conventional distributed circuit design. A novel distributed transversal filter that extends the range of filtering function using positive and negative filter tap gain weights was designed as a MMIC. Throughout this thesis, such filter is dubbed the triple-line transversal filter given that the circuit topology consists of two separate drain lines, a common-input (gate) line and an inverter to converts differential signals into single-ended bipolar signals.

This thesis provided an overview of the main multiplexing techniques for lightwave systems as well as the current limitation of the technology. Many of the advances of optical devices impact immediately the design of fibre networks. It is shown that in order to design flexible optical networks, the optical devices have to satisfy stringent requirements which cannot be solved with the existing technologies. This sets the example of novel (hybrid) multiplexing techniques that relaxes the technological requirements while provide multiple
access capability. Due to its great advantages, an FDM-CDM system was reported that enables multiple access without requiring network frequency or time management.

Chapter 3 has introduced encoding techniques for OCDMA fibre networks that eventually could be used with the distributed structures proposed in this thesis. In particular, it was shown that unipolar-bipolar correlator receiver and SIK technique can be used to improve the detection of CDMA signals. It is established that the limitation associated with the inability to detect bipolar signals can to some extent be counteracted by modifying the receiver structure in order to compensate for code imbalance effects. This opens up some prospects in possible techniques that can be implemented for time-domain encoding.

Chapter 4 carefully examined the principles of operation of distributed amplifiers and others related structures. It is noticed that a main factor in the limited tunability of adaptive filters is associated with the variation of the device parasitics when the filter is tuned to a predetermined filtering function. This issue was stressed throughout this chapter since the maintenance of electrical characteristics of the structures allows maintaining pulse shape integrity, which becomes a major concern in the design of transversal filters studied in following chapters. The distributed amplifier and transversal filter structures presented in this chapter share, as a principal characteristic, their construction by distributing the parasitic capacitance of active devices along lines. In fact the capacity of the distributed amplifier to attain wideband operations lies in the simplicity of the impedance of the distributed cells. The more complex the device impedance the lower its operation bandwidth is. This simple principle is also applied to the filter structures; however, with some restrictions given that in the process of tuning the response modifies the parasitics of active devices, thereby modifying the transmission characteristics of artificial transmission lines.

Chapter 5 is the first of three chapters detailing the research into the triple-line transversal filter and its application. This proposal presents two important features that allow wide bandwidth operations. First, the constant distributed characteristics for continuos tap gain weight settings becomes a fundamental condition to maintain pulse shape integrity. Secondly, the simplicity of the cell impedance (when compared with other filter structures) enables wideband impedance matching of distributed cells with additional delay sections. In this regard, capacitance division technique was established for the filter cell design as a method to enhance bandwidth when appropriate. Second order effects of the filter cell design were analysed using the electrical characteristics of a HEMT. In such approximated modelling, the input capacitance of the cell is increased whilst the parasitic input resistance and inductance are reduced given the parallel connection of two active devices sharing a common gate terminal. This straightforward model
Chapter 8 Concluding remarks

was considered for filter design in the following chapters as well as to explain the process of signal generation in the distributed structure, pointing out the description of mixed-modes of propagation on artificial transmission lines for filter analysis. A method of modelling filter sections was described, showing that the capacitive coupling technique can be used for effectively enhancing the bandwidth of the filter.

Chapter 6 began by obtaining the figure-of-merit of a HEMT process employed in the design. It was found that the employed HEMT device presents low isolation characteristics which in turn increases losses and lower the bandwidth. More complex models (than those provided by the manufacturer) need to be considered for determining the $f_T$ of the active device. It was corroborated that the $C_{gd}/C_{gs}$ ratio plays a major role in determining the response at high frequencies. An ‘intuitive’ model for the filtering of short pulses was established. Such model is in full agreement with simulation results obtained in the following chapter and establishes that the inherent bandwidth of the filter and the dissimilarities of the response of the filter taps make the optimisation in time and frequency domains two different processes. Delay line circuits suitable for on chip implementations were introduced as well as the parameters that permits their characterisation. The concept of Bragg cut-off frequency was used to predict accurately the delay per section of the transversal filter. This was used to derive a formula for the filter design in which two variables, the capacitive division ratio and input capacitance of the cell, can be chosen for achieving a specific bandwidth. A criterion was chosen to link the transmission lines Bragg-cut of frequency with the 3dB cut-off frequency of the transversal filter. The criterion establishes that in order to reduce the effect of impedance mismatch at the output and minimise the group delay distortion, the filter 3 dB cut-off frequency is set at half the Bragg cut-off frequency of the transmission lines. This methodology was assessed in a design based on a popular MMIC process. Simulation results showed that the above mentioned criterion was appropriate for filter implementation. A MMIC 7-tap transversal filter was designed using microstrip technology for 40 Gbit/s systems applications. Layout of the filter indicates the efficacy of the design techniques proposed in the thesis.

Chapter 7 provided an assessment of the MMIC transversal filter via computer simulations. It was shown that the MMIC transversal filter has an input VSWR ratio of 1:1.5 with output reflection parameters practically independent of the tap gains and better than -18 dB. Both parameters are sustained over a substantial part of the operation bandwidth (close to 35 GHz). In addition, the low 3 dB cut-off frequency was equal to 23 MHz and is independent of tap gain settings. Those parameters indicate broadband behaviour regardless the coding. Stability analysis of the MMIC transversal filter was also carried out, showing appropriate stability.
Chapter 8 Concluding remarks

parameters at the operating conditions. In addition, mixed-mode propagation analysis was developed for complete modelling of the designed three port structure. Using such analysis, it was confirmed that the MMIC transversal filter has suitable transmission characteristics for reconfigurable filters. A single-ended filter design was analysed showing that this approach could be considered for a MMIC implementation when a high unilateral and low loss HEMT process is used. A set of time domain simulations were carried out on the distributed structure. Simulation results show that transmission line characteristics play a significant role in the transient response of the filter. It was proved that inductor-based transversal filter in general presents lower attenuation per section and higher transient components, which in turn increase inter pulse interference. The use of small damping resistors in delay lines improves the overall response as low transient components and more symmetrical pulses are obtained at the output. In spite of the above, the MMIC transversal filter was designed without dissipative components since microstrip transmission lines introduce a high level of loss and the transient responses were appropriate for the application. In such distributed filters, the tap gain compensation results in an efficient method for reducing the inter pulse interference that results from the combination of frequency-dependent attenuation and pulse dispersion. The ability of the transmitter and receiver filters to operate together as sequence encoder and detector was tested. Through the use of reciprocal sequences, the potential of such filters for the reception / generation of waveforms featuring pre-specified correlation properties were demonstrated. The filter satisfies the first Nyquist criteria which is essential for time domain encoding. By testing with different sequences and showing its effective generation and reception, without compromising the performance, it is shown that the proposed filter is versatile and the developed methods are appropriate. The resulting structure can be used in a variety of applications where pulse shaping and filtering is needed at rates up to several tens of Gbit/s.

8.1 Suggestion for future research

The areas for future research that may be carried out comprise practical and theoretical work and can be summarised as follows.

- A practical assessment of the transversal filter concept is fundamental for further research. In implementations using the triple-line transversal filter topology, HEMTs can provide good gain per stage and matching conditions on artificial transmission lines. In fact, increasing the number of stages in implementations requires low losses and good matching among stages, therefore the use of active devices with high $f_T$, low source (input) resistance, $R_s$, and large $C_{gs}/C_{gd}$ ratios should result in lower losses per stage. Similarly, lower
source resistance can reduce the low cut-off frequency near dc. Additionally, the use of such devices can increase the number of filter stages as a consequence of reducing the oscillatory component in the response via the enhancement of the bandwidth. Other possibilities, which were not explored given the limitations of the process, were the use of cascode stages (the common-gate stage was prone to considerable feedback effects and instability) and active matching techniques in gate artificial transmission lines so as to reduce mismatching in terminal impedances. Explorations of designs based on the above may lead to modified designs with enhanced performance.

- The design presented in this thesis was based on microstrip technology, in which the via hole of the process has a large inductance which in turn reduces the delay per stage. Consequently, six delay sections (per stage) were needed to obtain the required response. This is a relatively high number when compared with other implementations. An important improvement on the overall filter performance could be attained by designing delay lines with bridged-T sections. This brings substantial benefits for transient responses since as the delay per section is increased, the phase delay becomes more linear over a substantial part of the operating bandwidth and the amplitude-frequency characteristic exhibits a Gaussian frequency dependency. Given that the delay per stage is increased, low number of stages is needed to achieve a specific delay; thereby the transient characteristics of pulses should present lower dissimilarities. In addition, the framework of differential scattering parameters derived in this thesis together with electromagnetic simulators could results in an interesting alternative to analyse parasitic couplings between transmission sections. Couplings in densely packed inductors (as in the case when bridged-T is used) can be analysed using electromagnetic simulation; the differential structure ensures good cancellation of parasitic couplings which could be conveniently exploited for compact implementations.

- Improvements on bandwidth and delay per stage may allow increasing the number of stages to twelve or beyond, whilst maintaining good performance. The design of a transversal filter with such number of stages ensures the use of certain codes (structured codes for instance) with suitable correlation properties. Other possibility is the use of multilevel codes as the filter structures allows for continuos gain settings. It is important to mention, that in the implementation of a specific sequence, the noise associated with transient components can reduce the number of amplitude signal levels. Prior to embarking on the signal design for OCDMA signal for this application, it could be convenient to test and characterise the performance of the filer using the techniques developed in this thesis. For implementations
Chapter 8 Concluding remarks

with a number of stages larger than eight, the use of multilevel \( m \)-sequences and their reciprocal partners can be used for testing.

Further exploration of the transversal filter and interconnection with external RF sources and equipment needs to be addressed. Aspects of particular importance are:

- The development of a methodology for measurements of practical structures in the frequency and time domains. For such an aim, suitable bias sources needs to be designed and their transient characteristics analysed carefully to avoid damage of the MMIC.

- The development of design techniques to cascade two or more transversal filters or consider more complex configurations could be envisaged. That could improve the number of filtering functions or when appropriate the detection of the coded signals.

- Investigation of the interconnection to or design of circuits suitable for driving Mach-Zehnder or electroabsorption modulators with suitable chirping characteristics and low extinction ratios.

In summary, the thesis has examined in detail circuit structures and design techniques for the realisation of circuits capable of handling high rate sequences. This has potential uses in a variety of applications where pulse shaping and filtering is required at high rates extending into the tens of Gbit/s regime.
Appendix A

MMIC Components and Parameter Extraction

This Appendix describes the basic components used in the HEMT foundry process used to implement the triple line filter and describes the process followed to re-extract the small signal equivalent parameters of the HEMT used. The details of the components in the chapter are given taken from the manufacturer’s data[63]

The ED02AH process is a GaAs MMIC multi-layered foundry with active and passive components characterised up 60 GHz and is well-suited for high-speed optical communication systems [63]. Enhanced and depleted-mode HEMT are obtained by metal organic molecular beam epitaxy (MOVPE) using an industrial multiwafer system. GaAs semi-insulating substrates with thickness of 100 μm ensure low losses in the buffer layer of HEMTs. The MMIC process has three metallisation layers, dielectric layers and mesa layer that enable the implementation of active devices (pseudomorphic-HEMT and Schottky-barrier diodes) and different passive components. The three metallisation layer can be combined to form a low-loss multi-metal transmission lines (multi-layered microstrip transmission lines). The first metal layer is used to form the bottom plate of metal-insulator-metal (MIM) capacitor, the under pass interconnection of resistors and recessed diodes. A second metallisation layer is used for a contact layer for NiCr resistors and for interconnection of HEMT gates with metal lines. A top metallisation layer is added to produce transmission lines, MIM capacitor top plates and bond pads.

In addition, the process has silicon nitride layer to form MIM capacitor dielectrics and Silicon oxide for interconnection of the second metal layer. The last is also used for passivation layer covering the whole die with the exception of the bonding areas (street dicing).
Appendix A  MMIC components and parameter extraction

A.1. Passive devices

All the devices in the process, except planar spiral inductors, have been modelled from DC up to 60 GHz, while spiral inductors can provide a free-resonance inductor behaviour up to 25 GHz. The models of such devices have been included in the ADS (Advanced design systems) simulation tools and layout generation. In particular, some passive devices considered for the final MMIC implementation were silicon nitride capacitors, NiCr resistors for gate biasing and GaAs resistors for precision resistance terminal impedances, via holes and single-metal transmission lines.

Transmission lines

The structure of the single-metal transmission line is depicted in Figure A.1. The upper layer is a 1.25µm thick gold metallization layer. Under that layer, the dielectric SiO2 and SiN layers provides isolation and size uniformity for interconnecting other passive devices. Following, the GaAs semi-insulating substrate with relative permittivity of 12.9 is backed by a 3.5µm thick gold layer to form the ground plane and ensure the electrical continuity with the front-side through the via holes.

The transmission lines have a typical loss of 0.03 Ω/□ and maximum current density of 6 mA/mm and the impedance ranging from 40Ω to 110Ω.

![Figure A.1. Structure of the single metal-layer transmission line](image)

For a specific frequency, \( f \) and length, \( l \) the electrical length of the transmission lines is:
where $c_0$ is the speed of the light ($3 \times 10^{11}$ mm/s) and $\varepsilon_{eff}$ is the effective permittivity of the material.

Different elements such as crosses, bends and coupled lines are modelled using the parameters supplied by the process. In our application, such devices were designed with values close to the dimensions recommended by the foundry. However, due to layout restrictions, the distance between some discontinuities becomes lower than 100 $\mu$m and the use of an electromagnetic simulation is recommended. Unfortunately, such simulations were not carried out given the inability to include frequency parameters of lumped devices such as transistors and small capacitors in the simulator.

**MIM capacitors**

Silicon nitride capacitors can provide values of capacitance in the range of 1 fF - 1 pF. Those capacitors were used in the implementation since they provide good design as their models take into account distributed effects. Figure A.2 shows the physical layout of the silicon nitride capacitor, in which the geometry and location of the access via to the top metallisation layer can be set during layout. Figure A.2 shows the equivalent circuit model valid up to 40 GHz, in which the distributed effects can be taken into account using the length, L and the width, W of the capacitor geometry.

For accurate dimensioning, the layout includes the capacitance to ground. The parameter of the model (area, S in $\mu m^2$ and perimeter, P in $\mu m$) are calculated to the target capacitance using the formula:

$$C [pF] = \alpha_2 \times S + 0.05 \times 10^{-3} \times P$$  \hspace{1cm} (A.2)
Appendix A MMIC components and parameter extraction

where $\alpha_2$ is the MIM parameter of the SiO$_2$ (pF/mm$^2$), typically close to 50 pF/mm$^2$. The second term is associated with the capacitance to ground contribution. Smart libraries of the process calculate the capacitor dimensions according to layout rules taking into account distributed capacitance.

**Resistors**

Two type of resistors were used in the MMIC implementation. Those are constructed using NiCr and GaAs N-layer metal films. GaAs resistors can produce more precise values of resistances (5% precision). However, they are prone to current saturation at high voltages and usually present more restrictions on the geometry. In addition, their layout consumes large MMIC area when high resistances are necessary. NiCr resistors, with a sheet resistance of 50 $\Omega$\,j, provide a wide range of resistances (from some ohms to 2 k$\Omega$). Figure A.3 shows the resistor layout and the equivalent circuit model of nichrome resistors, it includes a series resistance which is considered independent of the skin effect as the resistive layer is very thin. $C_p$ is the total ground capacitance. GaAs resistors were used for 50 $\Omega$ impedance terminal impedances whilst the nichrome resistors were used for biasing the voltage in the gate terminal of HEMTs.

\[ R(\Omega) = \frac{R \times MD \times L}{W} \]  

(A.3)

![Figure A.3 Nichrome resistor (right) and distributed model of the resistor at high frequencies (left)](image)
typically, the \( R_{MD} \) parameter of 40 \( \Omega \), \( L \) is the length and \( W \) the width of the resistor. The parasitic capacitance \( C_p \) in Figure A.3 corresponds to the capacitance created by the conductive pattern with ground plane. In our designs, this parasitic capacitance, however, does not change the behaviour of the resistor because most of the parasitics are reduced for high resistance values.

**Bond pads**

Bond pads are used for external connections to the MMIC. Those are constructed of two most external metallisation layers and are connected through vias in the silicon nitride and polyimide layers. A window is open in the upper silicon nitride to allow bonding to the upper metallisation layer. Bond pads are represented electrically by a capacitance to ground with a value proportional to the pad dimension.

**A.2 The HEMT process**

The active components are based on a GaAl/As - GaIn/As - GaAs heterostructure obtained by MOVPE, using an industrial multiwafer system. The pseudomorphic-HEMT has a 0.2\( \mu \)m gate length and can be produced as enhanced or depleted FET. Figure A.4 shows depleted mode HEMT. The effective HEMT area is defined in the mesa layer with the source and drain ohmic contacts. Wafer measurements up to 40GHz of a 6\( \times \)15\( \mu \)m depleted-HEMT gives an \( f_T \) equal to 60 GHZ as a typical value at \( I_{DSS} \) and \( V_{DS} = 3.0V \). The mean \( I_{DSS} \) figure is 202 mA/mm and threshold voltage, \( V_t = -0.8 \) V.

**Figure A.4 0.2\( \mu \)m gate length – depleted mode PHEMT**
Appendix A  MMIC components and parameter extraction

PHEMT has a typical intrinsic transconductance of 653 mS/mm. Allowed geometries for the device are an even number of fingers (from 2 to 8) and a gate finger ranging from 10 \( \mu \text{m} \) to 400\( \mu \text{m} \). The breakdown voltage of this process is equal to 5.0V, thus the HEMT process is well-suited for high power applications.

A.3 Small-signal model of the HEMT

The process manufacture proposes the small-signal model of HEMT process as the best available representation of the device at a given biasing point. Figure A.5 is the small-signal model in which the intrinsic (bias-dependent) small signal devices are separated from the external parasitic elements. OMMIC provides values of the electrical elements at different bias points. For our application, a VDS=3.0V was chosen as at this bias the output resistance of the transistor is large enough so as to ensure wide bandwidth coupling to the drain ATLs. Therefore, for the purpose of our design, the HEMT was characterised for different Gate-to-Source voltages at the fixed Drain-to-Source voltage, VDS=3.0V.

![Small-signal equivalent model of the HEMT](image)

**Figure A.5 Small-signal equivalent model of the HEMT**

The intrinsic parameters of the HEMT comprise the input capacitance or gate-to-source capacitance, \( C_{gs} \), the intrinsic transconductance, \( g_m \), the output resistance, \( R_{ds} \), the output capacitance, \( C_{ds} \) and the feedback capacitance, \( C_{gd} \). The gate-drain resistance, \( R_{gd} \) is used for modelling the device in the ohmic region at low drain voltages and is not considered (taken as zero) in our modelling. The intrinsic parameters are related to variations in the drain current and stored charge in the region of the active devices as a function of the gate and drain voltages. In HEMTs, the gate charging resistance, \( R_{gs} \) and the transconductance delay, \( t_d \) have low variations as functions of the gate-to-source voltage in the saturation region [63].

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The extrinsic circuit elements are bias-independent and those are:

1. The inductances, $L_g$, $L_d$ and $L_d$ associated with the size of the electrodes and interconnections.
2. The capacitances, $C_g$, $C_d$ and $C_s$ corresponding to the electrode capacitances.
3. The resistances $R_g$, $R_d$ and $R_s$ are the access resistances of the electrode contact resistances and the respective doped access regions.

As pointed out earlier, the geometry of the HEMT (finger number and gate width per finger) can be chosen to achieve some pre-specified design goals. The values of the intrinsic and extrinsic elements vary with the device area and finger number. Extrinsic elements are determined straightforwardly using simple relationship and referenced to the layout in Figure A.4.

The bias-dependent elements are also provided by OMMIC. However, inaccuracies were noticed when comparing the scattering parameters of the full HEMT model against the equivalent circuit in Figure A.5. For example, the capacitance ratio $\frac{C_{gs}}{C_{gd}}$ using the parameters extracted by the manufacturer and given in the manual is equal to 8.77, whilst further optimised parameter extraction yields a ratio equal to 6.0. Since our design is highly dependent on accurate modelling of the intrinsic elements (specifically; $C_{gs}$, $C_{gd}$, $g_m$ and $C_{ds}$) it was decided to re-extract the intrinsic small signal parameters of the circuit of Figure A4. This was done for different gate bias values. The extraction was done using optimisation techniques applied to the calculation of S-parameters. This is discussed below.

A.4 Extraction of small-signal model

The extraction process was based on obtaining the S-parameters from the HEMT non-linear model (which is geometry and bias dependent) and comparing these to S-parameters obtained by simulating the small signal model of Figure A4. This was done for 31 different gate bias points. Six intrinsic small signal parameters were extracted\(^1\). These optimised parameters are itemised in Table A.1 for two different gate-to-source voltages.

---

\(^1\) The transconductance delay $t_d$ did not exhibit changes with the bias voltages and was taken as 0.288ps in all optimisation routines. $R_{gd}$ was set to zero.
Table A.1 Optimised elements for two different bias conditions, HEMT 6×33μm

<table>
<thead>
<tr>
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<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>135.304</td>
<td>201.193</td>
<td>0.663511</td>
<td>142.604</td>
<td>45.1904</td>
<td>32.6133</td>
</tr>
<tr>
<td>-0.44</td>
<td>91.6451</td>
<td>174.855</td>
<td>0.765461</td>
<td>200.796</td>
<td>38.4683</td>
<td>38.471</td>
</tr>
</tbody>
</table>

The optimisation started by assigning values to the intrinsic elements equal to the small-signal parameters given in the manuals. The random optimisation routine was applied, in which the variables can acquire any value within a predefined range. In order to improve the optimisation, unequal weights \( (w_k's) \) were used in the minimisation of the error function below:

\[
E = \frac{1}{N} \sum_{f} w_1 |s_{11} - s_{11}^T|^2 + w_2 |s_{22} - s_{22}^T|^2 + w_3 |s_{21} - s_{21}^T|^2 + w_4 |s_{12} - s_{12}^T|^2
\]  

(A.4)

The range of frequencies for the optimisation was set from 1 to 50GHz, with \( N \) the number of frequencies at which the model was optimised. The factor \( w_3 \) weights the difference between the optimised and target forward gain parameters \( (s_{21}, s_{21}^T) \). That weight was set to a larger value than the other weighting factors as this improved the convergence of the error function. The re-extracted variables minimised the error function. Figure A.6 displays optimised scattering parameters and the target parameters for a VGS=0.0V and Figure A.7 displays parameters for VGS=-0.44V. In both cases the optimisation is in excellent agreement with the target response in the frequency range of interest.

Figure A.6 Optimised scattering parameters at VGS=0V
Appendix A MMIC components and parameter extraction

Figure A.7 Optimised scattering parameters at VGS=-0.6V

Based on those parameters, the input capacitance and the transconductance of the HEMT are depicted against applied voltage as shown in Figure A.8. Both parameters vary non-linearly with the gate to source voltage. From these figures we can estimate the cut-off frequency to be 93.5 GHz at VGS=0V and 63.5 GHz at VGS=-0.44 V

Figure A.8 Transconductance and input capacitance as functions of applied voltage
Finally, the intrinsic voltage gain of the HEMT $\frac{g_m}{g_o}$ and the capacitance ratio were displayed against the applied voltage. The voltage gain parameter is almost constant and results from the use of a high-gain and short gate length HEMT device (6 fingers $\times$ 33 $\mu$m). By contrast, the $\frac{C_{gs}}{C_{gd}}$ parameter is reduced for low applied voltages. The isolation characteristic of the device effectively deteriorate with reduced $V_{GS}$. 

Figure A.9. Capacitance ratio and intrinsic voltage gain of the HEMT
Appendix B

Derivation of transfer function of the image-impedance terminated transversal filter

In this Appendix we detail the derivation of generic transfer function for the distributed amplifier based of the transversal filter, described in Chapter 5. This is done by extending the work of Chen in [1] so that it can be applied to structures where the counterpropagating waves (on the ATLs) are taken into account. This is important for our application as the transmission characteristics of the structure provides responses which can be defined for distributed filter sections.

In [1] the application of the classical two-port cascaded network theory has been extended to the analysis of distributed amplification with sections with different propagation functions. This was based on the image-matched cascaded filter theory [2,3]. In such analyses, the artificial transmission lines are ideally terminated by the appropriate image impedances and the passive coupling between the two lines is neglected; i.e. the feedback elements of are not included in the analyses.

The Image Transfer Function theory can be extended to the analyses of the generalised transversal filter, in particular to the topology adopted in this thesis in which the delay in both drain and gate ATLs form an additive interstage filter delay. Such distributed amplifier topology terminated in image impedance $Z_f$ is represented in Figure B.1. The interconnecting transmission lines are designed to introduce additional time delay.
Derivation of the transfer function of the image-impedance terminated transversal filter

By assuming uniformity in both gate and drain ATLs, the distributed transversal filter can be analysed by using the chain (ABCD) parameters [2]. Each stage is characterised by a complex image propagation factor and matched to an image impedance. Hence, the voltage and current at any node within an ATL can be calculated. By taking into account the fact that both ATLs are actively coupled via transconductance current sources, the transfer function can be evaluated using the network parameter. The adopted block diagram which forms the basis of our analysis of the transversal filter is shown in Figure B.2

\[
\theta^g_{k} = \theta^g_{k,1} + \theta^g_{k,2} \\
\theta^d_{k} = \theta^d_{k,1} + \theta^d_{k,2}
\]  
(B.1)
Derivation of the transfer function of the image-impedance terminated transversal filter

The method of numbering network sections described in Figure B.2 is similar to that of Chen [1]; however, it utilises different indexes in both gate and drain lines. The methodology generated from this method is simple and resembles the filter network schemes pertaining to the distributed amplifiers [1,3]. As presented above, the controlled current sources distributed along the drain line are subscripted in correspondence with the wave paths formed in the image terminated transversal filter of Figure B.1. Using this method, the voltages and currents at both ports are related in a simple manner. The equations remain simple as derived in the application of image transfer function technique [3].

By applying the superposition theory to filter networks, it is possible to prove that for a network matched on image basis and characterised by its propagation constants as shown in Equation B.1, the current and voltage in each section of the drain line are given by:

\[ I_{k-1} = I_k \exp(\theta_k^d) - \sqrt{\frac{Z_x^d}{Z_T^d}} \cosh(\theta_k^{d,1}) I_{N-k+1}^d \]  

\[ E_{k-1} = E_k \exp(\theta_k^d) - \sqrt{\frac{Z_x^d Z_T^d}{Z_T^d}} \sinh(\theta_k^{d,1}) I_{N-k+1}^d \]

where \( I_k \) and \( E_k \) are the current and voltage phasors in drain line sections, respectively, \( I_k^d \) is the controlled current source phasor, \( Z_x^d \) and \( Z_T^d \) are the characteristic impedances of the \( T \) and \( II \) network formed from cascading two identical \( L \) sections in an image basis.

The above recursive equations are solved by using the boundary conditions in the drain line:

\[ E_N = Z_T^d I_N \]

\[ E_0 = -Z_T^d I_0 \]

The gate line has a more simple equation relating cascaded section voltages. A straightforward analysis yields the equation to the controlling voltage:

\[ V_k^g = \sqrt{\frac{Z_x^g}{Z_T^g}} \exp(-\theta_1^g - \theta_2^g - \ldots - \theta_k^g) \exp(\theta_k^{g,2}) V_{\text{inp}} \]

where \( Z_x^g \) and \( Z_T^g \) are the characteristic impedances of the \( T \) and \( II \) network in the gate line filter and \( V_{\text{inp}} \) is the input voltage provided by the source voltage.

Solving the recursive Equation B.2, gives:
Derivation of the transfer function of the image-impedance terminated transversal filter

\[ I_0 = I_N \exp \left( \sum_{i=1}^{N} \theta_i^d \right) - \sqrt{\frac{Z_n^d}{Z_T^d}} (I_N \cosh(\theta_1^{d,1}) + I_{N-1}^s \cosh(\theta_2^{d,1}) \exp(\theta_1^d) + \cdots + I_{N-2}^s \cosh(\theta_N^{d,1}) \exp(\theta_1^d + \theta_2^d + \cdots + \theta_{N-1}^d)) \]  
(B.7)

On the other hand, Equation B.3 can be rewritten as:

\[ E_k = E_{k-1} \exp(-\theta_k^d) + \sqrt{Z_n^d Z_T^d} \exp(-\theta_k^d) \sinh(\theta_k^{d,1}) I_{N-k}^s \]  
(B.8)

Equation B.3 is solved for all \( k \) index values:

\[
E_N = E_0 \exp \left( - \sum_{m=1}^{N} \theta_m^d \right) + \sqrt{Z_n^d Z_T^d} (I_N^s \sinh(\theta_1^{d,1}) \exp \left( - \sum_{m=1}^{N} \theta_m^d \right) + I_{N-1}^s \sinh(\theta_2^{d,1}) \times \exp \left( - \sum_{m=2}^{N} \theta_m^d \right) + \cdots + I_{N-k}^s \sinh(\theta_{N-k}^{d,1}) \exp \left( - \sum_{m=k+1}^{N} \theta_m^d \right) + I_N^s \sinh(\theta_{N}^{d,1}) \exp \left( - \theta_N^d \right) ) \]

(B.9)

The voltages and currents in the drain line sections are interrelated by the boundary condition \( E_0 = -Z_T^d I_0 \) and substituting Eq. B.7 into B.3, the output voltage is given by:

\[
E_N = -I_N Z_T^d + \sqrt{Z_n^d Z_T^d} (I_N^s \cosh(\theta_1^{d,1}) \exp \left( - \theta_1^d - \cdots - \theta_N^d \right) + I_{N-1}^s \cosh(\theta_2^{d,1}) \exp \left( - \theta_1^d - \cdots - \theta_N^d \right) + \cdots + I_{N-k}^s \cosh(\theta_{N-k}^{d,1}) \exp \left( - \theta_1^d - \cdots - \theta_N^d \right) + I_N^s \sinh(\theta_{N}^{d,1}) \exp \left( - \theta_N^d \right) ) \times \exp \left( - \sum_{m=2}^{N} \theta_m^d \right) + \cdots + I_{N-k}^s \sinh(\theta_{N-k}^{d,1}) \exp \left( - \sum_{m=k+1}^{N} \theta_m^d \right) + \cdots + I_N^s \sinh(\theta_{N}^{d,1}) \exp \left( - \theta_N^d \right) ) \]

(B.10)

The output voltage can be obtained in function of the controlled current sources by using the boundary condition \( E_N = Z_T^d I_N \) and the identity \( \sinh(\theta) + \cosh(\theta) = \exp(\theta) \). This yields the equation:

\[
E_N = \frac{1}{2} \sqrt{Z_n^d Z_T^d} (I_N^s \exp(\theta_1^{d,1}) \exp \left( - \theta_1^d - \cdots - \theta_N^d \right) + I_{N-1}^s \exp(\theta_2^{d,1}) \exp \left( - \theta_1^d - \cdots - \theta_N^d \right) + \cdots + I_{N-k}^s \exp(\theta_{N-k}^{d,1}) \exp \left( - \theta_1^d - \cdots - \theta_N^d \right) + I_N^s \exp(\theta_{N}^{d,1}) \exp \left( - \theta_N^d \right) )
\]

(B.11)

The current controlled source is described by:
Derivation of the transfer function of the image-impedance terminated transversal filter

\[ I_k^x = -g_k V_k^x = -g_k \sqrt{\frac{Z_f^x}{Z_T^x}} \exp(-\theta_1^x - \theta_2^x - \ldots - \theta_k^x) \exp(\theta_k^{x,2}) V_{inp} \]  

(B.12)

Substituting B.12 into B.11, the transfer function of the distributed transversal filter matched in image impedance basis and different propagation function is:

\[ \frac{E_N}{V_{inp}} = -\frac{1}{2} \sqrt{\frac{Z_T^x}{Z_f^x}} \sqrt{Z_T^d Z_f^d} (g_N \exp(\theta_1^{d,1} + \theta_N^{x,2}) \exp(-\theta_1^d - \ldots - \theta_N^d - \theta_1^x - \ldots - \theta_N^x) + \ldots + g_{N-k+1} \exp(\theta_k^{d,1} + \theta_{N-k+1}^{x,2}) \exp(-\theta_k^d - \ldots - \theta_N^d - \theta_k^x - \ldots - \theta_{N-k+1}^x) + \ldots + g_1 \exp(\theta_1^{d,1} + \theta_1^{x,2}) \exp(-\theta_1^d - \theta_1^x)) \]  

(B.13)

If the propagation constants of the filter sections are made to be equal in both drain lines, that is \( \theta_1^d = \ldots = \theta_N^d = \theta^d \), \( \theta_1^x = \ldots = \theta_N^x = \theta^x \), and if the mid-shunt propagation characteristics are equal, then \( \theta_k^{d,1} = \theta_k^{d,2} = \theta^d / 2 \) and \( \theta_k^{x,1} = \theta_k^{x,2} = \theta^x / 2 \). This results in the voltage transfer function below:

\[ \frac{E_N}{V_{inp}} = -\frac{1}{2} \sqrt{\frac{Z_T^x}{Z_f^x}} \sqrt{Z_T^d Z_f^d} \exp\left(\frac{\theta^d + \theta^x}{2}\right) \sum_{k=1}^{N} g_k \exp\left(-k\left(\theta^d + \theta^x\right)\right) \]  

(B.14)

References

Appendix C

Derivation of Mixed-Mode Scattering Parameters of Dual-Drain Line Transversal Filter

This Appendix details the derivation of mixed mode scattering parameters of the triple line filter. The set of equations derived is effectively an extension of the work reported by Bockelman and Eisenstadt in [1] and [2]. This extension is specific to the triple line filter as discussed in Chapter 7.

Scattering parameters that enable testing and measurement of differential circuits at microwave frequencies were proposed by Bockelman and Eisenstadt in [1,2] to fulfil the need of standardised parameters for characterising differential microwave circuits. A complete treatment of simultaneous propagation modes of any generic distributed differential circuit has been established in [1] enabling the conversion of single-ended measurement of differential circuits into mixed-mode parameters [2]. A generic two-port microwave circuit is depicted in Figure C.1; it consists of a distributed or lumped differential circuit, coupled lines and a ground plane. Such scheme is general since the device under consideration can be measured either by using uncoupled transmission lines or coupled and uncoupled pair sections as depicted in Figure C.1 [1]. Parasitic couplings between transmission lines can be treated more conveniently by considering signals that propagates between the lines using a two-port circuit description which is suitable to the analysis of differential circuits.

In the generic differential circuit, mutual couplings between transmission lines are accounted for by defining the propagation constants and impedances of both the common and differential-mode as parameters of the lines. In the analysis of the transversal filter considered here, propagating waves are induced by lumped active circuits periodically distributed along transmission lines. Filter analysis requires mixed-mode analysis in which nodal waves are
Derivation of mixed-mode scattering parameters of the dual-drain line transversal filter
defined by current and voltages in discrete sections of distributed circuits. Parasitic couplings
associated with electromagnetic couplings are not accounted for in the scattering parameter
analysis for simplicity, however; such source of parasitic couplings could be analysed using the
mixed-mode parameters when appropriate.

Figure C.1. Schematic of the differential s-parameter measurement

A conceptual diagram of the triple-line transversal filter is shown in Figure C.2. The
structure has a single-ended input port and two-ended output port with orthogonal modes of
propagation. The scattering parameters at the input port are defined by the power waves
\( \{a_{sm1}, b_{sm1}\} \) and signals referenced to a common-ground plane. At the output port (port 2),
the common-mode waves \( \{a_{cm2}, b_{cm2}\} \) are defined by common-mode components referenced to a
common-ground plane whilst the differential power waves \( \{a_{dm2}, b_{dm2}\} \) are defined by
differential-mode components, which are referenced to each other and a common-ground plane [3].

Figure C.2. Conceptual diagram of the dual-drain line structure
Derivation of mixed-mode scattering parameters of the dual-drain line transversal filter

The definition of s-parameters in the input port uses traditional scattering parameters. Conversely, parameters in the port 2 need differential and common-mode components as described in Section 7.2.1. Here, the transformations between traditional s-parameters and mixed-mode parameters for the triple-line structure are obtained using the theory of mixed-mode parameters [1,2].

At microwave frequencies, signals are treated as travelling waves on artificial transmission lines. A set of scattering parameters can describe completely the frequency performance of the two-port filter normalising propagating waves in both ports using the same impedance system. Input signals are referred to a common-ground; the scattering parameters are defined by an incident wave \( a_1 \) and reflected wave \( b_1 \) as shown in Figure C.3. The output port can be described by two incident waves \( a_2, a_3 \) and reflected waves \( b_2, b_3 \). Voltages and currents at the output port are referred to a common ground.

![Diagram to define the traditional scattering parameters](image)

**Figure C.3 Diagram to define the traditional scattering parameters**

The diagram of the transversal filter given in Figure C.3 serves to define the standard s-parameters of the structure considering three independent ports; i.e. assuming that voltages and currents at terminal ports do not maintain necessarily any relationship between them. The three port transversal filter can be defined by a set of conventional s-parameters given by:

\[
\begin{bmatrix}
    b_1 \\
    b_2 \\
    b_3
\end{bmatrix} =
\begin{bmatrix}
    s_{11} & s_{12} & s_{13} \\
    s_{21} & s_{22} & s_{23} \\
    s_{31} & s_{32} & s_{33}
\end{bmatrix}
\begin{bmatrix}
    a_1 \\
    a_2 \\
    a_3
\end{bmatrix}
\]

(C.1)
Derivation of mixed-mode scattering parameters of the dual-drain line transversal filter

Conventional parameters are defined by incident and reflected power waves and related to the voltage and current in each port by equations [4]:

\[ a_i = \frac{1}{2\sqrt{Z_0}} \left( V_i + I_i Z_0 \right) \quad \quad \quad \quad \quad \quad \quad \quad b_i = \frac{1}{2\sqrt{Z_0}} \left( V_i - I_i Z_0 \right) \]  (C.2)

where \( a_i \) and \( b_i \) are the normalised forward and reverse power waves at node \( i, i \in \{1,2,3\} \) and normalised to the resistive impedance \( Z_0 \).

Differential signals at port two and three have differential components. Equivalent relationships are drawn from conventional scattering parameters at the output port in accordance with the method given in [1]. The circuit under consideration is depicted in Figure C.4, the response to an input stimulus possesses mixed-mode components, common-mode current and voltage \( \{V_{cm}, I_{cm}\} \) and differential-mode voltage and current \( \{V_{dm}, I_{dm}\} \) appear at the output port. The efficiency to couple differential signals to both output ports is analysed by launching a common mode wave and setting the relative phase between generators \( \Theta = 0^\circ \) in the circuit shown in Figure C.4, whilst a differential mode wave is analysed by setting \( \Theta = 180^\circ \). Source voltages have resistive impedance \( Z_0 \) equal to the terminal resistance in all filter ports. Voltage generator applied at the input port gives rise to incident and reflected power waves \( a_i \) and \( b_i \).

**Figure C.4 Transversal filter structure to measure mixed-mode parameters**

In [1,2] the linear transformations required to convert the set of standard single-ended s-parameters into a set of mixed-mode parameters have been introduced. In order to obtain a better understanding of the analysis, the scattering parameters will be defined by adopting the differential two-port circuit configuration as described in [3]. For such an aim, each individual
Derivation of mixed-mode scattering parameters of the dual-drain line transversal filter

parameter follows a naming convention in the following order: mode response, mode stimulus, port response and port stimulus. Referring to the transversal filter structure of Figure C.3, the stimulus can be single-ended stimulus at the input (S), pure differential mode with 180°-phase shifted voltage sources (D) and common-mode stimulus with applied voltages in phase (C). The nine scattering parameters that enable a complete characterisation of the filter structure are described by the equation:

\[
\begin{bmatrix}
  b_{sm1} \\
  b_{sm2} \\
  b_{cm2}
\end{bmatrix} =
\begin{bmatrix}
  s_{SS11} & s_{SD12} & s_{SC12} \\
  s_{DS21} & s_{DD22} & s_{DC22} \\
  s_{CS21} & s_{CD22} & s_{CC22}
\end{bmatrix}
\begin{bmatrix}
  a_{sm1} \\
  a_{sm2} \\
  a_{cm2}
\end{bmatrix}
\]  

(C.3)

Based on the normalised power waves of equation C.2 and assuming uncoupled transmission lines in the output port, the normalised nodal waves \{a_{dm2}, a_{cm2}\} and \{b_{dm2}, b_{cm2}\} are obtained straightforwardly using the same reference impedance \(Z_0\) as that of the nodal waves referred to a common ground. It can be shown that the normalised nodal waves of the coupled lines are given by the set of relationships [1,2]:

\[
\begin{align*}
  a_{dm2} &= \frac{1}{\sqrt{2}} (a_2 - a_3) \\
  b_{dm2} &= \frac{1}{\sqrt{2}} (b_2 - b_3) \\
  a_{cm2} &= \frac{1}{\sqrt{2}} (a_2 + a_3) \\
  b_{cm2} &= \frac{1}{\sqrt{2}} (b_2 + b_3)
\end{align*}
\]  

(C.4)

Such relationships facilitate the establishment of the linear conversions required for scattering analysis using linear transformations applied to matrices C.1 and C.3. The set of scattering parameters can be itemised as gain parameters, input reflection parameters, reverse transmission parameters and conversion mode parameters.

The equations found by such method and utilised in our simulations of Chapter 7 are itemised as follows.

**Gain parameters \{s_{DS21}, s_{CS21}\}**

\[
\begin{align*}
  s_{DS21} &= \frac{1}{\sqrt{2}} (s_{21} - s_{31}) \\
  s_{CS21} &= \frac{1}{\sqrt{2}} (s_{21} + s_{31})
\end{align*}
\]  

(C.5)
Derivation of mixed-mode scattering parameters of the dual-drain line transversal filter

Input reflection coefficient

\[ s_{ss11} = s_{11} \]  \hspace{1cm} (C.6)

Reverse transmission parameters \( \{s_{sc12}, s_{sd12}\} \)

\[ s_{sc12} = \frac{1}{\sqrt{2}} (s_{13} + s_{12}) \]
\[ s_{sd12} = \frac{1}{\sqrt{2}} (s_{12} - s_{13}) \]  \hspace{1cm} (C.7)

Conversion mode parameters \( \{s_{cd22}, s_{dc22}\} \)

\[ s_{dc22} = \frac{1}{2} (s_{22} + s_{23} - s_{32} - s_{33}) \]
\[ s_{cd22} = \frac{1}{2} (s_{22} - s_{23} + s_{32} - s_{33}) \]  \hspace{1cm} (C.8)

Output reflections parameters \( \{s_{dd22}, s_{cc22}\} \)

\[ s_{cc22} = \frac{1}{2} (s_{22} + s_{23} + s_{32} + s_{33}) \]
\[ s_{dd22} = \frac{1}{2} (s_{22} - s_{23} - s_{32} + s_{33}) \]  \hspace{1cm} (C.9)
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