Radio over fiber system for wireless LAN

Gurprakash Singh Sandhu

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Radio over Fiber System for Wireless LAN

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School of Electrical & Electronic Engineering

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Abstract

Radio-over-Fiber (RoF) technology has been identified as a technique to increase the capacity of wireless data networks. The opportunity to install vast networks coupled with the possibility of designing simple and cost-effective base stations has gained extremely good reviews for RoF technology. This thesis investigates the use of a RoF network to transport IEEE802.11a Wireless LAN signals.

Orthogonal Frequency Division Multiplexing (OFDM) is rapidly being adopted as a modulation technique for broadband wireless networks and forms the basis of the 802.11a standard. The work done in this project started with a simulation based study of OFDM with emphasis placed on the salient features of OFDM modulation that give it a performance edge over other modulation techniques such as its immunity to multipath fading. This further led to the development of a complete 802.11a physical link simulation model to investigate the performance of 802.11a compliant OFDM transmission over a multipath channel.

Finally individual optical link components were modeled and integrated to simulate a RoF link used to transport 802.11a WLAN signals. The simulations conducted were aimed towards studying the performance of the RoF link with respect to the power level requirements of optical modulation and demodulation for acceptable transmission Bit-Error-Rates.

The research and development done in this project has led to the development of a complete simulation model for a RoF based WLAN network, that was used to simulate the transmission of 802.11a Wireless LAN signals under various wireless and optical channel conditions.

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List of Acronyms

RoF	Radio-over-Fiber
IEEE	Institute of Electrical and Electronics Engineers
SNR	Signal-to-Noise Ratio
CNR	Carrier-to-Noise Ratio
BER	Bit Error Rate
CS	Central Sites
RS	Remote Sites
MSC	Mobile Switching Centre
BS	Base Station
RAP	Radio Access Point
IF	Intermediate Frequency
WLAN	Wireless Local Area Network
SMF	Single Mode Fiber
WDM	Wavelength Division Multiplexing
DWDM	Dense Wavelength Division Multiplexing
OFDM	Orthogonal Frequency Division Multiplexing
ISI	Inter-symbol Interference
ICI	Inter-carrier Interference
PAPR	Peak-to-Average Power Ratio
PSK	Phase Shift Keying
QAM	Quadrature Amplitude Modulation
AWGN	Additive White Gaussian Noise
QoS	Quality of Service
GSM	Global System for Mobile Communications
UMTS	Universal Mobile Telecommunications System
FDDI	Fiber Distributed Data Interface
ATM	Asynchronous Transfer Mode
ESCON	Enterprise Systems Connection

MVDS	Multipoint Video Distribution Services
MBS	Mobile Broadband System
B-ISDN	Broadband Integrated Services Digital Network
ITS	Intelligent Transport Systems
RVC	Road-to-Vehicle Communication
IVC	Inter-Vehicle Communication
WBMCS	Wireless Broad-band Mobile Communication Systems
U-NII	Unlicensed National Information Infrastructure
ISM	Industrial, Scientific and Medical
NA	Numerical Aperture
LED	Light-emitting Diode
LD	Laser Diode
AM	Amplitude Modulation
PM	Phase Modulation
FM	Frequency Modulation
ASK	Amplitude Shift Keying
FSK	Frequency Shift Keying
OOK	On-off Keying
MZM	Mach Zehnder Modulator
EAM	Electro-Absorption Modulator
SONET	Synchronous Optical Network
SDH	Synchronous Digital Hierarchy
EDFA	Erbium-doped Fiber Amplifier
IM-DD	Intensity Modulation – Direct Detection
RHD	Remote Heterodyne Detection
DIM	Direct Intensity Modulation
CW	Continuous Wave
SOA	Semiconductor Optical Amplifier
OADM	Optical Add-Drop Multiplexers
SCM	Sub-carrier Multiplexing
FHSS	Frequency-Hopping Spread Spectrum

DSSS	Direct-Sequence Spread-Spectrum
DFT	Discrete Fourier Transform
IFFT	Inverse Fast Fourier Transform
FFT	Fast Fourier Transform
OSI	Open Systems Interconnection
MAC	Medium Access Control
COFDM	Coded Orthogonal Frequency Division Multiplexing
MCM	Multi-Carrier Modulation
DMT	Discrete Multi-tone
CP	Cyclic Prefix
MIMO	Multiple-input Multiple-output
CCI	Co-channel Interference
RMS	Root-Mean-Square
FEC	Forward Error Correction
PAN	Personal Area Network
CCK	Complementary Code Keying
PMD	Physical Medium Dependent
PLCP	Physical Layer Convergence Procedure
PLME	Physical Layer Management Entity
ARQ	Automatic Repeat Request
ML	Maximum-Likelihood
PSDU	Physical Service Data Unit
PPDU	Physical Protocol Data Unit
E/O	Electro-optic
O/E	Opto-electric
RIN	Relative Intensity Noise
PD	Photodetector
NF	Noise Figure
DR	Dynamic Range
DPSSL	Diode-pumped Solid-state Laser
DFB	Distributed-Feedback

FP	Fabry-Perot
IM	Intermodulation
ASE	Amplified Spontaneous Emission
CSMA/CA	Carrier-Sense Multiple Access with Collision Avoidance
HPA	High Power Amplifier

1. Introduction

A revolution in electronic communication has been rapidly transforming both our society and our expectations for the future. Communications innovation is starting to free men and women from their offices and is creating more choices as to where, and when, they choose to work. Telecommuting is now an every day reality for millions of people around the world. Employers are welcoming the trend since providing office space for every employee can represent an immensely high overhead cost. As a significant bonus, the Internet, video-conferencing and on-line information systems will radically reduce the need for business travel, thus reducing road congestion, pollution and some of the stresses of commuting. Rapid advances in telecommunications also offer the prospect of a massive, and relatively inexpensive, expansion of higher education. Universities everywhere are experimenting with distance-learning, usually via the Internet. There are now educational institutions with no physical presence, but rather existing as "virtual campuses".

Wireless Communication forms the backbone of this communications revolution in today's society. The proliferation of mobile and other wireless devices coupled with increased demand for broadband services are putting pressure on wireless systems to increase capacity. To achieve this, wireless systems must have increased feeder network capacity, operate at higher carrier frequencies, and cope with increased user population densities. However, raising the carrier frequency and thus reducing the radio cell size leads to costly radio systems.

Radio-over-Fiber (RoF) technology has emerged as a cost effective approach for reducing radio system costs because it simplifies the remote antenna sites and enhances the sharing of expensive radio equipment located at appropriately sited (e.g. centrally located) Switching Centers (SC) or otherwise known as Central Sites/Stations (CS).

1.1 Motivation

In the domain of telecommunications, two fields that have witnessed tremendous amounts of research activity and development over the past decade are wireless communications and fiber optic communications. The benefits of these has already been felt at our homes and offices, and with more innovations being introduced in the global communications infrastructure at an astonishing pace, both wireless and optical technologies are revolutionizing the industry and will undoubtedly dominate its future.

Wireless communications are driven by the desire for change from wired fixed place-to-place communications to wireless mobile person-to-person communications, and for the ability to share information around the world with anyone, anywhere, at any time. Wireless applications include cellular systems, wireless local area networks (LAN), home wireless networking, and audio/video broadcasting. Wireless communication systems can be classified based on the target data rate, achievable transmission range, and constraint on the power use. Wireless LAN systems development has always centered on achieving high data transmission rates over limited transmission distances. As WLAN systems became more widespread, cost-effectiveness also became a major consideration during network installation and utilization. One promising alternative to this issue is a Radio over Fiber based WLAN network since in this network functionally simple and cost-effective base stations are utilized in contrast to conventional wireless systems.

Fiber optics has been utilized since the late 1970s to provide long-range high-speed transmission at low cost. By the early 1980s, single-mode fiber (SMF) operating in the 1310 nm and later the 1550 nm wavelength windows became the standard fiber installed for these networks. Today, computers, information networks, and data communications also embrace fiber optic transmission. Particularly, due to the successful development of dense wavelength division multiplexing (DWDM) technology, fiber transmission capacity has grown by a factor of 100 in the last decade. In 1990, Bell Labs transmitted a 2.5 Gb/s signal over 7,500 km without regeneration. In 1998, Bell Labs researchers transmitted 100 simultaneous optical signals, each at a data rate of 10 Gb/s for a distance

of nearly 250 miles (400 km). This increased the total data rate on one fiber to one Tb/s [2].

Wireless communications as well as optical transmission provide many technological challenges, due to channel impairments, bandwidth limitations, energy limitations, and delay constraints. Multipath signal propagation refers to the phenomenon where the incoming radio waves arrive from different directions with different propagation delays. The signal received by a receiver at any point in space may consist of a large number of plane waves having randomly distributed amplitudes, phases, and angles of arrival. These multipath components combine vectorially at the receiver antenna, and can cause the signal received to distort or fade. Even when a receiver is stationary, the received signal may fade due to movement of surrounding objects in the radio channel. Multipath generates inter-symbol interference (ISI) in the received symbols and therefore poses significant challenges in the development of wireless systems.

To combat the multipath fading channels, a special multi-carrier modulation, namely orthogonal frequency division multiplexing (OFDM), has been proposed as an effective modulation scheme. It has recently been adopted as standard for high-speed wireless LANs, video/audio broadcasting, and 4G cellular systems. During the course of this research, a substantial amount of time was allocated to study OFDM with respect to its Multipath immunity, BER performance and high Peak-to-Average Power Ratio (PAPR) and the results and analysis are presented for the same.

As mentioned before, the major advantage in using Radio over Fiber systems for WLANs is from the system cost perspective. The fact that future wireless LANs will consist of a high density of small radio cells makes the issue of system cost a major one. It is imperative to have simple and easy to maintain base stations [3]. The complexity of the BSs is related to the RoF techniques employed. Therefore, the choice of the microwave generation method is important. The kind of feeder network infrastructure is another crucial one. While standard single-mode fiber offers the most bandwidth, it has high installation and maintenance costs associated with it [7]. The required RoF system must

not only meet the present demand for capacity, but it must also be suited to meet anticipated future bandwidth demands. The requirement for broadband wireless services translates into the requirement for increased capacity in the distribution and feeder networks. As stated above, optical fibre is the best candidate for this. To study the above mentioned and other system level performance requirements, it was decided that a complete system model for WLANs using RoF technology needs to be developed. Since OFDM is seen as the modulation/multiplexing technique under the IEEE 802.11a standard, the RoF system model to be designed must be capable of distributing Phase Shift Keying (PSK) and Quadrature Amplitude Modulation (QAM) signals as defined in the OFDM standard.

1.2 Scope

The work done in this research is to study a cost-effective RoF system employing single-mode optical fibers to distribute high frequency and high-speed Wireless LAN signals in conformance with the latest Industry Standards of IEEE 802.11a. The delivered radio signals must be appropriately modulated in addition to meeting other radio-system requirements.

The first major objective of the research is to investigate the use of OFDM modulation scheme for WLAN architectures as proposed in the IEEE 802.11a standard. This would require comprehensive computer simulation of OFDM modulation technique for both PSK modulation as well as QAM modulation. Three main performance criteria would be tested, which are OFDM's tolerance to multipath delay spread, channel noise and peak power clipping.

A simulation model would be developed for the 802.11a physical layer that supports all modulation techniques, data rates and corrective coding schemes stated by the standard. This would be used for studying BER performance of OFDM code transmission through an AWGN + Multipath channel. This model would be used to study the error susceptibility of PSK and QAM modulation as the transmission data rate increases to the benchmarks set by IEEE 802.11a.

Finally a model would have to be developed that combines the 802.11a PHY layer model and the proposed RoF system. Classification and modeling of proposed optical interfaces would have to be done for the successful generation of such a model. Other parameters to be modeled in this simulation would be the RoF link characteristics, requirements, RF signal generation and transportation techniques.

1.3 Radio-over-Fiber Architecture

The deployment of optical fiber technology in wireless networks provides great potential for increasing the capacity and QoS without largely occupying additional radio spectrum. By using Radio-over-Fiber (RoF) technology, the capacity of optical networks can be combined with the flexibility and mobility of wireless access networks. The Radio-over-Fiber concept refers to the transportation of information over optical fiber by modulating the light with the radio signal either at the carrier frequency or at an intermediate frequency. This technique can be used in the backbone of the wireless access networks.

RoF is an *analog* optical link transmitting modulated RF signals. It serves to transmit the RF signals down- and uplink, i.e. to and from Central Sites (CS) to Remote Sites (RS) via an optical fiber link or network. If the application area is in a GSM network, then the CS could be the Mobile Switching Centre (MSC) and the RS could be the Base Station (BS). For Wireless Local Area Networks (WLANs), the CS would be the headend while the Radio Access Point (RAP) would act as the RS.

1.3.1 Salient Features of Radio-over-Fiber Architecture

The frequencies of the radio signals distributed by RoF systems span a wide range (usually in the GHz region) and depend on the nature of the applications. Pioneer RoF systems were primarily used to transport microwave signals, and to achieve mobility functions in the CS. That is, modulated microwave signals had to be available at the input end of the RoF system, which subsequently transported them over a distance to the RS in the form of optical signals. At the RS the microwave signals are re-generated and radiated by antennas. The added value in using such a system lay in the capability to dynamically allocate capacity based on traffic demands. RoF systems of nowadays, are designed to perform added radio-system functionalities besides transportation and

mobility functions. These functions include data modulation, signal processing, and frequency conversion (up and down).

For a multifunctional RoF system, the required electrical signal at the input of the RoF system depends on the RoF technology and the functionality desired. The electrical signal may be baseband data, modulated IF, or the actual modulated RF signal to be distributed. The electrical signal is used to modulate the optical source. The resulting optical signal is then carried over the optical fiber link to the remote station. Here, the data is converted back into electrical form by the photodetector. The generated electrical signal must meet the specifications required by the wireless application be it GSM, UMTS, wireless LAN or other.

By delivering the radio signals directly, the optical fiber link avoids the necessity to generate high frequency radio carriers at the antenna site. Since antenna sites are usually remote from easy access, there is a lot to gain from such an arrangement [1]. However, the main advantage of RoF systems is the ability to concentrate most of the expensive, high frequency equipment at a centralized location, thereby making it possible to use simpler remote sites. Furthermore, RoF technology enables the centralizing of mobility functions such as macro-diversity for seamless handover.

1.3.2 Comparison of Fiber-Optic Wiring and Coaxial Cables

The use of Fiberoptic cables enables Radio-over-Fiber technology to offer a wide range of benefits as compared to RF coaxial cables. A fiber optic cable with the same bandwidth as a comparable coaxial cable is less than 2% in both the size and weight. Apart from this characteristic, optical fibers provide a variety of advantages like low attenuation and insensitivity to electronic interferences. Fibers do not use electric current or have interference. They use less energy, maintenance costs are lower and are easier to install. CATV is also adopting fiber optics. Only fiber has the necessary bandwidth for carrying voice, data and video simultaneously. Fiber is a secure medium, immune from tapping, because if the signal is tapped, light loss is unavoidable, and the connection is shut down.

A currently installed coaxial cable system with an expected lifespan of eight years will have to be re-cabled after only 3-4 years due to the increased bandwidth requirements. Fiber optic cable, on the other hand, is currently ready to accept protocols and interfaces such as FDDI (Fiber Distributed Data Interface), ATM (Asynchronous Transfer Mode), ESCON (Enterprise Systems Connection), Fiber Channel and perhaps any additional mode that may arrive or be proposed in the near future. Because of these capabilities it is believed that the lifespan of a fiber based LAN with multimode fiber goes up to thirty years. Table 1.1 [7] below provides a comparison between Fiberoptic cables and coaxial cables over the major points of concern for wired communication systems.

Fiber	Coaxial
1. Does not radiate Radio Frequency nor is susceptible to Interference.	1. Radiates signals capable of interfering (RFI) with other electronic equipment.
2. Fiber Optic networks operate at speeds up to 2.5 Gbps.	2. Copper wire networks operate at speeds up to 55 Mbps.
3. Fibers have much higher bandwidth. Single mode > 100GHz for 100m.	3. Bandwidth is lower. 100m – 100MHz.
4. Signals can be transmitted over much longer distances > 40,000m for single mode.	4. Signal attenuation and distortion is much higher hence distances are about 100m.
5. Cost is up to 100 times more, but maintenance is very cheap.	5. Cost is low, but maintenance is expensive.

Table 1.1: Fiber-Optic and Coaxial cable: step-by-step comparison

1.3.3 Applications of Radio-over-Fiber Technologies

Some of the applications of RoF technology include satellite communications, mobile radio communications, broadband access radio, Multipoint Video Distribution Services (MVDS), Mobile Broadband System (MBS), vehicle communications and control, and wireless LANs over optical networks. The main application areas are briefly discussed below.

Cellular Networks: The field of mobile networks is an important application area of RoF technology. The ever-rising number of mobile subscribers coupled with the increasing demand for broadband services have kept sustained pressure on mobile networks to offer increased capacity. Therefore, mobile traffic (GSM or UMTS) can be relayed cost effectively between the SCs and the BSs by exploiting the benefits of SMF technology. Other RoF functionalities such as dynamic capacity allocation offer significant operational benefits to cellular networks.

Satellite Communications: Satellite communications was one of the first practical uses of RoF technology. One of the applications involves the remoting of antennas to suitable locations at satellite earth stations. In this case, small optical fiber links of less than 1km and operating at frequencies between 1GHz and 15GHz are used. By so doing, high frequency equipment can be centralized. The second application involves the remoting of earth stations themselves. With the use of RoF technology the antennae need not be within the control area (e.g. Switching Centre). They can be sited many kilometers away for the purpose of improved satellite visibility or reduction in interference from other terrestrial systems. Switching equipment may also be appropriately sited, for say environmental or accessibility reasons or reasons relating to the cost of premises, without requiring being in the vicinity of the earth station antennas.

Video Distribution Systems: One of the major promising application areas of RoF systems is video distribution. A case in point is the Multipoint Video Distribution Services (MVDS). MVDS is a cellular terrestrial transmission system for video (TV) broadcast. It was originally meant to be a transmit-only service but recently, a small return channel has been incorporated in order to make the service interactive. MVDS can be used to serve areas the size of a small town. Allocated frequencies for this service are in the 40 GHz band. At these frequencies, the maximum cell size is about 5km. To extend coverage, relay stations are required [13].

Mobile Broadband Services: The Mobile Broadband System or Service (MBS) concept is intended to extend the services available in fixed Broadband Integrated Services Digital

Network (B-ISDN) to mobile users of all kinds. Future services that might evolve on the B-ISDN networks must also be supported on the MBS system. Since very high bit rates of about 155 Mbps per user must be supported, carrier frequencies are pushed into mm-waves. Therefore, frequency bands in the 60 GHz band have been allocated. The size of cells is in diameters of hundreds of meters (micro-cells). Therefore, a high density of radio cells is required in order to achieve the desired coverage. The micro-cells could be connected to the fixed B-ISDN networks by optical fiber links. If RoF technology is used to generate the mm-waves, the base stations would be made simpler and therefore of low cost, thereby making full scale deployment of MBS networks economically feasible [22].

Wireless LANs: As portable devices and computers become more and more powerful as well as widespread, the demand for mobile broadband access to LANs will also be on the increase. This will lead once again, to higher carrier frequencies in the bid to meet the demand for capacity. For instance, presently wireless LANs operating at the 2.4 GHz ISM bands (IEEE 802.11b) offer the maximum capacity of 11 Mbps per carrier. IEEE802.11a broadband wireless LANs are primed to offer up to 54 Mbps per carrier, and will require higher carrier frequencies in the 5 GHz band [5]. Higher carrier frequencies in turn lead to micro- and pico-cells, and all the difficulties associated with coverage discussed above arise. A cost effective way around this problem is to deploy RoF technology.

Vehicle Communication and Control: This is another potential application area of RoF technology. Frequencies between 63-64 GHz and 76-77 GHz have already been allocated for this service within Europe. The objective is to provide continuous mobile communication coverage on major roads for the purpose of Intelligent Transport Systems (ITS) such as Road-to-Vehicle Communication (RVC) and Inter-Vehicle Communication (IVC). ITS systems aim to provide traffic information, improve transportation efficiency, reduce burden on drivers, and contribute to the improvement of the environment [8]. In order to achieve the required (extended) coverage of the road network, numerous base stations are required. These can be made simple and of low cost by feeding them through RoF systems, thereby making the complete system cost effective and manageable.

1.4 Wireless LAN Standard: IEEE 802.11a

1.4.1 Frequency Bands and Parameters

The IEEE 802.11a standard is proposed for a range of data rates from 6 to 54 Mbps using the OFDM modulation technique in the 5 GHz band. The 5 GHz band is specifically designed for “Wireless Broad-band Mobile Communication Systems (WBMCS)”. The parameters of the IEEE 802.11a standard are given in Table 1.2 below [5].

Parameter	IEEE 802.11a Standard
Configurations	Centralized system with access points, or peer-to-peer networking.
Range	Up to 30m at 24Mbps and up to 60m at 6Mbps with omni-directional antennas.
Channel Access	CSMA (Carrier Sense Multiple Access) CA (Carrier Access), variable size data packets (up to 8192 bytes).
Frequency Bands	5.15 – 5.35 GHz; 5.725 – 5.825 GHz.
Duplexing	TDD (Time Division Duplexing)
Data Rate	6,9 Mbps (BPSK) 12, 18 Mbps (QPSK) 24,36 Mbps (16-QAM) 54 Mbps (64-QAM).

Table 1.2: Parameters of IEEE 802.11a Standard.

1.4.2 Services and Applications

The IEEE 802.11 WLAN standard is mainly intended for communications between computers, thus being an extension of wired LANS; nevertheless, it can support real-time voice and image signals, and users are allowed some mobility and can have access to public networks. The goal of the IEEE 802.11 standard is to provide for wireless connectivity between fixed and portable units within a local area and to offer a standard for use by regulatory bodies to standardize access to one or more frequency bands for the purpose of local area communications. For customers, the benefit will be a standard that will allow compatibility among multiple vendors’ products. IEEE 802.11a mainly

supports higher data rate transmission for wireless broadband mobile communication systems (WBMCS).

1.4.3 Very-High-Speed and High Scalability features

The IEEE 802.11a WLAN standard is superior compared with current technologies because of its greater scalability, better interference immunity and significantly higher speed, and at the same time allowing for higher bandwidth applications and more users. The 802.11a standard utilizes 300 MHz of bandwidth in the 5 GHz Unlicensed National Information Infrastructure (U-NII) band. The frequency range used currently for most enterprise-class unlicensed transmissions, including 802.11b, is the 2.4 GHz Industrial, Scientific and Medical (ISM) band. This highly crowded band offers only 83 MHz of spectrum for all wireless traffic, including cordless phones, building-to-building transmissions, and microwave ovens. In comparison, the 300 MHz spectrum offered in the U-NII band represents a nearly four-fold increase in spectrum; this is more impressive considering that there is limited wireless traffic in the band today.

The 802.11a standard uses Orthogonal Frequency Division Multiplexing (OFDM), a new encoding scheme that offers benefits over spread spectrum in channel availability and data rate. Channel availability is significant because the more independent channels that are available, the more scalable the wireless network becomes. The high data rate is accomplished by combining many lower-speed subcarriers to create one high-speed channel. At speeds of 54 Mbps and greater, 802.11a is faster than any other unlicensed solution. The 802.11a and 802.11b standards both have a similar range, but the 802.11a standard provides a higher data rate throughout the entire coverage area. The 5 GHz band in which it operates is not highly populated, so that there are fewer users to cause interference or signal contention. In addition, the 8 non-overlapping channels in this band allow for a highly scalable and flexible installation. The 802.11a standard is the most reliable and efficient medium by which to accommodate high-bandwidth applications for multiple users.

1.5 Organization of Thesis

Following this introduction, the Thesis is organized into a further five chapters. A brief description of the flow of information in the different chapters is given below:

Chapter 2: This chapter discusses the background information that was useful in understanding the communication principles that form the basis of this research. Firstly it presents the elements of a basic fiber-optic communication system and their operating principles, followed by a brief description of optical modulation of microwave signals. The chapter also summarizes the IEEE 802.11a standard and the OFDM modulation technique that are utilized for the generation and distribution of the WLAN signals.

Chapter 3: Since a basic understanding of OFDM is necessary to fully comprehend the research performed, chapter 3 details the fundamentals of OFDM generation and transmission. The chapter concludes with a description of the design and implementation of the Matlab-based simulation model created to study a few important performance measures for OFDM and also presents the simulation results.

Chapter 4: In chapter 4, emphasis is placed on the simulation modeling of the 802.11a physical layer transmission. A detailed description of the physical layer specifications is provided, which is utilized to create an end-to-end physical link simulation model. Lastly the simulation results are presented for the BER performance of the 802.11a link.

Chapter 5: This chapter offers an overview of a RoF-based microcellular WLAN system. The fundamental principles of RoF are presented for readers who may be unfamiliar with the technology. Each major component in a RoF link is briefly discussed, with an emphasis on the mathematical modeling of its operation for the purpose of simulation modeling of a RoF link. The RoF simulation model is presented next along with the simulation results that help determine the effectiveness of the RoF techniques proposed.

Chapter 6: Chapter 6 provides a summary of the major results and conclusions in this thesis. Some future research directions that may follow from this work are also discussed.

2. Background

A basic Radio-over-Fiber WLAN system consists of an integrated optical and wireless infrastructure capable of delivering broadband multimedia traffic to subscribers located in a pico-cell architecture (radius < 100m). In such a scheme, the fiber is used to route the broadband modulated optical signals to base stations where the RF signals are detected and transmitted to client stations.

Figure 2.1: Radio-over-Fiber for Wireless LAN

The use of RF over fiber allows a significant reduction in the complexity and costs of remote base stations. It also provides an inexpensive method for system upgrades, since

most of the signal processing functions would be done at the central office and not at each individual base station. Basic system architecture for a Radio-over-Fiber system for Wireless LAN is depicted in Figure 2.1. The use of fiber optics to transport digital signals is quite common. However, the transmission of analog RF signals had been limited by the linearity constraints in modulating/demodulating devices, and by the distortion effects created by the optical link [20]. Advances in fiber optic technology now allow modulating laser devices with RF signals beyond 10 GHz. Utilizing optical devices that operate at high frequency to carry WLAN 802.11a signals could enable delivery of broadband multimedia traffic to subscribers in many wireless scenarios. WLAN 802.11a systems can provide very high-speed Internet access (up to 54 Mbps) for indoor environment such as public buildings, shopping malls, and airports.

The advantages of 802.11a over wired LAN includes fast flexible radio deployment. Combining fiber distribution and a WLAN multiplies the capacity of the system by a large factor, and also solves several problems related to deploying WLANs in outdoor environments in economical and flexible ways. The maintenance and upgrades of the system are simplified since the processing units for many cells are gathered under one roof.

2.1 Optical Transmission

In this section we briefly discuss a general optical transmission link which is depicted in Figure 2.2 below. An optical transmission link consists of a few basic elements which are discussed below to give a better understanding of optical elements before proceeding with modeling a suitable optical network for our research.

Figure 2.2: Optical Transmission Link

2.1.1 Optical Fiber

Optical fiber is a dielectric medium for carrying information from one point to another in the form of light. Unlike the copper form of transmission, the optical fiber is not electrical in nature. To be more specific, fiber is essentially a thin filament of glass that acts as a waveguide. A waveguide is a physical medium or path that allows the propagation of electromagnetic waves, such as light. Due to the physical phenomenon of total internal reflection, light can propagate following the length of a fiber with little loss. Optical fiber has two low-attenuation regions. Centered at approximately 1300 nm is a range of 200 nm in which attenuation is less than 0.5 dB/km. The total bandwidth in this region is about 25 THz. Centered at 1550 nm is a region of similar size with attenuation as low as 0.3 dB/km. Combined, these two regions provide a theoretical upper bound of 50 THz of bandwidth [21]. By using these large low-attenuation areas for data transmission, the signal loss for a set of one or more wavelengths can be made very small, thus reducing the number of amplifiers and repeaters actually needed. In single channel long-distance experiments, optical signals have been sent over hundreds of kilometers without amplification. Besides its enormous bandwidth and low attenuation, fiber also offers lower error rates compared to wired networks.

Transmission in Optical Fiber: Light can travel through any transparent material, but the speed of light will be slower in the material than in a vacuum. The ratio of the speed of light in a vacuum to that in a material is known as the material's refractive index (n) and is given by $n = c/v$, where c is the speed in a vacuum and v is the speed in the material. When light travels from one material of a given refractive index to another material of a different refractive index (i.e. when refraction occurs), the angle at which the light is transmitted in the second material depends on the refractive indices of the two materials as well as the angle at which light strikes the interface between the two materials. According to Snell's law, we have $n_a \sin \theta_a = n_b \sin \theta_b$, where n_a and n_b are the refractive indices of the first substance and the second substance, respectively; and θ_a and θ_b are the angles from the normal of the incident and refracted lights, respectively.

Figure 2.3 shows that a fiber consists of a core completely surrounded by a cladding (both of which consist of glass of different refractive indices). Let us first consider a step-index fiber, in which the change of refractive index at the core-cladding boundary is a step function. If the refractive index of the cladding is less than that of the core, then the total internal reflection can occur in the core and light can propagate through the fiber as shown in Figure 2.4 [7]. The angle above which total internal reflection will take place is known as the critical angle and is given by θ_c .

$$\sin \theta_c = n_{\text{clad}} / n_{\text{core}}. \quad (2.1)$$

where n_{clad} and n_{core} are the refractive indices of the cladding and core respectively. Thus, for a light to travel down a fiber, the light must be incident on the core-cladding surface at an angle greater than θ_c . For the light to enter a fiber, the incoming light should be at an angle such that the refraction at the air-core boundary results in the transmitted light's being at an angle for which total internal reflection can take place at the core-cladding boundary. The maximum value of θ_{air} can be derived from

$$n_{\text{air}} \sin \theta_{\text{air}} = n_{\text{core}} \sin (90^\circ - \theta_c) \quad (2.2)$$

Here we can rearrange Eqn. 2.2 to

$$n_{\text{air}} \sin \theta_{\text{air}} = \text{sqrt}(n_{\text{core}}^2 - n_{\text{clad}}^2) \quad (2.3)$$

The quantity $n_{\text{air}} \sin \theta_{\text{air}}$ is referred to as the *numerical aperture* (NA) of the fiber and θ_{air} is the maximum angle with respect to the normal at the air-core boundary, so that the incident light that enters the core will experience total internal reflection inside the fiber.

Figure 2.3: Multimode(a) & Single mode(b) optical fibers (unit μm)

Figure 2.4: Total Internal Reflection in an optical fiber

Multimode versus Single-mode Fiber: A mode in an optical fiber corresponds to one of the possible multiple ways in which a wave may propagate through the fiber. It can also be viewed as a standing wave in the transverse plane of the fiber. More formally, a mode corresponds to a solution of the wave equation that is derived from Maxwell's equations and subject to boundary conditions imposed by the optical fiber waveguide [4]. Although total internal reflection may occur for any angle θ that is greater than θ_c , light will not necessarily propagate for all of these angles. For some of these angles, light will not propagate due to destructive interference between the incident light and the reflected light at the core-cladding interface within the fiber. For other angles of incidence, the incident wave and the reflected wave at the core-cladding interface constructively interfere in order to maintain the propagation of the wave. The angles for which waves do propagate correspond to modes in a fiber. If more than one mode propagates through a fiber, then

the fiber is called multimode. In general, a larger core diameter or high operating frequency allows a greater number of modes to propagate.

The advantage of multimode fiber is that, its core diameter is relatively large; as a result, injection of light into the fiber with low coupling loss can be accomplished by using inexpensive, large-area light sources, such as light-emitting diodes (LED's). The disadvantage of multimode fiber is that it introduces the phenomenon of intermodal dispersion. In multimode fiber, each mode propagates at a different velocity due to different angles of incidence at the core-cladding boundary. This effect causes different rays of light from the same source to arrive at the other end of the fiber at different times, resulting in a pulse that is spread out in the time domain. Intermodal dispersion increases with the distance of propagation, so that it limits the bit rate of the transmitted signal and the distance that the signal can travel. Thus, in RoF networks multimode fiber is not utilized much, instead, single-mode fiber is widely used. Single-mode fiber allows only one mode and usually has a core size of about 10 μm , while multimode fiber typically has a core size of 50-100 μm . It eliminates intermodal dispersion and hence can support transmission over much longer distances. However, it introduces the problem of concentrating enough power into a very small core. LED's cannot couple enough light into a single-mode fiber to facilitate long-distance communications. Such a high concentration of light energy may be provided by a semiconductor laser, which can generate a narrow beam of light.

Fiber Attenuation: Attenuation in an optical fiber leads to a reduction of the signal power as the signal propagates over some distance. When determining the maximum distance that a signal can propagate for a given transmitter power and receiver sensitivity, one must consider attenuation. Let $P(L)$ be the power of the optical pulse at distance L km from the transmitter and A be the attenuation constant of the fiber (in dB/km). Attenuation is characterized by

$$P(L) = 10^{-AL/10} P(0) \quad (2.4)$$

where $P(0)$ is the optical power at the transmitter.

Dispersion in Fiber: Dispersion is the widening of pulse duration as it travels through a fiber. As a pulse widens, it can broaden enough to interfere with neighboring pulses (bits) on the fiber, leading to intersymbol interference. Dispersion thus limits the bit spacing and the maximum transmission rate on a fiber-optic channel. As described earlier, one form of the dispersion is an intermodal dispersion. This is caused when multiple modes of the same signal propagate at different velocities along the fiber. Intermodal dispersion does not occur in a single-mode fiber. Another form of dispersion is material or chromatic dispersion. In a dispersive medium, the index of refraction is a function of the wavelength. Thus, if the transmitted signal consists of more than one wavelength, certain wavelengths will propagate faster than other wavelengths. Since no laser can create a signal consisting of an exact single wavelength, chromatic dispersion will occur in most systems. A third type of dispersion is waveguide dispersion. Waveguide dispersion is caused as the propagation of different wavelengths depends on waveguide characteristics such as the indices and shape of the fiber core and cladding.

At 1300 nm, chromatic dispersion in a conventional single-mode fiber is nearly zero. Luckily, this is also a low-attenuation window (although loss is higher than 1550 nm). Through advanced techniques such as dispersion shifting, fibers with zero dispersion at a wavelength between 1300-1700 nm can be manufactured [35].

Fiber Nonlinearities: Nonlinear effects in fiber may potentially have a significant impact on the performance of optical communications systems. Nonlinearities in fiber may lead to attenuation, distortion, and cross-channel interference. In a WDM system, these effects place constraints on the spacing between adjacent wavelength channels, limit the maximum power on any channel, and may also limit the maximum bit rate. The details of the fiber nonlinearities are very complex and beyond the scope of the dissertation. It should be emphasized that they are the major limiting factors in the available number of channels in a WDM system [21].

2.1.2 Optical Transmitters

The word “*laser*” is an acronym for light amplification by stimulated emission of radiation. The key word is stimulated emission, which is what allows a laser to produce intense high-powered beams of coherent light (light that contains one or more distinct frequencies). To understand stimulated emission, we must first acquaint ourselves with the energy levels of atoms. Atoms that are stable (in the ground state) have electrons in the lowest possible energy levels. In each atom, there are a number of discrete levels of energy that an electron can have, which are referred to as “*states*” To change the level of an atom in the ground state, the atom must absorb energy. When an atom absorbs energy, it becomes excited and moves to a higher energy level. At this point, the atom is unstable and usually moves quickly back to the ground state by releasing a “*photon*”, a particle of light.

There are certain substances, however, whose states are *quasi-stable*, which means that the substances are likely to stay in the excited state for longer periods of time without constant excitation. By applying enough energy (in the form of either an optical pump or an electrical current) to a substance with quasi-stable states for a long enough period of time, *population inversion* occurs, which means that there are more electrons in the excited state than in the ground state. This inversion allows the substance to emit more light than it absorbs.

Figure 2.5 [17] shows a general representation of the structure of a laser. The laser consists of two mirrors that form a cavity (the space between the mirrors), a lasing medium, which occupies the cavity, and an excitation device. The excitation device applies current to the lasing medium, which is made of a quasi-stable substance. The applied current excites electrons in the lasing medium, and when an electron in the lasing medium drops back to the ground state, it emits a photon of light. The photon will reflect off the mirrors at each end of the cavity and will pass through the medium again.

Figure 2.5: General Structure of a Laser

Stimulated emission occurs when a photon passes very close to an excited electron. The photon may cause the electron to release its energy and return to the ground state. In the process of doing so, the electron releases another photon, which will have the same direction and coherency (frequency) as the stimulating photon. Photons for which the frequency is an integral fraction of the cavity length will coherently combine to build up light at the given frequency within the cavity. Between normal and stimulated emission, the light at the selected frequency builds in intensity until energy is being removed from the medium as fast as it is being inserted. The mirrors feed the photons back and forth, so further stimulated emission can occur and higher intensities of light can be produced. One of the mirrors is partially transmitting, so that some photons will escape the cavity in the form of a narrowly focused beam of light. By changing the length of the cavity, the frequency of the emitted light can be adjusted.

The frequency of the photon emitted depends on its change in energy levels. The frequency is determined by Eqn. 2.5

$$f = (E_i - E_f) / h \quad (2.5)$$

where f is the frequency of the photon, E_i is the initial (quasi-stable) state of the electron, E_f is the final (ground) state of the electron and h is Planck's constant ($= 6.626 \times 10^{-34}$ J.s).

Semiconductor Diode Lasers: The most useful type of a laser for optical networks is the semiconductor diode laser. The simplest implementation of a semiconductor laser is the bulk laser diode, which is a p-n junction with mirrored edges perpendicular to the junction (Figure 2.6).

In semiconductor materials, electrons may occupy either the valence band or the conduction band. The valence band and conduction band are analogous to the ground state and excited state of an electron mentioned above. The valence band corresponds to an energy level at which an electron is not free from an atom. The conduction band corresponds to an energy level at which an electron has become a free electron and may move freely to create current flow. The region of energy between the valence band and the conduction band is known as the “*band gap*”. An electron may not occupy any energy levels in the band-gap region. When an electron moves from the valence band to the conduction band, it leaves a vacancy, or “*hole*”, in the valence band. When the electron moves from the conduction band to the valence band, it recombines with the hole and may produce the spontaneous emission of a photon. The frequency of the photon is given by the equation above, where $E_i - E_f$ is the band-gap energy.

Figure 2.6: Structure of a Semiconductor Laser Diode

A semiconductor may be doped with impurities to increase either the number of electrons or the number of holes. An n-type semiconductor is doped with impurities that provide extra electrons. These electrons will remain in the conduction band. A p-type

semiconductor is doped with impurities that increase the number of holes in the valence band. A p-n junction is formed by layering p-type semiconductor material over n-type semiconductor material.

In order to produce stimulated emission, voltage is applied across the p-n junction to forward bias the device and cause electrons in the 'n' region to combine with holes in the 'p' region, resulting in light energy being released at a frequency related to the band gap of the device. By using different types of semiconductor materials, light with various ranges of frequencies may be released. The actual frequency of light emitted by the laser is determined by the length of the cavity formed by mirrored edges perpendicular to the p-n junction [17].

Optical Modulation: To transmit data across an optical fiber, the information must first be encoded, or modulated, onto the laser signal. Analog techniques include amplitude modulation (AM), frequency modulation (FM), and phase modulation (PM). Digital techniques include amplitude shift keying (ASK), frequency shift keying (FSK), and phase shift keying (PSK). Of all these techniques, binary ASK currently is the preferred method of digital modulation because of its simplicity. In binary ASK, also known as on-off keying (OOK), the signal is switched between two power levels. The lower power level represents a 0 bit, while the higher power level represents a 1 bit.

In systems employing OOK, modulation of the signal can be achieved by simply turning the laser on and off (direct modulation). In general, however, this can lead to chirp, or variations in the laser's amplitude and frequency, when the laser is turned on. A preferred approach for high bit rates (≥ 10 Gb/s) is to have an external modulator that modulates the light coming out of the laser. To this end, the Mach Zehnder interferometer or electro-absorption modulator are widely utilized [1, 6].

2.1.3 Optical Receivers

Photodetectors: In receivers employing direct detection, a photodetector converts the incoming photonic stream into a stream of electrons. The electron stream is then amplified and passed through a threshold device. Whether a bit is a logical zero or one depends on whether the stream is above or below a certain threshold for bit duration. In other words, the decision is made based on whether or not light is present during the bit duration. The basic detection devices for direct-detection optical networks are the PN photodiode (a p-n junction) and the PIN photodiode (an intrinsic material is placed between p- and n- type material). In its simplest form, the photodiode is basically a reverse-biased p-n junction. Through the photoelectric effect, light incident on the junction will create electron-hole pairs in both the “n” and the “p” regions of the photodiode. The electrons released in the “p” region will cross over to the “n” region, and the holes created in the “n” region will cross over to the “p” region, thereby resulting in a current flow.

Optical Amplifiers: Although an optical signal can propagate a long distance before it needs amplification, both long-haul and local lightwave networks can benefit from optical amplifiers. All-optical amplification may differ from optoelectronic amplification in that it may act only to boost the power of a signal, not to restore the shape or timing of the signal. This type of amplification is known as 1R (regeneration), and provides total data transparency (the amplification process is independent of the signal's modulation format). 1R amplification is emerging as the choice for the transparent all-optical networks of tomorrow. Today's digital networks [e.g., Synchronous Optical Network (SONET) and Synchronous Digital Hierarchy (SDH)], however, use the optical fiber only as a transmission medium, the optical signals are amplified by first converting the information stream into an electronic data signal and then retransmitting the signal optically. Such amplification is referred to as 3R (regeneration, reshaping, and reclocking). The reshaping of the signal reproduces the original pulse shape of each bit, eliminating much of the noise. Reshaping applies primarily to digitally modulated signals but in some cases it may also be applied to analog signals. The reclocking of the signal synchronizes the signal to its original bit timing pattern and bit rate. Reclocking applies

only to digitally modulated signals [4]. Another approach to amplification is 2R (regeneration and reshaping), in which the optical signal is converted to an electronic signal, which is then used to modulate a laser directly. The 3R and 2R techniques provide less transparency than the 1R technique, and in future optical networks, the aggregate bit rate of even just a few channels might make 3R and 2R techniques less practical [3].

Optical amplification uses the principle of stimulated emission, similar to the approach used in a laser. The two basic types of optical amplifiers are semiconductor laser amplifiers and rare-earth-doped-fiber amplifiers.

Doped-Fiber Amplifier: Optical doped-fiber amplifiers are lengths of fiber doped with an element (rare earth) that can amplify light. The most common doping element is erbium, which provides gain for wavelengths of 1525-1560 nm. At the end of the length of fiber, a laser transmits a strong signal at a lower wavelength (referred to as the pump wavelength) back up the fiber. This pump signal excites the dopant atoms into a higher energy level. This allows the data signal to stimulate the excited atoms to release photons. Most erbium-doped fiber amplifiers (EDFA's) are pumped by lasers with a wavelength of either 980 or 1480 nm [7].

A limitation to optical amplification is the unequal gain spectrum of optical amplifiers. While an optical amplifier may provide gain across a range of wavelengths, it will not necessarily amplify all wavelengths equally. This characteristic- accompanied by the fact that optical amplifiers amplify noise as well as signal and the fact that the active region of the amplifier can spontaneously emit photons, which also cause noise- limits the performance of optical amplifiers. Thus, a multi-wavelength optical signal passing through a series of amplifiers will eventually result in the power of the wavelengths' being uneven [7].

2.2 *Optical Techniques for Distributing and Generating Microwave Signals*

Several techniques for distributing and generating microwave signals via optical fiber exist. The techniques may be classified into two main categories namely Intensity Modulation – Direct Detection (IM-DD) and Remote Heterodyne Detection (RHD) techniques [23]. The electrical signal at the headend (CS) of the optical link may be one of three kinds namely, baseband, modulated IF, or the modulated RF signal itself. Whatever the case, the aim is to produce appropriate RF signals at the remote station, which meet the specifications of the wireless application. This means that apart from signal purity (frequency), robustness against noise, and power issues, the generated RF signals must also contain data in appropriate modulation format. If only baseband data is available at the headend, the RoF system must also perform the modulation function in addition to transporting and frequency up-conversion of the data. Therefore, a RoF system may have to perform radio-system functions as well, apart from signal transportation.

Apart from functionality, there are other factors such as performance, complexity and power issues to consider when selecting a suitable optical modulation scheme. Overall, the RoF system must be cost-effective for the application concerned. This section briefly presents the various available optical modulation techniques.

2.2.1 **RF Generation by Direct Intensity Modulation (DIM)**

The simplest method for optically distributing RF signals is simply to directly modulate the intensity of the light source with the RF signal itself and then to use direct detection at the photodetector to recover the RF signal. This method falls under the IM-DD technique. There are two ways of modulating the light source. The laser diode can itself be modulated directly by using the appropriate RF signal to drive the laser bias current. The second option is to operate the laser in continuous wave (CW) mode and then use an external modulator such as the Mach-Zehnder Modulator (MZM), to modulate the intensity of the light. The two options are shown in Figure 2.7 [25]. In both cases, the

modulating signal is the actual RF signal to be distributed. The RF signal must be appropriately premodulated with data.

After transmission through the fiber and direct detection on a PIN photodiode the photocurrent will be a replica of the modulating RF signal applied either directly to the laser or to the external modulator at the transmitter. The photocurrent undergoes transimpedance amplification to yield a voltage that is in turn used to excite the antenna. The modulation format of the data is preserved. Since the RF signal itself must be present at the headend, this technique can be used for distribution purposes only as it provides no other radio-system functions [24].

Figure 2.7: RF Signal Generation using DIM a) Laser, b) External Modulator

2.2.2 Photodetector based Optical Heterodyning

Most RoF techniques rely on the principle of coherent mixing in the photodiode [13]. These techniques are generally referred to as Remote Heterodyning Detection (RHD) techniques. While performing O/E conversion, the photodiode also acts as a mixer thereby making it a key component in RHD based RoF systems. The principle of coherent mixing is briefly described below. Two optical fields of angular frequencies Ω_1 and Ω_2 can be represented as:

$$\begin{aligned}
 E_1 &= E_{01} \cos(\Omega_1 t) \\
 E_2 &= E_{02} \cos(\Omega_2 t)
 \end{aligned}
 \tag{2.6}$$

If both fields impinge on a PIN photodetector, the resulting photocurrent on the surface will be proportional to the square of the sum of the optical fields. That is the normalized photocurrent will be:

$$\begin{aligned}
 i_{\text{PIN}} &= (E_1 + E_2)^2 \\
 i_{\text{PIN}} &= E_{01} E_{02} \cos[(\Omega_1 - \Omega_2) t] \\
 &\quad + E_{01} E_{02} \cos[(\Omega_1 + \Omega_2) t] + \text{other terms}
 \end{aligned}
 \tag{2.7}$$

The term of interest is $E_{01} E_{02} \cos[(\Omega_1 - \Omega_2) t]$, which shows that by controlling the difference in frequency between the two optical fields, radio signals of any frequency can be generated. The only limit to the level of frequencies that can be generated by this method is the bandwidth limitation of the photodiode itself [25].

2.2.3 Optical FM-Filter System

The Optical FM-Filter technique is a single-laser technique that involves modulating the optical frequency by applying an electrical signal to one of the laser's terminals. This generates a series of optical spectral lines (sidebands) all spaced by the drive frequency as shown in Figure 2.8 [25]. Two sidebands are then selected. The sidebands must be separated by the required RF frequency, which is normally in the mm-wave range of frequencies. The selected sidebands subsequently impinge on the surface of the photodiode and mix coherently to generate the desired RF signal as discussed in Section 2.2.2.

There are three commonly used methods for selecting the required sidebands. These are:

- Simple Optical Filtering
- Semiconductor Optical Amplifier (SOA)
- Injection-locked Lasers

Using an optical filtering arrangement, the required sidebands are selected while the rest are rejected. On the other hand, the SOA selects and amplifies the desired sidebands above the rest.

Figure 2.8: FM-Laser based Optical Coherent Mixing

2.2.4 Wavelength Division Multiplexing

Wavelength Division Multiplexing (WDM) is also used to exploit fiber bandwidth in RoF systems. It is perhaps the most commonly studied Optical Modulation Technique. Carriers modulated with mm-waves are dropped from and added to a fiber ring using Optical Add-Drop Multiplexers (OADM) [26]. The OADM are placed at base stations and tuned to select the desired optical carriers to drop.

2.2.5 Sub-Carrier Multiplexing

The Sub-carrier Multiplexing (SCM) technique is a maturing simple and cost effective approach for exploiting optical fiber bandwidth in analog optical communication systems in general and in RoF systems in particular. In SCM, an RF or microwave signal (the sub-carrier) is used to modulate an optical carrier at the transmitter's side. This results in an optical spectrum consisting of the original optical carrier f_0 , plus two side-tones located at $f_0 - f_{sc}$ and $f_0 + f_{sc}$ where f_{sc} is the sub-carrier frequency. If the sub-carrier itself is modulated with data (analog or digital), then sidebands centered on $f_0 - f_{sc}$ and $f_0 + f_{sc}$ are produced as shown in Figure 2.9.

Figure 2.9: Sub-Carrier Multiplexing of data in ROF Systems

To multiplex multiple channels on to one optical carrier, multiple sub-carriers are first combined and then used to modulate the optical carrier as shown in Figure 2.9. At the receiver side the sub-carriers are recovered through direct detection and then radiated. The mobile station is tuned to select and receive the desired sub-carrier(s), which may then undergo down conversion and appropriate demodulation. Different modulation schemes may be used on the different sub-carriers. In addition, the data used to modulate the sub-carriers need not be of the same kind. One sub-carrier may carry digital data, while another may be modulated with an analog signal such as video or telephone traffic. In this way, SCM supports the multiplexing of various kinds of mixed mode broadband data. Modulation of the optical carrier may be achieved by either directly modulating the laser (as shown in the figure), or by using external modulators such as the Mach Zender Modulator (MZM) or Electro-Absorption Modulator (EAM) [27].

2.3 Wireless LAN Technology

A wireless Local Area Network (LAN) is a flexible data communications system implemented as either an extension to, or as an alternative to the conventional wired LAN. The bulk of wireless LAN systems use Radio Frequency (RF) transmission technology. However, the Infrared (IR) spectrum is also used by some systems, mainly

noncommercial systems [28]. The focus of this project is on RF wireless LANs. Wireless LANs are typically fed through the wired LAN as shown in Figure 2.10. The radio Access Point (AP) consists of a bridge and a base station, and acts as the interface between the wired LAN and the wireless LAN. The interface between the wireless LAN and the mobile computer or other device is provided by a Network Adapter, which is installed in the mobile device.

Figure 2.10: Feeder Network for Wireless LAN

2.3.1 Narrowband and Wideband Wireless LANs

Wireless LANs using the RF spectrum employ either narrowband or wideband radio technology. For narrowband wireless LANs the end user must obtain an FCC license. However, wideband wireless LANs normally use the Instrumentation (or Industrial), Scientific, and Medical (ISM) frequency bands of 915 MHz, 2.4 GHz, and 5 GHz, which do not require licensing. The ISM spectra are used not only by wireless LANs but other electronic devices as well, such as microwave ovens, and portable mobile devices (Bluetooth). Therefore, wireless LAN technologies utilizing the ISM bands must be equipped to deal with interferences of all kinds in these spectra. Commercially available wireless LAN technologies employ spread-spectrum technology to provide reliable and

secure transmission in the ISM bands. Spread spectrum technologies trade-off bandwidth efficiency for reliability and security.

2.3.2 Spread-Spectrum Technology

Two kinds of spread-spectrum technologies exist. These are Frequency-Hopping Spread Spectrum (FHSS) and Direct-Sequence Spread-Spectrum (DSSS). FHSS uses a narrowband carrier that changes frequency in a pattern known to both the transmitter and the receiver. When properly synchronized, the system behaves as though there is a single logical channel between the transmitter and the receiver. In DSSS each bit to be transmitted is represented by a redundant bit pattern called the chip, or chipping code. As a result more bandwidth is used to transmit information. The extra redundant bits are used to recover from bit errors encountered on the noisy ISM frequency bands, and also for code division multiplexing.

2.3.3 Orthogonal Frequency Division Multiplexing (OFDM)

Orthogonal Frequency Division Multiplexing (OFDM) is seen as the modulation technique for future broadband wireless communications because it provides increased robustness against frequency selective fading and narrowband interference, and is efficient in dealing with multi-path delay spread. To achieve this, OFDM splits high-rate data streams into lower rate streams, which are then transmitted simultaneously over several sub-carriers. By so doing, the symbol duration is increased. The advantage of this is that the relative amount of dispersion in time caused by multi-path delay spread is decreased significantly. OFDM can be considered as both a multiplexing method as well as a modulation method [29].

As stated above, OFDM uses multiple sub-carriers to transmit low rate data streams in parallel. The sub-carriers are modulated by using Phase Shift Keying (PSK) or Quadrature Amplitude Modulation (QAM) and are then carried on a high frequency microwave carrier (e.g. 5 GHz). This is similar to conventional Frequency Division Multiplexing (FDM) or Sub-Carrier Multiplexing, except for the stringent requirement of orthogonality between the sub-carriers. OFDM is processor intensive because the basic

OFDM signal is formed using the Inverse Fast Fourier Transform (IFFT), adding a cyclic extension and performing windowing to get steeper spectral roll-off. In the receiver, the sub-carriers are demodulated by using Fast Fourier Transformation (FFT). The requirement for the intensive computations accounts for the complexity of OFDM transmitters and receivers. In comparison to single-carrier modulation systems, OFDM is more sensitive to frequency offset and phase noise. Furthermore, OFDM has a relatively large peak-to-average power ratio, which reduces the power efficiency of the RF amplifier.

2.4 Summarizing IEEE 802.11a

In the Open Systems Interconnection (OSI) communications model, the physical layer supports the electrical or mechanical interface to the physical medium. For example, this layer determines how to transfer a stream of bits from the upper (data link) layer onto the pins for a parallel printer interface, an optical fiber transmitter, or a radio carrier. The physical layer is usually a combination of software and hardware programming and may include electromechanical devices. It does not include the physical media as such.

The IEEE 802.11 specification is a wireless LAN (WLAN) standard that defines a set of requirements for the physical layer (PHY) and a medium access control (MAC) layer. The IEEE 802.11a standard is the physical layer standard for WLANs in the 5 GHz radio band. It specifies eight available radio channels as shown in Figure 2.11 [30]. The maximum link rate is 54-Mbps per channel, but the maximum user data throughput will be approximately one half of this value, and the throughput is shared by all users of the same wireless channel.

Figure 2.11: OFDM PHY Frequency Channel Plan

The data rate decreases as the distance between the user and the wireless access point increases. Because of the availability of 8 radio channels with the IEEE 802.11a standard as compared to 3 in the IEEE 802.11b standard, there is better protection against possible interference from neighbouring access points. The 5 GHz band also offers three times the operating bandwidth over the available spectrum in the 2.4 GHz band (for the IEEE 802.11b standard). The 5 GHz band is also less susceptible to interference, unlike the 2.4 GHz unlicensed band which shares the spectrum with other wireless appliances such as Bluetooth devices. From Figure 2.11, it can be seen that 300 MHz of bandwidth is allocated in the 5 GHz band to WLANs under the rules of the Unlicensed-National Information Infrastructure (U-NII). The bandwidth is fragmented into two blocks that are noncontiguous across the 5 GHz band. In the U-NII band, eight carriers are spaced across 200 MHz in the lower portion of the spectrum (5.150 - 5.350 GHz) and four carriers are spaced across 100 MHz in the upper portion of the spectrum (5.725 - 5.825 GHz). The channels are spaced 20 MHz apart, which allows for high data transmission rates for each channel.

The IEEE 802.11a standard defines OFDM modulation for the high-speed physical layer (PHY) in the 5 GHz band. The OFDM process provides for a wireless LAN with data payload communication capabilities of 6, 9, 12, 18, 24, 36, 48 and 54 Mbps. The modulation type used for each data rate in 802.11a is highlighted later in Chapter 4.

2.5 RoF-based Wireless LAN System Requirements

In general, the RoF system to be designed must be able to distribute the wireless LAN signals efficiently and in conformity with all aspects of the appropriate standard, such as signal purity and power levels. The system must also be easy to upgrade for operation with future systems. The factors influencing the choice of Microwave generation technique has already been discussed in Section 2.2. Some of the other issues and parameters which influence the design of such a system are discussed below.

2.5.1 System Cost

The fact that future wireless LANs will consist of a high density of small radio cells makes the issue of system cost a major one. It is imperative to have simple and easy to maintain base stations (i.e. RAPs). The complexity of the RAPs is related to the RoF techniques employed. Therefore, the choice of the microwave generation method is important. The kind of feeder network infrastructure is another crucial one. While standard single mode fiber offers the most bandwidth, it has high installation and maintenance costs associated with it.

2.5.2 Data Modulation Formats

The RoF system to be designed must be capable of generating microwave signals with appropriate data modulation. As mentioned in Section 2.2, this functionality depends on the choice of the RoF technique used. Since OFDM is seen as the modulation/multiplexing technique for future wireless systems, the RoF system to be designed must be capable of distributing Phase Shift Keying (PSK) and Quadrature Amplitude Modulation (QAM) signals as defined in the OFDM standard (see Table 2.1).

Data Rate (Mbps)	6, 9, 12, 18, 24, 36, 48, 54
Modulation	BPSK, QPSK, 16-QAM, 64-QAM
Coding Rate	1/2, 2/3, 3 /4
Number of Sub-Carriers	52
Number of Pilots	4
OFDM Symbol Duration	4 μ s
Guard Interval	800 ns
Sub-Carrier Spacing	312.5 KHz
-3 dB Bandwidth	16.56 MHz
Channel Spacing	20 MHz

Table 2.1: Main Parameters of the OFDM Standard for 802.11a

2.6 OFDM as a Modulation Scheme for Wireless Communications

Orthogonal Frequency Division Multiplexing (OFDM) is a multi-carrier transmission technique, which divides the available spectrum into many carriers, each one being modulated by a low rate data stream. OFDM is similar to FDMA in that the multiple user access is achieved by subdividing the available bandwidth into multiple channels, which are then allocated to users. However, OFDM uses the spectrum much more efficiently by spacing the channels much closer together. This is achieved by making all the carriers orthogonal to one another, preventing interference between the closely spaced carriers. Coded Orthogonal Frequency Division Multiplexing (COFDM) is the same as OFDM except that forward error correction is applied to the signal before transmission. This is to overcome errors in the transmission due to lost carriers from frequency selective fading, channel noise and other propagation effects. For this discussion the terms OFDM and COFDM are used interchangeably, though the main focus of the project in this domain has been a study of OFDM, but it is assumed that any practical system will use forward error correction, thus would be COFDM.

In FDMA each user is typically allocated a single channel, which is used to transmit all the user information. The bandwidth of each channel is typically 10 kHz-30 kHz for voice communications. However, the minimum required bandwidth for speech is only 3 kHz. The allocated bandwidth is made wider than the minimum amount required to prevent channels from interfering with one another. This extra bandwidth is to allow for signals from neighbouring channels to be filtered out, and to allow for any drift in the centre frequency of the transmitter or receiver. In a typical system up to 50% of the total spectrum is wasted due to the extra spacing between channels. This problem becomes worse as the channel bandwidth becomes narrower, and the frequency band increases.

TDMA partly overcomes this problem by using wider bandwidth channels, which are used by several users. Multiple users access the same channel by transmitting their data in time slots. Thus, many low data rate users can be combined together to transmit in a single channel that has a bandwidth sufficient so that the spectrum can be used efficiently. There are however, two main problems with TDMA. There is an overhead associated with the change over between users due to time slotting on the channel. A change over time must be allocated to allow for any tolerance in the start time of each user, due to propagation delay variations and synchronization errors. This limits the number of users that can efficiently utilize each channel. In addition, the symbol rate of each channel is high (as the channel handles the information from multiple users) resulting in problems with multipath delay spread.

OFDM overcomes most of the problems faced by FDMA and TDMA. OFDM splits the available bandwidth into many narrow band channels (typically 100-8000). The carriers for each channel are made orthogonal to one another, allowing them to be spaced very close together, with no overhead as in the FDMA example. Because of this there is no great need for users to be time multiplex as in TDMA, thus there is no overhead associated with switching between users. The orthogonality of the carriers means that each carrier has an integer number of cycles over a symbol period. Due to this, the spectrum of each carrier has a null at the centre frequency of each of the other carriers in

the system. This results in no interference between the carriers, allowing them to be spaced as close as theoretically possible.

Each carrier in an OFDM signal has a very narrow bandwidth (i.e. 1 kHz), thus the resulting symbol rate is low. This results in the signal having a high tolerance to multipath delay spread, as the delay spread must be very long to cause significant inter-symbol interference.

3. Study of Orthogonal Frequency Division Multiplexing

3.1 Fundamentals of OFDM Modulation

The concept of using parallel data transmission by means of frequency division multiplexing (FDM) was published in mid 60s. The idea was to use parallel data streams and FDM with overlapping sub-channels to avoid the use of high speed equalization and to combat impulsive noise, and multipath distortion as well as to fully use the available bandwidth. The initial applications were in the military communications. In the telecommunications field, the terms of discrete multi-tone (DMT), multi-channel modulation and multi-carrier modulation (MCM) are widely used and sometimes they are interchangeable with OFDM.

The nature of WLAN applications demands high data rates. Naturally dealing with ever unpredictable wireless channel at high data rate communications is not an easy task. Investigations were made into the use of multi-carrier transmission to be used for combating the hostility of wireless channel at high data rate communications. OFDM is a special form of multi-carrier transmission where all the sub-carriers are orthogonal to each other. OFDM promises a higher user data rate transmission capability at a reasonable complexity and precision. OFDM as a multicarrier, digital communications technique is gaining popularity and is now being used in the IEEE 802.11A high-speed wireless LAN standard. A single carrier transmission system and a multi carrier transmission system are illustrated in Figure 3.1 [29].

In single carrier systems, transmitted data symbols are pulse shaped by the transmitter filter. After the pulse shaped data is passed through the multipath channel, a receiver filter (matched to the channel) is used in the receiver. This maximizes the signal to noise ratio and helps to extract the estimated original data symbols.

Figure 3.1: Single and Multi-carrier System

For higher data rate transmissions, ISI caused by multipath effects is the major obstacle in the normal single carrier data communication system. The basic principle of OFDM is to split the higher data rate stream into a large number of lower data rate streams that are transmitted simultaneously over a number of sub-carriers. Because the symbol duration increases for the lower rate parallel sub-carriers, the relative amount of dispersion caused by multipath delay spread is decreased. ISI is almost completely eliminated by introducing a guard time into every OFDM symbol. In the guard time, the OFDM symbol is cyclically extended to avoid intercarrier interference. The approach taken is to modulate the data onto a large number of carriers, which are very closely spaced in frequency. The symbol rate of each sub-carrier is very low, giving it a very narrow bandwidth. In the time domain, the signal is rectangular giving it a conventional $(\sin x/x)$ spectral shape.

In a classical parallel data system, the total signal frequency band is divided into N non-overlapping frequency sub-channels. Each sub-channel is modulated with a separate symbol and then the N sub-channels are frequency multiplexed. It seems useful to avoid spectral overlap of the channels to eliminate inter-channel interference. However this leads to inefficient use of the available spectrum. To cope with the inefficiency, the ideas

proposed were to use parallel data and FDM (frequency division multiplexing) with over-lapping sub-channels, in which, each carrying a signaling rate b is spaced b apart in frequency to avoid the use of high speed equalization and to combat impulsive noise and multi-path distortion, as well as to fully use the available bandwidth.

Figure 3.2: Orthogonal Multi-Carrier Technique vs conventional Multi-Carrier

As shown in Figure 3.2, by using the over-lapping multi-carrier modulation technique, almost 50 percent of the bandwidth is saved. To realize the over-lapping multi-carrier technique, however cross talk between sub-carriers (ICI) should be reduced; which means that orthogonality between sub-carriers should be effective.

An OFDM signal consists of the sum of sub-carriers that are modulated by phase shift keying (PSK) or quadrature amplitude modulation (QAM).

If $d_i, i = -N_s/2 \dots N_s/2-1$, are the complex QAM or PSK symbols, N_s is the number of sub-carriers and, T is the symbol duration, then one OFDM symbol starting at $t = t_s$ in the baseband can be written as [30]:

$$s(t) = \sum_{i=-\frac{N_s}{2}}^{\frac{N_s}{2}-1} d_{i+\frac{N_s}{2}} \exp(j2\pi \frac{i}{T}(t-t_s)), t_s \leq t \leq t_s + T$$

$$s(t) = 0, t < t_s \wedge t > t_s + T \quad (3.1)$$

The complex baseband OFDM signal defined in Eqn. 3.1 is the inverse Fourier transform of N_s QAM or PSK input symbols. For discrete time, the result is the inverse Fast Fourier Transform (IFFT). OFDM avoids the problem of ICI by making sure that the subcarriers are orthogonal to each other. Orthogonality is best illustrated in Figure 3.3 [31]. Here, three subcarriers are depicted, with data modulated on each. Together, these subcarriers make up one OFDM symbol that is then sent out into the channel. In an actual OFDM system, these waveforms would all have different phases or different amplitudes and phases, due to the use of either phase-shift keying (PSK) or quadrature amplitude modulation (QAM), respectively. For the sake of illustrative simplicity, all are shown here with the same amplitude and phase offsets.

Figure 3.3: Three Sub-carriers in an OFDM Symbol

The orthogonality of OFDM comes from the precise relationship between the subcarriers that make up one OFDM symbol. In an OFDM system, each subcarrier has exactly an integer number of cycles in a given T time interval. In other words, the number of cycles between any two adjacent subcarriers differs exactly by one. This implies that each subcarrier frequency is an integer multiple of a base frequency (that is, $f_1 = f_0$, $f_2 = 2 \times f_0$, $f_3 = 3 \times f_0$, and so on). These properties allow each subcarrier to be individually and independently demodulated from any other adjacent subcarriers.

Figure 3.4: Orthogonality in an OFDM Symbol

An equivalent frequency domain representation of Figure 3.3 is seen in Figure 3.4 [31]. Here, each subcarrier's frequency spectrum is represented by a *sinc* function, one of whose properties is to peak at its center frequency and go to zero at all integer multiples of this frequency. This property allows the OFDM receiver to effectively demodulate each subcarrier because, at the peaks of each of these *sinc* functions, the contributions from other subcarrier *sinc* functions are zero. It is the orthogonality that allows the subcarriers to be packed tightly against one another and to efficiently use the frequency spectrum, thus delivering unprecedented high data speeds.

3.1.1 OFDM Generation

As discussed in the previous section, OFDM can be seen as a parallel data scheme consists of 'N' sub-bands. The choice of technique used to separate these sub-bands determines the generation (and hence detection) scheme used in OFDM communication systems. Summarized below are three schemes that can be used to separate the sub-bands:

1. Use filters to completely separate the sub-bands. This method was borrowed from the conventional FDM technology. The limitation of filter implementation forces the bandwidth of each sub-band to be equal to $(1+\alpha) f_m$, where α is the roll-off factor and f_m is the Nyquist bandwidth. Another disadvantage is that it is difficult to assemble a set of matched filter when the number of carriers is large.
2. Use staggered QAM to increase the efficiency of band usage. In this way the individual spectra of the modulated carriers still use an excess bandwidth, but they are overlapped at the 3 dB frequency. The advantage is that the composite spectrum is flat. The separability or orthogonality is achieved by staggering the data (offset the data by half a symbol). The requirement for filter design is less critical than that for the first scheme.
3. Use Discrete Fourier Transform (DFT) to modulate and demodulate parallel data. The individual spectra are now *sinc* functions and are not band limited. The FDM is achieved, not by bandpass filtering, but by baseband processing. Using this method, both transmitter and receiver can be implemented using efficient Fast Fourier Transform (FFT) techniques that reduce the number of operations from N^2 in DFT, down to $N \log N$.

To summarize at this juncture, OFDM can be simply defined as a form of multi-carrier modulation where its carrier spacing is carefully selected so that each sub-carrier is orthogonal to the other sub-carriers. As is well known, orthogonal signals can be separated at the receiver by correlation techniques; hence, intersymbol interference among channels can be eliminated. Orthogonality can be achieved by carefully selecting carrier spacing, such as letting the carrier spacing be equal to the reciprocal of the useful symbol period.

To generate OFDM successfully the relationship between all the carriers must be carefully controlled to maintain the orthogonality of the carriers. For this reason, OFDM is generated by firstly choosing the spectrum required, based on the input data, and

modulation scheme used. Each carrier to be produced is assigned some data to transmit. The required amplitude and phase of the carrier is then calculated based on the modulation scheme (typically differential BPSK, QPSK, or QAM). The required spectrum is then converted back to its time domain signal using an Inverse Fourier Transform. In most applications, an Inverse Fast Fourier Transform (IFFT) is used. The IFFT performs the transformation very efficiently, and provides a simple way of ensuring the carrier signals produced are orthogonal.

The Fast Fourier Transform (FFT) transforms a cyclic time domain signal into its equivalent frequency spectrum. This is done by finding the equivalent waveform, generated by a sum of orthogonal sinusoidal components. The amplitude and phase of the sinusoidal components represent the frequency spectrum of the time domain signal. The IFFT performs the reverse process, transforming a spectrum (amplitude and phase of each component) into a time domain signal. An IFFT converts a number of complex data points, of length that is a power of 2, into the time domain signal of the same number of points. Each data point in frequency spectrum used for an FFT or IFFT is called a bin.

The orthogonal carriers required for the OFDM signal can be easily generated by setting the amplitude and phase of each frequency bin, then performing the IFFT. Since each bin of an IFFT corresponds to the amplitude and phase of a set of orthogonal sinusoids, the reverse process guarantees that the carriers generated are orthogonal.

Figure 3.5 below shows the configuration for a basic OFDM transmitter and receiver. The signal generated is at base-band and so to generate an RF signal the signal must be filtered and mixed to the desired transmission frequency.

Figure 3.5: Basic FFT, OFDM Transmitter and Receiver

3.2 Advantages of an OFDM Modulation System

3.2.1 Increased immunity to ISI & ICI

When signal passes through a time-dispersive channel, the orthogonality of the signal can be jeopardized. In OFDM systems the use of a Cyclic Prefix (CP) helps to maintain orthogonality between the sub carriers. Cyclic prefix is an extension of the idea of using a guard interval as the solution. Guard interval was defined by an empty space between two OFDM symbols, which serves as a buffer for the multipath reflection. The interval must be chosen as larger than the expected maximum delay spread, such that multi path reflection from one symbol would not interfere with another. In practice, the empty guard time introduces ICI. ICI is crosstalk between different sub-carriers, which means they are no longer orthogonal to each other [29]. This solution was later fine-tuned to use a cyclic extension of OFDM symbol or CP. CP is a copy of the last part of OFDM symbol which is appended to front the transmitted OFDM symbol [32].

CP still occupies the same time interval as guard period, but it ensures that the delayed replicas of the OFDM symbols will always have a complete symbol within the FFT Interval (often referred as FFT window); this makes the transmitted signal periodic which plays a very significant role as this helps maintaining the orthogonality.

Figure 3.6: Use of Guard Intervals and Cyclic Prefix in combating ISI & ICI

At the receiver side, CP is removed before any processing starts. As long as the length of CP interval is larger than maximum expected delay spread say τ_{DS} , all reflections of previous symbols are removed and orthogonality is restored. The orthogonality is lost when the delay spread is larger than length of CP interval. Inserting CP has its own cost, we lose a part of signal energy since it carries no information. The loss is measured as

$$\text{SNR}_{\text{loss_CP}} = -10 \log_{10} (1 - T_{\text{CP}} / T_{\text{SYM}}) \quad (3.1)$$

However, CP gives two fold advantages, first occupying the guard interval, it removes the effect of ISI and by maintaining orthogonality it completely removes the ICI. The cost in terms signal energy loss is not too significant.

3.2.2 Spectral Efficiency

In the case of OFDM, a better spectral efficiency is achieved by maintaining orthogonality between the sub-carriers. When orthogonality is maintained between different sub-channels during transmission, then it is possible to separate the signals very easily at the receiver side. Orthogonality makes it possible in OFDM to arrange the sub-

carriers in such a way that the sidebands of the individual carriers overlap and still the signals are received at the receiver without being interfered by ICI. The receiver acts as a bank of demodulators, translating each sub-carrier down to DC, with the resulting signal integrated over a symbol period to recover raw data.

3.2.3 Low cost Transmitters & Simple Receiver Structure

OFDM transmitters can be made low cost due to the ability to implement the mapping of bits to unique carriers via the use of IFFT [31]. The OFDM transmitter also simplifies the channel effect, thus a simpler receiver structure is enough for recovering transmitted data. If we use coherent modulation schemes, then very simple channel estimation (and/or equalization) is needed, on the other hand, we need no channel estimator if differential modulation schemes are used.

3.2.4 Use of Smart Antennas and Adaptive Modulation

Smart antennas can be integrated with OFDM. MIMO systems and space-time coding can be realized on OFDM and all the benefits of MIMO systems can be obtained easily. Adaptive modulation and tone/power allocation are also realizable on OFDM.

3.2.5 Resistance to Frequency selective Fading

OFDM is more resistant to frequency selective fading than single carrier systems. The orthogonality preservation procedures in OFDM are much simpler compared to CDMA or TDMA techniques even in very severe multipath conditions.

3.3 Disadvantages of an OFDM Modulation System

3.3.1 Synchronization Requirements

OFDM is highly sensitive to time and frequency synchronization errors. Demodulation of an OFDM signal with an offset in the frequency can lead to a high bit error rate. The source of synchronization errors are two; first one being the difference between local oscillator frequencies in transmitter and receiver, secondly relative motion between the transmitter and receiver that gives Doppler spread. Local oscillator frequencies at both points must match as closely as they can. For higher number of sub-channels, the

matching should be even more perfect. Motion of transmitter and receiver causes the other frequency error. So, OFDM may show significant performance degradation at high-speed moving vehicles [33]. To optimize the performance of an OFDM link, accurate synchronization is of prime importance. Synchronization needs to be done in three factors: symbol, carrier frequency and sampling frequency synchronization.

3.3.2 Peak to Average Power Ratio

Peak to Average Power Ratio (PAPR) is proportional to the number of sub-carriers used for OFDM systems. An OFDM system with large number of sub-carriers will thus have a very large PAPR when the sub-carriers add up coherently. Large PAPR of a system makes the implementation of Digital-to-Analog Converter (DAC) and Analog-to-Digital Converter (ADC) to be extremely difficult. The design of RF amplifier also becomes increasingly difficult as the PAPR increases, since this leads to reduced efficiency of the amplifier.

3.3.3 Co-channel Interference in Cellular Systems

In cellular communications systems, CCI is combated by combining adaptive antenna techniques, such as sectorization, directive antenna, antenna arrays, etc. Using OFDM in cellular systems will give rise to CCI. Similarly with the traditional techniques, with the aid of beam steering, it is possible to focus the receiver's antenna beam on the served user, while attenuating the co-channel interferers. This is significant since OFDM is sensitive to CCI.

3.4 OFDM System Design

System design always needs a complete and comprehensive understanding and consideration of critical parameters. OFDM system design is of no exception, it deals with some critical and often conflicting parameters. Basic OFDM philosophy is to decrease data rate at the subcarriers, so that the symbol duration increases, thus the multipaths are effectively removed. This poses a challenging problem, as higher value for CP interval will give better results, but it will increase the loss of energy due to insertion of CP. Thus, a tradeoff between these two must be obtained for a reasonable design.

3.4.1 OFDM System Design Requirements

The major requirements governing an OFDM system implementation are discussed below.

Available Bandwidth: Bandwidth is always the scarce resource, so a primary parameter for system design should be the available bandwidth for operation. The amount of bandwidth will play a significant role in determining number of subcarriers, because with a large bandwidth, we can easily fit in large number of subcarriers with reasonable guard space.

Required Bit rate: The overall system should be able to support the data rate required by the users. For example, to support broadband wireless multimedia communication, the system should operate at more than 10 Mbps at least.

Tolerable delay Spread: Tolerable delay spread will depend on the user environment. Measurements show that indoor environment experiences maximum delay spread of few hundreds of ns at most, whereas outdoor environment can experience up to 10 μ s. So the length of CP should be determined according to the tolerable delay spread.

Doppler values: Users on a high speed vehicle will experience higher Doppler shift where as pedestrians will experience smaller Doppler shift. These considerations must be taken into account.

3.4.2 OFDM System Design Parameters

The design parameters are derived according to the system requirements. The requirement of the system design must be fulfilled by the system parameters. Following are the design parameters for an OFDM system [29].

Number of Subcarriers: Increasing number of subcarriers will reduce the data rate via each subcarrier, which will make sure that the relative amount of dispersion in time

caused by multipath delay will be decreased. But when there are large numbers of subcarriers, the synchronization at the receiver side will be extremely difficult.

Symbol duration and Guard Time Implementation: A good ratio between the CP interval and symbol duration should be found, so that all multipaths are resolved and not significant amount of energy is lost due to CP. As a thumb rule, the CP interval must be two to four times larger than the *Root-Mean-Square* (RMS) delay spread. Symbol duration should be much larger than the guard time to minimize the loss of *SNR*, but within reasonable amount. It cannot be arbitrarily large, because larger symbol time means that more subcarriers can fit within the symbol time. More subcarriers increase the signal processing load at both the transmitter and receiver, increasing the cost and complexity of the resulting device.

Subcarrier Spacing: Subcarrier spacing must be kept at a level so that synchronization is achievable. This parameter will largely depend on available bandwidth and the required number of sub-channels.

Modulation type per Subcarrier: This is trivial, because different modulation scheme will give different performance. Adaptive modulation and bit loading may be needed depending on the performance requirement. It is interesting to note that the performance of OFDM systems with differential modulation compares quite well with systems using non-differential and coherent demodulation [34]. Furthermore, the computation complexity in the demodulation process is quite low for differential modulations.

Forward Error Correction: Choice of FEC code will play a vital role also. A suitable FEC coding will make sure that the channel is robust to all the random errors.

3.5 Channel Model

In general, a communication system consists of a transmitter, a channel and a receiver. In baseband transmission, the arriving baseband data pulses are not in the form of ideal transmitted pulses, each one occupying its own symbol time interval. Transmitter

filtering and channel characteristics cause the received signal to suffer the effects of additive white gaussian noise (AWGN) and intersymbol interference (ISI) due to multi-path fading. Thus the channel consists of both AWGN and multi-path effects; both effects will be considered in this theoretical study and the simulations.

3.5.1 AWGN Channel

In digital communications, the AWGN channel is the easiest channel model to analyze. In an AWGN channel model, it is assumed that there is no distortion or other effects other than the addition of white gaussian noise. White gaussian noise is a model for the thermal noise generated by random electron movements in the receiver. In an AWGN channel the received signal is given by:

$$R(t) = S(t) + N(t) \quad (3.2)$$

Where $S(t)$ is the transmitted signal, $R(t)$ is the received signal, and $N(t)$ is a zero-mean wide-sense stationary random process with power spectral density $S(f) = N_0/2$ with N_0 being the average noise power; the factor of 2 is due to a two-sided power spectral density. When the noise power has a uniform spectral density, the noise source is referred to as white noise. An AWGN channel is a channel in which the amplitude distortion due to noise has a normal distribution.

3.5.2 Multipath Channel

The multi-path channel is the most common channel used to simulate fading in indoor communication systems. Normally, OFDM-based packet transmission is simulated in an AWGN plus multi-path fading channel. OFDM, with long symbol periods and a guard interval, is robust relative to multi-path fading in the channel. Fading is caused by interference between two or more versions of the transmitted signal that arrive at the receiver at slightly different times. These signals, called multipath waves, combine at the receiver antenna to give a resultant signal that can vary widely in amplitude and phase, depending on the distribution of the intensities and propagation times for the individual waves and on the bandwidth of the transmitted signal.

Reflections, diffractions and scattering (due to the environmental objects) create a constantly changing environment or channel and distribute signal energy in amplitude, phase and over time; this creates multipath waves. The phases and amplitudes associated with different multi-path components tend to be random and cause fluctuations in the received signal strength, thereby inducing small-scale fading, signal distortion, or both. Multi-path propagation often lengthens the time required for the baseband signal to reach the receiver; this can cause signal smearing due to ISI (intersymbol interference). The transmitted signal is composed of a sequence of symbols, and ISI occurs when the current symbol is interfered with or overlaps previous symbols that are attenuated and time delayed in the multi-path channel.

Time Dispersion Parameters of Multipath Channels: A wide-band multi-path channel can be quantified in a gross sense by the mean excess delay spread ($\bar{\tau}$) and the rms delay spread (σ_τ) of the power spectrum profile. The mean excess delay is the first moment of the power delay profile and is defined as:

$$\bar{\tau} = \frac{\sum_k P(\tau_k) \tau_k}{\sum_k P(\tau_k)} \quad (3.3)$$

Where $P(\tau_k)$ is the peak power of the impulse response of the channel at $t = \tau_k$. The power delay profile shows the variation of the normalized received signal power (on a dB scale) with excess delay time (typically in nanoseconds for an indoor channel). The RMS multi-path time delay spread is the square root of the second central moment of the power delay profile and is given by:

$$\sigma_\tau = \sqrt{\bar{\tau^2} - (\bar{\tau})^2} \quad (3.4)$$

where

$$\bar{\tau^2} = \frac{\sum_k P(\tau_k) \tau_k^2}{\sum_k P(\tau_k)} \quad (3.5)$$

These time delays are measured relative to the first detectable signal arriving at $\tau_0 = 0$. Eqns. 3.3 through 3.5 depend on the relative amplitudes of the multipath components within $P(\tau)$. In this thesis, because the channel is defined by an indoor environment, it will be assumed that the *RMS* time delay spread is on the order of nanoseconds. The coherent bandwidth is a statistical measure of the range of frequencies over which the channel passes all spectral components with approximately equal gain and linear phase. In other words, the coherent bandwidth is the range of frequencies over which two frequency components have a strong likelihood for amplitude correlation. Two sinusoidal signals with a frequency separation greater than the coherent bandwidth (B_c), are affected quite differently by the channel. If the correlation function between two frequency components in the range of frequencies, over which the coherent bandwidth is defined, is above 0.9, then the coherent bandwidth is given by:

$$B_c = \frac{1}{50 \times \sigma_\tau} \quad (3.6)$$

If the correlation function is 0.5, then the coherent bandwidth is given by

$$B_c = \frac{1}{5 \times \sigma_\tau} \quad (3.7)$$

Fading Effects due to Multipath delay Spreads: The time delay dispersion due to the multi-path time delay spread causes flat or frequency selective fading for the transmitted signal. A baseband signal will undergo flat fading if its bandwidth, B_s , is much lesser than channel bandwidth, B_c . The signal will undergo frequency selective fading if $B_s \gg B_c$. Thus, flat fading occurs in narrow band channels, while frequency selective fading occurs in wide band channels.

3.5.3 Exponential Multipath Model

To produce an exact channel model for a particular environment, we would need to know the attributes of every reflector in the environment at each moment in time. For instance,

we would have to know the position of each reflector, how much of the signal was reflected back to the receiver, and whether or not the reflector was moving. Obviously, it is impossible to have this kind of information for a given environment at each point in time. Therefore, channel models have been developed that emulate the typical or average behavior of a channel. An example of an indoor power delay profile is shown in Figure 3.7 [36].

Figure 3.7: Indoor Power Delay Profile

The channel model that is recommended in the IEEE 802.11a WLAN specification is an exponential channel model [5], which has a path delay profile that drops off exponentially. This model represents a real world scenario in which the positions of the reflectors generate paths that are longer and longer. This channel model is more realistic than the 2-ray model, and, although it tends to be a bit pessimistic, it is a reasonably accurate representation of many real world indoor wireless environments [9]. As shown in Figure 3.8, the path delay profile for this model has the form:

$$P[\tau] = 1/\tau_d \exp(-\tau/\tau_d) \quad (3.8)$$

where τ_d completely characterizes the path delay profile.

Figure 3.8: Path Delay Profile for an Exponential Channel Model

For an exponential model, the mean excess delay is $\tau_u = \tau_d$, RMS delay spread is $\tau_\sigma = \tau_d$, and the maximum excess delay:

$$\tau_m = (A \times \tau_d) / 10 \times \log_{10}(e) \quad (3.9)$$

where A is the amplitude of the smallest “noticeable” amplitude given in dB relative to the amplitude of the 0 delay (line of sight) path.

Coincidentally, the mean excess delay spread is $\tau_u = \tau_d$, and the RMS delay spread is also $\tau_\sigma = \tau_d$. Thus, if the mean excess delay spread or the RMS delay spread is known, the path delay profile can be completely characterized.

3.6 Description of OFDM Simulation Model

3.6.1 Mathematical Description of OFDM

Before proceeding to discuss the OFDM communication system modeled for simulation in this project, it is valuable to discuss the mathematical definition of the modulation system. This allows us to see how the signal is generated and how receiver must operate, and it gives us a tool to understand the effects of imperfections in the transmission channel. As mentioned earlier, OFDM transmits a large number of narrowband carriers,

closely spaced in the frequency domain. In order to avoid a large number of modulators and filters at the transmitter and complementary filters and demodulators at the receiver, it is desirable to be able to use modern digital signal processing techniques, such as fast Fourier transform (FFT).

Figure 3.9: Examples of OFDM Spectrum (a) Single sub-channel (b) 5 carriers

At the centre frequency of each sub-channel, there is no crosstalk from other sub-channels [31].

Mathematically, each carrier can be described as a complex wave:

$$S_c(t) = A_c(t) e^{j [\omega_c(t) + \phi_c(t)]} \quad (3.10)$$

The real signal is the real part of $S_c(t)$. Both $A_c(t)$ and $\phi_c(t)$, the amplitude and phase of the carrier, can vary on a symbol by symbol basis. The values of the parameters are constant over the symbol duration period τ .

OFDM consists of many carriers. Thus the complex signals $S_s(t)$ (Figure 3.9) is represented by:

$$S_s(t) = \frac{1}{N} \sum_{n=0}^{N-1} A_n(t) e^{j [\omega_n(t) + \phi_n(t)]} \quad (3.11)$$

where,

$$\omega_n = \omega_0 + n\Delta\omega$$

This is of course a continuous signal. If we consider the waveforms of each component of the signal over one symbol period, then the variables $A_c(t)$ and $\phi_c(t)$ take on fixed values, which depend on the frequency of that particular carrier, and so can be rewritten:

$$\begin{aligned}\phi_n(t) &\rightarrow \phi_n \\ A_n(t) &\rightarrow A_n\end{aligned}$$

If the signal is sampled using a sampling frequency of $1/T$, then the resulting signal is represented by:

$$S_s(kT) = \frac{1}{N} \sum_{n=0}^{N-1} A_n e^{j[(\omega_0 + n\Delta\omega)kT + \phi_n]} \quad (3.12)$$

At this point, we have restricted the time over which we analyse the signal to N samples. It is convenient to sample over the period of one data symbol. Thus we have a relationship:

$$\tau = NT$$

If we now simplify eqn. 3.12, without loss of generality by letting $\omega_0 = 0$, then the signal becomes:

$$S_s(kT) = \frac{1}{N} \sum_{n=0}^{N-1} A_n e^{j\phi_n} e^{j(n\Delta\omega)kT} \quad (3.13)$$

Now eqn. 3.13 can be compared with the general form of the inverse Fourier transform:

$$g(kT) = \frac{1}{N} \sum_{n=0}^{N-1} G(n/NT) e^{j2\pi nk/N} \quad (3.14)$$

In eqn. 3.13, the function $A_n e^{j\phi_n}$ is no more than a definition of the signal in the sampled frequency domain, and $S(kT)$ is the time domain representation. Eqns. 3.13 and 3.14 are equivalent if:

$$\Delta f = \Delta\omega/2\pi = 1/NT = 1/\tau \quad (3.15)$$

Eqn. 3.15 also represents the condition required for orthogonality. Thus, one consequence of maintaining orthogonality is that the OFDM signal can be defined by using Fourier transform procedures.

3.6.2 OFDM Simulation Model

For the purpose of this research a communication system was modeled using Matlab to study some key performance criteria for OFDM under different channel conditions. The simulation model also allowed us to vary OFDM signal generation with regards to the modulation scheme. Figure 3.10 shows a simplified flowchart of the MATLAB simulation code followed by a brief description of the simulation.

Figure 3.10: OFDM Simulation Flowchart

Serial to Parallel Conversion: The transmitter first converts the input data from a serial stream to parallel sets. The input serial data stream is formatted into the word size and

shifted into parallel format required for transmission based on the modulation scheme, e.g. for QPSK, the data is formatted to 2 bits/word etc.

Data Modulation: The data to be transmitted on each carrier is then mapped into a Phase Shift Keying (PSK) format. The modulation scheme in an OFDM system can be selected based on the requirement of power or spectrum efficiency. In general, the selection of the modulation scheme applying to each sub-channel depends solely on the compromise between the data rate requirement and transmission robustness. Another advantage of OFDM is that different modulation schemes can be used on different sub-channels.

Inverse Fourier Transform: After determining the signal constellation of each sub-carrier, an inverse Fourier transform converts the frequency domain data set into samples of the corresponding time domain representation of this data. Specifically, the IFFT is useful for OFDM because it generates samples of a waveform with frequency components satisfying orthogonality conditions.

Parallel to Serial Conversion: Then, the parallel to serial block creates the OFDM signal by sequentially outputting the time domain samples.

In summary, at the transmitter, the signal is defined in the frequency domain. It is a sampled digital signal, and it is defined such that the discrete Fourier spectrum exists only at discrete frequencies. Each OFDM carrier corresponds to one element of this discrete Fourier spectrum. The amplitudes and phases of the carriers depend on the data to be transmitted. The data transitions are synchronized at the carriers, and can be processed together, symbol by symbol.

Guard Interval: For the guard interval in this simulation we use a cyclic extension of the symbol to be transmitted. This makes the total symbol duration as $T_{\text{total}} = T_g + T$, where T_g is the guard interval and T is the useful symbol duration. The ratio of the guard interval to useful symbol duration is application-dependent. Since the insertion of guard interval will reduce data throughput, T_g is usually kept to less than $T/4$.

Channel: The channel model allows examination of common wireless channel characteristics such as noise, multipath, and clipping. By adding random data to the transmitted signal, simple noise is simulated. Multipath simulation involves adding attenuated and delayed copies of the transmitted signal to the original. This simulates the problem in wireless communication when the signal propagates on many paths. Finally, clipping simulates the problem of amplifier saturation. This addresses a practical implementation problem in OFDM where the PAPR is high.

Receiver: The receiver basically does the reverse operation to the transmitter. The guard period is removed. The FFT of each symbol is then taken to find the original transmitted spectrum. The OFDM data are split from a serial stream into parallel sets. The Fast Fourier Transform (FFT) converts the time domain samples back into a frequency domain representation. The magnitudes of the frequency components correspond to the original data. The phase angle of each transmission carrier is then evaluated and converted back to the data word by demodulating the received phase. The data words are then combined back to the same word size as the original data.

3.6.3 OFDM Simulation Parameters

The simulation Model developed supports BPSK, QPSK, 16PSK modulation, allowing individual sub-carriers to carry 1, 2 or 4 bits of data. The number of sub-carriers loaded with data was 48.

Parameter	Value
Data Modulation Schemes	BPSK, QPSK, 16PSK
Data sub-carriers	52
OFDM Symbol Duration	64 samples = 3.2μs
Guard Interval duration	16 samples = 0.8μs
Guard Period Type	Cyclic Prefix
FFT Size	64

Table 3.1: OFDM Simulation parameters

The simulation compares the performance of PSK modulation techniques. The primary aim was to study the trade of between transmission capacities of the OFDM system with the transmission reliability under different channel conditions. Table 3.1 shows the simulation configuration for the OFDM signals.

3.7 OFDM Simulation Results

3.7.1 OFDM Performance in AWGN and Multipath Channels

This section depicts the SNR performance of OFDM signals in AWGN and multipath channels. We assume the exponential multipath profile, described in Section 3.5.3, known to characterize an indoor wireless channel well. The RMS delay spread is taken as 50 ns. It was found that the SNR performance of OFDM is similar to a standard single carrier digital transmission. This is to be expected, as the transmitted signal is similar to a standard Frequency Division Multiplexing (FDM) system. Figures 3.11- 3.13 depict the simulation results for BPSK, QPSK & 16PSK digital modulations. Please take note that for Multipath transmission channel, the simulation results were obtained firstly without using any guard interval and later with the use of a cyclic extension of the OFDM symbol as well.

Figure 3.11: SNR performance for OFDM using BPSK Modulation in AWGN & Multipath channels

Figure 3.12: SNR performance for OFDM using QPSK Modulation in AWGN & Multipath channels

Figure 3.13: SNR performance for OFDM using 16PSK Modulation in AWGN & Multipath channels

The results for AWGN transmission of OFDM show that, while using QPSK modulation the system can tolerate a SNR greater than 10-11 dB. The BER however deteriorates rapidly below 8 dB. The use of BPSK would be more suited while transmitting through a noisy channel or at a lower transmission power at the expense of transmission data capacity. To increase the data transmission capacity the use of 16PSK can be considered, but only in a low noise link where a SNR > 18-20 dB can be ensured. The simulation results for OFDM transmission through a Multipath channel show a BER higher than 0.01, regardless of the modulation scheme used, in the case where no guard interval was utilized. However the BER performance is much improved after utilizing the proposed cyclic extension as a part of the transmission.

3.7.2 Tolerance to Multipath Delay Spread

This simulation was carried out to study BER performance of OFDM signals with increasing Delay Spread. Once again BPSK, QPSK & 16PSK modulation was used for the transmitted digital data. The multipath signal consisted of a single reflected echo that

was 3dB weaker than the direct signal. From earlier measurements it was noted that a reflected echo that was more than 3 dB weaker than the transmitted signal did not cause any measurable errors. This was especially true for BPSK and QPSK modulation. Figure 3.14 shows the simulation results. The effective guard period used in this case is 16, below which the BER performance is extremely good.

Figure 3.14: Delay Spread Tolerance of OFDM

It can be seen from Figure 3.14 that the BER is very low for a delay spread of less than approximately 16 samples. For a delay spread that is longer than the effective guard period, the BER rises rapidly due to the inter-symbol interference.

3.7.3 Peak Power Compression

This simulation studies the effect of Power clipping on the BER performance of OFDM signals. As mentioned earlier OFDM signals have a high PAPR due to the large number of sub-carriers. This limits the efficiency of RF amplifiers, hence affecting the range of wireless devices. The simulation results for BPSK, QPSK & 16PSK modulation are shown in Figure 3.15.

Figure 3.15: Effect of Peak Power Clipping for OFDM

The simulation results show that OFDM signals can have the peak power clipped to reduce the PAPR. In fact for BPSK and QPSK modulated signals, the signal could be clipped by up to 10 dB without a significant increase in the BER.

3.8 Conclusions

It has been shown that the use of a cyclic prefix in a transmitted OFDM system leads to a BER improvement of about 2 dB in all the modulation schemes as shown in Figures 3.11-3.13 in a multipath channel. OFDM transmission has shown great robustness against multipath fading, again due to the use of the cyclic prefix. In a practical system the length of the guard period can be chosen depending on the required multipath delay spread immunity required. While investigating the problem of high PAPR of OFDM signals we have seen that OFDM can indeed have its peak power clipped by a significant amount to reduce the PAPR. Hence, clipping the peak power will also allow for a greater transmitted power. From the results we can also infer that the signal is highly resistant to clipping distortions caused by the power amplifier used in transmitting the signal.

4. Physical Layer Modeling of IEEE 802.11a

4.1 *High Speed Wireless LAN Options*

The Wireless Local Area Network (WLAN) industry has emerged as one of the fastest-growing segments of the communications industry. The extensive growth in deployment and coverage area of wireless LAN products was due, in large part, to the introduction of standards-based WLAN products. The first products to be introduced – based on the 802.11b standard – were faster, lower in cost, and simpler to setup and use than previous generation products. The majority of WLAN products today communicate at speeds up to 11 Mbps.

In 1999, the IEEE approved both the 802.11a and 802.11b standards. 802.11a specified radios transmission at 5 GHz and at speeds up to 54 Mbps using orthogonal frequency division multiplexing (OFDM) modulation technology. The 802.11b standard – now popularly known as Wi-Fi – specified operation in the 2.4 GHz band (also known as the ISM band) and could achieve speeds up to 11 Mbps using direct sequence spread spectrum (DSSS) technology. Because DSSS is easier to implement than OFDM, 802.11b products appeared on the market first, starting in late 1999. Since then, 802.11b products have been widely deployed in corporations, small offices/home offices (SOHO), in residential home and in public locations (Wi-Fi “hotspots”). In early 2002, the first end-users products based on the 802.11a standard were shipped.

4.1.1 IEEE 802.11b WLAN Standard

802.11b, which was approved by the IEEE in 1999, is an extension of the 802.11 DSSS system and supports higher 5.5 and 11 Mbps payload data rates in addition to the original 1 and 2 Mbps rates. Products are now widely available, and the installed base of systems is growing rapidly. 802.11b also operates in the highly populated 2.4 GHz ISM band (2.40 to 2.4835 GHz), which provides only 83 MHz of spectrum to accommodate a variety of other radiating products, including cordless phones, microwave ovens, other WLANs, and personal area networks (PANS). This makes susceptibility to interference a

primary concern. The occupied bandwidth of the spread spectrum channel is 22 MHz, so the ISM band accommodates only three non-overlapping channels spaced 25 MHz apart. To help mitigate interference effects, 802.11b designates an optional frequency agile or hopping mode using the three non-overlapping channels or six overlapping channels spaced at 10 MHz.

802.11b uses eight-chip complementary code keying (CCK) as the modulation scheme to achieve the higher data rates. Instead of the Barker codes used to encode and spread the data for the lower rates, CCK uses a nearly orthogonal complex code set called complementary sequences. The chip rate remains consistent with the original DSSS system at 11 Mchip/s, while the data rate varies to match channel conditions by changing the spreading factor and/or the modulation scheme.

4.1.2 IEEE 802.11a WLAN Standard

This standard uses 300 MHz of bandwidth in the 5 GHz unlicensed national information infrastructure (UNII) band. The spectrum is divided into two “domains,” each having restrictions imposed on the maximum allowed output power. The lower band occupies 5.15 to 5.35 GHz, While the upper band occupies 5.725 to 5.825 GHz. 802.11a specifies eight non-overlapping 20 MHz channels in the lower band; each of these are divided into 52 sub-carriers (four of which carry pilot data) of 300-kHz bandwidth each. Four non-overlapping 20 MHz channels are specified in the upper band. The receiver processes the 52 individual bit streams, reconstructing the original high-rate data stream. Four complex modulation methods are employed, depending on the data rate that can be supported by channel conditions between the transmitter and receiver. These include BPSK, QPSK, 16-QAM, and 64-QAM.

Quadrature amplitude modulation is a complex modulation method where data are carried in symbols represented by the phase and amplitude of the modulated carrier. 16-QAM has 16 symbols. Each represents four data bits. 64-QAM has 64 symbols with each representing six data bits. BPSK modulation is always used on the four pilot sub-carriers. Although it adds a degree of complication to the baseband processing, 802.11a includes

forward error correction (FEC) as part of the specification. FEC, which does not exist within 802.11b, enables the receiver to identify and correct errors made during transmission by sending additional data along with the primary transmission. This nearly eliminates the need for retransmissions when packet errors are detected.

4.1.3 Advantages and Disadvantages of 802.11b, 802.11a

A drawback of the 5 GHz band, which has received considerable attention, is its shorter wavelength. Higher-frequency signals will have more trouble propagating through physical obstructions encountered in an office (walls, floors, and furniture) than those at 2.4 GHz. An advantage of 802.11a is its intrinsic ability to handle delay spread or multipath reflection effects. The slower symbol rate and placement of significant guard time around each symbol, using cyclical extension, reduces the inter-symbol interference (ISI) caused by multipath interference. To contrast, 802.11b networks are generally range-limited by multipath interference rather than the loss of signal strength over distance.

When it comes to deployment of a wireless LAN, operational characteristics have been compared to those of cellular systems, where frequency planning of overlapping cells minimizes mutual interference support mobility and assists seamless channel handoff. The three non-overlapping frequency channels available for IEEE 802.11b are at a disadvantage compared to the greater number of channels available to 802.11a. The additional channels allow more overlapping access points within a given area while avoiding additional mutual interference.

Both 802.11b and 802.11a use dynamic rate shifting where the system will automatically adjust the data rate based on the condition of the radio channel. If the channel is clear, then the modes with the highest data rates are used. But as interference is introduced into the channel, the radio will fall back to a slower, albeit more robust, transmission scheme.

4.1.4 Summarizing IEEE Wireless LAN Standards

In early 2001, the FCC announced new rules allowing additional modulations in the 2.4GHz range. This allowed IEEE to extend 802.11b to support higher data rates,

resulting in the 802.11g standard. 802.11g defines new data rate, up to 54 Mbps, at 2.4 GHz using OFDM, while at the same time providing backward compatibility with 802.11b at speeds up to 11 Mbps using DSSS. The IEEE WLAN standards discussed in this chapter are summarized in the table below.

	802.11	802.11a	802.11b	802.11g
Standard Approved	July 1997	September 1999	September 1999	June 2003
Available Bandwidth	83.5 MHz	300 MHz	83.5 MHz	83.5 MHz
Unlicensed Frequencies of Operation	2.4-2.4835 GHz DSSS, FHSS	5.15-5.35 GHz OFDM 5.725-5.825GHz OFDM	2.4-2.4835GHz DSSS	2.4-2.4835GHz DSSS, OFDM
No. of non-Overlapping channels	3 (Indoor/Outdoor)	4 Indoor (UNII1) 4 Indoor/Out(UNII2) 4 Outdoor (UNII3)	3 (Indoor/Outdoor)	3 (Indoor/Outdoor)
Data Rate per channel	2, 1 Mbps	54, 48, 36, 24, 18, 12, 9, 6 Mbps	11, 5.5, 2, 1 Mbps	54, 36, 33, 24, 22, 12, 11, 9, 6, 5.5, 2, 1 Mbps
Modulation Type	DQPSK (2 Mbps DSSS) DBPSK (1 Mbps DSSS) 4GFSK (2Mbps FHSS) 2GFSK (1Mbps FHSS)	BPSK (6, 9 Mbps) QPSK (12, 18 Mbps) 16-QAM (24, 36 Mbps) 64-QAM (48, 54 Mbps)	DQPSK/CCK (11, 5.5 Mbps) DQPSK (2 Mbps) DBPSK (1 Mbps)	OFDM/CCK (6,9, 12,18,24,36,48,54) OFDM (6,9,12,18, 24,36,48,54) DQPSK/CCK (22, 33, 11, 5.5 Mbps) DQPSK (2 Mbps) DBPSK (1 Mbps)
Compatibility	802.11	Wi-Fi5	Wi-Fi	Wi-Fi at 11Mbps and below

Table 4.1: IEEE Wireless LAN Standards

4.2 Wireless LAN 802.11a Physical Layer Architecture

The 5 GHz 802.11a physical layer architecture is depicted in the layer model shown in Figure 4.1 [10]. The PHY specification contains three functional entities: the PMD (Physical Medium Dependent) function, the PHY convergence function (PLCP - Physical Layer Convergence Procedure) and the layer management function.

- PLCP Sublayer: In order to allow the IEEE 802.11 MAC to operate with minimum dependence on the PMD sublayer, a PHY convergence sublayer is

defined. This function simplifies the PHY service interface to the IEEE 802.11 MAC services.

- PMD Sublayer: The PMD sublayer provides a means to send and receive data between two or more stations. This specification is concerned with the 5 Ghz band using OFDM modulation.
- PHY Management Entity (PLME): The PLME performs management of the local PHY functions in conjunction with the MAC management entity.

Figure 4.1: Physical Layer Model (IEEE 802.11a)

4.3 Error Control Coding

Communications system performance can be improved significantly by implementing error control coding. There are two basic error control strategies, automatic repeat request (ARQ) and forward error correction (FEC) coding. In an ARQ system, the receiver checks for errors but does not correct them; it simply requests the transmitter to resend the incorrectly received data. Unlike an ARQ system, a FEC coding system does correct the data errors at the receiver. IEEE 802.11a PHY uses FEC coding.

4.3.1 Forward Error Correcting (FEC) Coding

The purpose of FEC coding is to increase the robustness of a channel by adding a certain number of redundant bits to the actual data bits in a particular pattern such that recovery of the actual data bits is enhanced. There are basically two types of FEC codes, convolution and block codes. With block codes, a block of k data bits is encoded into a block of n coded bits, where $n > k$. The $n - k$ redundant bits in a codeword are called *parity bits*, which are used to check the codeword for errors during the decoding process. If errors exist due to the presence of noise, the decoder will ideally detect and correct them. Convolution codes use linear shift-registers to add redundancy into the entire data stream. Convolution codes operate on serial data, one or a few bits at a time. Block codes operate on relatively large (i.e., up to a couple of hundred bytes) message blocks. Since IEEE 802.11a PHY uses convolution codes, block codes are not considered in this section.

4.3.2 Implementation of (FEC) Coding

In a system using FEC coding, for every k information data bits, n coded bits are transmitted such that $n > k$. Since the transmission time is the same for both coded and uncoded bits,

$$kT_b = nT_{bc} \quad (4.1)$$

where T_{bc} is the duration of a coded bit and T_b is the duration of a data bit. The coded bit rate is

$$R_{bc} = 1/T_{bc} \quad (4.2)$$

Substituting Eqn. 4.2 into Eqn. 4.1 we get

$$R_{bc} = (n/k).R_b = R_b/r \quad (4.3)$$

where $r = k/n$ is the code rate. Since FEC coding adds redundancy to the original data, the trade-off of using FEC coding is the need for increased bandwidth. As seen in Eqn. 4.3, this bandwidth expansion is a function of the code rate. In addition to the

transmission time, the average transmitted power is the same whether coded or uncoded bits are transmitted. Consequently,

$$P_c = E_{bc} \cdot R_{bc} = E_b \cdot R_b \quad (4.4)$$

where E_b is the average energy of the uncoded data bit, and E_{bc} is the average energy of the coded data bit. Eqn. 4.4 can be rewritten as

$$\begin{aligned} E_{bc} \cdot R_{bc} &= E_b \cdot R_b \\ E_{bc} &= (R_b / R_{bc}) E_b = r E_b \end{aligned} \quad (4.5)$$

since $r = R_b / R_{bc}$. Hence, with Eqns. 4.1, 4.3 and 4.5, we can obtain the relationship between coded and uncoded systems in terms of T_b , R_b or E_b .

4.3.3 Convolution Encoding

Convolution encoding is a technique that adds redundancy to the data systematically. The information bits are processed by the shift registers and the encoded output bits are obtained by the modulo-2 summation of the input bits and the contents of the shift registers. A general convolution encoder can be implemented with k shift registers and n modulo-2 adders. The convolution encoder specified by the IEEE 802.11a standard has six shift registers and two modulo-2 adders. The constraint length is $\nu = 7$. The constraint length represents the number of k data bit shifts over which a single data bit can influence the encoder output. The constraint length ν is defined as the length of the shift register plus one. Convolution codes are specified by two parameters, r and ν , where $r = k/n$ is the code rate, n is the total number of coded bits generated by k input bits, and ν is the constraint length of the code. The typical range of values for r and ν are

$$1/4 \leq r \leq 7/8 \quad (4.6)$$

and

$$2 \leq \nu \leq 9 \quad (4.7)$$

Note that higher coding gain is generally achieved by either increasing v or decreasing r .

4.3.4 Viterbi Decoding

In modern digital communication systems, the Viterbi decoding algorithm is a computationally efficient and easily achievable algorithm used for the optimum decoding of convolutional codes. The Viterbi algorithm decodes a convolutional code by choosing a path through the code trellis, which yields a code word that differs from the received code word in the fewest possible places. The Viterbi algorithm searches all possible paths in the trellis in order to compute the path metrics. Each state (node) in the trellis diagram is assigned a value that is determined from $s = 0$ at the time $t = 0$ to a particular state K at $t \geq 0$. At each state, the path with the best metric is the survivor, while the other entering paths are non-survivors. The selected metric represents the survivor path and the remaining metrics represent the non-survivor paths. The best value may be either the smallest or the largest, depending on hard or soft decision decoding and the metric chosen. The path selected by the Viterbi algorithm is the maximum-likelihood (ML) path. At any given state, we can only continue backward on a path that survived upon entry into that node. Since each node has only one entering survivor, our trace-back operation always yields a ML path.

Hard decision and soft decision decoding are the two possible ways to generate the branch metric for a Viterbi decoder. In hard decision decoding each received signal is examined and a decision is made as to whether the signal represents a transmitted bit zero or a bit one. For a hard decision decoding, the Viterbi algorithm is a minimum Hamming distance decoder. Hamming distance is obtained by choosing a path through the trellis which yields a codeword that differs from the received codeword in the fewest possible places. In soft decision decoding, the receiver takes advantage of the side information generated by the receiver quantization circuitry.

4.4 802.11a PLCP Sublayer

This sublayer is a convergence procedure for converting PSDUs (Physical service Data Units - the Data Part of the PLCP Frame in the Physical Layer) to and from PPDUs

(Physical Protocol Data Unit - the Training Sequence, Header and Data Parts of the PLCP Frame in the Physical Layer). During transmission, the PSDU will be provided with a PLCP preamble and header to create the PPDU. At the receiver, the PLCP preamble and header are processed to aid in demodulation and delivery of the PSDU.

4.4.1 PLCP Frame Fields

Figure 4.2 shows the frame format in the physical layer of the IEEE 802.11a standard. Each frame has a preamble (training sequence), header, data part (PSDU), tail and pad bits.

Figure 4.2: PPDU Frame Format

The header consists of the RATE, Reserved, LENGTH, Parity (even parity bit) and SERVICE fields. The RATE, Reserved, LENGTH and Parity fields constitute a separate single OFDM symbol, denoted as SIGNAL. This single OFDM symbol is transmitted with a very robust combination of BPSK modulation and a coding rate of $r = \frac{1}{2}$. The SERVICE field of the header and the PSDU field (with 6 “zero” tail bits and pad bits appended), denoted as DATA, are transmitted at the data rate defined in the RATE field and may constitute multiple OFDM symbols. The tail bits in the SIGNAL symbol enable

the decoding of the RATE and LENGTH fields immediately after the reception of the tail bits.

4.4.2 Rate Dependant Parameters

The modulation parameters that depend on the data rate will be set according to Table 4.2 below.

DATA RATE (Mbits/sec)	Modulation	Coding Rate (R)	Coded Bits per subcarrier (N _{BPSC})	Coded Bits Per OFDM Symbol (N _{CBPS})	Data Bits Per OFDM Symbol (N _{DBPS})
6	BPSK	1/2	1	48	24
9	BPSK	3/4	1	48	36
12	QPSK	1/2	2	96	48
18	QPSK	3/4	2	96	72
24	16- QAM	1/2	4	192	96
36	16- QAM	3/4	4	192	144
48	64-QAM	2/3	6	288	192
54	64- QAM	3/4	6	288	216

Table 4.2: Rate Dependent Parameters

FEC coding is applied in order to improve overall system performance; otherwise, the error probability will generally be determined by weaker sub-carriers due to fading conditions. The convolutional encoder employed by the IEEE 802.11a standard for the rate 1/ 2 code uses industry-standard generator polynomials $g_0 = 133_8$ and $g_1= 171_8$ with constraint length 7; that is, six linear shift registers comprise the memory components of the encoder as shown in Figure 4.3 [5].

Figure 4.3: Convolution Encoder $r = 1/2$ with constraint length $v = 7$

The bit denoted as “A” is output from the encoder before the bit denoted as “B.” Higher coding rates of $2/3$ and $3/4$ are obtained by puncturing the rate $1/2$ code as depicted in Figure 4.4. Puncturing is a procedure for omitting some of the encoded bits in the transmitter and inserting a dummy zero metric into the convolutional decoder on the receive side in place of the omitted bits. Puncturing reduces the free distance of the convolution code; however, the coding gain of the punctured codes is almost the same as that of the best code for that particular code rate. The rate-dependent parameters of IEEE 802.11a standard, which uses BPSK, QPSK, 16QAM, and 64QAM as sub-carrier modulation schemes in combination with rate $1/2$, $2/3$, and $3/4$ convolutional codes are listed in Table 4.2.

Figure 4.4: Puncturing Method for higher rates

4.4.3 Timing Related Parameters

A list of timing related parameters associated with OFDM PLCP in the IEEE 802.11a standard is given in Table 4.3.

Parameter	Value
N_{SD} : No. of Data Subcarriers	48
N_{SP} : No. of Pilot Subcarriers	4
N_{ST} : No. of Subcarriers, total	52 ($N_{SD} + N_{SP}$)
Δ_F : Subcarrier frequency spacing	312.5 KHz
T_{IFFT} : IFFT/ FFT Period	$3.2\mu s (= 1/ \Delta_F)$
$T_{PREAMBLE}$: PLCP preamble duration	$16\mu s$
T_{SIGNAL} : Duration of SIGNAL BPSK-OFDM symbol	$4.0\mu s (T_{IFFT} + T_{GI})$
T_{GI} : Guard Interval duration	$0.8\mu s (T_{IFFT} /4)$
T_{SYM} : Symbol Interval	$4.0\mu s (T_{IFFT} + T_{GI})$

Table 4.3: Timing Related Parameters

Number of Subcarriers: There are a total of 52 subcarriers for each OFDM symbol. However, four pilot subcarriers are used to assist timing and carrier tracking tasks during data symbols after the preamble. Hence, the remaining 48 subcarriers are used to carry the data stream.

Guard Interval: Guard interval T_{GI} is a very important parameter in IEEE 802.11a PHY because it provides robustness to rms delay spreads up to several hundreds of nanoseconds depending on the code rate and the modulation used in any indoor wireless application. In order to minimize ISI, which decreases orthogonality and has an effect that is similar to inter-channel-interference, T_{GI} is inserted between two consecutive OFDM symbols. T_{GI} should be larger than the expected rms delay spread; otherwise the impact of ISI will be significant. From [11], the reported rms delay spread values can be up to 200 ns for small/medium-size office buildings and 300 ns for large office buildings.

As can be seen from Table 4.3, the guard interval T_{GI} for each OFDM symbol is $0.8\mu s$ which is greater than 300 ns.

OFDM Symbol Duration and Subcarrier Spacing: As mentioned in the previous chapter, guard interval is an important parameter which eliminates ISI; however, there is an inverse relationship between T_{GI} and OFDM effective symbol duration; that is, when T_{GI} increases, the OFDM effective symbol duration decreases. Hence, T_{GI} cannot be chosen too large. The total symbol duration chosen in 802.11a is four microseconds. The inverse of the symbol duration less the inverse of the guard time of $0.8\mu s$ results in the sub-carrier spacing of 312.5 kHz.

4.4.4 PLCP Preamble Field

The PLCP preamble field is used for synchronization, frequency offset estimation and channel estimation. It consists of 10 short symbols and two long symbols as shown in Figure 4.5.

Figure 4.5: PLCP Preamble

Figure 4.5 shows the OFDM training structure (PLCP preamble), where the first frame consists of 10 short training symbols, and $T1$ and $T2$ denote long training symbols. The PLCP preamble is followed by the SIGNAL field and the DATA. The total preamble length is $16\mu s$.

4.4.5 SIGNAL Field

The OFDM training symbols in the PLCP Preamble will be followed by the SIGNAL field, which contains the RATE and the LENGTH fields. The RATE field conveys information about the type of modulation and the coding rate as used in the remainder of the packet. The encoding of the SIGNAL field's single OFDM symbol will be performed with BPSK modulation of the sub-carriers and using convolution coding at a coding rate of $r = \frac{1}{2}$. Contents of the SIGNAL field are not scrambled.

Figure 4.6: SIGNAL Field Bits

The SIGNAL field is to be composed of 24 bits, as illustrated in Figure 4.6. The four bits 0 through 3 will encode the RATE. Bit 4 is to be reserved for future use. Bits 5 through 16 will encode the LENGTH field, with the least significant bit (LSB) being transmitted first. Bit 17 will be used for the positive parity (even parity) bit for bits 0-16. Bits 18-23 constitute the SIGNAL TAIL field, and these 6 bits will be set to zero.

DATA Rate: The bits R1 through R4 will be set, dependent on RATE, according to the values given in Table 4.4.

R1-R4	Rate (Mbps)
1101	6
1111	9
0101	12
0111	18
1001	24
1011	36
0001	48
0011	54

Table 4.4: RATE bits content

LENGTH Field Bits: The PLCP length field will be an unsigned 12-bit integer that indicates the number of octets in the PSDU unit that the MAC is currently requesting the PHY to transmit. This value is used by the PHY to determine the number of octet transfers that will occur between the MAC and PHY after receiving a request to start transmission. The LSB is transmitted first in time. This field will be encoded using convolution encoding.

4.4.6 DATA Field

The DATA field contains the SERVICE field, the PSDU unit, the TAIL bits and the PAD bits, if needed, as described below. All bits in the DATA field are scrambled as described in Section 4.4.7.

SERVICE Bits: The IEEE 802.11a SERVICE field has 16 bits, which will be denoted as bits 0-15 as shown in Figure 4.7. Bit 0 will be transmitted first in time. Bits 0-6 of the SERVICE field are set to zero and are used to synchronize the descrambler in the receiver. The remaining 9 bits (7-15) of the SERVICE field are reserved for future use. All reserved bits will be set to zero.

Figure 4.7: SERVICE Field Bits

PPDU TAIL Bits: The PPDU tail bit field will consist of six "0's" which are required to return the convolution encoder to the "zero state". This procedure improves the error probability of the convolution decoder, which relies on future bits when decoding and which may not be available after the end of the message. The PLCP tail bit field is produced by replacing six scrambled "zero" bits following the end of the message with six non-scrambled "zero" bits.

PAD Bits: The number of bits in the DATA field will be a multiple of N_{CBPS} , the number of coded bits in an OFDM symbol (48, 96, 192 or 288 bits depending on the modulation type as shown in Table 4.2). To achieve this, the length of the message is extended so that it becomes a multiple of N_{DBPS} , the number of data bits per OFDM symbol. At least 6 bits are appended to the message in order to accommodate the TAIL bits, as described above. The number of OFDM symbols, N_{SYM} , the number of bits in the DATA field, N_{DATA} , and the number of pad bits, N_{PAD} are computed from the length of the PSDU (LENGTH) as follows:

$$N_{SYM} = \text{Ceiling}((16 + 8 \times \text{LENGTH} + 6)/N_{DBPS}) \quad (4.8)$$

$$N_{DATA} = N_{SYM} \times N_{DBPS} \quad (4.9)$$

$$N_{PAD} = N_{DATA} - (16 + 8 \times \text{LENGTH} + 6) \quad (4.10)$$

The function, *Ceiling* (.), is a function that returns the smallest integer value greater than or equal to its argument value. The appended bits ("pad bits") are set to "zeros" and are subsequently scrambled with the rest of the bits in the DATA field.

4.4.7 DATA Scrambler and Descrambler

All the bits transmitted by the 802.11a OFDM PMD in the data portion are scrambled using a frame synchronous 127 bits sequence generator. Scrambling is used to randomize the SERVICE, PSDU, PAD and data patterns, which may contain long strings of binary 1s or 0s. The tail bits are not scrambled. The octets of the PSDU are placed in the transmitted serial bit stream, bit 0 first and bit 7 last.

The frame synchronous scrambler uses the generator polynomial $S(x)$ as follows:

$$S(x) = x^7 + x^4 + 1 \tag{4.11}$$

The 127 bits sequence generated repeatedly by the scrambler is (leftmost used first)

**00001110 11110010 11001001 00000010 00100110 00101110 10110110 00001100
11010100 11100111 10110100 00101010 11111010 01010001 10111000 11111111**

when the "all ones" initial state is used. The same scrambler is used to scramble transmit data and to de-scramble receive data. When transmitting, the initial state of the 802.11a scrambler will be set to a pseudo random non-zero state. The seven LSBs of the SERVICE field will be set to all zeros prior to scrambling to enable estimation of the initial state of the scrambler in the receiver. The contents of the SIGNAL field of the 802.11a are not scrambled.

4.4.8 Data Interleaving

In a typical radio channel, the arriving OFDM subcarriers generally have different amplitudes due to frequency selective fading. Deep fades in the frequency spectrum may cause groups of subcarriers to be less reliable than others, thereby causing bit errors to

occur in bursts rather than being randomly scattered. Most forward error correction codes are not designed to deal with error bursts. Therefore, interleaving is applied to randomize the occurrence of bit errors prior to decoding. At the transmitter, the coded bits are permuted in a way that will assure that originally adjacent bits are separated by several bits after interleaving. At the receiver, the reverse permutation is performed before decoding.

In the 802.11a standard, a block interleaver interleaves all encoded data bits. The block size corresponds to the number of bits in a single OFDM symbol, N_{CBPS} . The interleaver is defined by a two steps permutation. The first permutation ensures that adjacent coded bits are mapped onto nonadjacent subcarriers. The second ensures that adjacent coded bits are mapped alternately onto less and more significant bits of the constellation and, thereby, long runs of low reliability (LSB) bits are avoided.

4.4.9 Modulation and Mapping

The OFDM subcarriers are modulated by using BPSK, QPSK, 16 QAM, or 64 QAM modulation, depending on the RATE requested. In the 802.11a the encoded and interleaved binary serial input data are divided into groups of N_{BPSC} (1, 2, 4, or 6) bits and mapped into BPSK, QPSK, 16 QAM, or 64 QAM constellation points. The constellation schemes are all shown in Figure 4.8.

Figure 4.8: Constellation for BPSK, QPSK, 16QAM and 64QAM

Following the constellation maps, the data bits can also be represented as complex numbers of the form $I + jQ$ by reading the integer values of I and Q ($\pm 1, \pm 3, \pm 5, \pm 7$) from the constellation maps.

The modulator outputs are formed by multiplying the resulting $(I + jQ)$ with a normalizing factor (K_{mod}). K_{mod} depends on the modulation type, as prescribed in Table 4.9. The modulation type can be different from the beginning to the end of the transmission, as the signal changes from SIGNAL to DATA. The purpose of the normalization factor is to achieve the same average power ($I^2 + Q^2$) for all mappings.

Input Bit (b0)	Inphase	Quadrature
0	-1	0
1	1	0

Table 4.5: BPSK modulation IQ mapping

Input Bit (b0)	Inphase	Input Bit (b1)	Quadrature
0	-1	0	-1
1	1	1	1

Table 4.6: QPSK modulation IQ mapping

Input Bits (b0 b1)	Inphase	Input Bits (b2 b3)	Quadrature
00	-3	00	-3
01	-1	01	-1
11	1	11	1
10	3	10	3

Table 4.7: 16QAM modulation IQ mapping

Input Bits (b0 b1 b2)	Inphase	Input Bits (b3 b4 b5)	Quadrature
000	-7	000	-7
001	-5	001	-5
011	-3	011	-3
010	-1	010	-1
110	1	110	1
111	3	111	3
101	5	101	5
100	7	100	7

Table 4.8: 64QAM modulation IQ mapping

Modulation	K_{mod}
BPSK	1
QPSK	$1/\sqrt{2}$
16QAM	$1/\sqrt{10}$
64QAM	$1/\sqrt{42}$

Table 4.9: Normalization Factor

4.5 Simulation Model for 802.11a Physical Layer

In accordance with the physical layer specifications, a Matlab simulation model was developed to simulate the performance of an end-to-end 802.11a physical link.

Figure 4.9: 802.11a PHY Simulation Model

The transmission channel was modeled both as a simple AWGN channel as well as a Multipath fading channel. The Multipath fading profile follows the exponential channel model (Section 3.5.3) as recommended by the IEEE 802.11a Standard specifications.

Figure 4.9 depicts baseband model of the 802.11a system. The encoding and decoding procedures are summarized as individual blocks depicting the flow of the simulation code. Each of these blocks has been discussed in detail in section 4.4. The sampling rate used for the 64-pt FFT is 20MHz (50nsec). Hence each OFDM symbol is 80 samples long.

4.6 Simulation Results for BER Performance of 802.11a PHY Layer

The performance of OFDM signals (compliant with 802.11a specifications) is examined in this section. As mentioned above the channel is modeled as an AWGN + Multipath Fading channel. The RMS delay spread is taken as 200nsec. Individual Bit Error performance results are presented for the sub-carrier modulation techniques for OFDM as specified in the IEEE 802.11a standard, which are BPSK, QPSK, 16QAM and 64QAM. The performance is simulated both with and without the use of FEC coding. As regards to system performance expectations, it is well known that a Bit Error Probability $P_b = 10^{-3}$ is a good benchmark required by many wireless communications systems for practical applications. The FEC scheme employed for the simulations uses a convolution encoder with a constraint length of 7 as recommended by the 802.11a standard. The Viterbi Decoding in the receiver uses Hard Decision Decoding to generate the branch metrics.

Figure 4.10 shows the performance of all the modulation schemes for 802.11a signal distribution without employing any FEC. The BER performance results with the use of FEC (constraint length 7) are shown in Figures 4.11, 4.12, 4.13 and 4.14 with the use of coding rates $r = \frac{1}{2}, \frac{2}{3}, \frac{3}{4}$. As shown in Table 4.2, BPSK modulation is used in 802.11a with coding rates $r = \frac{1}{2}$ and $\frac{3}{4}$ corresponding to transmission data rates of 6 Mbps and 9 Mbps. For QPSK, Coding rates $r = \frac{1}{2}$ and $\frac{3}{4}$ are used for transmission rates of 12 Mbps and 18 Mbps. 16QAM modulation also used the same coding rates, with 64QAM being the only case where $r = \frac{2}{3}$ is utilized for data rates of 48 Mbps.

Figure 4.10: Performance of 802.11a Modulation Schemes (AWGN + Multipath Fading Channel) without FEC coding

Figure 4.11: Performance of BPSK (AWGN + Multipath Fading Channel) with FEC and HDD

Figure 4.12: Performance of QPSK (AWGN + Multipath Fading Channel) with FEC and HDD

Figure 4.13: Performance of 16QAM (AWGN+ Multipath Fading Channel) with FEC and HDD

Figure 4.14: Performance of 64QAM (AWGN+ Multipath Fading Channel) with FEC and HDD

Simulation results show that there is a significant improvement in the performance with the use of convolution encoding. To achieve the targeted BER of 10^{-3} the performance advantage is around 15-20 dB with the use of FEC for all Modulation schemes.

Comparing the performance of BPSK on the basis of coding rate, we can see that for $r = \frac{1}{2}$ (6 Mbps), the advantage is around 6-7 dB as compared to $r = \frac{3}{4}$ (9 Mbps). QPSK Modulation for $r = \frac{1}{2}$ (12 Mbps) performs 7-8 dB better than $r = \frac{3}{4}$ (18 Mbps) for acceptable performance levels of $\text{BER} = 10^{-3}$. The performance difference is identical in the case of 16QAM with rates $r = \frac{1}{2}$ (24 Mbps) and $r = \frac{3}{4}$ (36 Mbps). For 64QAM the performance difference is only 3 dB between $r = \frac{2}{3}$ (48 Mbps) and $r = \frac{3}{4}$ (54 Mbps).

4.7 Conclusions

Using an end-to-end 802.11a physical layer model, we have obtained the required signal to noise levels permissible for transmission. As expected, with the implementation of FEC and Viterbi Decoding the BER performance makes a significant improvement. Also as expected, for a specific modulation type, as the code rate increases, the SNR required to achieve a fixed probability of bit error increases. The approximate coding gain for all

modulation types is around 15-20 dB. The use of the higher data rate transmission has also shown very acceptable performance results with the use of appropriate FEC coding rates (as suggested by 802.11a). The required SNR for acceptable bit-error-rates as well as the coding gain observed is similar to the expected performance values obtained from existing literature [40]. Similar Research into OFDM transmission systems has shown a further coding gain of around 2 dB through the use of Viterbi Soft Decision Decoding [46].

5. Transmission Performance of RoF Link for OFDM Signals

5.1 RoF based Broadband Data Networks

Conventional broadband wireless access systems use a cellular spectral reuse pattern similar to those of wireless telephony networks (Figure 5.1). The capacity of such a system is somewhat proportional to the geographical density of its component base stations. High-density systems require the deployment of a large number of base stations with each covering a relatively small area.

Figure 5.1: Conventional Large-cell Wireless Broadband Access System

Base station antennas are electromagnetically isolated from one another by physical separation, intervening terrain or structures and directional antennas. Signal attenuation by terrain and structures is greater when smaller coverage areas are used since lower antenna elevations are required. Greater attenuation facilitates improved frequency reuse which in turn helps to improve the overall system capacity. These capacity gains are over and above those provided simply by increased antenna density. From a capacity perspective it is therefore desirable to make the coverage area of each antenna as small as

possible. Since broadband data networks require a very large system capacity, very small antenna coverage areas are called for.

Deploying a very large number of full base stations along with their associated infrastructure would be prohibitively expensive. A more economical alternative is to use RoF to relay a radio signal from a distant base station or cluster of base stations to remote antennas. Such a configuration is illustrated in Figure 5.2.

Figure 5.2: Wireless Broadband Access System using micro- and pico-cells served by RoF

The primary elements of a RoF link are illustrated in Figure 5.3. A base station at some central location generates an RF signal $v_{sig}(t)$ that is to be transmitted to the subscribers in its prescribed service area. The signal is relayed to an antenna in the service area via an optical link that includes four major subsystems: a light source, an electro-optic (E/O) interface, a fiber optic cable and an opto-electric (O/E) converter. The output of the O/E converter is transmitted into the wireless medium by the antenna with which the converter is co-located. The antenna site is the actual WLAN Remote Access Point (RAP) for the subscribers. Though not shown in the figure, some basic control logic and RF circuitry may also be required at the RAP. The wireless signal is received and demodulated by antennas and transceivers located on the subscriber's premises or

wireless devices. The uplink path is essentially a mirror image of the downlink, though additional gain control and amplifier circuitry may be needed at the RAP to handle the large dynamic range and low power levels characteristic of uplink signals in a system such as this.

Figure 5.3: Major elements of a Radio-over-Fiber Link

5.2 Optical components of a RoF Link

An introduction to these basic components of optical transmission (optical fibers, transmitters and receivers) has been provided in Sections 2.1 and 2.2. Here we will examine these components and optical modulation techniques in detail for the purpose of simulation modeling of a RoF link.

5.2.1 Optical Link Light Source

While light sources such as light-emitting diodes (LEDs) can be used in some low-performance applications, lasers are more commonly used because of their greater intensity and because their narrow spectral line width allows the light to travel through fiber with much less dispersion.

A laser diode (LD) is the most common optical source used in optical communication links. Such a device emits coherent light at an intensity that is proportional to the

instantaneous current through the diode. A simplified plot illustrating the optical power emitted P_o as a function of the input current I_{LD} to the LD is shown in Figure 5.4.

Figure 5.4: Laser Diode Output Power P_o vs Input Current I_{LD}

Laser Diode Transfer Function: When the diode is forward biased ($I_{LD} > 0$ for simplicity) it emits optical power. For $I_{LD} < I_{Th}$ the diode acts as a common LED and the output light is incoherent. When $I_{LD} > I_{Th}$, the diode produces coherent lased light. The slope of the curve and therefore the output efficiency is much greater in the lasing region. The output intensity increases almost linearly with increasing I_{LD} in this region until compression begins to occur. The 1 dB compression point is defined as the point on the curve at which the optical power output is 1 dB less than it would be were the curve to continue to increase linearly. The input current corresponding to this point is defined as I_{1dB} . As the input current is increased past I_{1dB} the optical output power asymptotically approaches the maximum power that the laser is capable of delivering.

Laser Diode Noise and Distortion: An ideal laser diode generates a constant intensity coherent light beam with an infinitesimal linewidth, as the occupied bandwidth of an optical signal is called. A practical laser is subject to random fluctuations in the optical intensity output when the bias current is constant. These fluctuations appear as random noise to the detector. This noise is expressed as relative intensity noise (RIN), defined as

$$RIN = \{(\Delta P_o)^2\} / |P_o|^2 \quad (5.1)$$

where ΔP_o is the difference between the instantaneous optical intensity and the mean intensity $|P_o|$. RIN contributes to the total noise in the output of the detector.

Real lasers have a finite, but very small linewidth. Because light of different wavelengths travels at different speeds through a refractive medium, a laser pulse is dispersed by optical fiber. This can lead to ISI in a broadband baseband system or to fading in RoF systems as discussed later in Section 5.2.3.

5.2.2 Optical Modulator

An electrical information signal $v_{sig}(t)$ is transmitted over an optical link by using the electrical signal to modulate the intensity of the light injected into the fiber channel. This electro-optic (E/O) conversion can be accomplished by directly modulating the intensity of the light source or by using a constant-intensity source followed by an external modulator.

Direct Modulation: If a laser diode is biased with a constant input current I_{bias} , a time-varying signal current $i_{sig}(t)$ may be added to the constant bias to modulate the power delivered by the laser. This is known as direct modulation. The signal current is

$$i_{sig}(t) = v_{sig}(t) / R_{LD} \quad (5.2)$$

when a signal voltage $v_{sig}(t)$ is applied to an LD circuit that includes a series resistance R_{LD} . The total current input to the laser diode is then

$$i_{LD}(t) = I_{bias} + v_{sig}(t) / R_{LD} \quad (5.3)$$

Provided that $I_{Th} \leq i_{LD}(t) < I_{1dB}$, the optical power output (intensity) will vary nearly linearly with the signal voltage, so the signal suffers little nonlinear distortion as a consequence of the optical modulation operation.

When an analog signal directly modulates the intensity of a laser, the LD bias current is usually set to $I_{bias} = I_{Th} + (I_{1dB} - I_{Th})/2$, which is in the center of the near-linear region between the lasing threshold and the 1 dB compression point. However, if the biased input signal, $i_{LD}(t)$ is not strictly limited to that near-linear region, the resulting modulated

optical signal will exhibit significant nonlinear distortion. As $i_{LD}(t)$ approaches and exceeds I_{1dB} the output becomes compressed.

More severe distortion occurs when $i_{LD}(t) < I_{Th}$ at any time t during the signal. When this happens, the LD stops lasing and acts like an LED provided the diode is still forward biased. Referring to the simplified transfer function in Figure 5.4 we can see that the large difference in the slopes between the lasing and LED region would effectively cause the optical output to be clipped when $i_{LD}(t) \leq I_{Th}$. In practice an LD takes a finite amount of time t_{TR} to make the transition from operation in the LED region to the lasing region. When lasing, an LD has a finite bandwidth $B_{LD} = 1/T_{LD}$, where T_{LD} is the period of the highest-frequency sinusoid that falls within B_{LD} . In a typical LD, $t_{TR} \gg T_{LD}$ so if $v_{sig}(t)$ has a bandwidth $B_{sig} \gg 1/t_{TR}$ severe distortion that cannot be described by a simple clip will occur if the LD is driven into its LED region. This characteristic of laser diodes, combined with the fact that an LD may be damaged if reverse biased, make it undesirable to use direct modulation when the signal is not easily confined to the linear lasing region. In the case of OFDM, it would be necessary to confine the majority of the signal's power to a small portion of the lasing region to ensure that the large peaks do not drive the LD into its LED region. The only practical alternative would be to employ a high-bandwidth clipping or limiting circuit to prevent the input signal from dropping below the lasing threshold.

External Modulation: External modulators encode an information signal on an optical beam by modulating the output of a constant-intensity optical source. In an external modulator the constant-intensity optical beam is routed through an electro-optic material for which the optical properties can be varied by applying an electric field. External modulators are able to modulate signals that have bandwidths much greater than can be achieved with direct modulation. External modulators also tend to be less susceptible to damage caused by spurious high-voltage peaks in the drive signal $v_{sig}(t)$. Because OFDM has large peaks due to its high PAPR and is normally used at RF center frequencies above 5 GHz, external modulation is preferred for the transmission of the standard OFDM signals of interest.

Several varieties of external modulators are available. These can be broadly grouped into those that use electro-optic materials for which the phase of the light beam is varied through the application of an electric field and those which absorb, and therefore attenuate, the light beam by an amount that is a function of the electric field applied. Devices of the latter variety are known as electro-absorption modulators (EAMs).

Considering modulators based on optical phase modulation, the most commonly used external modulator is the phase modulation-based Mach-Zehnder interferometer, commonly referred to as a Mach-Zehnder modulator (MZM). A simplified diagram of the architecture of an MZM is shown in Figure 5.5.

Figure 5.5: Mach-Zehnder Modulator

A constant intensity light beam enters the MZM where it is split between two arms of equal length, each consisting of an electro-optic waveguide and a parallel electrode. A time-varying electrical signal $v_1(t)$ or $v_2(t)$ is applied to each electrode, and a constant bias voltage V_{bias} is also applied to one electrode. Each arm forms an optical phase modulator in which the instantaneous optical phase shift is a function of the electrode voltage. The light from the two arms is combined at the output of the device where constructive and destructive interference between the two phase-shifted beams modulates the intensity of the optical output.

The relationship between the electric field E_{in} of the lased light input to the device and the output electric field E_o is

$$E_o = E_{in} \cos(\pi (\Delta V/V_\pi)) \cdot \exp(j\pi(|V|/V_\pi)) \quad (5.4)$$

where

$$\Delta V = (v_1(t) - v_2(t)) / 2 \quad (5.5)$$

and

$$|V| = (v_1(t) + v_2(t)) / 2 \quad (5.6)$$

Note that when the voltage across the MZM $v_1(t) - v_2(t) = 2\Delta V = 0$, $|E_o| = |E_{in}|$ so the optical beam is not attenuated by the MZM. Conversely, when $2\Delta V = V_\pi$, $|E_o| = 0$ and the optical output of the device is extinguished. Therefore V_π , which is a constant for a particular device, is known as the switching voltage.

Since the power of an electromagnetic (EM) wave is proportional to the square of the magnitude of its electric field, the optical power at the output of the MZM is

$$\begin{aligned} P_{o,Tx}(t) &\propto |E_o|^2 \\ &\propto E_{in}^2 \cos^2(\pi(\Delta V/V_\pi)) \end{aligned} \quad (5.7)$$

The normalized transmitted optical power is plotted as a function of voltage applied to the device $v_{in}(t) = v_1(t) - v_2(t) = 2\Delta V$ in Figure 5.6 [17]. Note that the sinusoidal transfer function is phase shifted by π radians when $v_{in}(t)$ is changed by V_π .

Figure 5.6: Normalized MZM optical power output as a function of the voltage across the electrodes ($V_{bias} = -V_{\pi}/2$; $v_{in}(t)$ is normalized by V_{π})

It will be shown that the optical power transmitted is proportional to the electrical signal information $v_{sig}(t)$, so it is desirable to bias the MZM such that the relationship between $v_{sig}(t)$ and $P_{o,Tx}(t)$ is as linear as possible. Let us define $v_1(t) = v_{sig}(t)$ and set $v_1(t) = -v_2(t)$. When a non-zero bias voltage V_{bias} is applied the electrode driven by $v_1(t)$, we may define the total time-varying electrode drive voltage $v_1'(t) = v_{sig}(t) + V_{bias}$. If $v_1(t)$ is replaced with $v_1'(t)$ in Eqn. 5.5 under the conditions just stated we find that

$$\Delta V = v_{sig}(t) + V_{bias}/2 \quad (5.8)$$

Substituting the expression in Eqn. 5.8 into Eqn. 5.7 gives

$$\begin{aligned} P_{o,Tx}(t) &\propto E_{in}^2 \cos^2 (\pi (2v_{sig}(t) + V_{bias})/ 2V_{\pi}) \\ &\propto E_{in}^2/2 [\cos (\pi (2v_{sig}(t) + V_{bias})/ V_{\pi}) + 1] \end{aligned} \quad (5.9)$$

The identity $\cos^2(u) = \frac{1}{2} [\cos(2u) + 1]$ has been used to obtain the form used in Eqn. 5.9. From Eqn. 5.9 we can see that the relationship between the signal input $v_{sig}(t)$ to the optical link and the optical power transmitted $P_{o,Tx}$ is most linear when the argument to the cosine function is $\pm k\pi / 2$, where k is an integer. To minimize the nonlinear distortion introduced by the MZM it is desirable to operate the device such that the dc level V_{DC} of

$v_{sig}(t)$ is mapped to the center of the most linear region. Forcing the cosine argument in Eqn. 5.9 to $k\pi/2$ when $v_{sig}(t) = V_{DC}$ we find that

$$\pi (2V_{DC} + V_{bias})/ V_{\pi} = k\pi/2 \quad (5.10)$$

In RoF applications $v(t)$ is a bandpass IF or RF signal, so $V_{DC} = 0$. The optimal bias voltage is therefore

$$V_{bias} = k V_{\pi} /2 \quad (5.11)$$

To meet the requirements of Equation 5.11, a bias voltage of $V_{bias} = - V_{\pi}/2$ was used to create the power transfer function plot shown earlier in Figure 5.6. A negative bias with an odd-value of k is used here to ensure that the slope is positive at the dc level of the input signal point. The MZM may be biased such that the slope is negative at $v(t) = V_{DC}$ provided the receiver can detect and recover phase rotation.

In addition to centering the small signal in the most linear portion of the transfer function, the use of $V_{bias} = \pm kV_{\pi}/2$ also gives the transfer function odd symmetry about the dc level of $v(t)$. The other set of symmetrical bias points, found at $V_{bias} = \pm kV_{\pi}$, exhibits even symmetry resulting in full-wave rectification of the input signal $v(t)$.

5.2.3 Optical Fiber

The modulated optical signal propagates to the receiver over a silica fiber designed for low loss at the wavelength of the optical source. The signal suffers attenuation and distortion while traversing the optical channel.

Fiber Attenuation: Signal attenuation results from a variety of causes that are beyond the scope of this thesis. It is sufficient to note that in the 1550 nm wavelength window used in this study, fiber attenuation ranges from 0.3-0.5 dB/km for fiber lengths of up to 80 km. If an 802.11a-based RoF system were used with the base stations located in central offices not more than 5 or 6 km (serving an area of more than 75 km²) from the remote antennas the loss due to fiber attenuation would be not more than 3 dB.

Chromatic Dispersion: The main source of distortion in an optical fiber is chromatic dispersion. This type of dispersion arises because the index of refraction in a fiber is partly a function of optical wavelength. The modulation of a carrier produces a signal with a spectrum of width proportional to the data rate. Different frequencies in the spectrum correspond to different wavelengths, or colours, each of which propagates through a fiber at a unique speed. It can be shown that the group delay is actually linear, but not flat [7]. When broadband baseband signaling is used the group delay can vary sufficiently from one end of the band to the other, causing significant ISI if it is not equalized.

When the bandwidth of an analog signal is much less than the frequency of the optical carrier, the variation in group delay across the band of interest is negligible and results in no appreciable distortion. The variation in group delay between the two sidebands of an analog double-sideband modulated signal can be significant, particularly if the RF carrier frequency is large. The difference in group delay between the two sidebands causes their phases to rotate with respect to one another. The resulting constructive and destructive interference between the two sidebands causes the power in the detected signal to vary in a fading pattern that is a function of the fiber length traversed. The detected power after the signal has traveled through a length L_f of fiber is [7]

$$P_o(L_f)_{dB} \propto 10 \log \{ \cos^2 (\pi D L_f \lambda_o^{-2} f_c^2 / c) \} \quad (5.12)$$

where D is a dispersive parameter usually expressed in units of $ps/(nm.km)$, λ_o is the wavelength of the optical source, and c is the speed of light in free space. The fading profiles due to chromatic dispersion with $\lambda_o = 1550$ nm and $D = 16.5$ ps/ (nm.km) are plotted in Figure 5.7 for RF carrier frequencies f_c of 5.8 GHz. Practical RoF applications using 802.11a would use the 5.8 GHz ISM band. The first null in the fading profile at this frequency occurs when the signal has traveled over more than 100 km of fiber. There is less than 3 dB of attenuation when the distance traveled is not more than 50 km. The latter distance should be more than sufficient for most Local-Area RoF network applications.

Figure 5.7: Chromatic dispersion fading as a function of fiber length for 5.8 GHz RF center frequency

5.2.4 Photodetection

A photodetector (PD) is used to perform the O/E conversion. A photodiode is normally used as the photodetector in an optical communication system. A photodiode is a device that produces an electrical current output directly proportional to the optical power received $P_{o,Rx}$. Disregarding all noise in the link, the received signal generated by a PD driving a load of R_{out} ohms is [7]

$$v'_{Rx}(t) = R_{out} (\check{R}_{PD} G_f P_{o,Tx}(t) + N_{Rx}(t)) \quad (5.13)$$

where $\check{R}_{PD}$ is the responsivity of PD expressed in units amps/watt (A/W). $N_{Rx}(t)$ is a current term that represents the sum of all noise sources present in the received optical signal and within the PD itself.

Eqn. 5.13 confirms that the electrical output of the optical link is directly proportional to the transmitted optical power $P_{o,Tx}$ expressed in Eqn. 5.7. When a MZM is optimally biased and driven by complementary small-signal inputs, substituting Eqns. 5.7 and 5.11 into Eqn. 5.13 gives the electrical output of the link

$$\begin{aligned}
v'_{Rx}(t) &= \{K_1[\cos(\pi((2v_{sig}(t)/V_\pi) - 1/2)) + 1] + N_{Rx}(t)\} R_{out} \\
&= \{K_1[\sin(2\pi(2v_{sig}(t)/V_\pi)) + 1] + N_{Rx}(t)\} R_{out} \quad (5.14)
\end{aligned}$$

as a function of the electrical input. K_1 is a constant of proportionality. The dc offset may be removed via ac coupling and the expression further simplified to produce

$$v_{Rx}(t) = K_1[\sin(\pi(2v_{sig}(t)/V_\pi))] + N_{Rx}(t) R_{out} \quad (5.15)$$

Expanding the sin term into its Taylor series we find that

$$\begin{aligned}
v_{Rx}(t) &= K_1(\pi(2v_{sig}(t)/V_\pi) - (\pi(2v_{sig}(t)/V_\pi))^3/3! + (\pi(2v_{sig}(t)/V_\pi))^5/5! - \dots \\
&\quad + (-1)^n (\pi(2v_{sig}(t)/V_\pi))^{2n+1}/(2n+1)! + \dots) + N_{Rx}(t) R_{out} \\
&= K_2 v_{sig}(t) + D_{NL}(t) + N_{Rx}(t) R_{out} \quad (5.16)
\end{aligned}$$

where $K_2 = (2K_1\pi)/V_\pi$ and $D_{NL}(t)$ is the non-linear distortion contributed by all of the terms in the Taylor expansion of order greater than unity. Hence the original signal $v_{sig}(t)$ may be recovered but is degraded by noise and nonlinear distortion. It can be seen that the nonlinear distortion terms are small when the signal is small relative to V_π , in which case the system will be noise limited. As the magnitude of $v_{sig}(t)$ grows relative to V_π the nonlinear distortion term $D_{NL}(t)$ will increase and the system will eventually become distortion limited. The noise term $N_{Rx}(t)$ includes noise in the received optical signal arising from RIN and other sources as well as noise generated at the PD itself. PD noise sources include thermal noise, shot (or quantum) noise, dark current and surface leakage current.

5.3 Measures of IMDD-Link Performance

Section 5.2 discussed a RoF link based on Intensity-Modulation Direct-Detection optical transmission link. As shown in Figure 5.3, we define a link as consisting of all the hardware required to modulate the RF signal onto the optical carrier, the optical transmission medium, and the hardware required to recover the RF from the optical carrier. By limiting the discussion to amplifier-less links, we more easily see the effects that device performance parameters have on the link parameters. The process of

combining a link with amplifiers involves many tradeoffs which are beyond the scope of study of a basic optical transmission for RoF applications.

The discussion in this section is further limited to basic measures of link performance which affect a wide variety of applications: Gain, Noise Figure (NF), and Dynamic range (DR). Unless otherwise indicate, all parameters studied below are for links operating at optical wavelength of 1.55 μm , which is one of the dominant ones in use because of the availability of optical fiber with low and near-zero dispersion.

5.3.1 Link Gain

It has been shown [14] that the IMDD-link gain can be expressed as a function of the fiber-coupled modulation device and photodetector slope efficiencies (S_{MD} and S_{PD} , with dimensions of W/A and A/W, respectively), modulation device impedance (R_{MD}) and photodetector impedance (R_{PD}), namely

$$g = S_{MD}^2 \cdot S_{PD}^2 (R_{PD} / R_{MD}) \quad (5.17)$$

Expressions for direct and external modulation-link gain are derived by substituting into Eqn. 5.17 the appropriate values of the device slope efficiencies and resistances. In the case of a direct modulation-link, $R_{MD} = R_{LD}$, and S_{MD} is simply S_{LD} , the fiber-coupled external differential quantum efficiency of the directly modulated semiconductor laser; therefore,

$$g_{Dir} = S_{LD}^2 \cdot S_{PD}^2 (R_{PD} / R_{LD}) \quad (5.18)$$

In the case of an external modulation-link, $R_{MD} = R_{Mod}$ (the equivalent parallel resistance of the modulator), and S_{MD} is related to the fiber-to-fiber optical transmission efficiency of the external modulator t_{ff} , the CW input optical power to the modulator P_I , and the modulator switching voltage V_π , resulting in the following expression for gain:

$$g_{ext} = (t_{ff} \cdot P_I \cdot \pi / 2 V_\pi)^2 \cdot S_{PD}^2 \cdot R_{Mod} \cdot R_{PD} \quad (5.19)$$

An important distinction between Eqns. 5.18 and 5.19 is the dependence on average optical power: Eqn. 5.18 for direct modulation is independent of optical power, whereas Eqn. 5.19 for external modulation depends quadratically on optical power. This distinction in optical power dependency has been confirmed experimentally as shown by the direct and external link gain data shown in Figure 5.8 [15].

Figure 5.8: Analytically determined (lines) and measured (points) gain and NF at several values of average detector current for links using various optical modulators

Figure 5.8 depicts the Gain and Noise Figure for the following three different combinations of optical source and modulator:

- Diode-pumped solid-state laser (DPSSL) externally modulated by a LiNbO_3 MZI modulator.
- Directly modulated DFB laser.
- Directly modulated FP laser.

The direct modulation gain is independent of optical power, at least until average power levels are reached where the laser's P versus I curve begins to saturate. In the case of the direct modulation links in Figure 5.8, the gain is higher when a DFB rather than a FP laser is used, simply because the former had higher slope efficiency than the latter.

An important question while using IMDD techniques is whether it is better to use a given LD for direct modulation or to use the same LD as the CW source for an external modulator. At first thought it seems that it is always better to modulate directly, because the unavoidable loss of the modulator will always make external modulation have lower gain. However, Eqns. 5.18 and 5.19 show that if the average optical power is low, external modulation will have lower gain, even if the modulator has no optical insertion loss. Conversely, if the CW optical power is high, then external modulation will yield higher gain than direct modulation, in spite of the additional modulator loss. This idea can be expressed by the crossover optical power $P_{I,cross}$, which yields the same gain for direct and external modulation

$$P_{I,cross} = (2V_{\pi} \cdot S_{LD}) / (t_{ff} \cdot \pi \cdot \sqrt{R_{LD} \cdot R_{Mod}}) \quad (5.20)$$

As can be seen in Eqn. 5.20, $P_{I,cross}$ is not a fixed number, but rather depends upon characteristics of the direct and external modulation-link devices - i.e., the semiconductor laser and the external modulator, respectively. These characteristics, particularly S_L and V_{π} , depend in turn upon the RF frequencies at which the devices were designed to operate, so $P_{I,cross}$ is also frequency dependent to that extent.

5.3.2 Link Noise Figure

NF is a critical parameter in antenna-remoting applications, but less so in distribution applications such as CATV. It can be expressed as

$$NF = 10 \log [N_{out} / kTg] \quad (5.21)$$

where N_{out} is the total noise spectral density at the link output, k is Boltzmann's constant, and $T = 290$ K. Thus, to obtain the effect of an individual noise source on the link NF, the output noise due to that source is divided by the link gain.

In any IMDD link the primary sources of noise are the photo-detected optical intensity noise [usually dominated by laser relative intensity noise (RIN)], shot noise arising from the photo-detection process, and thermal noise arising from the ohmic impedances of the RF source, modulation device, photodetector, and from ohmic losses in the interface

circuitry between the source and modulation device and between the photodetector and the link output. The extent to which RIN, shot noise, and the various sources of thermal noise contribute to N_{out} are affected by the manner in which the modulation device and photodetector are impedance-matched to the link input and output impedances, respectively. However, only when the effects of RIN and shot noise are reduced to a level that is much lower than what has been achieved in most experimental IMDD links does thermal noise constitute a large portion of N_{out} [16].

For most direct modulation links, the laser RIN dominates over the laser shot noise and the thermal noises. The link output noise due to laser RIN can be expressed as

$$|N_{out,RIN}| = I_{PD}^2 \cdot RIN \cdot R_{PD} \quad (5.22)$$

where RIN is the laser relative intensity noise at the analog signal frequency and I_{PD} is the average photocurrent. Eqn. 5.22 shows that noise caused by laser RIN increases as the square of average optical power. In Eqn. 5.18, it was shown that direct modulation-link gain is independent of average optical power. Consequently, dividing a noise source which increases as the square of optical power by a gain which is independent of it yields an NF which increases as the square of optical power.

State-of-the-art external modulation links often use a solid-state laser as the CW optical source. The relaxation frequency of these lasers is typically only a few hundred kilohertz. A laser's RIN spectrum falls off as f^2 above its relaxation frequency. Consequently, the RIN from these lasers has dropped to negligible levels in the lower passband frequency of most links. Therefore, the dominant output noise power-spectral density is due to detector shot noise, which can be expressed as

$$|N_{out,shot}| = 2q \cdot I_{PD} \cdot R_{PD} \quad (5.23)$$

To obtain the effects of shot noise on the NF, N_{out} is divided by the link gain. Dividing a noise term (which depends linearly on average optical power) by a gain expression,

(which depends quadratically on optical power) yields an NF, which decreases linearly with increasing optical power. Thus, despite the fact that increasing the link average optical power generates more shot-noise power at the link output, the effect of this increased shot noise is decreased at the link input because the link gain is increasing faster than the shot noise. This trend can be seen in the external modulation-link data in Figure 5.8.

5.3.3 Intermodulation-free Dynamic Range

The IM-free DR is an important parameter in links where multiple RF frequencies are simultaneously present. The IM-free DR is defined as the maximum difference between the noise floor and the fundamental output which produces distortion terms of equal amplitude to the noise floor. The noise floor in turn depends on the link's instantaneous bandwidth (B), which varies by application. Consequently, to lend general applicability to the IM-free DR measurements, the results are often given in terms of a 1-Hz noise bandwidth. To use such results in a specific application simply requires scaling the 1-Hz data to the receiver's actual instantaneous bandwidth. The bandwidth scaling exponent depends on the order of the dominant distortion. Since many applications have an operational bandwidth less than an octave, the second-order distortion terms can be filtered out. Thus, for such applications, the lowest order in-band distortion is the third order, which causes the IM-free DR to scale as $B^{2/3}$.

Conventional wisdom has held that the IM-free DR of FP lasers is lower than that of DFB lasers - partially because the conventional view is that the FP's RIN is higher, and partially because in early DFB's the electrical isolation of the active region was used to achieve a more linear optical power versus laser-current curve.

The IM-free DR of an external modulation link is dominated by the type of external modulator and the average optical power. However, the IM-free DR for external modulation link using a standard MZM modulator is about the same as for a DFB-based direct modulation link.

Although in some links the photodetector or the fiber itself can contribute measurable distortion, in most links the distortion is dominated by the modulation device.

5.4 System Modeling for RoF Link

A system level diagram of the Radio over Fiber simulation considered in this Thesis is given in Fig. 5.9. The Simulation Model is designed with the considerations of a RoF system used to distribute 802.11a WLAN signals. A radio access point (RAP) is placed in each micro-cell or pico-cell to provide wireless access to the portables and is linked to the central base station via optical fiber as shown. The performance evaluation simulations were performed on the downlink (i.e. Base station to Portable Device) channel with the assumption that the uplink path is essentially a mirror image of the downlink, though additional gain control and amplifier circuitry may be needed at the RAP.

Figure 5.9: RoF System Model

5.4.1 Input Data

The OFDM symbol generation for this model is very similar to the one used in the simulation described in Section 4.5 and shown in Figure 4.9. The OFDM Modulation format employed is summarized in Table 5.1.

FFT Size	64
No. of Data Carriers	48
Modulation	16QAM
Guard Time	800 ns
Net Data Rate	24 Mbps
FEC rate(Convolution Encoder with Hard Decision Viterbi Decoding)	½
Gross Data Rate	48 Mbps
Sub-carrier Bandwidth	312.5 KHz
3dB RF Bandwidth	16.56 MHz

Table 5.1: OFDM Modulation Format

5.4.2 MZM Modulation

The mathematical modeling of the operation of a laser is described through its 'rate equations' [18]. LASER direct modulation results in frequency chirping. A common way to avoid this is to employ external modulation, which is to operate the laser in a continuous wave (CW) mode and then externally modulate the optical field.

A Mach-Zehnder (MZ) interferometer can be employed to modulate an optical signal, as shown in Figure 5.5. From Section 5.2.2, the MZM transfer function can be represented as

$$\psi(v_1(t), v_2(t)) = \frac{1}{2} \exp(j\pi v_1(t)/V_\pi) + \frac{\delta}{2} \exp(j\pi v_2(t)/V_\pi) \tag{5.24}$$

where :

δ is related to the extinction ratio r measured at the output of the MZM as: $(\sqrt{r}-1)/(\sqrt{r}+1)$. The extinction ratio is the ratio between maximum measurable power to minimum measurable power at the output of the device.

5.4.3 Fiber Propagation

Figure 5.10 below shows the detailed fiber propagation as well as photodetection simulation flowchart. In the fiber, the non-linear phenomena were neglected and only the group velocity dispersion was taken into account.

Figure 5.10: Simulation Flowchart Propagation and Photodetection

The effect of attenuation and dispersion in the fiber is applied on the signal in the Fourier domain using the transfer function of the fiber:

$$H_f(f) = \exp((j\pi D \cdot L \cdot (\lambda f)^2 / c) - (\alpha L / 2)) \quad (5.25)$$

where D is the dispersion parameter, L is the fiber length, λ is the transmission wavelength (i.e. 1550nm), c is the speed of light in vacuum and α is the attenuation coefficient.

Note that here we have used a configuration where the amplifier stage follows the fiber rather than the other way around. Although this configuration results in worse noise figures, however, with the adoption of this tactic we can better model a realistic dispersive system with suppressed nonlinear effects.

At the exit of the fiber, incident on the optical pre-amplifier, the optical field $r(t)$ is described by:

$$r(f) = m(f).H_f(f) \quad (5.26)$$

$$r(t) = m(t) * h_f(t) = s(t).\exp(j\phi_s(t)) \quad (5.27)$$

where, $r(f)$ and $m(f)$ are the fourier transforms of $r(t)$ and $m(t)$, respectively. Now, following an input coupling loss c_1 , the field is amplified by the optical preamplifier. Assuming that the amplifier gain spectrum is flat at the operating wavelength, a constant gain G over all frequencies is used. In addition, noise due to amplified spontaneous emission (ASE) is added to the signal. The noise process is described by two zero mean Gaussian distributed random variables, $n_c(t)$ and $n_s(t)$, corresponding to the in-phase and out-of- phase noise components of the bandlimited noise process, respectively. The power spectral density of the ASE noise, per polarization, is given by:

$$N_{ASE} = (G-1) h\nu n_{sp} \quad (5.28)$$

where G is the gain of the amplifier, $h\nu$ is the photon energy and n_{sp} the spontaneous emission factor. The variance (power) of the random variables is determined by N_{ASE} and the filtering process accompanying the amplification stage. At the output of the optical pre amplifier the variance is equal to the amplified spontaneous emission power, proportional to the noise equivalent bandwidth B_o of the optical filter.

$$\sigma^2 = P_{ASE} = (G-1) h\nu n_{sp} B_o \quad (5.29)$$

By adding these two effects of the optical amplifier, gain and noise, the optical field becomes:

$$y(t) = \sqrt{(Gc_1c_2)}.s(t).\exp(j\phi_s(t)) + \sqrt{c_2} (n_c(t) + jn_s(t)) \quad (5.30)$$

Note that c_2 corresponds to the output coupling loss.

Photodetection: The optical power $P(t)$ incident on the photodiode is derived from Eqn. 5.30 as:

$$\begin{aligned} P(t) &= |y(t)|^2 \\ &= 2 \sqrt{(Gc_1).c_2.s(t) (n_c(t)\cos\phi_s(t) + n_s(t)\sin\phi_s(t))} \\ &\quad + Gc_1c_2s^2(t) + c_2 (n_c^2(t) + n_s^2(t)) \end{aligned} \quad (5.31)$$

Finally, the photodetection current generated is incorporated into the model as a Poisson process with intensity:

$$\lambda(t) = (\eta/h\nu).P(t) + \lambda_0 \quad (5.32)$$

where η is the quantum efficiency of the power conversion process and λ_0 is the dark current. This intensity corresponds to the number of electron counts and its relation to the Moment Generating Function $\Phi(s)$ of the detection current is [22]

$$\Phi(s) = \exp\left(\int \lambda(\tau) [\exp(qh_\tau(t - \tau)s) - 1] d\tau\right) \quad (5.33)$$

By substituting Eqns. 5.29 – 5.32 into Eqn. 5.33 and evaluating the first two derivatives at $s = 0$, we get:

$$\Phi'(0) = E\left\{q \cdot \int \lambda(\tau) h(t - \tau) d\tau\right\} \quad (5.34)$$

$$\Phi''(0) = E\left[(q \cdot \int \lambda(\tau) h(t - \tau) d\tau)^2\right] + E\left[q^2 \cdot \int \lambda(\tau) h^2(t - \tau) d\tau\right]$$

Where $E[...]$ is the expectation, conditioned on the signal. The mean $E[I_d]$ of the decision variable is equal to Eqn. 5.34, whereas the variance is:

$$\sigma_{Id}^2 = \Phi''(0) - [\Phi'(0)]^2 = E\left[q^2 \cdot \int \lambda(\tau) h^2(t - \tau) d\tau\right] + \text{Var}\left[q \cdot \int \lambda(\tau) h(t - \tau) d\tau\right] \quad (5.35)$$

Now, by direct substitution of Eqn. 5.31 into Eqns. 5.34 and 5.35, the mean and variance of the decision variable are presented in Eqns. 5.36 and 5.37, respectively.

The mean equals to:

$$E[I_d] = \check{R}_{PD} G c_1 c_2 s^2(t) * h_e(t) + 2c_2 P_{ASE} R_s H_e(0) + q\lambda_0 H_e(0) \quad (5.36)$$

Where $\check{R}_{PD}$ is the responsivity of the diode, and equals $(q\eta/h\nu)$.

The variance is equal to:

$$\begin{aligned} \sigma_{Id}^2 = & q\check{R}_{PD} c_1 c_2 s^2(t) * h_e^2(t) + 2q\check{R}_{PD} c_2 P_{ASE} B_e \\ & + 4\check{R}_{PD}^2 G c_1 c_2^2 N_{ASE} \\ & \times \int |r(f).H_0(f) * H_e^*(f)e^{-j2\pi ft}.H_0(f)|^2 df \\ & + q^2 \lambda_0 B_e + 2\check{R}_{PD}^2 c_2^2 N_{ASE}^2 I_2 \end{aligned} \quad (5.37)$$

The contribution from shot noise and thermal noise has also been reflected in Eqn. 5.37.

5.5 Simulation Results

5.5.1 Link Performance with MZM Characteristics

With the signal output defined by Eqn. 5.15, R_{out} the average noise current, $N_{Rx}(t)$, was selected to give a relatively high SNR of 20 dB. With the input SNR being held constant, the average carrier amplitude $A_{sig,Av}$ of an OFDM input signal $v_{sig}(t)$ was swept through a range of levels. The same input signal was used at each input level. V_π was set to a typical value of 4V.

The MZM is biased at $V_{Bias,Nom} = V_\pi/2$ as in virtually every application in which the device is used. This is the point on the MZM transfer function that is most linear and provides the greatest input/output power efficiency. In Figure 5.11 the mean-squared constellation error MSE_{Const} is plotted as a function of the modulation index Γ_{MZM}

$$\Gamma_{MZM} = A_{sig,Av} / (0.5 V_\pi) \quad (5.38)$$

Figure 5.11: Constellation Error Curve at the Nominal MZM Bias Point

The curve remains essentially constant at a level slightly greater than 10^{-3} for low values of Γ_{MZM} . Constellation degradation increases near $\Gamma_{MZM} = 0.45$ when the MSE_{Const} increases rapidly. At this point the signal degradation caused by noise is at a minimum, so this is the optimal operating point for the link.

Figure 5.12 shows the average optical intensity at the Output of the MZM for a given Γ_{MZM} . The plot also shows the input signal amplitude as a function of the modulation index. From the results we can see that the received SNR would be significantly increased with an increase in the modulation index above the optimal operating point.

Figure 5.12: Signal Input Amplitude and Output Optical Intensity (nominal bias point)

5.5.2 Effect of Fiber Length on BER Performance

In Figure 5.13 the simulation results depict the dependence of BER on Fiber length of a SMF. Simulation results are depicted without the use of any error correction as well as by using FEC (constraint length 7).

Figure 5.13: BER Dependence on Fiber Length

For the simulated RoF system model, we can see that an Input Power of -14 to -16 dBm would be ideal to achieve a BER of 10^{-3} to 10^{-4} for a fiber length of less than 1 km, without the use of FEC. For the deployment of a local WLAN system, this would be a reasonable estimation of fiber length. From Figure 5.13 we can also expect little performance degradation (WRT Tx Power) for the longer fiber lengths of 2 and 4 km. While using FEC, a performance gain of around 4 dBm is achieved for fiber lengths of 1 km and 2 km. A similar performance improvement is achieved for a fiber length of 4 km.

5.5.3 Transmission Performance and Receiver Sensitivity

To study the effects of Fiber Dispersion on the signal received at the photodiode, the optical signal was simulated to pass through varying lengths of the SMF model with a Dispersive Parameter $D = 17$ ps/nm.km. At the receiver the signal was pre-amplified and detected. Figure 5.14 shows the measured receiver sensitivities versus the accumulated dispersion. The results shown in Figure 5.14 can also be read for fiber lengths varying from 500m (dispersion = 8.5 ps/nm) to 5 km (dispersion = 85 ps/nm).

Figure 5.14: Receiver Sensitivity (at BER 10^{-4}) vs Fiber Dispersion

The simulation results show that Fiber dispersion has a limited effect on the receiver sensitivity for fiber lengths < 2-3 km. Only when the unrepeatd fiber length exceeds 3 km do we see any reasonable performance degradation of the system.

The measured BER plotted as a function of received optical power level is shown in Figure 5.15.

Figure 5.15: Receiver Power Performance for 1km Fiber Transmission

The receiver simulation results (when compared to Figure 5.13) show a power loss of 3-4 dBm for transmission over 1 km length of SMF. From the simulations, it is evident that the receiver sensitivity can be further improved with the use of the FEC techniques of IEEE 802.11a. A receiver power level of -21 to -19 dBm is required for achieving the target BER performance from the system with the use of FEC.

5.5.4 Effects of MZM Nonlinear Distortion

Earlier the output of the photodetector was given in Eqn. 5.15. For the interest of observing the nonlinear distortion caused by the MZM we set $K_I = 1$ and $N_{Rx}(t) = 0$, allowing us to rewrite Eqn. 5.15 as

$$v_{Rx}(t) = \sin(2\pi v_{sig}(t)/V_\pi) \quad (5.39)$$

Considering a net voltage transition in the input signal between time t_0 and t_1

$$\begin{aligned} \Delta V_{Tx} &= v_{sig}(t_1) - v_{sig}(t_0) \\ &= V_1 - V_0 \end{aligned} \quad (5.40)$$

The received signal undergoes a net voltage change of

$$\Delta V_{Rx} = \sin(2\pi V_1/V_\pi) - \sin(2\pi V_0/V_\pi) \quad (5.41)$$

The transfer function defining the relationship between the transmitted and received voltage transitions is

$$G_{OL}(V_0, V_1) = \Delta V_{Rx}/\Delta V_{Tx} \quad (5.42)$$

Figure 5.16 shows the comparison between the linear and nonlinear MZM transfer functions.

Figure 5.16: MZM Nonlinear Transfer Function and Linear Approximation

For a fixed value of ΔV_{Tx} , the distortion is minimized when the transition is symmetrical about the bias point. By extension the entire signal will clearly suffer from the least distortion if it is symmetrical about the nominal bias point. This is because the transfer function is the most linear about the nominal bias point as shown in Figure 5.16.

Referring back to section 5.2.2, the optical power output $P_{o,MZM}$ by an MZM is a sinusoidal function of the MZM input voltage V_{MZM} having a period of $2V_\pi$. This optical output power P_o , plotted in Figure 5.6 for a nominally-biased MZM, may be expressed as

$$P_{o,MZM}(V_{MZM}) = 0.5P_{o,LD} (\sin(\pi V_{MZM}/V_\pi) + 1) \quad (5.43)$$

where $P_{o,LD}$ is the optical power output on the laser diode. If an input voltage V_{in} is applied directly to the MZM, $V_{MZM} = V_{in}$ so $P_{o,MZM}$ is a sinusoidal function of V_{in} as shown in Figure 5.17.

Figure 5.17: MZM Power Output $P_{o,MZM}$ emitted as function of Input Voltage V_{in}

When no other distortions are present on the optical link the output of the link V_{out} is directly proportional to $P_{o,MZM}$. V_{out} is therefore a sinusoidal, rather than linear, function of V_{in} .

5.6 Conclusions

For a RoF installation of WLAN we have seen that a linear fiber model causes only small distortions as the fiber length increases. Simulation of a basic RoF transmission system has helped us better understand the performance of such a system with respect to the power requirements of optical modulator and the photodetector.

6. Results and Discussions

The work done in this thesis has been focused towards merging two promising technologies that are presently deployed and have even greater future potential. If OFDM based 802.11a wireless networks can be successfully integrated with Radio-on-Fiber (RoF) technology it will be possible to deploy vast cost-effective wireless networks with performance that is superior than most WLAN networks available today. This study is concerned with development of strong simulation techniques to investigate the impact of fiber-radio systems on the performance of 802.11a Wireless LAN signals. This work has shown that the fundamental concept of transmitting WLAN signals over RoF is sound, so such networks may become a reality as both OFDM and RoF technologies continue to evolve.

6.1 *Summary of Research Results*

The basic principles and characteristics of OFDM signals and communication systems were reviewed in **Chapter 3** and an OFDM link has been confirmed to work by using computer simulations. So far three main performance criteria have been tested, which are OFDM's tolerance to multipath delay spread, channel noise and peak power clipping. The modulation techniques for OFDM that were investigated in this section included BPSK, QPSK and 16PSK.

It was shown that with properly designed guard interval, OFDM is capable of handling very strong echoes. The BER improvement, which resulted from the single echo, was indicated by computer simulations. With the capability of withstanding strong multipath propagation, OFDM would allow the continued use of omni-directional antennas in urban WLAN systems where C/N is sufficiently high. The improvement in BER performance through a Multipath fading channel was also demonstrated especially in the case of using a cyclic extension for the transmitted OFDM symbols. The preference for using a cyclic extension structure for the guard period originates from its ability to maintain receiver carrier synchronization and the fact that cyclic convolution can still be applied between the OFDM signal and the channel response to model the transmission system.

Simulation results also showed that OFDM signals can have the peak power clipped to reduce the PAPR without a significant increase in the BER, hence allowing for a greater transmitted power. In common practice, signals could be clipped because of limited quantization levels, rounding and truncating during the FFT computation as well as other distribution parameters after D/A conversion. It is safe to say that the Gaussian model can be used as the upper bound for the OFDM signals.

One important area, which hasn't been investigated, concerns the problems that may be encountered when OFDM is used in a multi-user environment. One possible problem is that the receiver may require a very large dynamic range in order to handle the large signal strength variation between users.

Chapter 4 details the simulation model developed to reflect the 802.11a PHY layer transmission. The model is able to simulate the mandatory IEEE 802.11a compliant data rates using four different modulation techniques (BPSK, QPSK, 16QAM, 64QAM). The BER performance accounts for errors in the data portion of the packet under the assumption that header are always correctly received. The simulations were performed for multipath fading channels having an rms time delay spread of 200 nsec and convolution encoders with constraint length 7. For performance evaluation without the use of FEC, we discovered that the required system SNR was greater than 25 dB.

For the performance with convolution coding and Viterbi HDD, the 802.11a system performance is improved significantly by adding FEC coding. The coded system shows a direct performance advantage of 10-15 dB as compared to the uncoded results. When expressed in terms of absolute performance at $P_b = 10^{-3}$, which represents the low end of acceptable performance for WLAN applications, the absolute performance in SNR ranges from 6 to 20 dB approximately. Also, as expected, for a specific modulation type, regardless of the channel conditions, the SNR required for achieving a fixed probability of bit error increases as the code rate increases.

Difficulty was encountered in developing Matlab code for Viterbi decoding algorithms. The implementation and testing time for developing the Matlab code exceeded the initial estimates and led to delays in completing the simulations. After software code was developed for a constraint length of $K = 3$, extension of the algorithm for other, higher constraint lengths became easier. It was also felt that the computational complexity of the Viterbi algorithm is very high for the convolution encoder prescribed by the IEEE 802.11a standard.

The simulation results shown in **Chapter 5** concentrate on the performance of a basic RoF transmission system for 802.11a WLAN signals. The performance was studied under the presence of common optical noise contributors such as RIN, shot noise, fiber dispersion and thermal noise. From the simulation of MZM modulation we were able to identify the optimal operating modulation index for the MZM. For the deployment of a local WLAN distributions system the results for transmit power requirement and receiver sensitivity show that a fiber span of 2-3 km would be able to deliver expected BER performance results for the system. A simulation of fiber dispersion shows that it is not the primary factor in performance degradation of the system. With a predicted receiver sensitivity of -16 to -14 dBm and only fiber attenuation to contend with, the transmit optical power required for a BER of 10^{-4} would be in the range of -10 to -12 dBm.

In summary the research activities in this project have led to the development of a complete Simulation Model to study the performance of a RoF link to deliver IEEE 802.11a Wireless LAN signals under various channel conditions and prescribed digital modulation formats.

6.2 Impact of Radio-over-Fiber Links on Wireless LAN Protocols (The Problem of Extra Propagation Delay)

In a RoF network utilized in a wireless access system, the optical link behaves as an 'analog transparent' distribution system that shall not modify the radio signal format, but deliver it with as high a performance as possible at the remote antenna location, which

acts as a dummy analog repeater-interface between the optical and the radio media. Since, the radius of coverage in radio systems is mainly set by power constraints (maximum transmit power, propagation losses, interference and receiver sensitivity), these systems are usually built on the basis of these limitations, and it is the medium access, link control and other upper layer protocols that set the delay and timing boundaries, in order to make an optimal usage of the radio resources.

In addition to the generation and transport of the microwave signals from the Central Site (CS) to the Remote Access Point (RAP) and vice versa, the impact of inserting an optical path within the wireless system protocol stack becomes a highly relevant issue from a network system point of view. The RoF link depicted below (Figure 6.1) treats the optical domain as an extension of the radio access domain.

Figure 6.1: Wireless Protocol Stack for RoF Link

When inserting a fiber link between the CS and the RAP to support a Wireless LAN network, multiple radio access and radio duplexing parameters become key requirements for the design of the RoF link, since the extra propagation delay added by the fiber path,

might outrun the timing boundaries of the medium access protocols and the round trip delay.

Wireless LAN standard IEEE 802.11a specifies a distributed medium access control protocol called Carrier-Sense Multiple Access with Collision Avoidance (CSMA/CA), in which the medium is seized by one station during a contention period performing a back-off procedure. The medium is held by this station until the transmission of one frame and its acknowledgement (ACK), if required have finished. After transmitting a frame that requires an ACK response, the source station (CS) shall wait for an ACK Timeout interval; if the ACK response does not occur during this interval, the CS concludes that the transmission of the frame has failed, and this station will invoke its back-off procedure. When a long fiber path is included in the WLAN link, an extra propagation delay occurs for all frame transfers and this may lead to the expiration of the ACK timeout before a transmitted ACK response is received by the Central Site. To overcome this timing impairment, the ACK timeout interval can be incremented so that the station waits a longer period to receive the acknowledgement.

According to the 802.11a PHY specifications, the back-off procedure has the required slotted behaviour as long as the propagation delay ($aAirPropagationTime \ll 1 \mu s$) remains much less than the slot time ($aSlotTime = 9 \mu s$). An operating limit can be placed on the propagation delay as $0.1 \mu s$. For introducing an optical path in the RAP, the parameter $aAirPropagationTime$ needs to be shared for the delays of both radio and optical channels. Assuming the limit on propagation delay ($0.1 \mu s$) the maximum coverage distance allowed in the radio interface is 30 m. In the optical domain, the maximum fiber length allowed for the same propagation delay will be approx. 20m, for a refractive index typical value of 1.5. If a fiber path of e.g. 1 km is inserted, it cannot be considered as a slotted access behavior anymore, because the total propagation delay is not much less than the slot time defined by the standard. If the fiber path is 2 km long, the access point cannot perform its backoff procedure properly, since its attempts to seize the medium will appear one slot too late for the rest of stations, and the other way around, if one station seizes the medium, the access point will be able to see the medium is 'busy'

one slot too late, thus, increasing the probability of collision. Combining both radio and fiber delays, under present propagation time specifications, yields very short fiber lengths and radio coverage, which may not be enough for networking applications. In order to increase the radio coverage and optical path length of the system, a solution would be increasing the allowed propagation delay over the radio and fiber paths. However, this will imply a greater time interval between frames and hence a reduction of the effective transmitting time and the mean user data throughput.

6.3 *Future Directions*

Based on the experience and knowledge acquired during this research work, several areas for future research can be suggested. These are:

- Performing a real-time feasibility test using off-the-shelf radios using an MZM-based optical link. It is anticipated that such a test should be successful provided the power of the signal driving the MZM is set appropriately. Similar tests have been successfully conducted with 802.11b already, and it should be possible to incorporate a wireless portion to the complete link. Since 802.11a differs from the 802.11b standard only at the physical layer, the network topology already used for RoF deployments [19] should work for the new standard.
- Extension of the 802.11a PHY simulation model to include Soft Viterbi Decoding which can provide 2 dB better performance than Hard Decision Decoding. However, the computational complexity is even higher for soft decision decoding. The 802.11a PHY simulation presented in this thesis incorporates the effects of such impairments as AWGN and fading multipath channel while assuming perfect timing synchronization and no phase offset. The simulation model could be extended to include other impairments such as frequency offset, synchronization errors, phase imbalance and phase noise in the future.
- Extension of the RoF simulation model to further investigate the impact of fiber-radio system non-linearities on the performance of OFDM signals. Powerful

analytical techniques would be required to model both the frequency independent and frequency dependent non-linear behaviour typical of fiber-radio system nonlinearities. Some research effort has already gone into the development of analytic techniques to assess the impact of nonlinearities on the performance of OFDM signals [37-39]. However, the techniques discussed there focus on the performance of OFDM signals distorted by clipping devices and high power amplifiers (HPA), which exhibit frequency independent non-linear behaviour. This generally renders them incomplete to determine the performance of OFDM signals distorted by fiber-radio system elements, where frequency dependent non-linear behaviour is of greater concern. Accordingly, new developments in the area of modeling and analysis of OFDM-RoF communication systems in the presence of nonlinearities are required.

- Gaining a more solid understanding of the influence of nonlinearities on the performance of 802.11a OFDM signals, to help study and simulate the BER performance of a RoF OFDM link. Studies have somewhat surprisingly shown that, in the presence of nonlinearities, the error probability performance of OFDM signals with higher PAPR can be better than the error probability performance of OFDM signals with lower PAPR, because the probability of error of an OFDM signal is ultimately related to the Intermodulation Distortion (IMD) behaviour and not to the PAPR behaviour [41, 42]. This suggests that it may be more beneficial to rely on IMD minimization techniques [45] instead of PAPR reduction techniques [43, 44] to improve the performance of nonlinearly distorted OFDM signals. The feasibility of development of IMD reduction codes could be a subject of further work. These codes introduce a set of redundant symbols in the set of data symbols to cancel out different Intermodulation Products (IMPs) upon OFDM signal non-linear distortion.
- Investigating the use of Dynamic Cyclic Prefix length for OFDM signal transmission. The current standards are too robust with the channel impairments since the channel is strictly taken as stationary indoor channel. By using the term

stationary it is meant that between few OFDM symbols, the channel is expected to become static, or not changing. Thus, it may be wise to measure the channel maximum delays and insert the CP according to the measurement. In that case, a technique has to be found to measure the maximum channel delays. This can be done using the pilots that are inserted for channel estimation and synchronization process.

- Further investigation into solutions for the propagation delay limitations placed by 802.11a, which severely limits the length of deployable optical links. The solution for this problem may well lie in the modification of WLAN standards to optionally tolerate longer propagation delays to facilitate their direct interface with RoF systems for applications such as these. Another possibility for further research is the use of IEEE 802.16a that is more tolerant than 802.11a of the delay spreads and propagation times because of its intended application for long-haul wireless links. This suggests that future integration of RoF and WLAN technologies may wish to focus on the OFDM physical layer of the IEEE 802.16a standard. However, the compatibility of 802.11a with existing wireless LANs indicates that it should not be dismissed in favour of 802.16a if the former standard can cope with application-specific channel parameters. The delay spread can be minimized by using a very large number of remote antennas to keep the wireless links short.

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