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UMI
A DIRECT MICROWAVE M-QAM
ADAPTIVE TRANSMITTER FOR FIXED
WIRELESS ATM NETWORKS

A Thesis
Submitted to the College of Graduate Studies and Research
In Partial Fulfilment of the Requirements
for the Degree of
Doctor of Philosophy
in the Department of Electrical Engineering
The University of Saskatchewan
Saskatoon, Saskatchewan

By
Abbas Mohammadi
Fall 1998
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SUMMARY OF DISSERTATION
Submitted in partial fulfillment
of the requirements for the

DEGREE OF DOCTOR OF PHILOSOPHY

by

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ABSTRACT

Wireless ATM plays a key role in the realization of broadband wireless networks. The transmission of various classes of traffic and the provision of bandwidth on demand over a wireless channel poses a number of new technical challenges. This thesis addresses the design of a low cost adaptive transmitter for fixed wireless ATM/B-ISDN systems with emphasis on optimum use of wireless network resources.

A new architecture for a direct microwave wireless ATM transmitter is proposed. The transmitter capacity adaptation is implemented by using an admission control metric and an M-QAM modulator. The two main components of the transmitter are: an M-QAM control unit and a direct microwave QAM modulator unit. The M-QAM control unit is used to select an optimum modulation level for the QAM modulator. The modulation level is adjusted based on the bandwidth demand, QoS requirements, and outage conditions of the wireless ATM link. The direct microwave QAM modulator unit transforms the broadband traffic to a modulated microwave signal that is suitable for transmission over a wireless network.

The required bandwidth of the broadband traffic is estimated using an effective bandwidth metric. An analytical relation, called the capacity reduction factor, is derived to represent the performance degradation due to the wireless channel and channel fading in a B-ISDN network. Using the effective bandwidth metric and the capacity reduction factor, a QoS metric for the wireless broadband network is introduced. This metric is termed as, modified effective bandwidth. This metric is used to adapt the M-QAM modulator.
Another significant contribution of this research work is a new architecture for the direct QAM modulator. This is based on use of PIN diode reflection attenuators. The PIN diodes operate in forward bias condition thereby overcoming the speed limitation problem due to charge storage. Using residue theory, analytical results to model the large signal forward bias operation of PIN diodes are presented. This theory also examines the transition time of a PIN diode with bias changes from a reverse bias to a forward bias.

The direct microwave QAM modulator implementation using MIC and silicon MMIC technologies is examined. While a realization using MIC is simple and straightforward, a silicon MMIC realization offers a very cost effective solution.

A system study was conducted to examine the operation of the adaptive direct microwave M-QAM modulator in the wireless channel with ATM traffic. The operation has been examined for different wireless channels and for various classes of traffic. The call acceptance and outage performance are compared with those for a fixed QAM modulator. The results show that the proposed system can be used for implementation of cost effective adaptive transmitters for broadband wireless applications.
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DEDICATION

To the Loving Memory
of my father, Mr. Saeed Mohammadi
for his unlimited love.
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<td>Available Bit Rate</td>
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<tr>
<td>ATM</td>
<td>Asynchronous Transfer Mode</td>
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<td>AWGN</td>
<td>Additive White Gaussian Noise</td>
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<tr>
<td>BER</td>
<td>Bit Error Rate</td>
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<td>B-ISDN</td>
<td>Broadband Integrated Services Digital Network</td>
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<td>BPSK</td>
<td>Binary Phase Shift Keying</td>
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<td>CDV</td>
<td>Cell Delay Variation</td>
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<td>Local Multipoint Distribution Service</td>
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<td>LRD</td>
<td>Long-Range Dependency</td>
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<tr>
<td>MBS</td>
<td>Maximum Burst Size</td>
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<tr>
<td>MIC</td>
<td>Microwave Integrated Circuit</td>
</tr>
<tr>
<td>MMIC</td>
<td>Monolithic Microwave Integrated Circuit</td>
</tr>
<tr>
<td>MoD</td>
<td>Multimedia on Demand</td>
</tr>
<tr>
<td>MPEG</td>
<td>Motion Picture Expert Group</td>
</tr>
<tr>
<td>MPSK</td>
<td>M-ary Phase Shift Keying</td>
</tr>
<tr>
<td>Acronym</td>
<td>Description</td>
</tr>
<tr>
<td>---------</td>
<td>-------------------------------------------------</td>
</tr>
<tr>
<td>M-QAM</td>
<td>M-ary Quadrature Amplitude Modulation</td>
</tr>
<tr>
<td>PAM</td>
<td>Pulse Amplitude Modulation</td>
</tr>
<tr>
<td>PCR</td>
<td>Peak Cell Rate</td>
</tr>
<tr>
<td>QAM</td>
<td>Quadrature Amplitude Modulation</td>
</tr>
<tr>
<td>QoS</td>
<td>Quality of Service</td>
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<tr>
<td>QPSK</td>
<td>Quadrature Phase Shift Keying</td>
</tr>
<tr>
<td>RF</td>
<td>Radio Frequency</td>
</tr>
<tr>
<td>SCR</td>
<td>Sustainable Cell Rate</td>
</tr>
<tr>
<td>SONET</td>
<td>Synchronous Optical NETwork</td>
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<tr>
<td>SPDT</td>
<td>Single Pole Double Through</td>
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<tr>
<td>SNR</td>
<td>Signal to Noise Ratio</td>
</tr>
<tr>
<td>TDMA</td>
<td>Time Division Multiple Access</td>
</tr>
<tr>
<td>UBR</td>
<td>Unspecified Bit Rate</td>
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<tr>
<td>VC</td>
<td>Virtual Channel</td>
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<td>VoD</td>
<td>Video on Demand</td>
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<td>VP</td>
<td>Virtual Path</td>
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<tr>
<td>WAN</td>
<td>Wide Area Network</td>
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1. INTRODUCTION

The beginning of a new millennium has coincided with the emergence of a new generation of wireless systems collectively called broadband wireless communication services. Following the emergence of the first generation of wireless communication services, which covers the car phone and analog cellular era, a second generation, characterised by the development of digital cellular technology and cordless telephony came into being. A new generation is now under development with the aim of supporting broadband communication services over a wireless channel. The broadband wireless system must use an efficient networking technique to transmit the different classes of traffic, such as computer data, voice, and video, that have different traffic characteristics.

The introduction of the Asynchronous Transfer Mode (ATM) networking technique has been an important contributing factor in the realization of wireless broadband communications [1]. The ATM is an efficient networking technique that is designed to transmit different classes of traffic over the same network and can be used as a platform for broadband wireless communications. Moreover, the ATM technique could also provide the capability for delivering bandwidth-on-demand to portable terminals. Further, the ATM network has the feature of statistical multiplexing which maximises the use of the available wireless network capacity when traffic originates from multiple sources.

In a broadband wireless network, an optimum transmitter has to take
into account both radio communication models and broadband traffic characteristics. This has opened a new challenge for the telecommunications community. There is presently a considerable amount of research on wireless broadband networking using the ATM technique. Following some early research before 1995, this subject is now entering the mainstream in terms of research, technology, and standardisation.

1.1 Broadband Wireless Communications

1.1.1 ATM Networks

The ATM technique, called a fast packet switching method, has been proposed to carry different classes of traffic in the same medium by using short, fixed length packets. The basic idea of the ATM is to segment the digital data of different sources into a sequence of blocks, called cells, that are transported and routed through a telecommunications network. The length of the cell is 53 bytes, with the first 5 bytes consisting of header information and the remaining 48 bytes consisting of the information payload. Both real time traffic, such as voice and video, which may tolerate some loss but not delay, and non real time traffic such as computer data, which may tolerate some delay but not loss, should be efficiently transmitted using the same medium.

An ATM traffic source may demand a bandwidth with a high peak for a short duration. These sources are generally referred to as bursty traffic sources. A main feature of the ATM is the statistical multiplexing. Using the statistical multiplexing concept, different classes of traffic are carried in the same channel with a bandwidth smaller than the aggregate peak rate of the sources. The ATM network has been designed to use a high quality transmission system, in which transmission bit errors are rare and randomly distributed. This is a reasonable assumption for optical fiber based systems. ATM networks using optical fibers have been realized successfully.
Figure 1.1 The typical model for radio ATM communications

1.1.2 Radio ATM

Radio ATM is an evolving technology that has not been tested in the market yet. Although in the past two years several companies have proposed different methods for implementing radio ATM technology, an overall architecture has yet to be designed. A typical architecture for a radio ATM network is shown in Fig.1.1. The model considers the ATM network using ATM switches, optical fiber, and a wireless channel as transmission systems [2], [3]. High speed wireless ATM links (e.g., STM-1 or OC-3) are installed between the nodes that are not connected by fiber. These radio links may be used for ease of deployment or to gain system portability, and to realize a network backbone. As shown in Fig.1.1, lower speed ATM traffic is delivered to the multimedia terminals by radio ATM cellular networks. A key element in this design is the radio ATM transmitter. Two different versions of this unit have to be used in the network architecture shown in Fig.1.1 as its operation has to be optimised to maximise the utilisation of the wireless network resources. The ATM radio transmitter for the fixed point to point link will be different from the cellular radio type transmitter for the multimedia transmission to the mobile terminals.

A reference model for the fixed radio ATM link is shown in Fig.1.2. As
Figure 1.2  The reference model for fixed radio ATM communication services

shown in this figure, a number of sources with different traffic characteristics compete to send their traffic through a high speed point to point radio link. Each source uses a dedicated path to connect to the high speed radio ATM site. It is assumed that the connections from traffic sources to the radio ATM site have enough bandwidth. The paths can be wired or wireless. Moreover, the traffic sources may have different traffic characteristics with different quality of service (QoS) requirements. As a typical application for this system, one may think of the individual sources as a number of video suppliers trying to deliver their movies to a lot of houses in a geographically diverse area (e.g., an island) using a high speed radio ATM link. Such models can be used to study the recently introduced local multipoint distribution services (LMDS). An important issue in this model is the estimation of the radio bandwidth for an ATM network. This is discussed in the next section.

1.1.3 Bandwidth Estimation in Radio ATM Network

ATM bursty sources are characterised by their peak rate, mean bit rate, burstiness, as well as their required QoS [4]. An essential task in the ATM network design is to develop a realistic traffic model to estimate the required
bandwidth of ATM connections. This requirement for bandwidth has to be matched with the capacity of the radio system which depends on the available spectrum and wireless channel characteristics. A metric based upon a realistic traffic model is obviously needed. Such a metric should be based on:

- The QoS requirements of various connections,
- The traffic model,
- The wireless channel characteristics.

These requirements are studied in this thesis and a metric, called a Modified Effective Bandwidth [5], is derived in Chapters 3 and 4.

1.1.4 Adaptive Modulation

As stated above, in a radio ATM network, various types of traffic with different quality of service (QoS) requirements are carried over a wireless channel. The variable bandwidth demand for diverse traffic can be efficiently transmitted over radio only if the radio ATM transmitter operates as an adaptive unit that is controlled by the ATM traffic characteristics. Due to the demand for the spectrum, novel adaptive techniques that can dynamically allocate resources are highly desirable. The adaptation process may introduce some additional delay which must be kept below an acceptance level.

To ensure an overall QoS, radio channel performance must also be taken into account. Because of spectrum congestion and bandwidth limitations, the adaptive transmitter has to use comparatively high level modulations that are spectrally efficient. This makes the design of such an adaptive transmitter a challenging task [3]. The requirements for the adaptive transmitter may be summarised as:
• An adaptation method based on ATM traffic characteristics and wireless channel performance,

• A highly bandwidth-efficient modulation method,

• A real time adaptation process.

These requirements are studied in this thesis and an adaptive multilevel quadrature amplitude modulator (M-QAM) is proposed and analysed in Chapters 5, 6, and 7.

1.1.5 Modulator Implementation

As mentioned above, the radio ATM transmitter has to work at high bit rates using a highly bandwidth-efficient modulation technique. A low-cost realization for the modulator is highly desirable [6]. A low-cost and high performance necessitate an architecture that is suitable for an integrated circuit implementation [7], [8]. It may be concluded that the modulator must be implemented with the following features:

• Operation at high speed data rates,

• Low cost hardware realization,

• Implementation as an adaptive modulator.

A new high performance direct microwave multi-level quadrature amplitude modulation (M-QAM) modulator that is suitable for integrated circuit design is proposed [6], [9-12]. The modulator implementation is discussed in Chapters 5, 6, and 7.

1.2 Literature Review

During the last few years, several research groups in universities and industry have initiated research on radio ATM. While an overall architecture
has yet to be designed, various architectures for radio ATM have been proposed in the open literature [13-19]. While some architectures have been introduced to handle terminal mobility, others are focused on a point to point radio ATM link [20-24]. An experimental point to point high speed link (2.488 Gbps) using a 1 GHz bandwidth in a 19 GHz band has been realized for LuckeyNet network [20]. The link used a QPSK modulator with sub-harmonic pumping.

The performance of a point to point ATM link has been analysed by Gans et al. [21]. The main conclusion of this work is that it is important to make the radio link error rate performance similar to the SONET network. The use of an error code with an ATM interleaver was recommended in this work as well. The transmission of high speed ATM cells in SDH frames has also been studied in [22]. It is suggested that a high level QAM modulator is the most suitable candidate for future high speed communications systems. The measurement of cell loss ratio of ATM traffic over radio in a 20 GHz band has been reported in [25]. The experimental results of this study are in agreement with the results in [2].

In an ATM network, a transmission link is divided into a number of different virtual paths (VP), each with a unique identifier. Each of these VPs may be then subdivided into a number of separate virtual channels (VC) with a separate identifier. A VC is a communication channel that provides for transport of ATM cells. A VP is a bundle of VCs. Two switching concepts namely VP and VC switching, have been used based upon these logical connections in ATM architecture [26-29]. Both switching concepts effectively provide an efficient and flexible capacity allocation mechanism. However, the VP switching needs a simpler ATM switch but is less flexible [27], [28].
In a broadband integrated network, a capacity controlled radio may be designed using the virtual path switching concept. The concept of VP switching to design a radio ATM was introduced in the early 1990's in [24]. A specific ATM switch design for radio ATM communications was proposed in [23]. Based upon previously published theoretical results [24], the use of an individual QAM modulator that corresponds to each virtual path was recommended in [23].

The effective bandwidth as a metric of QoS is a metric that has been commonly considered to estimate the capacities of virtual paths. The effective bandwidth (in bits per second) of a time varying traffic source (e.g., a VP) represents the minimum required bandwidth that it needs to meet QoS guarantees. The idea of an effective bandwidth for single and aggregate traffic sources was originally proposed using a stochastic theory for data sources [30], [31]. Other methods such as large deviation approximations [32] and uniform arrival and service models [33] have also been used to compute the effective bandwidth of single and aggregate traffic sources.

An accurate estimation of effective bandwidth requires a good statistical model. A commonly used assumption has been to consider the source as having exponentially distributed burst and idle periods and to approximate the real source with an n-state Markovian source [34-37]. The autocorrelation functions of these conventional traffic models drop off exponentially. However, certain recent measurements of broadband traffic, e.g., local area network (LAN), wide area network (WAN) and VBR compressed video, reveal long-range dependency (LRD). An important class of stochastic models that can account for LRD is self-similar models [38-40]. The autocorrelation function in a LRD model drops off slowly (typically as a power function). These interesting observations have changed views about traffic characteris-
tics. It is known that the effective bandwidth based upon Markovian models underestimates the cell blocking probabilities when a source exhibits only moderate long-range dependence [40]. Consequently, the effective bandwidth of a source or aggregate of sources must be computed based upon a self-similar traffic model. This is a challenging problem that has been addressed in networking theory [41-43]. A function to estimate the effective bandwidth based upon the self-similarity has been recently reported [44], [45].

As previously mentioned, the adaptive realization of radio ATM transmitter is very desirable [46]. On one hand, an adaptive transmitter may dynamically increase the capacity of the link when there are demands for more bandwidth; on the other hand, the adaptive transmitter may improve the radio link availability and decrease the outage when the bandwidth demands are not at peak. A capacity controlled multilevel QAM (M-QAM) modulator can be used to realize an adaptive radio ATM transmitter. The idea of an adaptive QAM modulator was originally introduced for fading channels [47], [48], where an M-QAM modulator was used to improve the system gain and throughput [49]. The M-QAM is also considered to absorb traffic variations in a system in which a spectrum band is assigned to each virtual path [23].

Direct microwave modulator implementation has also been studied extensively in the literature. In the early 1990’s, the idea of a direct microwave modulator was presented [50]. The conventional method of generating the modulation at IF, followed by an up-conversion to the transmit frequency, is not efficient in power, size, and cost [51]. An extensive reported study shows the advantages of a direct microwave modulator over the conventional method [52], [53]. In [54], a direct microwave QAM modulator was implemented by directly mixing an RF signal with a baseband signal. This method
of direct modulation has also been used for QPSK and vector modulators [55], [56]. An alternative method of implementing a direct microwave modulator is to use a variable attenuator. A direct conversion modulator for \(\pi/4\) QPSK using a variable attenuator in the 1.9 GHz band was reported in [57]. This method has also been used to realize a vector modulator [58], [59]. Direct microwave modulation techniques can be used for various modulation techniques [58], [60], [61], and [62].

To implement a direct microwave modulator with a variable attenuator, a PIN diode is a suitable electronic device. A PIN diode can be used as a suitable electronically tuned resistor in the attenuator due to its power-handling advantage [63]. Recent progress in MMIC technology for silicon microwave devices [64], [65], as well as the growing demand for high speed communications, makes high speed modulator implementation using PIN diodes a viable option [58].

As stated in Section 1.1.2, the objective of this thesis is to design an cost-effective wireless ATM transmitter for fixed applications with an efficient bandwidth utilization. The above literature review shows that a good architecture for a radio ATM transmitter has not yet been reported. There is a great demand for a direct microwave modulator due to the rapidly increasing demand for high bit rate wireless ATM systems. An accurate adaptation metric to adapt the modulator according to incoming traffic has not yet been reported. A real time adaptation metric will improve the throughput of the link and increase radio system availability. In addition, a direct microwave high speed QAM modulator implementation using a variable attenuator has not been reported yet. This implementation will be very useful if a variable attenuator can be implemented using PIN diodes. However, a suitable model for analysing a PIN diode operation in a high speed application has not been
addressed in literature [66]. Each of these issues will be addressed in this thesis.

1.3 Research Objectives

The objectives of the research work reported in this thesis are as follows:

- To design an optimum architecture for a fixed wireless ATM transmitter
- To derive a metric to estimate the required bandwidth of the ATM traffic sources in the wireless network
- To design an adaptive transmitter that optimises the utilisation of the wireless resources
- To design a high performance QAM modulator for high speed data rates
- To develop a hardware implementation for the proposed modulator, leading to a cost effective radio ATM
- To model and characterise a PIN diode to use in the high bit rate modulator.

1.4 Thesis Organisation

In addition to the introductory chapter, this thesis contains seven chapters.

In Chapter 2, design concepts for a radio ATM transmitter are developed. The focus is to introduce an architecture to realize an optimum fixed radio ATM transmitter. Following a model extraction for an effective bandwidth estimation, the physical layer issues in ATM networking, adaptive transmitter design, and direct microwave QAM modulator are described.
In Chapter 3, an accurate metric to estimate the required bandwidth of broadband traffic in wireless channels is introduced. The metric is established by examining the required bandwidth for empirical video and data traffic traces, and developing a statistical model to estimate the effective bandwidth in a wireless channel. The statistical results are compared with those obtained from computer simulations.

Following extraction of a capacity reduction factor for multipath fading, a model to realize an adaptive M-QAM transmitter, which optimises the wireless resource utilisation, is described in Chapter 4. The transmitter performance is analysed for different ATM traffic loads in Ricean and Rayleigh channels.

In Chapter 5, a new direct microwave QAM architecture is introduced. An analytical method is developed to investigate the operation of direct microwave QAM modulators using PIN diodes. This requires an analytical model to characterise the PIN diode for high speed operation.

Microwave integrated circuit (MIC) and silicon monolithic microwave integrated circuit (Si-MMIC) techniques for implementation of the direct microwave M-QAM modulator are investigated in Chapter 6. The microstrip and co-planar waveguide (CPWG) line circuits are used to realize the modulator. While realization using the MIC is simple and straightforward, the Si-MMIC realization offers a low cost and high performance implementation method.

A system study based on analytical, simulation and measurement results
is reported in Chapter 7. Finally, conclusions are presented in Chapter 8.

Some of the results of this research work have been patented [7], published [2], [5], [6], [8], [10], [121], will publish [12], and submitted to [3], [9], [11] various journals and conferences.
2. FUNDAMENTAL CONCEPTS FOR WIRELESS ATM TRANSMITTER

The design of a broadband wireless network using ATM techniques requires that networking issues and radio transmission issues be considered simultaneously. A practical design must take into account the ATM networking characteristics and wireless transmission parameters [67]. In this chapter a brief discussion of some fundamental concepts in ATM networking and high speed wireless communications are presented. The chapter is organised as follows. Following a review of ATM networking, a model is developed to estimate the required bandwidth of a high speed ATM link. The impact of the cell loss ratio due to the wireless channel on the end-to-end QoS is studied in the next section. This is followed by a discussion of an adaptive modulator for a fixed wireless link.

2.1 Fundamentals of ATM Networking

2.1.1 ATM Basic Concepts

The basic idea of the ATM technique is to segment a digitally coded information stream (such as data, voice, and video) into a sequence of elementary blocks, called cells, that are transported and routed through the telecommunication network. The cell is the basic element in ATM networking. It is a short and fixed length packet of information. Cells pertaining to independent connections and with different generation rates can be carried on the same link and are accepted as they are generated. This uses statistical multiplexing instead of fixed-frame time division multiplexing. The ATM technique is based on packet processing thus enabling the implementation of
basic network functions using a high bit rate [68].

The cell, which is 53 bytes long, is made of two main fields: the header used for switching, and the information field, which transports the user information. The content and structure of the header are different at the user network interface (UNI) and at the network node interface (NNI) [3]. The cell header is 5 bytes long and contains two identifiers to route two separate logical entities: the virtual channel (VC) and the virtual path (VP). An NNI cell header is shown in Fig.2.1. The header is divided into virtual path identifier (VPI), virtual channel identifier (VCI), payload type, cell loss priority (CLP) and header error correction (HEC). HEC is employed to protect the header from transmission errors. It can correct one bit error in the header and detect a large class of multiple bit errors. In the sender, the eight bit cyclic redundancy check (CRC) across the first four octets of the cell header are calculated and the results are inserted in the HEC field which is the 5th byte of the header. It is specified by the generator polynomial [68]

\[ g(x) = x^8 + x^2 + x + 1. \]  \hspace{1cm} (2.1)
The physical layer is divided into a number of different virtual paths, each with a unique VPI. Each of these VPs may be then subdivided into a number of separate virtual channels, each with a separate VCI. The physical configuration is illustrated in Fig. 2.2.

2.1.2 ATM Services

The services to be provided by ATM networking may be categorised into interactive services and distribution services. Interactive services include video and multimedia conferencing, transmission of high-resolution images, document browsing as well as other applications [69]. Distribution services involve distribution from centralised service providers to network users e.g., files, stored video, movies, images and so on. These services have various QoS requirements. Different services in an ATM network are classified in the following classes:

- CBR: Constant bit rate connections require a fixed bandwidth channel with tight delay variation constraints (e.g., voice),
- VBR-rt: Real time variable bit rate (VBR-rt) services that require a small delay and fixed timing relationship (e.g., audio/video for real time conference),
- VBR-nrt: Non-real time variable bit rate connections that are used for bursty applications which do not require real-time performance and in
which the delay does not have a heavy impact on the service (e.g., data transfer),

- ABR: Available bit rate services that do not support bounded delay for delivery but support variable bit rate traffic. The network provides a best effort service which is not suitable for real time applications,

- UBR: The unspecified bit rate connections that are not specified by any parameters allowing the user to send any amount of the data over the network. However, using this service, the network gives no guarantee on cell loss rate, delay, and delay variations.

Typically a large amount of the traffic in an ATM network is VBR services.

### 2.1.3 Bandwidth Demand and QoS Requirements

Different ATM services require various bandwidth and QoS performance. These requirements are specified by traffic and QoS related parameters. A traffic parameter, e.g., peak cell rate (PCR), is specified for the source, and a QoS parameter, e.g., cell loss ratio (CLR), describes the expected QoS. According to the ATM Forum [69], various QoS and traffic requirements include cell loss ratio (CLR), cell transfer delay (CTD), cell delay variation (CDV), cell delay variation tolerance (CDVT), peak cell rate (PCR), sustainable cell rate (SCR), maximum burst size (MBS), and minimum cell rate, among others. An ATM call is admitted under a traffic contract consisting of a subset of these parameters. The QoS related parameters (CLR, CTD, CDV) and the peak bandwidth requirements for voice, interactive video, media playback (e.g., HDTV), and file transfer are shown in Table 2.1 [70].

### 2.2 ATM Multiplexer

An ATM multiplexer is shown in Fig.2.3. In ATM terminology, a source is a terminal. This could be a telephone handset, a video player, or a multi-
<table>
<thead>
<tr>
<th>Table 2.1</th>
<th>QoS Requirements</th>
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</thead>
<tbody>
<tr>
<td>Voice</td>
<td>Video</td>
</tr>
<tr>
<td>CLR</td>
<td>$10^{-3} - 10^{-6}$</td>
</tr>
<tr>
<td>CDV(msec)</td>
<td>0-5</td>
</tr>
<tr>
<td>CTD$_{max}$(msec)</td>
<td>10-150</td>
</tr>
<tr>
<td>PCR(Mbps)</td>
<td>8-64 kbps</td>
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</table>

media computer. When a source wants to transmit information, it requires the establishment of a virtual channel (VC). Once a source has established a VC, it generates a stream of cells, each cell consisting of 53 bytes. A typical cell stream generated by a VC consists of silent periods, during which no cells are generated, and activity periods, during which cells are generated at a variable rate. The group of cells generated during an activity period is called a burst.

An ATM multiplexer consists of a data buffer and a high speed link; the buffer receives the cells generated by the established VC and transmits these cells, one after another, onto the high speed link [37]. A VC's allowable cell delay and cell loss are specified by its quality of service (QoS) requirements. To guarantee that all established VCs meet their QoS requirements, the multiplexer may have to deny certain VC establishment requests.

When a VC demands capacity to be transmitted using an ATM multiplexer, if there is enough capacity in the high speed link, its request is accepted; otherwise the call will be rejected. An admission policy is needed to control the call acceptance process. A simple admission control is the peak rate admission policy [69]. The peak rate admission constrains the aggregate peak rate to be less than the transmission capacity of the high speed link. It enforces these constraints by rejecting a VC establishment request when the
VC’s peak rate added to the sum of the peak rates for the established VCs exceeds the transmission capacity. As an example, suppose a source has an average bandwidth of 20 Mbps and a peak bandwidth 55 Mbps. Peak bandwidth allocation requires that 55 Mbps be reserved at the high speed link for this specific source, independent of whether or not the source transmits 55 Mbps continuously.

Let $C$ denote the transmission capacity of a high speed link; $K$ denote the number of services, and $b_1, \ldots, b_K$ denote the peak rates for the $K$ services. The VC profile is $(n_1, \ldots, n_k)$, where $n_k$ is the current number of established VCs of service-$k$. Since VCs arrive and depart, the VC profile changes with time. Peak rate admission admits a new service-$k$ VC if and only if

$$b_k + \sum_{l=1}^{K} b_l n_l \leq C. \quad (2.2)$$

Consequently, at all times the VC profile $(n_1, \ldots, n_k)$ satisfies

$$\sum_{k=1}^{K} b_k n_k \leq C. \quad (2.3)$$

On the other hand, statistical multiplexing permits the aggregate peak
rate to exceed the transmission capacity. This technique can utilise the link more efficiently, allowing the link to transmit at its maximum rate even when some of the established VC's are silent. To implement the statistical multiplexing, we must determine whether a given collection of established VC's meets the QoS requirements.

An effective bandwidth admission is an admission policy which is easy to implement when the ATM multiplexer operates in the statistical multiplexing mode. This policy is characterised by an effective bandwidth vector \((b^e_1, ..., b^e_K)\) and it admits a new service-\(k\) VC if and only if

\[
b^e_k + \sum_{i=1}^{K} b^e_i n_i \leq C, \tag{2.4}
\]

where \((n_1, ..., n_k)\) is the current VC profile. However, determining suitable values for \(b^e_1, ..., b^e_K\) is a challenging problem in ATM networking theory.

### 2.3 Bandwidth Estimation

In an ATM network, a virtual path is an information transport that makes a logical direct link between two nodes and accommodates a number of virtual channels simultaneously. A predefined route is usually defined for each virtual path in the physical layer. Each virtual path has a bandwidth, in other words, "capacity", which defines the upper limit for the total virtual channel bandwidth carried by it. Virtual paths are also multiplexed on physical transmission links using cell multiplexing [71]. While mapping various services to the VPs is an active research subject [72],[73], one proposal suggests having a separate VP for each service class or even for each different set of QoS requirements within the same service class. Although QoS control is easier, the total number of VPs could become large.
Figure 2.4 A Finite Buffer with VP as an output

Let us again consider the multiplexer model shown in Fig.2.3. Each high speed link in Fig.2.3 can be considered a virtual path. This is illustrated in Fig.2.4. As may be seen, different VCs have to compete to use the available bandwidth of a VP.

The above model may be used to examine the capacity of total VPs in a physical layer. Fig. 2.5 shows this model with $M$ VPs feeding into a communication system. Considering the peak rate admission for this model, one could compute the capacity for $VP_1$ as follows:

$$\sum_{k=1}^{K_1} b_{k,1} n_{k,1} \leq C_1, \quad (2.5)$$

where $(n_{1,1}, n_{2,1}, \ldots, n_{K_1,1})$ is the current VC profile for $VP_1$. Similarly, the VC profile for $VP_i$ will be

$$\sum_{k=1}^{K_i} b_{k,i} n_{k,i} \leq C_i, \quad (2.6)$$

Total capacity of the physical connection is obtained using superposition of the individual VPs. Thus, the total admission policy of the link can be written as follows:

$$\sum_{k=1}^{K_1} b_{k,1} n_{k,1} + \ldots + \sum_{k=1}^{K_M} b_{k,M} n_{k,M} \leq C, \quad (2.7)$$
where $C$ is the total capacity of the high speed link. This equation can also be written as

$$\sum_{i=1}^{M} \sum_{k=1}^{K_i} b_{k,i} n_{k,i} \leq C. \quad (2.8)$$

For the statistical multiplexing mode, the above equation is modified as follows:

$$\sum_{i=1}^{M} \sum_{k=1}^{K_i} b_{k,i}^e n_{k,i} \leq C, \quad (2.9)$$

where $(n_{1,i}, n_{2,i}, ..., n_{K_i,i})$ is the VC profile for $VP_i$, and $(b_{1,i}^e, b_{2,i}^e, ..., b_{K_i,i}^e)$ are the effective bandwidths of VCs in $VP_i$. It is assumed that each output port is connected to a VP.

In a nonblocking switch, as shown in Fig.2.6.a, a buffer has been used in each output port and the point of connection occurs at the output ports [73]. A switch fabric is nonblocking if it can deliver all packets to the requested output ports. A simplified model for bandwidth estimations of a VP is also shown in Fig.2.6.b. In a network node where the switch routes VPs, this model can be used as well. Thus, the effective bandwidth of the $ith$ VP can be shown to be:
Figure 2.6  a) A Nonblocking Switch b) Equivalent circuit for capacity estimation

\[ BW_i^e = \sum_{k=1}^{K_i} b_{k,i} n_{k,i}. \]  \hspace{1cm} (2.10)

Then, the capacity of the total link will have the following relation with the effective bandwidth of the individual VP's:

\[ \sum_{i=1}^{M} BW_i^e \leq C. \]  \hspace{1cm} (2.11)

As can be seen, the effective bandwidth of the VP's are related to available capacity of the link. Moreover, this relation shows that the required bandwidth in a high speed link is bound by the linear superposition of the VP's' effective bandwidth. These interesting results will be used to design an adaptive transmission system in the following chapters.
2.4 End-to-End QoS Guarantees in Radio ATM Networks

End-to-end performance with multiple traffic types and multiple QoS requirements needs the derivation of an overall bound that always guarantees the operation of a wireless ATM network with desired QoS [69]. Estimation of this bound resulting in an optimum network utilisation is a very difficult task. While extraction of an end-to-end delay bound is in itself an interesting networking problem that has attracted much attention [74], [75], the focus of this research is on the cell loss performance bound. The cell loss ratio is defined as being the ratio of discarded cells to transmitted cells.

A reference model is considered to estimate a bound for cell loss ratio in a fixed radio ATM link. This model is shown in Fig.2.7. As may be seen, two different mechanisms may cause cell loss ratio in the wireless ATM communications. While congestion results in cell loss ratio at nodes, the radio link type of the physical layer introduces bit errors resulting in an increase of the cell loss ratio. The bit errors in an optical fiber link are random and occur rarely. However, for the radio channel the conditions change with time.
Such a channel could suffer from severe degradation for some fraction of time [74].

2.4.1 Cell Loss Ratio Due to Congestion

In an ATM network, congestion could occur at various switches and/or buffers when buffers overflow and cells are dropped resulting in a decrease in the throughput. In a nonblocking switch, if there are more requests to use a $VP_i$ than it can handle, the cells in the $ith$ buffer will start to drop.

2.4.2 Cell Loss Ratio in a Radio Channel

The desired channel for the transmission of ATM traffic is one with a low bit error rate. The study of the problems associated with transmission of ATM traffic over a wireless channel is important because the original ATM network was designed to use a high performance physical layer such as optical fiber. In this section, the impact of a wireless channel on ATM traffic is examined. In the following analysis, only the CLR resulting from the corruption of the header in the ATM cells is considered.

Quality of Service for the Random Error Channel

When errors occur randomly, the error distribution is binomial. The header error correction (HEC) has been designed to correct a single bit error in the 40 bit header [76]. Using binomial distribution, the CLR in this case is given by:

$$CLR_R = \sum_{n=2}^{40} \binom{40}{n} p^n(1-p)^{40-n},$$

(2.12)

$$CLR_R \approx 780 \times p^2,$$

(2.13)

where $p$ is the bit error rate and $CLR_R$ is the cell loss ratio in the random error channel. Equation (2.13) relates $CLR_R$ to the bit error rate in the
random channel and shows that CLR is a second order function of the bit error rate. In Fig.2.8.a the relation between bit error rate and CLR for a random channel is presented.

**Quality of Service in the Burst Error Channel**

Burst errors increase the probability of cell discard [5]. A suitable probability model in this case is the Newman-A contagious model [77]. The two parameters \(m_1\) and \(m_2\) in this probability model depend on the average burst length \(\beta\), the header length \(l\), and the probability of bit error \(p\). These parameters are given by \(m_1 = \frac{lp}{\beta}\) and \(m_2 = \beta\). The probability of \(k\) bit errors \(P(X = k)\) is given by:

\[
P(X = k) = \frac{\beta^k}{k!} e^{-ip/\beta} \sum_{n=0}^{\infty} \frac{(lp/\beta)^n}{n!} n^k/n!,
\]

(2.14)

Considering that the cell loss ratio occurs due to uncorrected header error, the cell loss ratio for the bursty channel is given by:

\[
CLR_B = 1 - P(X = 1) - P(X = 0),
\]

(2.15)

An approximation to Eq.(2.15) can be given by:

\[
CLR_B = 1 - e^{-40p/\beta} \left[1 + \frac{40p}{\beta} (1 + \beta) e^{-\beta}\right].
\]

(2.16)

In Fig.2.8.a the relation between bit error rate and CLR for the bursty channel is plotted with burst width as the parameter. For a given BER a higher burst width actually results in slightly lower CLR. This may be attributed to the fact that when the errors are lumped together they affect fewer cells. These analytical results are compared with the previously published measurements in Fig.2.8.b [25]. As may be seen, the agreement is quite good for a burst width of two bits.
Figure 2.8  a) CLR versus BER for random and bursty channels b) A comparison between cell loss estimation results for the bursty channel with the previously published results [25].
2.5 Adaptive Radio ATM Transmitter

An important characteristic of an ATM network is the bandwidth on demand feature. In this network, the total number of connections varies with time, and thus, the bandwidth demand varies with time. As stated before the ATM sources require different QoS. For instance, while the voice traffic is mostly delay sensitive and relatively loss insensitive, the data traffic is quite sensitive to loss although some loss can be tolerated. Further, even in a single class of traffic, different applications require different QoS. For example, with video traffic, TV broadcasts and video on demand (VoD) services have more relaxed delay requirements than video conferences and multimedia on demand (MoD) where a strict delay bound is required to support interactivity. Also, while TV broadcast may require high picture quality, video conference and VoD/MoD services may trade off, at times, picture quality for high bandwidth efficiency [78]. All these issues make an adaptive system that can provide bandwidth on demand highly desirable.

In addition, a radio channel is essentially a time varying channel. The channel variation results in a different bit error rate (BER) and cell loss ratio (CLR) variations. Thus, the radio channel must be monitored so as to maintain the QoS. A simple method to adapt the system to the channel variation is to use a channel estimator in conjunction with a feedback channel where feedback channel only carries the carrier [79]. This is shown in Fig.2.9.

2.6 Modulation Techniques for Fixed Radio ATM

Due to the spectrum limitation in a radio channel, a high bandwidth efficiency modulation method is required. As discussed above, an adaptive realization of the modulator is also highly desirable. M-level modulation techniques such as M-QAM, and M-PSK are possible candidate modulation methods for fixed radio ATM because of their high bandwidth efficiency and
Figure 2.9 An adaptive modulator in a time varying channel

Figure 2.10 Constellations for QAM signals

their adaptability features [79]. When the radio bandwidth efficiency and power requirements are examined, the M-QAM offers a better trade-off. The QAM modulation is discussed in the next few subsections.

2.6.1 Quadrature Amplitude Modulation

Quadrature amplitude modulation is a two-dimensional linear modulation technique and the QAM modulated signal may be expressed as:

\[ s_m(t) = A_{mc}u(t)\cos2\pi f_c t - A_{ms}u(t)\sin2\pi f_c t; \quad m = 1, 2, ..., L, \quad (2.17) \]

where \( u(t) \) is a signal pulse, \( f_c \) is carrier frequency, \( A_{mc} \) and \( A_{ms} \) are the inphase and quadrature signal amplitudes, and \( L = \sqrt{M} \). Constellations of this modulation in the I-Q planes are shown in Fig.2.10.
2.6.2 Spectral Efficiency of QAM

The lowpass equivalent of the QAM signal is given by:

\[ v(t) = \sum_n (a_n + jb_n)u(t - nT); \quad a_n, b_n = \pm 1, \pm 3, \ldots, \pm L - 1, \]     \hspace{1cm} (2.18)

where \( T \) is the symbol rate, and \( u(t) \) is pulse shape. The bandpass signal \( s_m(t) \) can be related to the lowpass signal \( v(t) \) through

\[ s_m(t) = Re[v(t)e^{j2\pi f_c t}], \]     \hspace{1cm} (2.19)

The autocorrelation function of \( s_m(t) \) is given by:

\[ \phi_{ss}(\tau) = Re[\phi_{vv}(\tau)e^{j2\pi f_c \tau}], \]     \hspace{1cm} (2.20)

where \( \phi_{vv}(\tau) \) is the autocorrelation function of the equivalent low pass signal \( v(t) \). Then, the power spectral density of \( \Phi(f) \) is given by:

\[ \Phi_{ss}(f) = \frac{1}{2}[\Phi_{vv}(f - f_c) + \Phi_{vv}(-f - f_c)] \]     \hspace{1cm} (2.21)

where \( \Phi_{vv}(f) \) is the power spectral density of \( v(t) \). It can be shown [80] that the power spectral density of \( v(t) \) is given by:

\[ \Phi_{vv}(f) = \frac{1}{T}|U(f)|^2\Phi_{ii}(f), \]     \hspace{1cm} (2.22)

where \( U(f) \) is the Fourier transform of \( u(t) \) and \( \Phi_{ii}(f) \) is the power spectral density of the information sequence. The QAM signal may be considered to consist of two PAM signals conveyed by the cosine and sine carriers. Thus, the spectral efficiency of the QAM will be twice that its of PAM components in the inphase and the quadrature channels. Substitution for \( \Phi_{ii}(f) \) in the above equation gives the power spectral density the PAM signal carries by
cosine carrier as [80]:

\[ \Phi_{uv}(f) = \frac{\sigma^2}{T} |U(f)|^2 + \frac{\mu^2}{T^2} \sum_{m=-\infty}^{m=\infty} |U\left(\frac{m}{T}\right)|^2 \delta(f - \frac{m}{T}), \]  

(2.23)

where \( \mu \) and \( \sigma^2 \) are the mean and variance of the information sequence. As may be seen, the mean value of the PAM symbols is zero \( (\mu = 0) \), and, thus, Eq.(2.23) can be simplified as being

\[ \Phi_{uv}(f) = \frac{\sigma^2}{T} |U(f)|^2. \]  

(2.24)

Thus, the spectral efficiency of the PAM signal is controlled by the pulse shape \( u(t) \). The raised cosine pulse is a commonly used pulse shape for the digital communications in bandlimited channels. The signal \( u(t) \) for this pulse shape is given by:

\[ u(t) = \frac{\sin(\pi t/T) \cos(\pi \alpha t/T)}{\pi t/T \left(1 - (2\alpha t/T)^2\right)} \]  

(2.25)

where \( \alpha \) is called as roll off factor. The Fourier transform \( U(f) \) is given by:

\[ U(f) = \begin{cases} \frac{T}{2} & \text{if } 0 \leq |f| \leq \frac{1-\alpha}{2T} \\ \frac{T}{2}[1 - \sin(f - \frac{1-\alpha}{2T}/\alpha)] & \text{if } \frac{1-\alpha}{2T} \leq |f| \leq \frac{1+\alpha}{2T} \end{cases} \]

The normalised spectrum of the raised cosine pulse shape is shown in Fig.2.11.

Thus, the bandwidth efficiency of QAM modulator can be adjusted by parameter \( \alpha \). As may be seen from Fig.2.11, the channel bandwidth, \( W \), is approximately equal to \( (1+\alpha)/T \) and, since \( 1/T = C/log_2(M) \) symbols/sec, we obtain the result:

\[ \eta = \frac{C}{W} = \frac{log_2 M}{1 + \alpha}, (bps/Hz) \]  

(2.26)
where $\eta$ is M-QAM spectral efficiency, and $C$ is the M-QAM transmission capacity.

### 2.6.3 Power Efficiency

Although spectral efficiency may be the primary criterion, the power efficiency of the selected method is also an important factor in system design. Power efficiency is a measure of the received power needed to achieve a specified bit error rate (BER). For an additive white Gaussian noise (AWGN) channel, the probability of symbol error in the M-QAM system is estimated as follows[80]:

$$P_{MQAM} = 2(1 - \frac{1}{\sqrt{M}})erfc\left(\sqrt{\frac{3}{2(M-1)k\gamma_b}}\right) \left[1 - \frac{1}{2}(1 - \frac{1}{\sqrt{M}})erfc\left(\sqrt{\frac{3}{2(M-1)k\gamma_b}}\right)\right]$$

where $k$ is even ($k$ is bits per symbol and $M = 2^k$), $\gamma_b$ is the average signal-to-noise ratio (SNR) per bit, and $M$ is the modulation level. BER results for coherent demodulation with 2, 4, 6 and 8 bits per symbol and assuming
perfect clock and carrier recovery in a Gaussian channel [81] are shown in Fig.2.12.

2.7 QAM Modulator Implementation

The QAM modulator implementation has a considerable impact on the radio ATM performance and cost. Different design approaches used in the past could be categorised into the two categories of heterodyne and direct modulation method. Each approach has its specific advantages and disadvantages, and a selection is based upon system requirements and design objectives.

2.7.1 Classic Implementation

The QAM modulator has been traditionally realized using the heterodyne method shown in Fig.2.13. The QAM mapper in the figure divides a binary sequence with rate $f_b$ into two binary symbol streams, each with a rate $f_b/2$. 
Figure 2.13 Block diagram of a heterodyne QAM modulator

A 2-to-$\sqrt{M}$ level baseband converter is used to convert the streams into $\sqrt{M}$-level baseband PAM signals in the inphase or the quadrature paths. These symbols are shaped by the raised cosine filters. The filtered I and Q baseband signals are used to modulate the inphase and quadrature outputs of an IF oscillator. The modulated signal at the IF is up-converted to the desired transmit frequency in one or more steps. As may be seen, mixers are commonly used in this realization. A power amplifier is used to boost the modulated signal at the transmit frequency to the required power.

2.7.2 Direct Microwave Implementation

The classic heterodyne design described above suffers from disadvantages such as design complexities, high costs, RF filter requirements, and data rate limitation if IF frequency is low. The direct implementation methods could effectively overcome these limitations. The basic idea of direct microwave implementation is to use the baseband signal to modulate a carrier at the desired transmitted frequency. A basic direct microwave modulator implementation is illustrated in Fig.2.14.

2.8 Radio Microwave Link Design

It is well known that multipath fading has a significant adverse effect on the performance of a wireless communications system. In a broadband
microwave system, the fading generally consists of a combination of flat and frequency selective components. The flat component is a time-varying, frequency independent attenuation of the channel response and, hence, of the signal. This attenuation can be thought of as the slow variation of the channel loss over a broad range of frequencies. The frequency selective component of the fading could be considered to be the frequency selective variation of the power response. It shows itself either as a monotonic gain change (or slope) across the radio channel bandwidth, or as a dip (or notch) within the bandwidth. Multipath fading arises from the fact that the signal propagates along several paths, each of different electrical length. At the receiver, the relatively delayed signal components interfere with each other, and this leads to the frequency selective effects described above. In a line of sight (LOS) microwave link, the received signal generally consists of a strong direct wave and some reflected waves. A Ricean distribution model for fading can be used in this case [82].

The radio link tolerance to multipath fading is usually specified through the radio link outage probability. This metric is closely related to the radio link fade margin and system gain; the system gain being the difference between transmit output power and receiver threshold sensitivity for a given
bit error rate. The system gain is given by the following relation [83]:

\[ FM = G_s - L_P - L_F - L_B + G_T + G_R, \] (2.27)

where \( G_s \) is the system gain, \( FM \) is the fade margin, \( L_P \) is the free space path loss(dB), \( L_F \) is the feeder loss, \( L_B \) is the branching loss, and \( G_T \) and \( G_R \) are gains of transmitter and receiver antennae. The outage probability is related to the fade margin through the following relation [83]:

\[ 10 \log(1 - R) = 30 \log(d) + 10 \log(6Af) - 70 - FM, \] (2.28)

where \((1 - R)\) is the outage probability, \(d\) is the path length in km, \(f\) is the carrier frequency in GHz, and \(A\) is the terrain roughness factor.

A typical example of a short haul radio link for an 18 GHz system with 64QAM modulation is given in [5]. The path length is 10 km. The same link is used for the performance study in Chapter 7. The height of transmit and receive antennae is selected as 7 meters so as to achieve an LOS path of 14.3 km. Both antennae are parabolic dishes with diameters of 2 feet and a nominal gain of 38.9 dB. The HPA was assumed to be a commercial MMIC with power output of 23 dBm. The results of this link design are summarised in Table 2.2.

2.9 Summary

In this chapter different issues to realize a transmitter for fixed wireless ATM applications were studied. The study examined both ATM networking and radio transmission issues. A relation was derived to relate the effective bandwidth of the VPs to the available capacity of the link. It was shown that CLR in wireless ATM networks results from buffer congestion and radio channel BER. A relation between BER and CLR was derived for a radio
<table>
<thead>
<tr>
<th>Transmitter</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Transmitter Power</td>
<td>23 dBm</td>
</tr>
<tr>
<td>Transmitter Antenna Gain</td>
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</tr>
<tr>
<td>Antenna Height</td>
<td>7 m</td>
</tr>
<tr>
<td>Space Link</td>
<td></td>
</tr>
<tr>
<td>Loss</td>
<td>142.19 dB</td>
</tr>
<tr>
<td>Total Loss</td>
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</tr>
<tr>
<td>Multipath Level</td>
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</tr>
<tr>
<td>Receiver</td>
<td></td>
</tr>
<tr>
<td>Receiver Antenna Gain</td>
<td>38.9 dB</td>
</tr>
<tr>
<td>Noise Figure</td>
<td>3 dB</td>
</tr>
<tr>
<td>Receive SNR&lt;sub&gt;θ&lt;/sub&gt;</td>
<td>46.81 dB</td>
</tr>
<tr>
<td>Theoretical SNR&lt;sub&gt;θ(10^{-8})&lt;/sub&gt;</td>
<td>21 dB</td>
</tr>
<tr>
<td>Implementation Loss</td>
<td>5 dB</td>
</tr>
<tr>
<td>Fade Margin</td>
<td>20.81 dB</td>
</tr>
<tr>
<td>Outage (for 14.3 km)</td>
<td>.03&lt;sub&gt;θ&lt;/sub&gt;/° (Availability 99.97&lt;sub&gt;θ&lt;/sub&gt;/°)</td>
</tr>
</tbody>
</table>

channel. The concept of an adaptive modulator was presented to provide bandwidth on demand over a wireless channel. The QAM modulator was selected as a suitable modulation technique to realize an adaptive modulator in a wireless channel. Different implementation techniques to realize an M-QAM modulator were investigated and a direct microwave implementation was selected as a better choice for fixed wireless ATM applications. The following chapter discusses the concept of bandwidth estimation in a wireless broadband network.
3. BANDWIDTH ESTIMATION OF BROADBAND TRAFFIC

The bandwidth estimation of the broadband traffic sources is an important issue in realizing a wireless broadband network. The concern is how a unique traffic parameter should be defined to represent the required capacity in a wireless broadband network. This issue is important in wireless ATM design as the capacity is closely related to the cost and complexity of the radio ATM.

A straightforward design is to develop a radio to provide enough bandwidth for the peak rate of existing traffic sources. Although this solution is simple, it is neither an optimum nor a practical design. It is not optimum because at a given time using all available resources, it only provides service to some of the users. It is not practical because a very broadband radio system needs a very broad spectrum band and a rather complex system design.

A better alternative method is a demand based design. In this method, the bandwidth is assigned to the different traffic sources based upon their traffic characteristics. This method, however, requires that the traffic sources describe their traffic and QoS parameters before a link can be established. In this chapter an accurate metric to characterise the required bandwidth in a wireless ATM network is introduced. The metric is established by examining the required bandwidth for empirical video and data traffic traces, and developing a statistical model to estimate the effective bandwidth in a broadband network. The results of the statistical model are compared with
Figure 3.1 a) First sixteen seconds of Jurassic Park video-
trace b) Total trace(27 minutes)

simulation in Section 3.4.

3.1 Characterisation of Broadband Traffic

It is well known that a broadband traffic source, e.g., video or data, may
be described using statistical models. The investigation of the empirical
traffic traces can be the first step to build a suitable statistical model. This
is the subject for the next subsection.

3.1.1 Video Traffic

As an example of a video traffic source, the MPEG-1 encoded trace of the
Jurassic Park movie is examined. In the MPEG-1 encoding for this trace,
the individual packets correspond to the data in video frames produced at a
constant frequency of 25 frames per second [84]. The trace record consists
of the number of bits per frame in a sequence of 40,000 consecutive frames.
Fig.3.1 shows the bits per frame of the trace. A strong periodic component
can be seen in the trace which arises from the MPEG-1 encoding format. In
this format groups of 12 frames are encoded into a fixed pattern (known as the Group of Pictures (GOP)) of I-frames, P-frames and B-frames according to varying amounts of motion compensation between frames. For this trace, I-frames carry the most data and occur every 12 frames.

As may be seen, the number of bits per frame is a time varying function. Thus, the trace requires a variable channel capacity for transmission. On the other hand, if a peak rate allocation system is used, the bandwidth has to be reserved according to the maximum bit rate. For this trace, the maximum frame size is 119632 bits, and for the frame rate 25 frames per second, a link with 1.1 Mbps capacity will be required if peak rate is used as the allocation criterion.

3.1.2 Ethernet Traffic

The data trace of the Ethernet traffic source is also examined. These data were collected on different Bellcore LAN networks in August'89 by Bellcore researchers [85]. The first 500 seconds of this traffic is presented in Fig.3.2. The trace contains the time stamps and length of packets in bytes. As may be seen, the required capacity is once again a time varying function with the average lower than peak rate of 10 Mbps of Ethernet traffic.

3.2 Estimation of the Required Bandwidth of Broadband Traffic

The above discussion of the video and Ethernet data traces shows that the required bandwidth for the broadband network is time variable. It is also clear that the peak rate bandwidth allocation is not a bandwidth efficient mechanism. An efficient bandwidth allocation policy may be designed if one relates the required capacity to the desired quality of service. In this section, a series of simulations is conducted to estimate the required bandwidth of
the real traffic traces based on their desired QoS.

3.2.1 Simulation Model

As mentioned earlier, the objective is to separate the ATM switch operation from the cell transmission system. A model for a non-blocking ATM switch was discussed in Chapter 2. As explained in Section 2.3, the switching operation is done by an ATM fabric that is independent from the capacity estimation section. The operation of the ATM switch fabric has been extensively investigated in the literature [69]. A model for cell transmission is presented in Fig.3.3.
3.2.2 Bandwidth Estimation for Empirical Video Traffic

A basic multiplexer was simulated to estimate the required bandwidth of the video traces. It is assumed that all the data of the frame arrive back to back at the incoming link rate (Table 3.1), and that they have been converted into ATM cells with 48 bytes of frame data. No header was added for sake of simplicity. Cells that do not fit into the buffer at the time of arrival are discarded. The only performance metric of interest is the cell loss ratio (CLR) at the multiplexer’s output buffer. The delay constraints were not considered. The simulator is used to estimate the required bandwidth of an MPEG-1 encoded Jurassic Park movie. The total trace contains 40000 frames (approximately 27 minutes of video). The data trace constituted different frame sizes in bits [84]. The results of the simulations are presented in Fig.3.4. As may be seen, by increasing the buffer size, the required bandwidth decreases for a constant cell loss ratio. The simulation was also repeated for the MPEG-1 encoded Silence of the Lambs and Star Wars movies with 40000 frame traces. The results are shown in Fig.3.5 and Fig.3.6.

The next simulation was conducted to estimate the required bandwidth of the aggregate of the three MPEG-1 encoded movies, namely Jurassic Park, Silence of the Lambs, and Star Wars. The result is shown in Fig.3.7. A phase shift 160 msec was selected between the videotraces to make sure that the I frames are not all in synchronization with each other. As may be seen, the required bandwidth of the aggregate of the movies is less than the required bandwidth of the superposition of the individual traces. This demonstrates the statistical multiplexing gain for the aggregate traffic.

3.2.3 Bandwidth Estimation for Ethernet Traffic

A similar simulation was also conducted to estimate the required bandwidth of the Ethernet data trace. The trace of Ethernet LANs measured by
Figure 3.4 Simulated results for the required bandwidth of a VBR videotrace *Jurassic Park* movie with buffer size as a parameter.

Figure 3.5 Simulated results for the required bandwidth of a VBR videotrace *Silence of the Lambs* movie with buffer size as a parameter.
Figure 3.6 Simulated results for the required bandwidth of a VBR videotrace Star Wars movie with buffer size as a parameter

Bellcore in August 1989 was used for the simulation. The required bandwidth as a function of the QoS for this trace is shown in Fig.3.8 with buffer size as a parameter. The simulated results presented in this section show that an optimum network requires a traffic bandwidth estimation. An analytical model which can be used for this estimation is developed in the next section.

3.3 Bandwidth Estimation Using Statistical Models

The ATM network is a connection oriented network meaning that a traffic source requires to establish a VC before transmission. This is achieved by first introducing the traffic parameters and the QoS parameters to the network from the traffic sources. These parameters can be used to develop a statistical model to estimate the required bandwidth of the broadband traffic.
Figure 3.7  Simulated results for the required bandwidth of aggregate VBR videotrace movie with buffer size as a parameter (phase shift 160 msec for each source)

Figure 3.8  Required bandwidth of Bellcore Ethernet traffic with buffer size as a parameter
The required bandwidth of the broadband traffic is generally characterised by an effective bandwidth. The effective bandwidth of a time varying traffic source represents the minimum required bandwidth that it needs to meet QoS guarantees. Assigning an effective bandwidth function to each traffic source, which depends not only on its mean bandwidth but also on its burstiness, gives a suitable tool to estimate the required bandwidth of that source [5],[86].

### 3.4 Effective Bandwidth using the Fractional Brownian Motion Approach

In a series of papers [38],[39],[40] researchers from Bellcore have reported the measurement results of real traffic sources. Their results show that the different bursty traffic sources have a self-similar or long-range dependent behaviour. For a long range dependent input process or self-similar process the tail of the queue has the following large buffer asymptotic behaviour [40]:

$$P[Q > \Pi] \sim e^{-\gamma \Pi^\epsilon},$$

where \( \Pi \) is the buffer size, \( \gamma \) is a constant and \( \epsilon = 2 - 2H \), is between zero and one. \( H \) lies between .5 and 1, and is the Hurst parameter or self-similar parameter. As compared to this, the large buffer asymptotic behaviour of the queue tail in a conventional Markovian model with no long range dependent behaviour is asymptotically exponential [87]. This may be approximated by:

$$P[Q > \Pi] \sim e^{-\eta \Pi}$$

where \( \eta \) is a positive constant. As may be seen, for \( H = .5 \) the two asymptotic relations (3.1) and (3.2) are identical. The estimation of the effective bandwidth using exponentially distributed idle and burst models underestimates the cell blocking probabilities for real traffic sources [40],[86]. This error is considerable even for moderate long range dependence, e.g., \( H = .7 \).
The estimation of the effective bandwidth of a self-similar traffic source is a challenging problem.

It is known that if traffic arrival could be considered as being a Gaussian process, a Weibull distribution may be used to estimate the buffer overflow [44]. Based upon the fractional Brownian motion approach, a theory has been recently proposed to deal with effective bandwidth of self-similar sources [45]. According to this theory the following equation holds for self-similar traffic:

\[
\frac{(1 - \rho)}{\rho^{1/2H}} BW_{eff}^{(H-1/2)/H} \sigma^{-1/(2H)} \Pi^{(1-H)/H} = \Omega
\]

(3.3)

where the constant \( \Omega \) depends on \( H \), the mean burst size, and the quality of service but is independent of the utilisation and buffer size. An explicit relation for the constant \( \Omega \) is not straightforward. However, an approximation of the upper bound may be used to estimate the effective bandwidth as [45]:

\[
BW_{eff} = R_{mean} + \left( K(H) \sqrt{-2 \log(P_{CLR})} \right)^{1/H} \sigma^{1/(2H)} \Pi^{-(1-H)/H} R_{mean}^{1/(2H)},
\]

(3.4)

where

\[
K(H) = H^H (1 - H)^{1-H},
\]

(3.5)

and \( P_{CLR} \) is the cell loss ratio, and \( \sigma \) is the variance coefficient defined to be the variance over the mean (bit-sec), \( H \) is the Hurst parameter of the stream (a dimensionless measure of long range dependence between .5 and 1), and \( \Pi \) is the buffer size (in bits). This effective bandwidth equation is used to estimate the required bandwidth of the Ethernet traffic. The effective bandwidth as a function of buffer size and Hurst parameter for a utilisation factor of \( u = .7 \) and a CLR of \( 10^{-6} \) is shown in Fig.3.9 (The utilisation factor is defined as \( u = R_{mean}/R_{peak} \)). As may be seen, increasing the \( H \) parameter generally increases the required bandwidth for a given quality of service and buffer size [88]. The effective bandwidth is also examined as a
function of the $H$ parameter and the utilisation for a CLR of $10^{-6}$. As shown in Fig.3.10 increasing utilisation also increases the effective bandwidth for a given quality of service and an $H$ parameter.

<table>
<thead>
<tr>
<th>The Video traffic sources</th>
<th>Mean [Mbps]</th>
<th>Var. Coeff. [b/s]</th>
<th>Hurst</th>
</tr>
</thead>
<tbody>
<tr>
<td>Jurassic Park</td>
<td>.326950</td>
<td>4.928e4</td>
<td>.8448</td>
</tr>
<tr>
<td>Silence of the Lambs</td>
<td>.182788</td>
<td>6.484e4</td>
<td>.8959</td>
</tr>
<tr>
<td>Star Wars</td>
<td>.232830</td>
<td>7.250e4</td>
<td>.8458</td>
</tr>
<tr>
<td>Aggregate stream</td>
<td>.742568</td>
<td>6.039e4</td>
<td>.8959</td>
</tr>
</tbody>
</table>

### 3.4.1 Comparison of Results for VBR Video Traffic

The traffic parameters of a VBR entertainment video source (MPEG encoded "Jurassic Park" movie) was estimated using this method and the results are shown in Table 3.1. The Group of Pictures (GoP) format used for the encoding is IBBPBBPBBPBB [84] and the frame rate is 25 frames per second. The trace contains 40000 frames (approximately 27 minutes of video). For this movie, the Hurst parameter was estimated as being $H = .845$. The effective bandwidth of the videotrace was estimated using Eq.(3.4) and the results are shown in Fig.3.11. As may be seen, increasing the buffer size decreases the required bandwidth for a given quality of service. A comparison of Fig.3.4 and Fig.3.11 clearly establishes that the effective bandwidth estimation method offers a close upper bound for the required bandwidth of this videotrace. This comparison is also repeated for the Silence of the Lambs and Star Wars movies in Figures 3.12 and 3.13. As may be seen, the effective bandwidth estimation using Eq.(3.4) again offers a suitable upper bound estimation for the required bandwidth. On the other hand, Fig.3.14 compares the effective bandwidth with the peak rate allocation. As may be seen, using the effective bandwidth can efficiently save bandwidth.
Figure 3.9  Effective bandwidth of Ethernet traffic source with \( U = 0.7 \) and \( CLR = 10^{-6} \)

Figure 3.10 Effective bandwidth of Ethernet traffic source with buffer size 20 KBytes and \( CLR = 10^{-6} \)
In the next step, video traffic consisting of an aggregate of three MPEG
videotraces (Jurassic Park, Silence of the Lambs, and Star Wars) was ex-
amined. Each trace contains 40000 frames (approximately 27 minutes of
video). Their individual and aggregate traffic characteristics are shown in
Table 3.1. The required bandwidth based on ATM switch simulation for ag-
ggregate traffic using 200 and 500 Kbits shared buffer is shown in Fig.3.7. The
aggregate traffic is also examined using Eq.(3.4). The Hurst parameter of the
aggregate traffic is assumed to be the highest H parameter of the individual
videotraces [88]. The effective bandwidth of the aggregate of the videotraces
as calculated by Eq.(3.4), is shown in Fig. 3.15. As may be seen, the effective
bandwidth again offers a close upper bound on the required bandwidth of
real aggregate video traffic sources.

3.4.2 Comparison of Results for Ethernet Traffic

Eq.(3.4) is also examined to estimate the required bandwidth of Ethernet
data traces collected by the Bellcore researchers. The mean rate for this
data trace was estimated to be \( m = 1.3620 \text{Mbps} \), the variance coefficient
\( \sigma = 264410 \) bit-sec, and the Hurst parameter \( H = .8 \) [85]. The results are
shown in Fig.3.16. The required bandwidth of this data trace was already es-
timated using an ATM multiplexer simulation and the results were presented
in Fig.3.8. As may be seen from Figures 3.8 and 3.16, Eq.(3.4) offers an accu-
rate estimate of the required bandwidth of these video traces. However, this
estimate is not very close for the data traces. Indeed, the fractional Brown-
nian approach offers a better approximation when used in a highly utilised
network using large buffer size [45],[88]. The utilisation factor of this data
trace is about 14 percent.
Figure 3.11 Effective bandwidth of a VBR videotrace *Jurassic Park* movie using Eq.(3.4) (Buffer size as a parameter)

Figure 3.12 Effective bandwidth of a VBR videotrace *Silence of the Lambs* using Eq.(3.4) (Buffer size as a parameter)
Figure 3.13 Effective bandwidth of a VBR videotrace *Star Wars* using Eq.(3.4) (Buffer size as a parameter)

Figure 3.14 A comparison between the effective bandwidth and the peak rate for different videotraces
Figure 3.15 Effective bandwidth of the aggregate VBR video traffic using Eq. (3.4) (Buffer size as a parameter).

Figure 3.16 Effective bandwidth of the Bellcore Ethernet data traces using Eq. (3.4)
3.4.3 Experimental Estimation of the Effective Bandwidth

An experimental estimation of the effective bandwidth can be done using computer simulation. The ATM-TN simulator was used to obtain a relation between the desired quality of service and the effective bandwidth [41],[89]. The simulation model is shown in Fig.3.17.

Network Model of a Single Ethernet Source in ATM-TN

As shown in Fig.3.17.a, this network consists of a source, two end-nodes, and a per-port switch. The traffic source is a self-similar Ethernet source. The source is connected to a buffer of size B. The objective is to study the quality of service based on the cell loss ratio. As the source is a LAN source which is insensitive to delay, the delay variation is not considered here.

Network Model of Multiple Ethernet Sources in ATM-TN

The estimation of the effective bandwidth of multiple Ethernet sources has been done using the topology shown in Fig.3.17.b. The network consists of 10 Ethernet sources. The sources are characterised by utilisation factors as
Figure 3.18 The experimental results for the effective bandwidth of a self-similar source with 80 percent utilisation with three H values.

well as Hurst parameters. This network has 11 end nodes. All of the sources are connected to a shared buffer of size B. The capacity of the connected links of the multiplexed sources is 10Mbps. The capacity of the output link was a variable in the experiment.

Experiment Results

The results of the experiments for a single Ethernet source are presented in Figures 3.18, 3.19, and 3.20. In Fig.3.18 the quality of service for a self-similar Ethernet source with utilisation of 80 percent is shown for different values of the Hurst parameter. It may be seen that increasing the self-similarity (higher H) increases the required bandwidth for the same quality of service. This behaviour was also predicted by the analytical model. Fig.3.19 shows the effective bandwidth for different utilisation with the same Hurst parameter $H = .8$. 

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Figure 3.19 The experimental results for the effective bandwidth of a self-similar source with different utilisation for $H=.8$.

Figure 3.20 The effective bandwidth of 10 self-similar Ethernet source with $U=.8$. 

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Fig. 3.20 illustrates the effective bandwidth of multiple sources. In this experiment, utilisation of all sources were considered to be 80 percent. The $H$ parameter for different sources is similar. The buffer size of shared memory is 100 cells and the simulation time is fixed at 10 seconds. Again the effect of $H$ parameter variation was studied in homogeneous sources. Increasing the $H$ parameter of sources increases the required bandwidth, provided that the quality of service is kept constant. The ATM-TN simulator was also used to estimate the effective bandwidth of the self-similar traffic sources [88].

3.5 Modified Effective Bandwidth for Radio ATM

The cell loss resulting from buffer overflow is independent of the cell loss resulting from channel bit errors. The effect of cell loss due to a non-ideal physical layer such as the radio channel is an increase in the required bandwidth. Thus, a modified effective bandwidth function has to be defined. The total CLR ($CLR_T$) is related to the CLR due to buffer overflow ($CLR_o$) and the CLR due to the non-ideal physical channel ($CLR_c$). Thus, the total $CLR_T$ can be estimated as follows:

$$CLR_T = CLR_o + (1 - CLR_o)CLR_c.$$  \hspace{1cm} (3.6)

Using Eq.(3.7), the CLR due to buffer overflow can be expressed as

$$CLR_o = \frac{CLR_T - CLR_c}{1 - CLR_c}.$$ \hspace{1cm} (3.7)

Fig.3.21 shows that the desired $CLR_T$ of the network is equivalent to the $CLR_o$ for the random error channel. However, for a burst error channel, the tolerable $CLR_o$ decreases as a function of the BER and the burst length. Using this relation, the estimation of the effective bandwidth for a non-ideal channel is straightforward. The required modification is to use Eq.(3.6) in Eq.(3.4), where $CLR_T$ is the desired CLR for a guaranteed quality of service.
Figure 3.21 Tolerable CLR due to buffer overflow as function of BER for $QoS = 10^{-5}$ as a function of burst size.

Figure 3.22 Effective bandwidth for random error and bursty channel for utilisation 50 percent and $H = .7$. 
The modified effective bandwidth for an Ethernet source with $U = .5, H = .7, BER = 10^{-5}$, and burst length of two bits is shown in Fig.3.22. It should be pointed out that the burst error effect can be efficiently resolved using an interleaving technique. However, the extra delay due to interleaving should be considered when an end-to-end delay bound is studied.

3.6 Summary

In this chapter it was established that an optimum ATM network requires a mechanism for estimation of the traffic bandwidth. An analytical model as well as computer simulation results for effective bandwidth estimation were presented for various types of traffic. The effective bandwidth procedure was modified to accommodate the bursty behaviour of the radio channel. In the next chapter another important issue, namely, the adaptive modulator for the radio ATM network is addressed.
4. ADAPTIVE M-QAM FOR RADIO ATM IN MULTIPATH FADING CHANNELS

In a broadband wireless network, an optimum transmitter has to take into account both radio channel and broadband traffic characteristics. Newer services are being continuously introduced and the communications network is becoming more dynamic and complex. These services require various bandwidths and QoS. As discussed in the last chapter, the QoS requirements can be described by an effective bandwidth metric and different traffic calls have different effective bandwidth requirements.

On the other hand, in a radio transmission system, multipath fading results in the reception of a multitude of reflected and delayed signal components [90],[91]. Although the performance degradation due to the multipath fading is more severe for the mobile network, it affects the performance of fixed radio links as well. It is well known that multipath fading in mobile communications has a Rayleigh distribution, and its distribution in the line of sight link is Ricean [82],[92]. In a fixed channel, the simplest method to overcome the fading effect is to make the transmitter power large enough, or the bit rate low enough so as to get a satisfactory error probability during a specific fraction of the time. Although this is simple, these methods result in low efficiency [90].

For an optimum broadband radio system, a more sophisticated design is clearly required. Overall, the system should provide an adaptive capacity to maximise the available spectrum usage while maintaining the radio link
performance. Further, this system must be implemented in a cost effective manner.

In this chapter, AWGN channel, and Ricean and Rayleigh fading channels are examined. The results outlined in this chapter are applicable to severe fading scenarios, e.g., a short haul link in a crowded city (LMDS applications) or a land mobile radio. It is shown that a dynamic bandwidth demand system can be realized by using an adaptive transmitter. Finally, a multilevel QAM modulator is introduced to implement the adaptive transmitter.

4.1 Requirements for an Adaptive Transmitter

From the previous discussion it is clear that a broadband wireless ATM system that is designed to carry multimedia traffic has to adapt to the QoS requirements as well as to the radio channel variations [69],[93]. The following subsections address the capacity variation due to QoS and radio channel performance.

4.1.1 Capacity Variation Due to QoS

The number of users in an ATM network is time varying and these users also have various QoS requirements which may be specified by the cell loss ratio and delay constraints. The effective bandwidth as a metric of QoS was derived in the last chapter. Eq.(3.4) is rewritten below:

\[ \text{BW}_{\text{eff}} = R_{\text{mean}} + (K(H)\sqrt{-2\log(p_{\text{CLR}})})^{1/H} \sigma^{1/(2H)} \Pi^{-(1-H)/H} R_{\text{mean}}^{1/(2H)}, \]

where

\[ K(H) = H^H (1 - H)^{1-H}, \]

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and $\sigma$ is the variance coefficient. The above equation has to be further modified to take into account the delay limitations of ATM sources. If the maximum tolerable delay is $D_{\text{max}}$, the relation between maximum delay and buffer size may be written as:

$$\frac{\Pi}{BW_{\text{eff}}} \leq D_{\text{max}}. \quad (4.3)$$

Using (4.3), Eq. (4.1) can be modified as:

$$BW_{\text{eff}} = R_{\text{mean}} + (K(H)\sqrt{-2\log(p_{\text{CLR}})})^{1/H} \sigma^{1/(2H)}(BW_{\text{eff}}D_{\text{max}})^{-(1-H)/H} R_{\text{mean}}^{1/(2H)}. \quad (4.4)$$

This equation can be rearranged to collect the effective bandwidth terms at the left hand side as:

$$BW_{\text{eff}}^{1/H}(1 - R_{\text{mean}}/BW_{\text{eff}}) = (K(H)\sqrt{-2\log(p_{\text{CLR}})})^{1/H} \sigma^{1/(2H)}(D_{\text{max}})^{-(1-H)/H} R_{\text{mean}}^{1/(2H)}. \quad (4.5)$$

As may be seen, this is a nonlinear equation that relates the effective bandwidth to the required cell loss ratio, $p_{\text{CLR}}$, and maximum delay, $D_{\text{max}}$. This equation has to be solved numerically to obtain the effective bandwidth value for a given cell loss ratio and maximum delay.

As an example, for an ATM traffic source with the $R_{\text{mean}} = 2$ Mbps, $\sigma = 10^4$ bit-sec, and $H = .7$, the effective bandwidth as a function of $p_{\text{CLR}}$ and $D_{\text{max}}$ was computed using Eq.(4.5) and results are presented in Fig.4.1. As may be seen, the required effective bandwidth is a function of cell loss requirement and delay constraint. This suggests that if optimum transmission of multimedia traffic is desirable, a flexible transmitter design is essential.
4.1.2 Capacity Reduction Due to Channel Fading

A model has been introduced for an ATM transmission system in Chapter 2. As already discussed in Section (2.11), the effective bandwidth of a high speed link, $C$, has the following relation with the effective bandwidth of individual VPs.

$$\sum_{i=1}^{N} BW_{effi} \leq C, \quad (4.6)$$

where $N$ is the number of VPs and $BW_{effi}$ is the effective bandwidth of the $ith$ VP.

The available capacity in a radio ATM link is a function of channel performance as well. To estimate the capacity reduction due to fading, the Shannon capacity is examined first. The well known Shannon limit for an AWGN channel is given by [80],[94]:

$$C_w = W\log_2(1 + \gamma), \quad (4.7)$$
where $W$ is channel bandwidth and $\gamma$ is signal to noise ratio. For a fading channel, the Shannon limit can be modified as \[95\]

$$C_f = \int_{\gamma} W \log_2(1 + \gamma)p(\gamma)d\gamma,$$  
(4.8)

where $p(\gamma)$ is the probability density function of the received CNR. Thus, the Shannon limit for a fading channel decreases by a factor:

$$\xi^c = \frac{C_f}{C_w} = \frac{\int_{\gamma} \log(1 + \gamma)p(\gamma)d\gamma}{\log_2(1 + \gamma)},$$  
(4.9)

where $\xi^c$ is called the capacity reduction factor.

For LOS radios the fading distribution is Ricean. For a Rice channel, the probability density function of the received signal to noise ratio is given by \[81\]:

$$p(\gamma) = \frac{1 + K}{\Gamma} \exp(-K - \frac{\gamma}{\Gamma}(1 + K)) I_o(2\sqrt{\frac{\gamma}{\Gamma}(K^2 + K)}),$$  
(4.10)

where $\Gamma$ is the average power of $\gamma$ and $K$ is the Rice parameter ($K \rightarrow 0$ for a weak direct wave and $K \rightarrow \infty$ for a strong direct wave \[96\]). $I_o$ is the zero order modified Bessel function.

For Rayleigh fading, the probability density function of the received CNR, $p(\gamma)$, is given by:

$$p(\gamma) = \frac{1}{\Gamma} e^{-\frac{\gamma}{\Gamma}},$$  
(4.11)

where $\Gamma$ is the average power of $\gamma$. The capacity reduction factor due to Rice and Rayleigh fading channel is computed and the results are presented in Fig.4.2. An approximate analytical model for a Rayleigh channel has been used in \[70\],\[95\]. The capacity reduction factor using this approximation is also included to show a comparison with exact results.

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Figure 4.2  Shannon capacity reduction factor due to channel fading for Ricean ($K = 5$) and Rayleigh channels

For a fading channel, the available capacity in an ATM link using Eq.(4.6) and Eq.(4.9) can be modified as:

$$\sum_{i=1}^{N} BW_{effi} \leq \xi^c C_w.$$  \hspace{2cm} (4.12)

A dynamic bandwidth demand could be accommodated by a variable capacity transmitter as long as Eq.(4.12) is satisfied.

4.2 Adaptive M-QAM Modulator

According to Eq.(4.12), a variable capacity transmitter is required to realize an optimum radio ATM transmitter. An M-QAM modulator can provide high spectral efficiency and is suitable for implementation as an adaptive transmitter for radio ATM. The following subsections describe the performance of M-QAM as an adaptive modulator in the wireless channel.
4.2.1 M-QAM in AWGN Channel

According to Eq.(2.26), the spectral efficiency of the M-QAM modulator using a raised cosine filter with roll-off factor $\alpha$ is given by:

$$\eta = \frac{C}{W} = \frac{\log_2 M}{1 + \alpha},$$  \hspace{2cm} (4.13)

where $M$ is the modulation level, $C$ is the channel capacity, and $W$ is the channel bandwidth. Thus, the number of possible VP connections in an AWGN channel for an M-QAM modulator is given by:

$$\sum_{i=1}^{N} BW_{eff_i} \leq \frac{W \log_2 M}{1 + \alpha}.$$  \hspace{2cm} (4.14)

This shows that the modulation level can be selected using the following criterion:

$$\log_2 M \geq \frac{(1 + \alpha) \sum_{i=1}^{N} BW_{eff_i}}{W}.$$  \hspace{2cm} (4.15)

As a typical example for calculation of the constellation size, M-QAM is studied in an AWGN channel using a raised cosine filter with a roll-off factor $\alpha = .5$ with a 40 MHz channel bandwidth for a number of similar video-traces, namely, *Jurassic Park* with $BW_{eff} = 824.23$ Kbps, and *Silence of the Lambs* with $BW_{eff} = 626.96$ Kbps. A buffer size of 200 KBits, and cell loss ratio $10^{-6}$, are used. The results of this calculation are shown in Fig.4.3.

4.2.2 M-QAM in Multipath Fading Channel

In the previous section, a capacity reduction factor was derived for the Shannon limit. This factor shows that the capacity is reduced in the presence of fading. Using a bit error rate upper bound for the M-QAM modulation, the capacity reduction factor can also be computed for M-QAM modulator. A bit error rate upper bound for the M-QAM modulator has been derived in
The constellation size based on the number of identical videotapes [74]. The bit error rate for an AWGN channel for the M-QAM modulation can be bounded by:

\[ p_b \approx 2\exp\left(-\frac{3\gamma}{2(M - 1)}\right). \]

This upper bound has been compared with the exact results in Fig.4.4 and as may be seen from this figure, the bound is quite tight. From Eq.(4.16), the modulation level is obtained as follows:

\[ M \approx 1 - \frac{3\gamma}{2\log(5p_b)}. \]

By substitution for M in Eq.(4.13) the capacity reduction factor in the multipath channel for M-QAM modulator is given by:

\[ \xi_{MQAM} = \frac{f_\gamma \log_2(1 - 3\log(5p_b)^{-1})p(\gamma)d\gamma}{\log_2(1 - 3\log(5p_b)^{-1})}. \]
A closed form relation for Rayleigh fading is derived in Appendix A and is given by:

$$\xi_{MQAM} = \frac{-\exp\left(-\frac{2\ln(5p_b)}{3F}\right)(E + \ln\left(\frac{-2\ln(5p_b)}{3F} + \frac{2\ln(5p_b)}{3F}\right))}{\ln(1 - 3\gamma[2\ln(5p_b)]^{-1})},$$  \hspace{1cm} (4.19)

where $E=.577215$ is the Euler constant. The computed results for the capacity reduction factor for the Rayleigh and Ricean channels are presented in Fig.4.5 for different values of BER. Using this factor for bandwidth allocation, Eq.(4.15) can be modified as:

$$\sum_{i=1}^{N} BW_{effi} \leq \frac{\xi_{MQAM}W\log_2 M}{1 + \alpha}.$$  \hspace{1cm} (4.20)

This leads to the following criterion for the constellation size selection:

$$\log_2 M \geq \frac{(1 + \alpha)\sum_{i=1}^{N} BW_{effi}}{W\xi_{MQAM}}.$$  \hspace{1cm} (4.21)
Figure 4.5  Capacity decrease due to fading in MQAM (K=5 in Rice channel)

The required M-QAM constellation size in AWGN and Rayleigh fading channels using Jurassic Park movie videotracers for a cell loss ratio 10^{-6} with a raised cosine filter with roll-off factor $\alpha = 0.5$ with 40 MHz channel bandwidth is presented in Fig.4.6. As expected, the possible number of connections for a specific level of $M$ in a fading channel is less than that for an AWGN channel. It may also be seen that accepted connections increase when the average of the received CNR is increased. As shown in Fig.4.5, the capacity reduction due to multipath fading is a function of the required BER. The required constellation size for similar videotrace connections, for different values of BER with a CNR=30 dB and Rayleigh fading is presented in Fig.4.7.
4.2.3 M-QAM for ATM Cell Transmission in Multi-path Fading

Using Eq.(2.13), an approximate relation between BER and CLR for a random channel is given by:

\[ p_b^2 \approx \frac{p_{CLR}^r}{780} \]  \hspace{1cm} (4.22)

where \( p_{CLR}^r \) is the cell loss ratio due to random error channel.

Using (4.22) in (4.18), the capacity reduction factor for a random error channel is obtained as:

\[ \xi_{MQAM}^{cr} = \frac{f_\gamma \log_2(1 - 3\gamma[\log(p_{CLR}^r/31.2)]^{-1})p(\gamma)d\gamma}{\log_2(1 - 3\gamma[\log(p_{CLR}^r/31.2)]^{-1})}, \]  \hspace{1cm} (4.23)

where \( \xi_{MQAM}^{cr} \) is the capacity reduction factor for the M-QAM in a random channel. A similar relation can be derived for a burst error channel. Using Eq.(2.16), the bit error rate and cell loss ratio relation in a burst channel is
Figure 4.7 The constellation size vs. number of connections of Jurassic Park movie for CNR=30 dB with a CLR = 10^{-6} ratio for a Rayleigh fading channel given by:

$$ p_b \simeq \frac{\beta p_{CLR}^b}{40}, \quad (4.24) $$

where $\beta$ is the burst length and $p_{CLR}^b$ is the cell loss ratio due to burst channel. Substitution of Eq.(4.24) in Eq.(4.23) gives the capacity reduction factor for the burst channel as:

$$ \xi_{M_{QMAM}}^{cb} = \frac{\int_{0}^{\gamma} \log_2(1 - 3\gamma [\log(\beta p_{CLR}^b/40)]^{-1}) p(\gamma) d\gamma}{\log_2(1 - 3\gamma [\log(\beta p_{CLR}^b/40)]^{-1})}, \quad (4.25) $$

where $\xi_{M_{QMAM}}^{cb}$ is the capacity reduction factor for the M-QAM in a burst channel. The capacity reduction factor for a given cell loss ratio in the random and burst error channels are presented in Fig.4.8 and 4.9 for Ricean and Rayleigh channels.
Figure 4.8 Capacity reduction factor in a random error channel due to fading ($K=5$ for Ricean Channel)

Figure 4.9 Capacity reduction factor in a burst error channel due to fading
4.3 Adaptive M-QAM Modulator for Radio ATM Applications

As shown in previous sections, the assigned bandwidth is a function of the cell loss ratio. As explained in Section 2.4, the cell loss ratio may result from cell blocking or radio channel impairment. In the following section, a multimedia radio modulator is designed according to these characteristics.

4.3.1 Modified Effective Bandwidth for M-QAM

A general relation between the transmission capacity and the number of the permitted connections may be written as:

\[
\frac{1}{\xi_{MQAM}(P_{CLR}^C)} \sum_{i=1}^{N} BW_{effi}(p_{CLR}^B) \leq C,
\]

(4.26)

where \( p_{CLR}^C \) is the cell loss ratio due to the channel and \( p_{CLR}^B \) is the cell loss ratio due to cell blocking. A new metric is needed for use in bandwidth allocation in radio ATM. This new metric, termed as the modified effective bandwidth, has been originally defined as:

\[
\sum_{i=1}^{N} BW_{meffi}(p_{CLR}^B, p_{CLR}^C) \leq C,
\]

(4.27)

where

\[
BW_{meffi}(p_{CLR}^B, p_{CLR}^C) = \frac{1}{\xi_{MQAM}(P_{CLR}^C)} \sum_{i=1}^{N} BW_{effi}(p_{CLR}^B),
\]

(4.28)

is the modified effective bandwidth of each traffic connection. As an example, the modified effective bandwidth for an Ethernet source with \( R_{mean} = 5 \) Mbps, \( H = .7 \), and \( \sigma = 1 \) Mb-s is illustrated in Fig.4.10. The buffer size is 10 Kbits and channel bit error rate is \( 10^{-6} \). As may be seen, the highest bandwidth is required for a Rayleigh channel. The required bandwidth in a Ricean channel is a function of Rice parameter, \( K \), and lies between the
Figure 4.10 The modified effective bandwidth for different channels

level selection for the M-QAM can be obtained using Eq. (4.14):

\[ \log_2 M \geq \frac{(1 + \alpha) \sum_{i=1}^{N} BW_{\text{meff}_i}(p_{\text{CLR}}^B, p_{\text{CLR}}^C)}{W}, \]

where \( BW_{\text{meff}_i} \) is the modified effective bandwidth of \( VP_i \).

4.3.2 Adaptive M-QAM Architecture

Based on the above theoretical results, an architecture for a radio ATM system using an M-QAM modulator is proposed in Fig.4.11. The function of the adaptive M-QAM control unit is to select an optimum modulation level. The adaptation is based on an ATM call admission task and wireless channel condition. The demodulator can be informed about the level of modulation by different techniques [97],[98]. The control unit for an adaptive M-QAM control unit is presented in Fig.4.11. The number of virtual paths is \( N \) and \( n_i \) connections are demanding to use \( VP_i, i = 1, ..., N \) at time \( t_j \). The operation
of the control unit is described as follows:

- Different connections, to be transmitted using $VP_i, i = 1, \ldots, N$, introduce their traffic and QoS parameters (e.g., tolerable cell loss ratio $p_{CLR}$ and maximum delay $D_{max}$) to the connection gate control block.

- The effective bandwidth of $VPs$ are evaluated by using the traffic parameters of the new connection and information from the buffer size manager.

- A buffer size is selected for each $VP$ according to minimum tolerable delay. Eq.(4.5) is solved to find the effective bandwidth. The buffer size is selected according to the delay constraints of each $VP$ using Eq.(4.3).

- This is followed by a modified effective bandwidth estimation of the individual $VP$s. To do this, the capacity reduction factor, $\xi_{MQAM}$, is computed according to Eq.(4.19).

- The total value of the modified effective bandwidth is computed according to the new values of the modified effective bandwidth of individual $VPs$, respectively.

- The modulator capacity is checked with the instantaneous bandwidth demand, and the modulation level is controlled according to the instantaneous bandwidth demand.
Figure 4.12 MQAM modulator for radio ATM using one VP

- The outage estimator examines the outage percentage according to the new modulation level. The outage estimation is to be computed using a link design program which is initialised according to the specific radio link.

- The $k(M = 2^k)$ is increased so long as the desired system outage is achieved.

- A positive acknowledgement from the outage estimator, implies that the gate control will permit the new connections to use the $VP_i$ and radio link.

As may be seen, the M-QAM control unit is designed to operate in harmony with a non-blocking ATM switch. The performance of this unit has been studied in Chapter 7.
Figure 4.13 The M-QAM control unit
4.4 Summary

To optimise the ATM radio transmitter performance, an adaptive M-QAM modulator has been proposed and analysed in different wireless channels. An M-QAM control unit was introduced to select an optimum modulation level. An implementation method for such a unit was presented. The hardware architecture for a direct microwave M-QAM modulator is described in the next chapter.
5. DIRECT MICROWAVE M-QAM MODULATOR HARDWARE

An adaptive transmitter for radio ATM applications using M-QAM was introduced in the last chapter. The proposed modulator could be used for a system with high bit rate multimedia traffic. The cost efficiencies of the modulator is another important design consideration. The conventional method of QAM modulation is not efficient in power, size, and cost. In addition to the IF sections, such a method also requires a transmitter power amplifier which may include a linearizer circuit. This could have an impact on the cost of the radio. Direct microwave modulation at transmission frequency offers an inexpensive, efficient solution which leads to a compact circuit that is suitable for MMIC implementation.

In this chapter a novel structure is proposed for direct microwave QAM modulator implementation without a power amplifier. The proposed design uses high speed attenuators implemented with PIN diodes. A new analytical model to analyse a forward biased PIN diode is presented in Section 5.2. This is followed by a detailed examination of the direct M-QAM modulator performance.

5.1 Modulator Architecture

A new architecture of the direct microwave QAM modulator is presented in Fig.5.1. As shown, the power oscillator output is divided equally between the inphase and the quadrature paths using a quadrature hybrid. A 180 degree hybrid in each path is used to modulate the oscillator signal with
Figure 5.1 Direct Microwave QAM

the sign of the baseband signals. The positive sign in the baseband signal selects the inphase output of the 180 degree hybrid and the negative sign in the baseband signal selects the out-of-phase output of the 180 degree hybrid. The rectified baseband signal in the inphase path is used to adjust the level of an attenuator in the inphase path. Similarly, the rectified baseband signal in the quadrature path is used to adjust the attenuator level in the quadrature path. Pulse-shaping is realized using a raised cosine filter in each path. Using this modulator, the filtering can be done on the baseband signal before applying the baseband signals to the variable attenuators. Moreover, due to the nonlinear relation between the control voltage and the attenuation for the PIN diode attenuators, baseband pre-distortion becomes essential. The pre-distortion and pre-filtering can be implemented by using the same memory circuit [10]. An inphase power combiner is used to combine the inphase and quadrature signals.
5.2 Theoretical Model for Direct QAM Modulator

The main block in the proposed direct modulator is a high speed variable attenuator. A common method to realize a variable attenuator is to use PIN diodes. The PIN diode has traditionally been used for high power microwave switching applications as well as a variable attenuator. However, to use a PIN diode circuit as a high speed modulator certain design issues have to be resolved. The main problem is the hysteresis effect when a PIN diode is used in the ON-OFF mode [66]. In this mode of operation, when the diode condition changes from forward bias to reverse bias, a minimum time is required to clear out the stored charge accumulated during the forward bias. This time is a function of the carrier life time as well as the ratio of the forward to reverse bias current [99].

This speed limitation can be overcome by operating the PIN diode in the forward bias mode only. Hence, an accurate PIN diode model is required to characterise high speed forward biased operation at microwave and millimeter-wave frequencies. While the PIN diode has been extensively studied, an accurate model that describes the PIN diode operation in response to high speed data is not available. In this section, the PIN diode operation in response to high speed multilevel data is studied. The residue theorem and complex inversion formula are used to obtain the electron density function. A device model is derived from this density function. The method results in a closed form formula for the PIN diode resistor under forward bias.

5.2.1 Analysis

The fundamental equations for PN junction analysis are the current density and the continuity equations [100]. The current density equations consist
of a drift component and a diffusion component and are given by:

\[ \vec{J}_p = q(\mu_p n \vec{E} - D_p \nabla p), \]
\[ \vec{J}_n = q(\mu_n n \vec{E} + D_n \nabla n), \]

(5.1)

(5.2)

where \( J_p \) and \( J_n \) are hole and electron current density, \( \mu_p \) and \( \mu_n \) are the hole and electron mobility factors, and \( D_p \) and \( D_n \) are diffusion constants of holes and electrons respectively. The well known continuity equations are:

\[ -\nabla \cdot \vec{J}_p = q\left( \frac{\partial p}{\partial t} + \frac{p - p_n}{\tau_p} - g_p \right), \]
\[ \nabla \cdot \vec{J}_n = q\left( \frac{\partial n}{\partial t} + \frac{n - n_p}{\tau_n} - g_n \right). \]

(5.3)

(5.4)

where \( p \) and \( n \) are hole and electron minority carrier densities, \( q \) is electron charge, \( p_n \) and \( n_p \) are equilibrium hole and electron densities, \( \tau_p \) and \( \tau_n \) are hole and electron carrier life times, and \( g_p \) and \( g_n \) are hole and electron generation factors. The other important equation is Gauss’ law:

\[ \nabla \cdot (\epsilon \vec{E}) = \rho, \]

(5.5)

\[ \rho = q(p - n + N_d - N_a). \]

(5.6)

where \( \rho \) is the net charge density, \( \epsilon \) is permittivity and \( N_d \) and \( N_a \) are donor and acceptor densities. The generation term is usually very small and it can be ignored. Moreover, the assumption of heavily injected holes and electrons is usually valid (\( n \gg n_n \) and \( p \gg p_n \)). Using these assumptions, the continuity equations may be simplified to:

\[ -\nabla \cdot \vec{J}_p = q\left( \frac{\partial p}{\partial t} + \frac{p}{\tau_p} \right), \]
\[ \nabla \cdot \vec{J}_n = q\left( \frac{\partial n}{\partial t} + \frac{n}{\tau_n} \right). \]

(5.7)

(5.8)
These equations have to be solved to obtain the electron and hole densities. Assuming that the electric field \( \vec{E} \) has only transverse components, i.e., \( \frac{\partial \vec{E}}{\partial x} = 0 \), and the electron and hole densities are a function of \( x \) and \( t \), the electron and hole densities equations may be written as follows:

\[
\frac{\partial n}{\partial t} = D_n \frac{\partial^2 n}{\partial x^2} - \frac{n}{\tau_n} \tag{5.9}
\]

\[
\frac{\partial p}{\partial t} = D_p \frac{\partial^2 p}{\partial x^2} - \frac{p}{\tau_p} \tag{5.10}
\]

Assuming a symmetrical structure, electron and hole carrier densities may be considered to be the same. It is also possible to approximate \( \tau_p = \tau_n = \tau \) and define an ambipolar diffusion constant \( D = 2D_nD_p/(D_p + D_n) \). Using these assumptions and taking the Laplace transform of (5.9) result in:

\[
\frac{\partial^2 N(x, s)}{\partial x^2} = \left( \frac{1 + s\tau}{\tau D} \right) N(x, s), \tag{5.11}
\]

where \( N(x, s) \) is the Laplace transform of \( n(x, t) \). Eq.(5.11) has the following solution:

\[
N(x, s) = k_1e^{-\sqrt{\frac{1 + s\tau}{\tau D}}x} + k_2e^{\sqrt{\frac{1 + s\tau}{\tau D}}x}. \tag{5.12}
\]

Using this solution, and considering a symmetrical PIN diode as shown in Fig.5.2, we can assume \( N(x, s) = P(x, s) \). For a symmetrical PIN diode the carrier density is minimum at the center of the intrinsic region (\( x=0 \) in Fig.5.2). The RF carrier density is mostly concentrated in the boundary of n and p region [101], so that:

\[
\frac{\partial}{\partial x}[P(x, s) + N(x, s)]|_{x=0} = \frac{\partial}{\partial x}2N(x, s)|_{x=0} = 0. \tag{5.13}
\]

This condition results in \( k_1 = k_2 \) in (5.12); and this equation may be written as:

\[
N(x, s) = 2k_1cosh\left(\sqrt{\frac{1 + s\tau}{\tau D}}x\right). \tag{5.14}
\]
Using (5.14), \( k_1 \) may be solved in terms of the charge density at \( x = \frac{W}{2} \), so that:

\[
k_1 = \frac{N\left(\frac{W}{2}, s\right)}{2 \cosh\left(\sqrt{\frac{1+s\tau}{\tau D}} \frac{W}{2}\right)}.
\]

(5.15)

Using \( k_1 \) in (5.14), the charge density in the Laplace domain is obtained as:

\[
N(x, s) = \frac{N\left(\frac{W}{2}, s\right)}{\cosh\left(\sqrt{\frac{1+s\tau}{\tau D}} \frac{W}{2}\right)} \cosh\left(\sqrt{\frac{1+s\tau}{\tau D}} x\right).
\]

(5.16)

Next, the solution for \( N(x, s) \) in response to the bias current is examined. At \( x = \frac{W}{2} \), the hole current is zero, so that the following condition is valid:

\[
D \frac{\partial P(x, s)}{\partial x}\bigg|_{x=\frac{W}{2}} = \mu_p \bar{E} \cdot P\left(\frac{W}{2}, s\right) = \mu_n \bar{E} \cdot N\left(\frac{W}{2}, s\right).
\]

(5.17)

The ambipolar mobility factor is defined as \( \mu = 2\mu_n \mu_p / (\mu_n + \mu_p) \), so that:

\[
D \frac{\partial P(x, s)}{\partial x}\bigg|_{x=\frac{W}{2}} = \mu E \cdot N\left(\frac{W}{2}, s\right).
\]

(5.18)

When the PIN diode is forward biased with high speed data, three current components have to be considered. The first component is a DC bias which makes diodes operate under the forward bias condition. The second component arises from the incoming data signal that is used to control the level of the attenuator. The last current component is the RF current component at the frequency \( \omega_0 \). Under higher power operation, the RF signal harmonics have to be considered as well. Thus, the total current in the PIN diode may be expressed as:

\[
\frac{I_0}{s} + I_D(s) + I_R(s) = Aq(\mu N \bar{E} + D \frac{\partial N}{\partial x})\bigg|_{x=\frac{W}{2}}.
\]

(5.19)
Using Eq.5.18, Eq.5.19 may be written as:

$$\frac{I_o}{s} + I_D(s) + I_R(s) = Aq(2D \frac{\partial N(x, s)}{\partial x})|_{x = \frac{W}{2}}. \quad (5.20)$$

Substituting from (5.16) in (5.20), results in:

$$\frac{I_o}{s} + I_D(s) + I_R(s) = 2AqD \cdot \frac{N\left(\frac{W}{2}, s\right)\sqrt{\frac{1 + sT}{\tau_D}}}{\cosh\left(\sqrt{\frac{1 + sT}{\tau_D}} \cdot \frac{W}{2}\right)} \sinh\left(\sqrt{\frac{1 + sT}{\tau_D}} \cdot x\right)|_{x = \frac{W}{2}}. \quad (5.21)$$

From this equation, $N\left(\frac{W}{2}, s\right)$ may be written as:

$$N\left(\frac{W}{2}, s\right) = \frac{\cosh\left(\sqrt{\frac{1 + sT}{\tau_D}} \cdot \frac{W}{2}\right)}{2AqD\sqrt{\frac{1 + sT}{\tau_D}} \sinh\left(\sqrt{\frac{1 + sT}{\tau_D}} \cdot \frac{W}{2}\right)} \left[\frac{I_o}{s} + I_D(s) + I_R(s)\right]. \quad (5.22)$$

From Eq.(5.16) and (5.22), the electron density function can be given by:

$$N(x, s) = \frac{\cosh\left(\sqrt{\frac{1 + sT}{\tau_D}} x\right)}{2AqD\sqrt{\frac{1 + sT}{\tau_D}} \sinh\left(\sqrt{\frac{1 + sT}{\tau_D}} \cdot \frac{W}{2}\right)} \left(\frac{I_o}{s} + I_D(s) + I_R(s)\right). \quad (5.23)$$

The charge density, $N(x, s)$, may be separated into three components as
\[ N(x, s) = N_1(x, s) + N_2(x, s) + N_3(x, s), \quad (5.24) \]

where,

\[ N_1(x, s) = \frac{\cosh(\sqrt{\frac{1+\sigma t}{\tau_D}}x)}{2AQD\sqrt{\frac{1+\sigma t}{\tau_D}}\sinh(\sqrt{\frac{1+\sigma t}{\tau_D}}W/2)} \frac{I_o}{s}. \quad (5.25) \]

\[ N_2(x, s) = \frac{\cosh(\sqrt{\frac{1+\sigma t}{\tau_D}}x)}{2AQD\sqrt{\frac{1+\sigma t}{\tau_D}}\sinh(\sqrt{\frac{1+\sigma t}{\tau_D}}W/2)} I_D(s). \quad (5.26) \]

\[ N_3(x, s) = \frac{\cosh(\sqrt{\frac{1+\sigma t}{\tau_D}}x)}{2AQD\sqrt{\frac{1+\sigma t}{\tau_D}}\sinh(\sqrt{\frac{1+\sigma t}{\tau_D}}W/2)} I_R(s). \quad (5.27) \]

The three time domain components of densities are obtained by taking the inverse Laplace transform, i.e.,

\[ n_1(x, t) = L^{-1}(N_1(x, s)), \quad (5.28) \]

\[ n_2(x, t) = L^{-1}(N_2(x, s)), \quad (5.29) \]

\[ n_3(x, t) = L^{-1}(N_3(x, s)). \quad (5.30) \]

In order to obtain the time domain components, the residue theorem and complex inversion formula can be used.

**DC Response**

The DC current is used to ensure that the diode operates in the forward bias condition only. The Laplace transform of electron density resulting from the DC current \( N_1(x, s) \) is given by:

\[ N_1(x, s) = \frac{\cosh(\sqrt{\frac{1+\sigma t}{\tau_D}}x)}{2AQD\sqrt{\frac{1+\sigma t}{\tau_D}}\sinh(\sqrt{\frac{1+\sigma t}{\tau_D}}W/2)} \frac{I_o}{s}. \quad (5.31) \]
This equation may be written as:

\[
N_1(x, s) = \frac{I_o}{2AqD} F_1(x, s),
\]  
(5.32)

where,

\[
F_1(x, s) = \frac{\cosh(\sqrt{\frac{1+s\tau}{\tau D}} x)}{s \sqrt{\frac{1+s\tau}{\tau D}} \sinh(\sqrt{\frac{1+s\tau}{\tau D}} \frac{W}{2})}.
\]  
(5.33)

This function has a pole at the origin \(s = 0\), a second order pole at \(s = -\frac{1}{\tau}\), and infinite poles where the \textit{sinh\_hyperbolic} function is equal to zero, i.e., at

\[
\sinh(\sqrt{\frac{1+s\tau}{\tau D}} \frac{W}{2}) = 0,
\]  
(5.34)

or,

\[
2 \sqrt{\frac{1+s\tau}{\tau D} \frac{W}{2}} = j2k\pi \quad k = \ldots, -2, -1, 1, 2, \ldots
\]  
(5.35)

These poles are given by:

\[
s_k = -\frac{1}{\tau} - D\left(\frac{2k\pi}{W}\right)^2 \quad k = 1, 2, \ldots
\]  
(5.36)

As shown in Fig. 5.3, to obtain the residues, an integration contour with radius \(Z_m\) may be considered where,

\[
Z_m = (m + \frac{1}{2}) \frac{1}{\tau} + D(\frac{2m\pi}{W})^2 \quad m = \text{integer}
\]  
(5.37)

This choice for \(Z_m\) insures that the contour does not pass through any of the poles. Now, the residues for different poles can be obtained.

For \(s = 0\), the residue \(R_o\) is given by:

\[
R_o = \sqrt{\tau D} \frac{\cosh(\frac{x}{\sqrt{\tau D}})}{\sinh(\frac{W}{2\sqrt{\tau D}})}.
\]  
(5.38)
The residue for poles $s = s_k, k = 1, \ldots$, are given by:

$$R_k = \lim_{s \to s_k} (s - s_k)F_1(x, s)e^{st}. \quad (5.39)$$

Now

$$\lim_{s \to s_k} \sqrt{\frac{1 + st}{\tau D}} = j\frac{2k\pi}{W}, \quad (5.40)$$

so that the residue $R_k$ can be expressed as:

$$R_k = \frac{4D\cos\left(\frac{2k\pi}{W}x\right)e^{-\frac{1}{\tau} + D\left(\frac{2k\pi}{W}\right)^2t}}{W\cos(k\pi)(-\frac{1}{\tau} - D\left(\frac{2k\pi}{W}\right)^2)^{k = 1, 2, \ldots, m}} \quad (5.41)$$

The only pole left in the equation is the second order pole at $s = -\frac{1}{\tau}$. Its residue is given by:

$$R' = \lim_{s \to -\frac{1}{\tau}} \frac{d}{ds}\left((s + \frac{1}{\tau})^2F_1(x, s)e^{st}\right). \quad (5.42)$$

$\frac{W}{2\sqrt{\tau D}} \ll 1$ is a good practical assumption [66], which can be used in the above equation so as to give the residue:

$$R' = \frac{2D}{W}\lim_{s \to -\frac{1}{\tau}} \frac{d}{ds}\left[(s + \frac{1}{\tau})\frac{\cosh\left(\frac{1+sx}{\tau D}\right)}{s}e^{st}\right]. \quad (5.43)$$
After taking the limit, a simple result for $R'$ is obtained as:

$$ R' = -\frac{2D\tau}{W} e^{\frac{t}{\tau}}. \quad (5.44) $$

Using the integration contour shown in Fig.5.3,

$$ \frac{1}{2\pi j} \oint_{C_m} e^{st} F_1(x, s) ds = R_o + R' + \sum_{k=1}^{M} R_k. \quad (5.45) $$

Using the residue theorem to estimate the above integral, the electron density function corresponding to the DC current is obtained as:

$$ n_1(x, t) = \frac{I_o}{2AqD} \left( \frac{\sqrt{\tau}D cosh \left( \frac{x}{\sqrt{\tau}D} \right)}{\sinh \left( \frac{W}{2\sqrt{\tau}D} \right)} - \frac{2D\tau}{W} e^{-\frac{t}{\tau}} + \sum_{k=1}^{\infty} \frac{4Dcos \left( \frac{2k\pi}{W}x \right) e^{-\frac{1}{2}D \left( \frac{2k\pi}{W} \right)^2 t}}{W \cos \left( k\pi \right) \left[ -\frac{1}{\tau} - D \left( \frac{2k\pi}{W} \right)^2 \right]} \right). \quad (5.46) $$

As may be seen from (5.46), as time $t$ tends to infinity, the second and third components approach zero and the electron density after a transition time can be obtained from the first term. This is consistent with the well known steady state relation [66], [102]. The second and third terms also show that diode operation in the on-off mode is limited to slow speed operation only.

**Data Response**

The $N_2(x, s)$ component in Eq.5.26 corresponds to the data signal and is defined as:

$$ N_2(x, s) = \frac{1}{2AqD} F_2(x, s), \quad (5.47) $$

where,

$$ F_2(x, s) = \frac{\cosh \left( \sqrt{\frac{1+\tau s}{\tau D}} x \right)}{\sqrt{\frac{1+\tau s}{\tau D} \sinh \left( \sqrt{\frac{1+\tau s}{\tau D}} \frac{W}{2} \right)}} I_D(s). \quad (5.48) $$

A PAM sequence with period $T$ is considered as the data signal. Such a signal is shown in Fig.5.4 and it can be expressed as:

$$ i_D(t) = \sum_{k=0}^{\infty} A_k(u(t - kT) - u(t - (k + 1)T)), \quad (5.49) $$
where $A_k$ is the amplitude of signal. The Laplace transform of this signal is given by [103]:

$$I_D(s) = \sum_{k=0}^{\infty} \frac{2A_k}{s} e^{-kT_s} \sinh\left(\frac{sT}{2}\right).$$  \tag{5.50}

Substituting (5.50) in (5.48), $F_2(x, s)$ may be written as:

$$F_2(x, s) = \frac{\cosh\left(\sqrt{\frac{1+TS}{TD}}x\right)}{\sqrt{\frac{1+TS}{TD}} \sinh\left(\sqrt{\frac{1+TS}{TD}}W/2\right)} \sum_{k=0}^{\infty} \frac{2A_k}{s} e^{-skT} \sinh\left(\frac{sT}{2}\right).$$ \tag{5.51}

Hence, this component of density function is given by:

$$N_2(x, s) = \frac{1}{2AcD} \frac{\cosh\left(\sqrt{\frac{1+TS}{TD}}x\right)}{\sqrt{\frac{1+TS}{TD}} \sinh\left(\sqrt{\frac{1+TS}{TD}}W/2\right)} \sum_{k=0}^{\infty} \frac{2A_k}{s} e^{-skT} \sinh\left(\frac{sT}{2}\right).$$ \tag{5.52}

Using the final value theorem in Eq.(5.52), $\lim_{t \to \infty} n_2(x, t) = \lim_{s \to 0}sN_2(x, s)$, the electron density function because of the data signal is given by:

$$\lim_{t \to \infty} n_2(x, t) = \lim_{s \to 0}s \frac{1}{2AcD} \frac{\cosh\left(\sqrt{\frac{1+TS}{TD}}x\right)}{\sqrt{\frac{1+TS}{TD}} \sinh\left(\sqrt{\frac{1+TS}{TD}}W/2\right)} \sum_{k=0}^{\infty} \frac{2A_k}{s} e^{-skT} \sinh\left(\frac{sT}{2}\right).$$ \tag{5.53}

It may be easily seen that the electron density goes to zero when $t \to \infty$. 
This shows that the forward bias operation highly decreases the limitations on the PIN diode operation speed. These interesting results can be used to design high speed multilevel digital modulators [7], [10], [11] and [12].

**RF Response**

The electron density for the third component of the current, i.e., RF current can also be computed in a similar way. For simplicity an RF signal with only a sinusoidal component is considered, so that:

\[
i_{RF}(t) = B \cos(\omega t), \tag{5.54}\]

The Laplace transform of this signal is:

\[
I_r(s) = \frac{Bs}{s^2 + \omega^2}, \tag{5.55}
\]

where \(B\) is the amplitude of the sinusoidal and \(\omega\) is its frequency. This component of current results in the third component of electron density, \(N_3(x, s)\), and it can be written as:

\[
N_3(x, s) = \frac{\cosh(\sqrt{\frac{1+\tau}{\tau_D}}x)}{2AqD\sqrt{\frac{1+\tau}{\tau_D}} \sinh(\sqrt{\frac{1+\tau}{\tau_D}} \frac{W}{2})} \frac{Bs}{s^2 + \omega^2}, \tag{5.56}
\]

or:

\[
N_3(x, s) = \frac{B}{2AqD} F_3(x, s), \tag{5.57}
\]

where:

\[
F_3(x, s) = \frac{s \cosh(\sqrt{\frac{1+\tau}{\tau_D}}x)}{\sqrt{\frac{1+\tau}{\tau_D}} \sinh(\sqrt{\frac{1+\tau}{\tau_D}} \frac{W}{2})(s^2 + \omega^2)}. \tag{5.58}
\]

This function has a second order pole at \(s = \frac{-1}{\tau}\), two poles at \(s = j\omega\) and \(s = -j\omega\) and an infinite number of poles because of \(\sinh\) function zeros. For
the poles at \( s = j\omega \) and \( s = -j\omega \), the residues are:

\[
L_1 = \lim_{s \to j\omega} [(s - j\omega)F_3(x, s)e^{st}], \tag{5.59}
\]

\[
L_2 = \lim_{s \to -j\omega} [(s + j\omega)F_3(x, s)e^{st}], \tag{5.60}
\]

or:

\[
L_1 = \frac{\cosh(\sqrt{\frac{1+j\omega}{\tau_D}x})e^{j\omega t}}{2\sqrt{\frac{1+j\omega}{\tau_D} \sinh(\sqrt{\frac{1+j\omega}{\tau_D} \frac{W}{2}})}}, \tag{5.61}
\]

\[
L_2 = \frac{\cosh(\sqrt{\frac{1-j\omega}{\tau_D}x})e^{-j\omega t}}{2\sqrt{\frac{1-j\omega}{\tau_D} \sinh(\sqrt{\frac{1-j\omega}{\tau_D} \frac{W}{2}})}}, \tag{5.62}
\]

As before, the residues corresponding to the poles resulting from the \( \sinh \) function zeros can be obtained as:

\[
L_k = \lim_{s \to s_k} [(s - s_k)F_3(x, s)e^{st}]. \tag{5.63}
\]

This equation may be written as:

\[
L_k = \frac{4D(\frac{1}{\tau} - D(\frac{2k\pi}{W})^2)\cos(\frac{2k\pi}{W}x)}{W\cos(k\pi)[(\frac{1}{\tau} - D(\frac{2k\pi}{W})^2)^2 + \omega^2]}e^{-(\frac{1}{\tau} + D(\frac{2k\pi}{W})^2)t}. \tag{5.64}
\]

The other pole of \( F_3(x, s) \) is at \( s = -\frac{1}{\tau} \), and this pole is a second order pole:

\[
L' = \lim_{s \to -\frac{1}{\tau}} \frac{d}{ds} [(s + \frac{1}{\tau})^2F_3(x, s)e^{st}]. \tag{5.65}
\]

Using the approximation:

\[
\sinh(\sqrt{\frac{1+s\tau W}{\tau D}}) \approx \sqrt{\frac{1+s\tau W}{\tau D}}, \tag{5.66}
\]

The residue is obtained as:

\[
L' = -\frac{2D}{W\tau} \frac{e^{\frac{s\tau}{\tau}}}{(\frac{1}{\tau})^2 + \omega^2}. \tag{5.67}
\]
Now, the inverse Laplace transform of $F_3(x, s)$ can be obtained as:

$$\frac{1}{2\pi j} \oint_{C_m} e^{st} F_3(x, s) ds = L_1 + L_2 + L' + \sum L_k. \quad (5.68)$$

As a result, the $n_3(x, t)$ term can be written as follows:

$$n_3(x, t) = \frac{B}{2AqD} \left( \frac{\cosh(\sqrt{\frac{1+j\omega \tau}{\tau D}} x) e^{j\omega t}}{2\sqrt{\frac{1+j\omega \tau}{\tau D}} \sinh(\sqrt{\frac{1+j\omega \tau}{\tau D}} \frac{W}{2})} + \frac{\cosh(\sqrt{\frac{1-j\omega \tau}{\tau D}} x) e^{-j\omega t}}{2\sqrt{\frac{1-j\omega \tau}{\tau D}} \sinh(\sqrt{\frac{1-j\omega \tau}{\tau D}} \frac{W}{2})} \right)$$

$$\frac{2D}{W\tau (\frac{1}{\tau})^2 + \omega^2} + \sum_{k=1}^{\infty} \frac{4D(\frac{1}{\tau} - D(\frac{2k\pi}{W})^2) \cos(\frac{2k\pi}{W} x)}{W \cos(k\pi)(\frac{1}{\tau} - D(\frac{2k\pi}{W})^2 + \omega^2)} e^{-\frac{1}{\tau} + D(\frac{2k\pi}{W})^2 t}.$$ 

As may be seen, this density component includes a harmonic term and two transient terms. The factor $\omega$ in the denominator for the transient terms shows that at the higher frequency operation these two transient terms disappear rapidly.

**PIN Diode Voltage**

The total electron density, $n(x, t)$, consists of three components and is given by:

$$n(x, t) = n_1(x, t) + n_2(x, t) + n_3(x, t). \quad (5.69)$$

The conductivity equation gives:

$$\sigma(x, t) = q(\mu_n n(x, t) + \mu_p p(x, t)). \quad (5.70)$$
Assuming \( p(x, t) = n(x, t) \), the conductivity can be written as:

\[
\sigma(x, t) = 2q\mu n(x, t).
\]  \( \tag{5.71} \)

Using Ohm's law, the resistance of a cylindrical conductor of electrical conductivity \( \sigma \), length \( W \) along the current path, and cross section \( A \) is:

\[
R = \frac{W}{\sigma A}.
\]  \( \tag{5.72} \)

Using this relation, the PIN diode voltage may be written as:

\[
v(t) = \frac{i(t)}{A\mu q} \int_0^W \frac{dx}{n(x, t)}.
\]  \( \tag{5.73} \)

### 5.2.2 Numerical Simulation

**Forward Biased Resistance**

As a typical example, the model was applied to study the response of a cylindrical PIN diode with radius \( r = 0.78 \) mm, \( \mu = 0.061 m^2/V \cdot s, D = 15.6 \times 10^{-4} cm^2/s, \tau = 5 \mu sec, W = 28 \mu m \). As a first step, the PIN diode is considered to operate in its linear region without any bias variation (only a constant DC bias), and it is assumed that no RF signal is applied to the diode. The DC current values are assumed as \( 0.1 mA \) and \( 1 mA \). The results obtained from the above model were compared with the previously reported results [66],[104]. As may be seen in Figures 5.5 and 5.6, the final value of the PIN diode resistance approaches the well known results [66], [104]. However, the resistance value during the transition is much higher than its final value. It is an important factor that limits the PIN diode speed as a data modulator for on-off operation modes. For instance, for the diode under study, the maximum speed of data modulation in on-off mode is less than 100 kbps. The numerical results are also compared with the measure-
Figure 5.5 The PIN diode forward bias resistance with I = .1 mA

ment result of the commercially available PIN diodes, MA4P4000 from M/A Com and SMP-1304 from Alpha. As may be seen from Fig.5.7, there is a good agreement between M/A Com measurement [105] and the results of the above model. The comparison between this model and the results in [106] for Alpha's diodes is shown in Fig.5.8. Again, a good agreement was obtained.

PIN Diode as a High Speed PAM Modulator

The model developed above was used to study a basic multilevel PAM modulator. This circuit is shown in Fig.5.9. An RF signal with peak value of 1 volt and frequency of 2.5 GHz is applied to the PIN diode ($V_{in} = \cos(2\pi \times 2.5 \times 10^9 t)$). The data rate was selected as 4 Msym/sec and data level symbols correspond to the drive currents of .1, .3, .5, and 1 mA. To avoid the charge storage problem the diode was always operated in the forward bias mode. In Fig.5.10, the drive current for the PIN diode is shown. To avoid the transition time region, the diode operation was examined after .04 msec. The PIN diode resistance because of data drive is shown in Fig.5.11. The PIN diode voltage is obtained using Eq.(5.73) and is shown in Fig. 5.12. As
Figure 5.6 The PIN diode forward bias resistance with I=1 mA

may be seen, the net DC level of current controls the resistance of the PIN diode in the forward bias condition, and the level variation does not impose any limitation on modulator performance.

5.2.3 Direct Microwave QAM Modulator

The above theoretical results were applied to analyze a direct QAM modulator structure introduced in Fig.5.1. For practical implementation, the variable modulator section and sign modulator section can be replaced by a reflection type PIN diode attenuator and an SPDT switch, respectively. Such a direct microwave modulator is shown in Fig.5.13.

High Speed Reflection PAM Modulator Using PIN Diode

The key block in the direct microwave modulator structure is a reflection multilevel PAM modulator. The circuit block consists of two PIN diodes driven by high speed data and a hybrid as shown in Fig.5.14. A multilevel baseband PAM signal is used as the drive signal for the PIN diode while an
Figure 5.7  Comparison between the model and M/A-Com measurement for MA4P4000 at 100 MHz and 100 mA RF current

Figure 5.8  Comparison between the model and measurement for SMP 1304 from Alpha at 100 mA
Figure 5.9  Basic structure used in the study of the PIN diode modulator

Figure 5.10  Driver current
Figure 5.11 PIN diode forward biased resistance as a function of time

Figure 5.12 PIN diode voltage in the basic structure as a function of time
Figure 5.13 An architecture for direct microwave QAM using the variable attenuator in I and Q paths
RF signal pumps the RF power in the modulator. If the PIN diode resistance is denoted by \( R_D \), the output signal of the attenuator is given by [102]:

\[
S_{21} = 2S_{21}^c \rho_T(t) S_{41}^c,
\]

where

\[
\rho_T(t) = \frac{R_D(t) - Z_o}{R_D(t) + Z_o},
\]

and \( S_{21}^c \) is the coupled voltage transmission coefficient of the coupler, \( S_{41}^c \) is the direct voltage transmission coefficient of the coupler, and \( Z_o \) is the characteristic impedance. By using an ideal quadrature coupler, Eq.(5.75) describes the reflection modulator response. Again, in this experiment, the data rate was selected to be 4 Msym/sec and pump frequency was selected to be 2.5 GHz with a peak amplitude 1 volt. A four level data signal corresponding to the drive currents of .1, .3, .5, and 1 mA was applied to the diode. The response is shown in Fig.5.14.

In the next experiment an eight level PAM signal was applied to the modulator. The drive current and PIN diode resistance are shown in Figures 5.15.a and 5.15.b. The modulator output for a lower carrier frequency is also shown in Figure 5.15.c. As may be seen, a high speed modulator could be implemented at 40 Msym/sec. Higher speed is also feasible. The use of predistortion for further improvement of the modulator performance is described in the following chapters.

This experiment was repeated by using prefiltered baseband data. The filter was a raised cosine filter with roll-off factor \( \alpha = .5 \). The drive current, the PIN diode resistance and the modulator output are shown in Fig.5.16. As may be seen, a high speed modulator was realized. However, predistortion must be implemented to linearise the output.
Figure 5.14 a) A reflection multilevel PIN diode PAM modulator b) Modulator response for the drive currents of 0.1, 0.3, 0.5 and 1 mA.
Figure 5.15 Reflection Modulator without prefiltering and predistortion

- a) Input current
- b) PIN diode resistance
- c) Modulator output
Figure 5.16 Reflection Modulator using presfiltering ($\alpha = .5$) and without predistortion a) Input current
b) PIN diode resistance c) Modulator output
Direct Microwave QAM Modulator

An architecture to realize a direct microwave QAM modulator was presented in Fig.5.13. The analog signal processing section prior to the reflection modulator realizes a BPSK modulator in each path where the symbol rate is $1/T$ sym/sec. Then, the input signal to the PIN diode modulator can be written as

$$S_f(t) = \sum_{k=-\infty}^{\infty} a_k u(t - kT) \cos(2\pi f_c t),$$  \hspace{1cm} (5.76)

where $a_k = 1$ when the symbol has positive polarity, and $a_k = -1$ when the symbol has negative polarity; and $u(t) = 1$ in interval $-T/2 \leq t \leq T/2$ and zero elsewhere. A similar expression can be written for the Q path. Therefore, the direct microwave modulator can be modelled using Fig.5.17. The model is used to realize a 64QAM direct microwave modulator. The input data rate was selected to be equal to OC-3 (155.52 Mbps). While an eight level PAM signal with symbol rate of 25.92 Msymbol/sec was used in each path, a four level rectified PAM signal was applied to each attenuator. Figures 5.18.a and 5.18.b show the bias currents for the I and Q attenuators. As may be seen, these signals have been filtered using a raised cosine filter with roll-off factor $\alpha = .5$. The PIN diode forward bias resistances in response to these biases are shown in Figures 5.18.c. and 5.18.d. The PIN diode voltage responses to bias voltages are shown in Figures 5.19.a and 5.19.b, and the outputs of the modulator in I and Q paths are presented in Figures 5.19.c and 5.19.d. The time and frequency domain responses of the 64QAM modulator are shown in Fig.5.20.a and Fig.5.20.b. According to Fig.5.20.b, the band-
Figure 5.18  a) Bias in I path b) Bias in Q path c) PIN diode forward bias resistance in I path d) PIN diode forward bias resistance in Q path
Figure 5.19  a) PIN diode voltage in I path  b) PIN diode voltage in Q path  c) PIN modulator output in I path  d) PIN modulator output in Q path

Figure 5.20  a) Time domain response of 64QAM  b) Power spectrum of 64QAM modulator
width of the proposed direct microwave 64QAM modulator for transmission of OC-3 signal is about 40 MHz.

5.3 Distortion in the PIN Diode Modulator

The PIN diode electron density in the steady state case was obtained as:

\[
n(x, t) = \frac{I_o}{2AqD} \left( \frac{\sqrt{\tau D} \cosh\left(\frac{x}{\sqrt{\tau D}}\right)}{\sinh\left(\frac{W}{2\sqrt{\tau D}}\right)} + \sum_{i=1}^{M} \frac{B_i}{2AqD} \left( \frac{\cosh\left(\frac{1+j\omega t}{\tau D} x\right)e^{j\omega t}}{2\sqrt{1+j\omega t} \sinh\left(\frac{W}{2\sqrt{1+j\omega t}}\right)} + \frac{\cosh\left(\frac{1-j\omega t}{\tau D} x\right)e^{-j\omega t}}{2\sqrt{1-j\omega t} \sinh\left(\frac{W}{2\sqrt{1-j\omega t}}\right)} \right) \right),
\]

where \( B_i \) is the amplitude of the sinusoidal signal. The PIN diode voltage is obtained using Eq.(5.73), where \( i(t) \) is as follows:

\[
i(t) = I_o + \sum_{i=1}^{M} B_i \cos \omega_i t. \quad (5.77)
\]

The Harmonic distortion and power spectrum for the PIN diode, basic attenuator, and reflection attenuator using an input with peak value of 1 volt and frequency of 2.5GHz with drive current .1 are compared and shown in Figures 5.21. As may be seen, the level of harmonic distortion is always better than 50 dBc. Also, by increasing the drive current, the harmonic distortion decreases. The measurement results for nonlinear PIN diode operation has already been reported [107]. A good agreement has been found between these results and measurement results. The second and third inter-modulation distortions for PIN diode, basic attenuator, and reflection attenuators are also shown in Figures 5.22. As may be seen with the higher drive current, the level of inter-modulation distortion decreases. Moreover, the level of inter-
Figure 5.21 Harmonic distortion of PIN diode at f=2.5 GHz

a) Harmonic distortion in PIN diode at $I_o = .1$ mA
b) Harmonic distortion in basic modulator at $I_o = .1$ mA
c) Harmonic distortion for the reflection modulator at $I_o = .1$ mA
d) Harmonic distortion for the PIN diode vs. current

e) Harmonic distortion for the basic modulator vs. current

f) Harmonic distortion for the reflection modulator vs. current
Figure 5.22 Inter-modulation distortion of PIN diode (frequencies are $f=2.25$ and $2.75$ GHz) a) IMD in PIN diode at $I_o = .1$ mA b) IMD in basic modulator at $I_o = .1$ mA c) IMD in PIN diode at $I_o = .3$ mA d) IMD in basic modulator vs. current e) IMD in PIN diode vs. current f) IMD in reflection modulator vs. current
modulation distortion is always better than 50 $dB_c$.

5.4 Summary

In this chapter, an accurate model to characterise a PIN diode as a variable resistor driven by high speed digital data and RF pump signals was presented. The model extracts the electron density function resulting from these signal drives. This is used to obtain a closed form expression for the PIN diode resistance. The residue theorem and complex inversion formula are used to solve the PN junction equations. A high speed reflection pulse amplitude modulator which is an essential block for implementing a more complex digital modulator is also examined using this model. Moreover, a new architecture for an M-QAM modulator using PIN diode attenuators was presented. A detailed discussion of the modulator hardware implementation is presented in the next chapter.
6. HARDWARE IMPLEMENTATION OF DIRECT MICROWAVE QAM MODULATOR

An architecture for a direct microwave QAM modulator was proposed and analysed in the last chapter. In this chapter a hardware implementation of such a modulator is addressed. The main objective is to design a direct microwave QAM modulator. The modulator implementation using microwave integrated circuits (MIC) and silicon monolithic microwave integrated circuits (Si-MMIC) techniques is described. The MIC realization is simple and straightforward. The carrier frequency used to implement the MIC version was selected to be 2.5 GHz. This implementation uses very low cost PIN diodes, making this version an attractive solution for this band of frequency. Although silicon monolithic microwave integrated circuit (Si-MMIC) is still an evolving technology particularly for the higher microwave and millimeter wave bands, it offers a very low-cost realization making it a very desirable option [108]. The direct modulator described in the last chapter is very suitable for implementation using this technology. The Si-MMIC version was implemented at a center frequency of 18 GHz. This implementation can be scaled to millimeter wave frequencies as well.

6.1 Direct Microwave M-QAM Implementation using MIC Technique

6.1.1 Subsystem Architecture

The architecture of a high power direct QAM modulator is shown in Fig.6.1. A Lange coupler is used to create the required RF signals for the I
and Q paths [4]. In each path, a rat-race coupler is used to get the inphase and out of phase RF signal. The subsequent circuit blocks in each path consist of a single pole double throw switch (SPDT) switched by the polarity of incoming data, a reflection attenuator implemented by using forward biased PIN diode terminations and controlled by the absolute values of a baseband pulse amplitude modulation (PAM) signal, and a variable reflection phase shifter to adjust the phase error. To realize the modulated signal filtering and to compensate for nonlinearities, a pre-filtering and pre-distortion block may be used to shape the baseband PAM signal in each path [16],[7]. Finally an inphase power combiner is used to combine signals from two paths into one high power QAM modulated and filtered output at microwave frequency. Although the center frequency is selected to be 2.5 GHz in a 500 MHz bandwidth for MIC implementations, a similar design may be repeated for different bandwidth and center frequencies. The various parts of this modulator are described in the following sections.

6.1.2 Quadrature Coupler

As may be seen in Fig.6.1., the power oscillator signal must be divided into inphase and quadrature components. A Lange coupler, shown in Fig.6.2,
is a suitable circuit to realize this block [63]. An ideal Lange coupler is a quadrature hybrid that equally divides input signals between direct and coupled ports with quadrature phase shift. Moreover, using a Lange coupler, a very broadband quadrature hybrid can be realized. In the Lange coupler, bonding wires are used to interconnect the microstrip lines. The length of coupler should be \( \lambda/4 \) in the operating frequency and the width of the fingers and their gap are estimated according to the coupling factor. A microstrip Lange coupler was implemented using a 25 mil thick substrate with \( \varepsilon_r = 10 \). The width of fingers, \( W \), their gaps, \( S \), and length of coupler at 2.5 GHz are \( W = 1.827 \) mil (1 mil = 0.0254 mm), \( S = 1.970 \) mil, and \( L = 481.352 \) mil, respectively. The coupler arms are 50 \( \Omega \) lines with \( W_1 = 23.57 \) mil. The circuit was designed using Libra-EEsof [109]. The layout and performance of the Lange coupler is shown in Fig.6.3. As may be seen from Fig.6.3.a, over the 500 MHz band, the through and coupled ports have almost the same amplitudes. Moreover, Fig.6.3.c shows that the two outputs have a 90° phase shift. Thus, the Lange coupler can operate efficiently as a quadrature hybrid. As may be seen from Fig.6.3.b the return loss and isolation performance of this coupler is very good over the 500 MHz band.
Figure 6.3 Lange Coupler a) S14 and S12 as a function of frequency
b) S11 and S13 as a function of frequency
c) Phase S14 and S11 as a function of frequency
6.1.3 Ring Coupler

As seen in Fig.6.1, an 180° coupler is required to provide the inphase and out of phase components to the RF signal in each path. A suitable circuit to realize this block in MIC technology is the rat race (ring) hybrid. Such a coupler is shown in Fig.6.4. When a signal is applied to port 1, the outputs at ports 2 and 3 are inphase, and when a signal is applied to port 4, the outputs at ports 2 and 3 will have a 180 degree phase shift. The characteristic impedance of the coupler arms is 50 Ω, and the characteristic impedance of the ring is 70.7 Ω. A standard design process was followed to realize this coupler [63]. The coupler was implemented on microstrip with a 25 mil thick substrate with $\varepsilon_r = 10$. The circuit was designed and optimised using Libra EEsOf [109]. In Fig 6.4, the physical dimensions of the coupler are $W1 = 23.57$ mil corresponding to 50 Ω lines, $W = 10.14$ mil corresponding to 70.7 Ω lines, and $R = 455$ mil. As may be seen from Fig.6.4.a, the coupling and through outputs are over 3 ± .2 dB range over a 500 MHz bandwidth. A good return loss and isolation (better than 20 dB) were obtained over the band. According to Fig.6.4.c, the phase shift between coupling and through ports is close to 180°.
Figure 6.4 A Rat-race or ring hybrid a) $S_{14}$ and $S_{12}$ as a function of frequency b) $S_{11}$ and $S_{13}$ as a function of frequency c) Phase $S_{14}$ and $S_{11}$ as a function of frequency
6.1.4 SPDT Switch

An SPDT switch must be implemented to provide the polarity modulation for the RF signal. The SPDT is controlled by the sign of the incoming data. Using an SPDT, 0° and 180° outputs of the ring coupler are selected based on the positive and negative polarities of the incoming data. A series configuration, shown in Fig.6.5, was used to realize this switch as the series configuration has the advantage of wider bandwidth compared with parallel topology. The switch uses a PIN diode in each arm. The diode bias was controlled by the data polarity.

To realize a high performance QAM modulator, an SPDT switch with high isolation is required. While traditional implementation of series SPDT switches on microstrip suffer from low isolation, a coplanar waveguide (CPWG) implantation could solve this problem[110]. Although CPWG implementation highly improves the isolation of the switch, it increases the insertion loss slightly. Fig.6.5 shows the insertion loss, isolation and return loss of an ideal line, as well as those of a microstrip, and a CPWG series SPDT switch. The PIN diodes are SMP1304 from Alpha industries [106], biased using a 20 mA current and unbiased with a zero current. The measured scattering parameters of the diodes provided by the manufacturer are used in the simulation. For the microstrip line implementation, the best performance is obtained with the following dimensions in Fig.6.5: \( W_1 = W_2 = W_3 = 23.57 \text{ mil} \), and \( A_1 = 18.65 \text{ mil} \), \( A_2 = A_3 = 100 \text{ mil} \), and \( L_1 = 1000 \text{ mil} \), \( L_2 = L_3 = 682.51 \text{ mil} \), \( D_1 = 344.70 \text{ mil} \), \( D_2 = D_3 = 0 \). As may be seen from Fig.6.5, the optimized microstrip switch characteristics are close to an ideal line switch performance. The microstrip implementation has an insertion loss of about .5 dB and an isolation of about 16 dB. A return loss of about 20 dB is also feasible with this implementation. The optimized dimensions for a coplanar waveguide implementation are: \( W_1 = W_2 = W_3 = 23.57 \text{ mil} \), and \( A_1 = 66.7 \text{ mil} \),
Figure 6.5 A series SPDT switch implemented using Alpha PIN 1304 and a forward bias current 20 mA
a) Insertion loss b) Isolation c) Return loss
\( A_2 = A_3 = 100 \text{ mil} \), and \( L_1 = 698.5 \text{ mil} \), \( L_2 = L_3 = 332.1 \text{ mil} \), \( D_1 = 128.2 \text{ mil} \), \( D_2 = D_3 = 100 \text{ mil} \) with a constant gap space of 1 mil. As may be seen from Fig.6.5, an insertion loss of about 2.2 dB is significantly more than the microstrip version. The isolation of around 25 dB is much better than the microstrip version. The insertion loss can be lowered by using a higher current than 20 mA for the forward bias arm. The return loss is acceptable within the desired bandwidth. Better performance could be achieved using higher cost diodes.

6.1.5 Variable Attenuator with Phase Shift Compensation

Reflection Attenuator

A voltage-controlled attenuator must be used to attenuate the signal to the levels required for QAM implementation. The control signal for the attenuator in each path is the absolute value of the multi-level PAM signal. A reflection attenuator using two PIN diode terminations at the output of a Lange coupler was used. In a reflection attenuator, the return loss depends on the hybrid performance. This allows an independent design for the desired insertion loss characteristics over the operating bandwidth. This attenuator is shown in Fig.6.6. The PIN diodes are SMP1304 from Alpha industries. The measured scattering parameters from the manufacturer were used to design the circuit. The attenuator response for different PIN diode bias currents is shown in Fig.6.6.a. The attenuator return loss is shown in Fig. 6.6.b. If PIN diode impedance is denoted by \( Z_D \), the output signal of the attenuator is given by [111]:

\[
S_{21} = 2S_{21}^C \rho T S_{41}^C,
\]

where

\[
\rho T = \frac{Z_D - Z_o}{Z_D + Z_o},
\]

120
and $S_{21}$ is the coupled voltage transmission coefficient of the coupler, $S_{41}$ is the direct voltage transmission coefficient of the coupler, and $Z_o$ is the characteristic impedance. Thus, the attenuation (in dB) is given by:

$$\text{Att} = -10\log(|S_{21}|^2). \quad (6.3)$$

As seen in Section 6.1.2, a Lange coupler has very good quadrature hybrid characteristics. Eq.(6.3) may be simplified as:

$$\text{Att} = -10\log(|\rho_T|^2). \quad (6.4)$$

In the forward bias operation, the PIN diode may be approximated by a variable resistor with impedance, $Z_d = R_d \{102\}$. Using Eq.(6.2), the attenuation for a PIN diode attenuator is given by:

$$\text{Att} = -10\log\left(\left|\frac{R_d - Z_o}{R_d + Z_o}\right|^2\right). \quad (6.5)$$

The PIN diode forward bias resistance as a function of bias current for an Alpha SMP1304 PIN diode is shown in Fig.6.7.a. In this measurement $R_d$ represents the real value of $Z_d$. Because of package parasitic elements, the results of Chapter 5 cannot be directly used to estimate $R_d$. A curve fitting method was used to model the diode forward bias resistance:

$$R_d = \frac{.0009}{I_d^3} - \frac{.787}{I_d^2} + \frac{39.3}{I_d} + 1.26. \quad (6.6)$$

In Fig.6.7.a, curve fitting results are compared with the measured results. From Eq.(6.6) and Eq.(6.5), it is clear that the attenuation (in dB) has a nonlinear relation with the bias current. On the other hand, a linear relation between the control voltage, $V_c$, and the attenuation of the reflection
Figure 6.6  A reflection attenuator using Alpha PIN 1304 (without pre-distortion circuit), a) Attenuation, b) Return Loss, c) Phase shift.
modulator is required. This requirement is critical when a filtered baseband signal is used as the attenuator control signal [112],[113]. The nonlinear relation between the forward bias current (in mA) and attenuation (in dB) is shown in Fig.6.7.b. A pre-distortion circuit has to be used to achieve a linear relation between the control voltage and attenuation. For the 25 dB dynamic range of the attenuator a seventh order polynomial provides the required predistortion. The required predistortion characteristics are shown in Fig.6.7.c. After the linearization the attenuation variation as a function of control voltage is shown in Fig.6.7.d.

Reflection Phase Shifter

Due to parasitic elements such as a junction capacitor and a package inductor, the phase shift of the reflection attenuator varies with the bias current. The phase shift for a reflection attenuator using $I_d$ as a parameter is shown in Fig.6.6.c. This phase shift has to be compensated. In Fig.6.1, a reflection phase shifter was included to compensate for the phase error associated with different levels of attenuation. The phase shifter was realized using a Lange coupler and varactor diode terminations.

The phase shifter uses a Lange coupler as a hybrid and two varactor diodes (SMV 120412 from Alpha industries) [106]. The phase shift, insertion loss, and return loss of the phase shifter are shown in Fig.6.8 as function of frequency. As may be seen from Fig.6.8, the attenuation level of the phase shifter is bounded between $-5 \pm 0.1$ dB. The return loss is also within an acceptable range.

The phase shift as a function of control voltage is shown in Fig.6.9. As may be seen, the phase shifter has an almost linear phase relation (with accuracy $\pm 2$ degree) with control voltage. This makes it quite suitable to compensate the phase error associated with different levels of attenuation.
Figure 6.7  a) PIN diode resistance versus forward bias current, b) Attenuation of reflection attenuator versus PIN diode bias current, c) The required predistortion characteristics d) Attenuation versus control voltage
Figure 6.8 A reflection phase shifter using varactor SMV120412 from Alpha a) Phase shift 
b) Insertion Loss c) Return loss
6.1.6 Power Combiner

To combine the inphase and quadrature path signals, an inphase power combiner is needed. A Wilkinson power combiner was used for this purpose. The layout of such a combiner is shown in Fig.6.1.a. The performance of this power combiner is presented in Fig.6.10 as well.

6.1.7 QAM Modulator

An MIC version of a direct microwave QAM modulator at a center frequency of 2.5 GHz and bandwidth 500 MHz was realized using the above circuits. The I and Q paths attenuation is shown in Fig.6.11. As may be seen from the results in Fig.6.11, a high performance and broadband M-QAM modulator was realized using these circuits. The number of levels is selected by the control voltage, \( V_c \). The constellation diagram of the direct microwave QAM modulator over a 500 MHz bandwidth is presented in Fig.6.12 for the
Figure 6.10 A Wilkinson power combiner a) Insertion Loss
              b) Return loss  c) Phase shift
Figure 6.11 I-Q path attenuation for M-QAM Modulation

a) 4QAM  b) 16QAM  c) 64QAM

4QAM, 16QAM, and 64QAM cases. An ideal demodulator was assumed for obtaining these constellation diagram. As may be seen from these diagrams the vector error magnitude is quite low and the proposed method results in a high performance modulator.

6.2 Direct Microwave M-QAM Implementation Using Si-MMIC Technique

As stated before, a cost effective realization is an important design objective for the direct microwave QAM modulator. This objective can be realized by using a monolithic microwave integrated circuit (MMIC) implementation. The selection of particular MMIC technology is also very im-
Figure 6.12 Constellation diagram for the direct M-QAM modulator a) 4QAM modulator. b) 16QAM modulator. c) 64QAM modulator
important. The GaAs MMIC technology has been commonly used due its excellent microwave circuit characteristics. On the other hand, Si-MMIC is an evolving technology that promises low cost implementation of microwave and millimeter circuits [114], [115]. The direct microwave QAM modulator has an architecture that is quite suitable for this technology as it is easy to design a shunt PIN diode in Si-MMIC technology. This section presents an Si-MMIC realization of the modulator at 18 GHz.

6.2.1 PIN Diode in Si-MMIC Technology

The side view of a grounded PIN diode realization using Si-MMIC is shown in Fig.6.13. As an example, the substrate thickness was selected as 100 $\mu m$ with an I region width of 96 $\mu m$. For a 50 $\Omega$ microstrip line on a $h=100 \, \mu m$ thick silicon substrate, the first hybrid cut off mode, $f_{c,HE_1}$, is 200 GHz [116]. Using Eq.(5.73), the PIN diode resistance is presented in Fig.6.14. As discussed in the last chapter, the results of this equation are accurate and can be efficiently used for Si-MMIC circuit designs.

6.2.2 Subsystem Design on Si-MMIC Technology

Lange Coupler

Two important parameters concerning the circuit design using Si-MMIC are permittivity and loss tangent $\tan\delta$. The measured values for silicon are $\varepsilon_r = 11.68(\pm0.7\text{percent})$, and $\tan\delta = 1.3 \times 10^{-3}(\pm30\text{percent})$. With this
Figure 6.14 PIN diode resistance as a function of forward bias current in Si-MMIC

Figure 6.15 A Lange coupler in Si-MMIC implementation a) $S_{12}$ and $S_{14}$ [dB] b) $S_{11}$ and $S_{13}$ [dB]
Figure 6.16 A Rat-race coupler in Si-MMIC implementation

a) $S_{34}$ and $S_{24}$ [dB] b) $S_{44}$ and $S_{14}$ [dB] c) $S_{24}$ and $S_{34}$ [degree]

substrate the measured attenuation of 50 $\Omega$ microstrip line is 0.6 dB/cm. A microstrip Lange coupler was designed and optimised on a silicon substrate with a 100 $\mu m$ thickness. The width of fingers, their gaps, and coupler length are $W=7.88 \mu m$, $S=7.12 \mu m$, and $L=1504 \mu m$, respectively. The results are presented in Fig.6.15. As can be seen, a high performance coupler over 4 GHz bandwidth is realized using this technology.

Ring Coupler

A ring coupler was also designed and optimised for a silicon substrate. The design method is similar to the previous rat-race coupler and the dimensions are $W1=78.6 \mu m$ for the 50 $\Omega$ lines, $W=30.86 \mu m$ for 70.7 $\Omega$ line, and radius of the ring $R=1514 \mu m$. The results are presented in Fig.6.16. As may be seen, the ring coupler operates as a high performance 180° coupler.
over a wide bandwidth. If the miniaturisation of the circuit is desired, a lumped-element equivalent circuit or a reduced-size technique can be used instead of distributed implementation [58],[117],[118]. However, using these techniques decreases the bandwidth of the coupler.

**SPDT Switch on Si-MMIC**

As discussed, a shunt PIN diode can be easily implemented in Si-MMIC technology. The implementation of SPDT switch using shunt diodes is shown in Fig.6.17 [63]. A current \( I = 30 \) mA (corresponding to \( .4 \) \( \Omega \) forward bias resistance) was selected to bias the diode. A simple equivalent circuit, shown in Fig.6.17, is suitable as a model for Si-MMIC PIN diode. Due to high relative dielectric constant for silicon, the fringing capacitance (in air) around the \( I \) region is relatively small, and the capacitance calculated using the parallel plate capacitance formula below provides a useful estimate of the junction capacitance \( C_J \).

\[
C_J \approx \frac{\varepsilon_0 \varepsilon_r A}{W},
\]

(6.7)

where \( A \) is the junction area. For a square junction, \( A = 78.6^2 \mu m^2 \), resulting in a \( C_J = 6.7fF \). The insertion loss, isolation, as well as the return loss of shunt SPDT switch are shown in Fig.6.17. As may be seen, high performance SPDT switch over a very wide bandwidth is achieved.

**Reflection Attenuator Using Si-MMIC Technology**

By using Eq.(6.5) and the estimated \( R_d \) from Fig.6.14, a reflection attenuator can be realized in Si-MMIC technology. As may be seen from Fig.6.18, a given attenuation value can be realized using two bias current values that correspond to PIN diode resistance being smaller or bigger than 50 \( \Omega \). To avoid ambiguity, the bias current was limited from \( .3 \) mA to 6 mA. As may be seen in Fig.6.18, this current interval provides a 24 dB dynamic range for the attenuator.
Figure 6.17 An SPDT switch in Si-MMIC implementation
a) Switch topology b) Pin diode equivalent circuit
c) Insertion loss d) Isolation e) Return loss of switch
As for the MIC realization, a predistortion is required to linearise the control voltage versus attenuation. The predistortion realization is quite straightforward. As may be seen from Fig.6.14, when the bias current is limited to the .3 mA to 6 mA interval, the PIN diode forward bias resistance is always less than 50 Ω. Then, Eq.(6.2) may be written as:

\[ |\rho_T| = \frac{Z_o - R_d}{Z_o + R_d}, \]  

or

\[ R_d = \frac{Z_o(1 - |\rho_T|)}{1 + |\rho_T|}. \]  

Defining control voltage for an ideal diode as the interval .05 \( \leq V_c \leq .9 \) V, one may develop a predistortion relation assuming \(|\rho_T| = V_c\). The PIN diode resistance could be approximated as [102],

\[ R_d = \frac{W^2}{2\mu T I_d}. \]  

A simple relation between \( V_c \) and \( I_d \) is obtained as

\[ I_d = \frac{W^2}{2\mu T Z_o(1 - V_c)} \left( 1 + \frac{V_c}{Z_o(1 - V_c)} \right). \]  

This predistortion was used for the implementation of the modulator. Its performance is reported in the next chapter.

**Wilkinson Power Combiner in Si-MMIC**

To provide an inphase power combining, a single section Wilkinson power combiner was designed and optimised with a center frequency of 18 GHz and a bandwidth of 4 GHz. Similar to the MIC version, the combiner consists of a 50 Ω line input and output microstrip lines (78.6 μm wide), two uncoupled 70.7 μm lines (30.85 μm wide), and a 100 Ω isolation resistance. Thin film
Figure 6.18 Attenuation of reflection attenuator as a function of bias current

Figure 6.19 A Wilkinson combiner in Si-MMIC implementation a) Insertion loss, $S_{12}$ and $S_{13}$ [dB] b) Return loss, $S_{11}$ [dB] c) Phase shift, $S_{12}$ and $S_{13}$ [degree]
resistors have a resistance given by [115]:

\[ R = \frac{l}{w} \rho_s \]  
(6.12)

where \( l \) is the length of the resistor, \( w \) is the width of the resistor, and \( \rho_s \) is the sheet resistance of resistive material in \( \Omega/\text{square} \). For Si-MMIC, \( \rho_s = 342 \ \Omega/\text{square} \) results in a width of 192 \( \mu m \) for a length of 56 \( \mu m \). As may be seen from Fig.6.19, the insertion and return loss in two arms are equal.

**M-QAM Modulator Using Si-MMIC**

In addition to its lower cost, a Si-MMIC M-QAM modulator offers a number of performance advantages as well [12]. The parasitic and package elements have low values and the junction capacitor has a very small value as well. These characteristics result in low phase shift variation in the multilevel attenuator. Moreover, the phase variation for a single frequency operation can be compensated by using open circuit stubs. This method is discussed in Appendix B. Thus, the phase compensation circuit is not needed provided that the operation is limited to a narrow band operation.

A direct microwave QAM modulator was implemented using Si-MMIC technology [12]. As before, the modulation was generated using polarity modulation and attenuation control in the I and Q paths. The layouts of the polarity modulator section and the attenuator section are presented in Figures 6.20 and 6.21. The bias circuit design is described in Appendix C. The performance of the modulator is presented in the next chapter.

As stated before, a distributed implementation of the couplers and power combiner have been used to realize the modulator. This technique results in a higher bandwidth. A lumped-element equivalent circuits may be used
to miniaturise the couplers and power combiner. However, the design of the lumped-element circuits must be somewhat empirical, and it needs precise inductor models based on a specific foundry process. Moreover, the design becomes difficult at frequencies above 20 GHz [118].

6.3 Summary

In this chapter microwave integrated circuit (MIC) and silicon monolithic microwave integrated circuit (Si-MMIC) implementations of the direct microwave QAM modulator were presented. The MIC implementation uses microstrip and co-planar waveguides as well as discrete silicon PIN diodes. The constellation study over the 500 MHz bandwidth shows an error vector magnitude better than 10 percent for the MIC modulator. The Si-MMIC implementation uses microstrip line. PIN diodes were realized in silicon substrate. While realization using the MIC is simple and straightforward, the Si-MMIC realization offers a low cost and high performance modulator.
Figure 6.20 Layout of the polarity modulator section of direct microwave QAM modulator
Figure 6.21 Layout of the attenuator section of direct microwave QAM modulator

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7. PERFORMANCE OF FIXED WIRELESS ATM TRANSMITTER

The architecture and hardware for an adaptive direct microwave M-QAM modulator was described in the last two chapters. In this chapter the operation of the radio ATM transmitter for different wireless channels and under various traffic loads is examined. The Ricean and Rayleigh wireless channels are considered. The traffic sources are VBR computer data and video traffic.

7.1 Adaptive Direct Microwave M-QAM Transmitter for Radio ATM

A model for the adaptive direct microwave QAM modulator is presented in Fig.7.1. As may be seen, this transmitter consists of two units. The M-QAM control unit is used to provide an accurate estimate of the required bandwidth in a wireless channel. The direct microwave QAM modulator unit transmits the multimedia traffic through a wireless network. A control signal from the M-QAM control unit to the modulator unit assures that an optimum response to the bandwidth demand can be accommodated in a wireless channel. A detailed architecture for the transmitter is shown in Fig.7.2. Different sections of this modulator were described in the previous chapters. In this chapter the functionality as well as performance of different sections are examined.

7.1.1 Advantages of using Effective Bandwidth Metric

The effective bandwidth estimation is done in the M-QAM control unit. Using the effective bandwidth metric, a statistical model to estimate the re-
required bandwidth was used. This model is a real time model, that provides a close upper bound estimation for various traffic. The model was discussed in Chapters 3 and 4. The accuracy of this model is evident from Fig.7.3.a where the actual required bandwidth of the StarWars movie is upper bounded with its effective bandwidth using buffer size as a parameter. The advantage of using the effective bandwidth instead of the peak rate bandwidth for different videotraces is quite clear from Fig.7.3.b for cell loss ratio $10^{-6}$. As can be seen from this figure, a bandwidth saving between 25 to 40 percent for video transmission can be easily achieved. The saving percentage is a function of the buffer size. Figure 7.3.c compares the actual required bandwidth of the aggregate of three MPEG-1 coded movies (using Table 3.1) with its effective bandwidth. As can be seen from Fig.7.3.b and 7.3.c, using the effective bandwidth also shows a statistical multiplexing gain for aggregate video traffic. For instance, while the effective bandwidth of the videotraces are 760, 590, and 680 Kbps for a buffer size 350 Kbits, respectively, their aggregate traffic requires only 1500 Kbps bandwidth using 1 Mbits buffer size. This shows a statistical multiplexing gain of about 1.35. The peak rate and effective bandwidth are compared for aggregate video traffic in Fig.7.3.d. As may be seen the peak rate requires about 3 Mbps capacity, while the effective bandwidth reduces it to around 1.5 Mbps. These results show that a high bandwidth saving can be achieved using the effective bandwidth metric. It should be pointed out that this metric provides an accurate as well as a simple solution.
Figure 7.2 An adaptive direct QAM modulator for wireless ATM
to a very complicated bandwidth estimation problem.

As pointed out before, besides the different bandwidth requirements for various traffic sources, an extra factor arises due to the wireless network. As discussed in Chapter 4, a wireless channel capacity decreases due to fading. A capacity reduction factor due to fading was introduced to take this into account. A new metric, termed the modified effective bandwidth, was introduced to include the effect of channel performance on capacity allocation policy. Estimation of this metric is the main task of the M-QAM control unit. Fig.7.3.d compares the modified effective bandwidth with peak rate and effective bandwidth for aggregate video traffic for cell loss ratio $10^{-6}$ in Ricean channel. As may be seen, using the modified effective bandwidth, a higher bandwidth has to be reserved. This results a guaranteed QoS in the wireless channel.

7.2 Adaptive M-QAM Performance

In the previous section it was shown that using effective bandwidth instead of peak rate allocation highly saves the bandwidth. In this section, a simulation study is conducted to evaluate the performance of the adaptive M-QAM modulator. The M-QAM control unit uses modified effective bandwidth to adapt the M-QAM modulator. The objective of the study is to evaluate the advantages of using an adaptive structure over a conventional fixed QAM modulator. The adaptation was realized using an M-QAM control unit. The operation of the M-QAM control unit is summarised as follows.

According to Fig.7.2, the ATM connections introduce their traffic and QoS parameters to the gate control port of the M-QAM control unit. According to their VP addresses, the connections are directed to different VPs. Then, the traffic characteristics of these VPs are recalculated to include the
Figure 7.3  

(a) Experimental estimation versus effective bandwidth with buffer size as a parameter for StarWars movie.  
(b) Effective bandwidth and peak rates of different video traces  
(c) Experimental estimation versus effective bandwidth with buffer size as a parameter for aggregate video.  
(d) The peak bandwidth, effective bandwidth, and modified effective bandwidth for aggregate video traffic. The cell loss ratio is $10^{-6}$ in a Ricean channel.
characteristics of the new connections. The buffer sizes are selected by the buffer size manager according to delay constraints. This is followed by an estimation of the effective bandwidth and capacity reduction factor. Using these parameters, the modified effective bandwidth metric is calculated. A new modulation level is considered by the M-QAM control unit if the new level of modulation satisfies the outage conditions. This is used to adjust the level of the direct microwave M-QAM.

7.2.1 LAN Traffic

In the first experiment, a large number of ATM connections, with LAN traffic, are applied sequentially to an adaptive transmitter. The system uses three VPs. A connection should be directed to a VP according to its VP address. The connection selects a particular VP randomly in this experiment. Although the delay is not a critical QoS parameter for data transmission, a maximum delay bound is used during simulation to perform a buffer assignment policy. Each call must introduce its traffic as well as its QoS parameters. The required traffic parameters in this experiment are mean rate, variance coefficient and Hurst parameter. The QoS parameters are cell loss ratio and maximum tolerable delay. These parameters are generated using the MATLAB random number generator. The range for different parameters are limited to: \(1 \leq R_{\text{mean}} \leq 10 \text{ Mbps}, \ 10^5 \leq \sigma \leq 10^6 \text{ bit-sec} \), \(.5 \leq H \leq 1, 10^{-6} \leq p_{\text{CLR}} \leq 10^{-3}, \text{ and } .001 \leq D_{\text{max}} \leq .01 \text{ sec.}\)

A VP initially carries traffic having a mean rate between 1 kbps and 15 Mbps, using a buffer size between 100 kbits and 1 Mbits, and having variance coefficient, \(\sigma\), between \(10^4\) and \(5 \times 10^5\) bit-sec, respectively. The Hurst parameters of VPs is also random between .5 and 1. When a new LAN connection is applied to system, the initial values for VPs are regenerated. The channel is considered with burst errors. The burst width of the channel is a uniform random variable between 2 and 10.
The traffic is to be transmitted through a LOS radio link. The radio link discussed in Section 2.8 was used in the simulation. The link design parameters, presented in Table 2.2, were used in the outage block. As a result of the traffic varieties of the ATM calls, the network requires a variable bandwidth. The M-QAM control unit estimates this demand by using the modified effective bandwidth metric. This demand is satisfied by M-QAM level variation. The feasibility of using a new modulation level is examined by the outage block. According to bandwidth demand, the constellation size is a variable between \( M = 2 \) and \( M = 256 \). This results in a constellation size \( k \) of 2 to 8. The objective of simulation is to extract the distribution of the effective bandwidth, \( BW_{eff} \), the modified effective bandwidth, \( BW_{meff} \), the constellation size \( k \), and the outage. The experiments are conducted for Ricean and Rayleigh channels.

**Operation in Ricean Fading**

A large number of ATM calls are sent to the M-QAM control unit. The constellation size, \( k \), outage, \( BW_{eff} \), and \( BW_{meff} \) variations are shown in Fig.7.4.a. As expected, the constellation size, outage, effective bandwidth, and modified effective bandwidth change with the bandwidth demand. The probability of using any of the constellation sizes is presented in Fig.7.4.b. As may be seen, the traffic can be often transmitted using a constellation size \( k = 2 \) and \( k = 3 \). A 256-QAM(\( k = 8 \)) would be used if a conventional design had to be selected. The average and standard deviation of different parameters are shown in Table 7.1. These values only have statistical meanings. Fig.7.5 shows the acceptance improvement compared over that for the different fixed QAM modulators. As may be seen, in a Ricean channel a good improvement compared with the fixed QAM can be achieved. The outage performance of an M-QAM is also compared with the different fixed
QAM modulators in Fig.7.6. As may be seen, while a slight outage deficiency is tolerated compared with a low level fixed QAM, the adaptive modulator offers considerable performance advantage over the fixed high level QAM modulators.

**Table 7.1** Average and standard deviation for modulation level, outage, effective bandwidth, and modified effective bandwidth for Ricean Channel

<table>
<thead>
<tr>
<th></th>
<th>Modulation level</th>
<th>Outage</th>
<th>$Bw_e[Mbps]$</th>
<th>$Bw_{em}[Mbps]$</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Average</strong></td>
<td>9.9632</td>
<td>.00125</td>
<td>56.166</td>
<td>60.8545</td>
</tr>
<tr>
<td><strong>std</strong></td>
<td>18.2773</td>
<td>.00255</td>
<td>22.028</td>
<td>26.0022</td>
</tr>
</tbody>
</table>

**Operation in Rayleigh Fading**

Although a Ricean channel model is valid for a LOS link and a fixed wireless ATM link must be studied under this category, a performance study under Rayleigh channel conditions has some use, e.g., for comparison with the Ricean channel or expansion of the model to land mobile communications. Fig.7.4.c shows the constellation size, outage, effective bandwidth and modified effective bandwidth variations for Rayleigh channel. The constellation size distribution for the Rayleigh channel is shown in Fig.7.4.d. The average and standard deviation of these parameters are also shown in Table.7.2. As can be seen, compared with the Ricean channel, while the effective bandwidth is almost the same (traffic characteristics are independent from the channel), the constellation size and modified effective bandwidth have a higher average and variance. This behaviour is predictable due to the nature of the Rayleigh channel.

Moreover, by comparing Table 7.1 and 7.2, one can see that the average outage in the Rayleigh channel has been increased. The call acceptance per-
formance is shown in Fig.7.5. As can be seen, an M-QAM generally increases
the traffic acceptance in different channels; however, a higher performance
can be seen in Rayleigh fading. This is due to higher channel capacity vari-
ation in Rayleigh fading. The outage performance for the Rayleigh channel is
also compared with the different fixed QAM modulators in Fig.7.6. The su-
perior outage performance using the adaptive M-QAM modulator is evident.

| Table 7.2 | Average and standard deviation for modulation
level, outage, effective bandwidth, and modified
effective bandwidth for Rayleigh Channel |
<table>
<thead>
<tr>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Modulation level</td>
</tr>
<tr>
<td>Average</td>
<td>13.5944</td>
</tr>
<tr>
<td>std</td>
<td>28.8993</td>
</tr>
</tbody>
</table>

7.2.2 VBR MPEG Video Traffic

In the next experiment, a number of MPEG-1 encoded videos are used as
the traffic sources. These video sources and their traffic characteristics are
shown in Table 7.3.[84], [119]. The number of video sources simultaneously
directed to a VP was a random number between 0 and 30, and the number
of copies from a single video is between 0 and 5. As in the previous example,
the QoS parameters $p_{CLR}$ and $D_{max}$, were obtained using a random number
generator with range $1 \leq D_{max} \leq 10$ msec, and $10^{-9} \leq p_{CLR} \leq 10^{-6}$. Again,
three VPs, each initially carrying traffic with a mean bit rate between 100
kbits and 15 Mbits, a buffer size between 100 kbits and 1 Mbits and a
variance coefficient between $10^4$ and $5 \times 10^5$ bit-sec were considered. The
Hurst parameter of VPs was also random between .75 and .95. The channel is
considered to be a bursty channel with the burst width uniformly distributed
between 2 and 10. The radio link characteristics were similar to the link
studied in Section 2.8. The outage block examines the link with parameters
listed in Table 2.2 to estimate the outage probability. According to the
Figure 7.4  a) Simulation results for M-QAM in a Ricean channel for LAN traffic based on the number of connections b) Constellation size distribution in a Ricean channel for LAN traffic c) Simulation results for M-QAM in a Rayleigh channel for LAN traffic based on the number of connections d) Constellation size distribution in a Rayleigh channel for LAN traffic
Figure 7.5  ATM call acceptance improvement compared with fixed QAM modulator in Ricean and Rayleigh channels for computer data traffic

Figure 7.6  The outage improvement compared with fixed QAM in Rayleigh and Ricean channels for computer data traffic
bandwidth demand, the constellation size can be varied between $M = 2$ and $M = 256$.

<table>
<thead>
<tr>
<th>Traffic characteristics of various video traffic</th>
<th>Mean [Mbps]</th>
<th>Var. Coeff. [b/s]</th>
<th>Hurst</th>
</tr>
</thead>
<tbody>
<tr>
<td>Jurassic Park</td>
<td>.326950</td>
<td>4.9277e4</td>
<td>.8448</td>
</tr>
<tr>
<td>Silence of the Lambs</td>
<td>.182788</td>
<td>6.4843e4</td>
<td>.8959</td>
</tr>
<tr>
<td>Star Wars</td>
<td>.232830</td>
<td>7.25e4</td>
<td>.8458</td>
</tr>
<tr>
<td>Terminator II</td>
<td>.272625</td>
<td>2.7334e4</td>
<td>.89</td>
</tr>
<tr>
<td>Mr. Bean</td>
<td>.441175</td>
<td>1.0852e5</td>
<td>.85</td>
</tr>
<tr>
<td>Soccer</td>
<td>.678225</td>
<td>9.0173e4</td>
<td>.91</td>
</tr>
</tbody>
</table>

Operation in Ricean Fading

A large number of ATM video connections were sent sequentially to the M-QAM control unit. The number of videotraces in a call was limited to 30. The constellation size, outage, $BW_{eff}$, and $BW_{meff}$ variations are shown in Fig.7.7.a. As may be seen, as bandwidth demand changes, the constellation size, outage, effective bandwidth, and modified effective bandwidth change. However, the traffic variation for video are less than that of the LAN type traffic. The probability to use any of the constellation sizes is presented in Fig.7.7.b. The mean and variance of different parameters are shown in Table 7.4. Fig.7.8 shows the acceptance improvement compared with the fixed QAM modulators. The outage performance of an M-QAM is also compared with the different fixed QAM modulators in Fig.7.9. As may be seen, while a slight outage probability has to be tolerated compared with that of the low level fixed QAM, the adaptive modulator offers a much better performance compared with that for the fixed high level QAM modulators.

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Figure 7.7  a) Simulation results for M-QAM in a Ricean channel for VBR video traffic based on the number of connections b) Constellation size distribution in a Ricean channel for VBR video traffic c) Simulation results for M-QAM in a Rayleigh channel for VBR video traffic based on the number of connections d) Constellation size distribution in a Rayleigh channel for VBR video traffic
Figure 7.8  ATM call acceptance improvement comparison of the adaptive modulator and the fixed QAM modulator in Ricean and Rayleigh channel for video traffic

Figure 7.9  The outage improvement of the adaptive modulator compared with fixed QAM in Rayleigh and Rice channels for video traffic
Table 7.4  Average and standard deviation for modulation level, outage, effective bandwidth, and modified effective bandwidth for Ricean Channel for video traffic

<table>
<thead>
<tr>
<th></th>
<th>Modulation level</th>
<th>Outage</th>
<th>BWe [Mbps]</th>
<th>BWem [Mbps]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Average</td>
<td>4.8280</td>
<td>.000535</td>
<td>40.262</td>
<td>43.584</td>
</tr>
<tr>
<td>std</td>
<td>2.684</td>
<td>.000375</td>
<td>11.506</td>
<td>14.172</td>
</tr>
</tbody>
</table>

Operation in Rayleigh Fading

The M-QAM control unit operation was also examined for the Rayleigh fading channel for the video traffic. Fig.7.7.c shows the constellation size, outage, effective bandwidth and modified effective bandwidth variation of dynamic bandwidth demand. The mean and variance of these parameters are also shown in Table.7.5. Comparing Table 7.4 and 7.5, one can see that the average outage in the Rayleigh channel has been increased. The constellation size distribution for the Rayleigh channel is shown in Fig.7.7.d. As can be seen from Fig.7.7.d, a higher modulation level is required to transmit similar traffic in a Rayleigh fading. The call acceptance performance is shown in Fig.7.8. The outage performance for the Rayleigh channel is also compared with the different fixed QAM modulators in Fig.7.9. The outage performance in the high level QAM is evident.

Table 7.5  Average and standard deviation for modulation level, outage, effective bandwidth, and modified effective bandwidth for Rayleigh Channel for video traffic

<table>
<thead>
<tr>
<th></th>
<th>Modulation level</th>
<th>Outage</th>
<th>BWe [Mbps]</th>
<th>BWem [Mbps]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Average</td>
<td>5.488</td>
<td>.000627</td>
<td>40.6725</td>
<td>48.2593</td>
</tr>
<tr>
<td>std</td>
<td>2.831</td>
<td>.000396</td>
<td>11.2664</td>
<td>14.4715</td>
</tr>
</tbody>
</table>
7.2.3 Multiple Sources

In the previous experiments, a single ATM call was directed to a VP at a time. A general model is to apply the multiple inputs to each VP. The performance of the M-QAM control unit in Ricean and Rayleigh fading channels for Ethernet traffic and video traffic is examined in this section.

Multiple LAN Traffic Sources

In the first experiment, a number of Ethernet sources were simultaneously directed to the different VPs. The number of connections for each VP was between 1 and 4. The range for different parameters are limited to: $2 \leq R_{\text{mean}} \leq 5$ Mbps, $10^5 \leq \sigma \leq 10^6$ bit-sec , $.5 \leq H \leq 1$, $10^{-6} \leq p_{\text{CLR}} \leq 10^{-3}$, and $.001 \leq D_{\text{max}} \leq .01$ sec. There are three VPs and each VP initially carry traffic having a mean rate between 1 kbps and 10 Mbps. The other traffic parameters of VPs are similar to the previous experiments. The experiments were done for Ricean and Rayleigh fading channels. Fig.7.10.a shows the constellation size, outage, effective bandwidth and modified effective bandwidth variations in a Ricean channel. As can be seen, the effective bandwidth and modified effective bandwidth variations can also be seen in multiple source experiment. Fig. 7.10.b shows the constellation distribution in a Ricean channel. As may be seen, the M-QAM control unit selects the constellation size $k = 4$ with higher probability. Fig.7.10.c shows the constellation size, outage, effective bandwidth and modified effective bandwidth variations in a Rayleigh channel. A comparison between Fig.7.10.a and 7.10.c shows a higher variation in Rayleigh channel compared with Ricean channel. Fig.7.10.d shows that the most common constellation size in Rayleigh channel should be $k = 5$ compared to $k = 4$ for Ricean channel.
Figure 7.10 a) Simulation results for M-QAM in a Ricean channel for multiple LAN traffic sources b) Constellation size distribution in a Ricean channel for multiple LAN traffic sources c) Simulation results for M-QAM in a Rayleigh channel for multiple LAN traffic sources d) Constellation size distribution in a Rayleigh channel for multiple LAN traffic sources
Multiple Video Traffic Sources

An experiment was also conducted to study the multiple video traffic sources in Ricean and Rayleigh channels. In this experiment, each VP was requested to transmit between 0 and 30 MPEG-1 videotraces, and the number of copies from a single video is between 0 and 5. Table 7.3 shows the traffic characteristics of various video traffic sources. The initial values of VPs are similar to previous example. Fig.7.11.a shows the constellation size, outage, effective bandwidth and modified effective bandwidth variations in a Ricean channel. As may be seen, there is less variation compared with LAN traffic. This is due to lower mean rate of video traces compared to LAN traffic. Fig. 7.11.b shows the constellation distribution in a Ricean channel for multiple video sources. Fig.7.11.c shows the constellation size, outage, effective bandwidth and modified effective bandwidth variations in a Rayleigh channel. A comparison between Fig.7.11.a and Fig.7.11.c shows that a higher bandwidth is required in Rayleigh fading. This can also be seen by comparison between Fig.7.11.b and 7.11.d where operation in a Rayleigh channel in average requires a higher modulation level to transmit the similar traffic. Other interesting results can be seen from outage results. While two experiments used similar modulator and same input traffic, a Ricean channel resulted the better outage performance compared with Rayleigh channel. The average and standard deviation of the different parameters for multiple video sources are shown in Table 7.6. As can be seen, while the effective bandwidth is similar in Rayleigh and Ricean channels due to same input traffic, the average of the modulation level in Rayleigh channel is higher than that in Ricean channel. This is due to higher modified effective bandwidth requirements in a Rayleigh channel, and is a function of the parameter $K$ in a Ricean model. This experiment shows that a Rayleigh channel requires 12 percent more bandwidth compared to a Ricean model with $K = 6$. 

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Figure 7.11 a) Simulation results for M-QAM in a Ricean channel for multiple video traffic sources b) Constellation size distribution in a Ricean channel for multiple video traffic sources c) Simulation results for M-QAM in a Rayleigh channel for multiple video traffic sources d) Constellation size distribution in a Rayleigh channel for multiple video traffic sources
Table 7.6 Statistical Parameters for multiple MPEG-1 Video Traffic in a Ricean Channel

<table>
<thead>
<tr>
<th>Modulation level</th>
<th>Outage</th>
<th>BW_e [Mbps]</th>
<th>BW_{em} [Mbps]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Mean(Ricean)</td>
<td>16.8052</td>
<td>0.002208</td>
<td>76.6070</td>
</tr>
<tr>
<td>std(Ricean)</td>
<td>21.9679</td>
<td>0.003069</td>
<td>23.4240</td>
</tr>
<tr>
<td>Mean(Rayleigh)</td>
<td>22.0846</td>
<td>0.00295</td>
<td>78.2070</td>
</tr>
<tr>
<td>std(Rayleigh)</td>
<td>24.0069</td>
<td>0.00335</td>
<td>23.6688</td>
</tr>
</tbody>
</table>

7.2.4 Discussion of Results of the Experiments

The previous experiments were conducted to study the performance of the M-QAM modulator for a fixed wireless broadband network. These experiments presented the following results:

- The M-QAM control unit highly saves the bandwidth while providing a guaranteed QoS in wireless networks.

- The M-QAM control unit provides a flexible transmission system that highly improves the call acceptance and outage condition in wireless networks.

The advantages of using the effective bandwidth allocation compared with the peak rate allocation were discussed in Section 7.1. It was found that a bandwidth saving between 25 to 40 percent for a single video MPEG-1 trace can be easily achieved. This saving is close to 50 percent for the aggregate of three video traces due to statistical multiplexing gain. In a LAN type traffic, depends on utilisation and buffer size a high bandwidth saving is also achievable. For example, a bandwidth saving better than 60 percent was obtained for Bellcore August '89 Ethernet traces in Chapter 3. Moreover, using M-QAM control unit results in a high acceptance and outage improvement compared with conventional QAM modulators. As an example, in Fig.7.4.a one sees that the demand to use a more than 100 Mbps in a
Ricean channel is very limited. Thus, an adaptive modulator frequently operates in lower modulation level (e.g., \( k=2, \text{ and } 3 \)) and it improves the system outage. The outage improvement is due to using a lower modulation level compared with a 256-QAM conventional choice. Moreover, the system uses the high modulation level when there is a high bandwidth demand. As seen from Fig.7.5 and 7.6 as well as 7.8 and 7.9 a high acceptance and outage improvement can be achieved by using the M-QAM control unit. These results are also valid for multiple traffic sources.

### 7.3 Adaptive Direct Microwave M-QAM modulator

The second unit of the adaptive transmitter is a direct microwave QAM modulator. As shown in Fig.7.1, the modulation level of direct microwave M-QAM modulator is varied by using a control signal from the M-QAM control unit. When bandwidth demand is low, a low level modulation, e.g. 4QAM, is used. The use of low level M-QAM improves the system gain and decreases the outage. A low level modulation should also be used when a high QoS is required. On the other hand, when the bandwidth demand is high, a high level direct microwave QAM can be used provided that the required QoS can be met. The simulated constellation diagrams for different levels of QAM using two adaptive modulator implementations are presented below to illustrate the high adaptive performance over a broad bandwidth.

#### 7.3.1 Adaptive Direct Microwave M-QAM Implementation using MIC technique

This section outlines a performance study for a direct microwave QAM modulator using MIC. The constellation diagrams of the direct microwave M-QAM modulator with \( M=4, 16, 64, \text{ and } 256 \) and a center frequency 2.5 GHz are shown in Fig.7.12 over 500 MHz bandwidth. As can be seen, there is an error vector magnitude (EVM) in the constellation diagram. This error
is created due to broadband operation of the modulator. As may be seen, the error vector magnitude is less than 6 percent over the full bandwidth for 4QAM. This increased to 7 percent for 16QAM. For 64QAM this error is about 8 percent. The error vector magnitude for 256QAM is less than 10 percent. This increase is due to the attenuator response at higher attenuation levels. As may be seen in Fig.6.11, the attenuation characteristic is not completely flat. The error vector magnitude is in the acceptable level for practical systems.

7.3.2 Direct Microwave M-QAM Modulator Implementation using Si-MMIC Technique

As stated before, direct microwave QAM modulator realization using Si-MMIC technique is a highly cost effective implementation. Such a direct microwave modulator design was described in the last chapter. A performance study was conducted for this modulator using the model presented in Fig.5.17. In the forward bias mode, the shunt PIN diodes could be considered as a resistances. The PIN diode I region width was 100 \( \mu m \). The carrier life time was \( \tau = 5 \mu sec \). A predistortion circuit was designed using Eq.(6.12). The input data were selected to be at OC-3 (155.52 Mbps) rate. An eight level PAM signal with symbol rate 25.92 Msymbol/sec was used in each path. This translated to a four level PAM signal to each attenuator.

The power spectrum of the direct microwave 64QAM modulator and a theoretical 64QAM modulator without using baseband filter are compared in Fig.7.13. As may be seen the performance of the direct microwave QAM modulator is very close to an ideal QAM modulator. In addition, the level of the out of band spectrum and channel bandwidth are similar to an ideal 64QAM modulator. The degradation in the power spectrum of the direct microwave 64QAM modulator is mainly due to the predistortion circuit.
Figure 7.12 Constellation diagram for direct QAM modulator
a) 4-QAM. b) 16-QAM. c) 64-QAM. d) 256-QAM
Figure 7.13 The power spectrum of the Si-MMIC direct microwave and an ideal 64QAM modulator without using filter

In the next experiment, the bandwidth efficiency of a direct microwave MQAM modulator implemented using silicon MMIC was compared with that of a theoretical M-QAM modulator. The data rate is OC-3 and no pulse shaping filter was used. As can be seen from Fig. 7.14, a lower channel bandwidth can transmit an OC-3 data rate by using a higher modulation level. A theoretical channel bandwidth 38.88, 25.92, and 19.44 MHz is required to transmit an OC-3 signal using 16, 64, and 256 QAM modulator. As can be seen from Fig. 7.14, the performance of direct microwave M-QAM modulator is very close to the theoretical M-QAM modulator.

A raised cosine filter with roll-off factor $\alpha = .5$ was used in the next experiment. Fig. 7.15 shows the control voltage in I and Q paths as well as the PIN diode resistances corresponding to these signals. As may be seen, the PIN diode resistances vary between 0 and 50 $\Omega$. The power spectrum of the
Figure 7.14 Baseband Power Spectrum of M-QAM modulator implemented using Si-MMIC without using baseband filter for different levels of modulation.
direct microwave and an ideal 64QAM modulator using a predistortion and a prefilter with roll-off factor $\alpha = .1, \alpha = .5, \alpha = .7,$ and $\alpha = .95$ are presented in Fig.7.16 for OC-3 data rate (155.52 Mbps). As may be seen, the power spectrum in this figure demonstrates that prefiltering limits the out of band spectrum. As expected, the best out of band performance requires a higher channel bandwidth. As may be seen from Fig.7.16.a, although a filter using roll-off factor $\alpha = .1$ results in a narrower channel spectrum, it has a high out of band power spectrum. The best out of band power spectrum performance can be achieved with $\alpha$ close to one. However, as can be seen from Fig.7.16.d, a higher channel bandwidth is required in this case. A suitable trade-off is to use a roll-off factor $\alpha = .5$. As may be seen from Fig.7.16, while the pass band characteristics of the theoretical QAM and direct QAM are similar, the level of out of band spectrum is higher in the direct QAM modulator. This is due to the nonideal predistortion circuit.

7.4 Summary

To examine the operation of the adaptive direct microwave QAM transmitter in a wireless channel a performance study was conducted. The operation in Ricean and Rayleigh fading channels were compared for computer data and video traffic. It is shown that using modified effective bandwidth metric highly saves the bandwidth and improves the ATM call acceptance and wireless channel outage. It was also found that the modified effective bandwidth metric demands more bandwidth in a Rayleigh channel compared with a Ricean channel for similar traffic load. Moreover, the performance of the direct QAM modulator implemented using MIC and Si-MMIC techniques was studied in this chapter. A comparison between the direct modulator and theoretical QAM modulator shows a very good agreement.
Figure 7.15 Input current and pin diode forward bias resistance for I and Q paths in 64QAM modulator using a filter with roll-off factor $\alpha = .5$
Figure 7.16 Baseband power Spectrum for 64-QAM using direct microwave Si-MMIC implementation for 155(OC-3) Mbps data with different roll-off factor
8. CONCLUSIONS AND FUTURE RESEARCH

As stated in Chapter 1, this research work has been done to fulfill the following objectives:

- To design an architecture for a fixed wireless ATM transmitter with a focus on broadband applications.
- To derive a metric to estimate the required bandwidth of the ATM traffic sources in the wireless network scenario.
- To design an adaptive transmitter that adapts to the bandwidth requirements and the channel variation. An efficient use of the channel bandwidth was the overall goal for such a transmitter.
- To develop a suitable architecture for the transmitter.
- To develop a hardware implementation for the proposed modulator.
- To model and characterise the performance of the proposed transmitter for different types of ATM traffic and for different radio channels.

8.1 Conclusions

Based on the research work reported in the previous chapters, it can be stated that all these objectives were realized. The results are summarised as follows:

It is shown that for design of an optimum radio ATM transmitter, both
the broadband traffic and channel performance have to be monitored. Taking these requirements into account a new architecture for an adaptive transmitter has been proposed.

A statistical model was developed to estimate the required bandwidth of the broadband traffic. The model uses a nonlinear relation to estimate the required bandwidth as a function of cell loss ratio and maximum tolerable delay using a self-similar traffic model. This is followed by a study using empirical traces to investigate the required bandwidth of computer data and video traffic in an ATM network. Adequacy of the proposed effective bandwidth metric was established by comparison with the computer simulation results for different types of traffic. Moreover, the effective bandwidth allocation was compared with the peak rate allocation. It is found that a bandwidth saving close to 50 percent (depends on buffer size) can be achieved for aggregate MPEG-1 video traffic. In a LAN type traffic, depends on utilisation and buffer size a high bandwidth saving is also achievable.

A metric, called Modified Effective Bandwidth, has been introduced for the wireless network. It has been shown that the capacity requirement for such a network is a function of QoS, e.g., cell loss ratio and maximum delay; as well as channel characteristics, e.g., fading behaviour. A Capacity Reduction Factor that takes into account the channel conditions was introduced. Using this factor a mathematical equation for the modified effective bandwidth has been proposed as a QoS metric for wireless ATM.

The concept of modified effective bandwidth was applied to Ricean and Rayleigh channels and it was found that the modified effective bandwidth metric demands more bandwidth in Rayleigh channels compared with Ricean channels for similar traffic load. This is due to the nature of the Rayleigh
channel. An experiment with multiple video sources showed that a similar traffic load requires 12 percent more bandwidth in a Rayleigh channel compared to the Ricean channel with Rice parameter of $K = 6$.

An M-QAM modulator was proposed as an adaptive modulator for radio ATM. It was shown that the modified effective bandwidth can be used to adapt the M-QAM modulator to traffic characteristics as well as channel conditions thus resulting in an optimum transmitter.

An M-QAM control unit has been designed. This unit assigns a buffer size to each VP according to the delay constraints. By using the modified effective bandwidth metric, the required bandwidth for each VP was calculated in this unit and a modulation level was specified in the M-QAM modulator. It was shown that such a unit monitors the instantaneous bandwidth demand continuously while maintaining the outage conditions. Simulation studies showed that the call acceptance improvement is close to 20 percent compared to the average fixed QAM modulator. An outage improvement better than .02 was obtained compared to the highest level fixed QAM Modulator in all experiments.

A novel method to realize a QAM modulator has been introduced. The proposed method is suitable for high speed and high bandwidth operation.

A key issue in the design of the proposed modulator was a model for the PIN diode as a data controlled attenuator. An accurate model to characterise a PIN diode as a variable resistor controlled by high speed digital data and RF pump signals was developed. Using the residue theorem and complex inversion formula, an analytical model was derived. This model was
employed for computation of the switching time for PIN diodes. The theoretical results obtained using the proposed model were compared with the measurement results of the commercial PIN diodes. An excellent agreement was achieved. Expected performance of the QAM modulator was computed using circuit models and it was shown that the proposed method can be used to realize an adaptive modulator.

An implementation for the MIC version of the modulator was proposed using silicon diodes, microstrip and coplanar lines. A predistortion filter to provide a linear relation between the control voltage and reflection attenuator was proposed. The constellation study over the 500 MHz bandwidth shows an error vector magnitude better than 10 percent for the MIC version of the modulator.

A highly cost effective direct QAM modulator can be realized using Si-MMIC technology. Keeping this in view, subsystems were designed using Si-MMIC technology. Using these designs, a highly cost effective high bit rate adaptive M-QAM modulator for radio ATM can be realized. A comparison between the proposed direct QAM modulator and theoretical QAM modulator shows an excellent agreement. This new method to implement the QAM modulator has great potential.

Finally, a system study was carried out to examine the operation of adaptive direct microwave QAM modulators in the wireless channel for transmission of ATM traffic. The operations in Ricean and Rayleigh fading channels have been examined for computer data and video traffic. The call acceptance and outage characteristics were compared with those of the fixed QAM modulator. The results show that the proposed adaptive method can efficiently meet the expected high demand for a low cost transmitter for broadband wireless applications. Some results of this research work have been patented.
[7], published [2], [5], [6], [8], [10], [121], will publish [12], and submitted to [3], [9], [11] various journals and conferences.
8.2 Contributions of this Research Work

- An adaptive direct microwave M-QAM transmitter for wireless ATM applications has been designed. The transmitter uses an M-QAM modulator capacity controlled by modified effective bandwidth. The modulation control unit estimates the required bandwidth of traffic by considering traffic densities and channel performance.

- A new metric, called Modified Effective Bandwidth, has been introduced to assign the actual bandwidth to an ATM connection in a wireless channel and an analytical relation has been derived to calculate this metric.

- A new factor, called Capacity Reduction Factor for M-QAM, has been introduced in Rice and Rayleigh fading channels and an analytical relation is derived to calculate this factor in various fading channels.

- A new closed form relation based on residue theory has been derived to study the operation of PIN diodes. The relation is the only available relation that accurately estimates the transition time of the PIN diode when its bias is changing from reverse bias to forward bias.

- A new architecture for the direct microwave QAM modulator has been introduced that results in a high performance and low cost implementation of the QAM modulator.

- A microwave integrated circuit has been implemented for the direct microwave QAM modulator that is scalable in frequency.

- A Si-MMIC has been implemented for the direct microwave QAM modulator which results in a simple, low cost, and high performance modulator for broadband applications.
8.3 Future Work

The fabrication of the hardware and field trials in fixed wireless ATM deployment is the next step. The direct microwave QAM modulator introduced in this research work can be realized using MIC and Si-MMIC technologies. A power oscillator can be designed on the same MMIC chip as the modulator using an IMPATT diode oscillator that can be easily implemented in Si-MMIC technology. Thus, the complete transmitter can be implemented on a single MMIC chip. Meanwhile, the other architectures to realize a QAM modulator as well as Si-MMIC implementation issues should be studied. The design reported in this thesis should be examined at different microwave and millimeter-wave frequencies.

The proposed system can be developed to operate with various access techniques, e.g., FDMA and TDMA. The results of this method can be extended to other high speed services such as the new millimeter wave band local multipoint distribution services (LMDS).
REFERENCES


[8] S. Kumar, A. Mohammadi, and D. Klymyshyn, “A direct microwave frequency modulation for a 155.52 Mbps 17.7 to 19.7 GHz radio for ATM


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A. CAPACITY REDUCTION FACTOR IN M-QAM IN RAYLEIGH CHANNEL

The capacity reduction factor in a Rayleigh channel for M-QAM modulator is defined by Eq.(4.18) as follows:

\[ \xi_{MQAM}^c = \frac{\int_\gamma \log_2(1 - 3\gamma[2ln(5p_b)]^{-1})p(\gamma)d\gamma}{\log_2(1 - 3\gamma[2ln(5p_b)]^{-1})}. \]  

(A.1)

where \( p(\gamma) \) is the probability density function of the received CNR in Rayleigh channel and is obtained as follows:

\[ p(\gamma) = \frac{1}{\Gamma}e^{-\frac{\gamma}{\Gamma}}, \]  

(A.2)

where \( \Gamma \) is the average power of \( \gamma \). The numerator of Eq.(A.1) can be written as:

\[ I_{MQAM} = \int_\gamma \log_2(1 - \frac{3\gamma}{2ln(5p_b)})p(\gamma)d\gamma. \]  

(A.3)

Replacing Eq.(A.2) in Eq.(A.1) results in

\[ I_{MQAM} = \frac{\log_5}{\Gamma} \int_\gamma ln(1 - \frac{3\gamma}{2ln(5p_b)})e^{-\frac{\gamma}{\Gamma}}d\gamma. \]  

(A.4)

This integral can be solved as follows [120]:

\[ I_{MQAM} = -\log_5exp(-\frac{2ln(5p_b)}{3\Gamma})E_i(\frac{2ln(5p_b)}{3\Gamma}), \]  

(A.5)

where

\[ E_i(x) = E + ln(-x) + \sum_{k=1}^{\infty} \frac{x^k}{k.k!}, \]  

(A.6)
and $E=577215$ is the Euler constant. The Eq.(A.5) can be written as:

$$I_{MQAM} = -\log_2 \exp(-\frac{2\ln(5p_b)}{3\Gamma})(E + \ln(-\frac{2\ln(5p_b)}{3\Gamma}) + \frac{2\ln(5p_b)}{3\Gamma} + \frac{2\ln(5p_b)^2}{3\Gamma})/2.2! + ...$$

(A.7)

As can be seen, the higher order terms can be ignored if $\Gamma$ is big enough.

Then, Eq.(A.7) can be approximated as follows:

$$I_{MQAM} = -\log_2 \exp(-\frac{2\ln(5p_b)}{3\Gamma})(E + \ln(-\frac{2\ln(5p_b)}{3\Gamma}) + \frac{2\ln(5p_b)}{3\Gamma}).$$

(A.8)

Using Eq.(A.8), the capacity reduction factor in Rayleigh fading using M-QAM modulation can be obtained as:

$$\xi_{MQAM} = \frac{-\exp(-\frac{2\ln(5p_b)}{3\Gamma})(E + \ln(-\frac{2\ln(5p_b)}{3\Gamma}) + \frac{2\ln(5p_b)}{3\Gamma})}{\ln(1 - 3\gamma[2\ln(5p_b)]^{-1})}.$$  

(A.9)
B. PHASE SHIFT COMPENSATION IN A REFLECTION ATTENUATOR

A reflection attenuator is shown in Fig.B.1. As may be seen, this attenuator uses a quadrature hybrid and PIN diode terminations. A general equivalent circuit for a PIN diode was presented in Chapter 6. Using Eq.(6.2), the phase characteristics of the reflection attenuator may be obtained as:

\[
\phi(R_j) = \tan^{-1}\left(\frac{\omega L_s - \frac{\omega R_j^2 C_j}{1 + \omega^2 R_j^2 C_j^2}}{R_s - Z_o + \frac{R_j}{1 + \omega^2 R_j^2 C_j^2}}\right) - \tan^{-1}\left(\frac{\omega L_s - \frac{\omega R_j^2 C_j}{1 + \omega^2 R_j^2 C_j^2}}{R_s + Z_o + \frac{R_j}{1 + \omega^2 R_j^2 C_j^2}}\right)
\]

(B.1)

As may be seen in Eq.(B.1), the phase characteristic varies with the variation of \(R_j\). This can be attributed to the effect of the parasitic elements \(C_j, L_s\). This effect can be compensated using an open stub circuit [121]. Fig.B.2 shows such a phase compensation method using an open stub in the PIN diode circuit [122]. The impedance of an open stub is given by:

\[
Z_{oc} = \frac{Z_o}{j\tan(\theta)}
\]

(B.2)

Thus, the attenuator phase shift for attenuator with an open circuit stub is modified as follows:

\[
\phi(R_j) = \tan^{-1}\left(\frac{\omega L_s - \frac{\omega R_j^2 C_j}{1 + \omega^2 R_j^2 C_j^2} - \frac{Z_o}{\tan(\theta)}}{R_s - Z_o + \frac{R_j}{1 + \omega^2 R_j^2 C_j^2}}\right) - \tan^{-1}\left(\frac{\omega L_s - \frac{\omega R_j^2 C_j}{1 + \omega^2 R_j^2 C_j^2} - \frac{Z_o}{\tan(\theta)}}{R_s + Z_o + \frac{R_j}{1 + \omega^2 R_j^2 C_j^2}}\right)
\]

(B.3)

The stub length, \(\theta\), can be obtained by solving this equation with \(\phi(50)\) set to \(\phi(0)\).
Figure B.1 Reflection attenuator and equivalent circuit for the PIN diode termination.

Figure B.2 PIN diode equivalent circuit with an open circuited stub in series
C. BIAS CIRCUIT FOR Si-MMIC DIRECT MICROWAVE QAM MODULATOR

The Si-MMIC implementation of the various subsystems for direct microwave QAM modulator was described in Chapter 6. The SPDT switches and variable attenuators in the direct microwave QAM modulator circuit use PIN diodes. The PIN diode has to be suitably biased. This bias circuit must apply the high speed data to the SPDT switches and variable attenuators. This circuit can be realized by a low pass filter. The DC isolation between subsystems can be provided using coupling capacitors.

C.1 Lowpass Filter

The filter design depends on the data bandwidth and carrier frequency of the modulator. A second order Butterworth lowpass filter with cutoff frequency $f_c = 3.6$ GHz can provide the desired frequency characteristics. The required bandwidth of high speed data is less than 100 MHz and carrier frequency is about 18 GHz. The filter and its frequency characteristic are shown in Fig.C.1.

C.1.1 Lowpass Filter Implementation on Si-MMIC Substrate

To implement the lowpass filter on Si-MMIC substrate, a spiral inductor may be used [123]. The dimensions of the inductor are shown in Fig.C.2.a. This spiral inductor provides a 3.15 nH inductance in desired bandwidth. This inductance value can be easily realized on Si-MMIC technology [124]. A 1.26 pF capacitance was used in the lowpass filter.
Figure C.1  a) A lowpass filter as a bias circuit for the PIN diode b) Bias circuit frequency response

A metal-insulator-metal (MIM) capacitor in Si-MMIC can use $SiO_2$ as dielectric. Such a MIM capacitor is shown in Fig.C.2.b. A dielectric with thickness $d=2000\ A^o$ provides a capacitance $0.199\ fF/\mu m^2$. Thus, a 1.26 pF capacitor requires metal electrodes with dimensions $80 \times 80\ \mu m^2$. The lowpass filter, shown in Fig.C.3, is realized using MIM capacitor and spiral inductor. The frequency response has been compared with the ideal filter in Fig.C.1. A capacitor $C = 2.95\ pF$ may be used as a coupling capacitor to provide the DC isolation. Using an MIM capacitor on silicon substrate, a capacitor with dimensions $122 \times 122\ \mu m^2$ should provide the desired capacitance. The SPDT switch and variable attenuator circuits with their bias circuit are shown in Fig.C.4 and Fig.C.5.
Figure C.2 A MIM capacitor and a spiral inductor on Silicon substrate

Figure C.3 Lowpass filter bias circuit implemented using Si-MMIC technology
Figure C.4  The attenuator layout along with the bias circuit

Figure C.5  The SPDT layout with the bias circuit