Rate-Adaptive Code Combining over Time and Space for Wireless Radio Frequency and Infrared Communication Systems

A Thesis in
Electrical Engineering
by
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ABSTRACT

This dissertation deals primarily with the design and analysis of various rate-adaptive packet transmission schemes intended for wireless radio and infrared channels. To make the communications reliable and robust, all the proposed schemes take advantage of some form of diversity combining. Diversity, in general, can be provided in many dimensions including time and space. Usual ways of creating time diversity is through retransmission, coding and interleaving, whereas the most common way of achieving space diversity is by using antenna arrays at the transmitter and/or receiver. Although increasing the diversity order in only one dimension is sufficient for increasing the system capacity, it does not always result in an ideal solution as there are many practical considerations such as system complexity, underlying channel fading characteristics, power consumption, delay, etc. that favor use of one diversity scheme over another. It seems therefore reasonable to combine diversity schemes of different dimensions with one another to arrive at a practical solution. This is the primary focus of this thesis. In addition, to further increase the capacity, we consider some practical rate-adaptive coding (modulation) schemes to be used along with the proposed space and time diversity schemes. The first part of this thesis is dedicated to applying the above concepts to some well-known radio channels, having the goal of providing a guaranteed Quality-of-Service (QoS) under all channel conditions. The second part of this work is dedicated to indoor wireless infrared channels. In these channels, it is possible to provide space diversity in a manner somewhat different from the way it is provided in radio systems, and for this reason, it is commonly referred to as direction (or angle) diversity. We first investigate the properties of our proposed infrared channel and look into some of its
deterministic as well as statistical characteristics. Based on these characteristics we propose efficient modulation, coding and combining schemes to achieve the desired goal of QoS provisioning at all possible positions within the coverage area. Coding in time as well as in space, in this case, allows for having a low power wireless local-area network that can operate at very high bit rates.
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<td>ABR</td>
<td>Available Bit Rate</td>
</tr>
<tr>
<td>ARQ</td>
<td>Automatic Repeat Request</td>
</tr>
<tr>
<td>ASIC</td>
<td>Application-Specific Integrated Circuit</td>
</tr>
<tr>
<td>ATM</td>
<td>Asynchronous Transfer Mode</td>
</tr>
<tr>
<td>AWGN</td>
<td>Additive White Gaussian Noise</td>
</tr>
<tr>
<td>$A_{det}$</td>
<td>Area of the Photodetector</td>
</tr>
<tr>
<td>$a(d)$, $a_d$</td>
<td>Distance Spectrum of a Convolutional Code</td>
</tr>
<tr>
<td>$a[n]$</td>
<td>Binary Information Sequence</td>
</tr>
<tr>
<td>$\alpha$</td>
<td>Roll-off Factor of a Raised-Cosine Function</td>
</tr>
<tr>
<td>BER</td>
<td>Bit Error Rate BPSK Binary Phase Shift Keying</td>
</tr>
<tr>
<td>BW</td>
<td>Bandwidth</td>
</tr>
<tr>
<td>CBR</td>
<td>Constant Bit Rate</td>
</tr>
<tr>
<td>CGH</td>
<td>Computer Generated Hologram</td>
</tr>
<tr>
<td>CLR</td>
<td>Cell Loss Ratio</td>
</tr>
<tr>
<td>CRC</td>
<td>Cyclic Redundancy Check</td>
</tr>
<tr>
<td>CSI</td>
<td>Channel State Information</td>
</tr>
<tr>
<td>$C$</td>
<td>Channel Capacity</td>
</tr>
<tr>
<td>$c_d$</td>
<td>Number of Paths having a Hamming Distance $d$ with the All-zero Path</td>
</tr>
<tr>
<td>$c[n]$</td>
<td>Sequence of Coded Symbols</td>
</tr>
<tr>
<td>DFE</td>
<td>Decision Feedback Equalizer</td>
</tr>
<tr>
<td>D-LOS</td>
<td>Directed Line-of-Sight</td>
</tr>
<tr>
<td>$D$</td>
<td>Delay Spread</td>
</tr>
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$d_{\text{free}}$  
Free Distance

$d_{\text{min}}$  
Minimum Hamming Distance

$\delta$  
Number of Zeros in Puncturing Array

$\delta(t)$  
Impulse Function

EGC  
Equal Gain Combining

$E_b$  
Information Bit Energy at the Receiver

$E_s$  
Symbol Energy at the Receiver

$e$  
Number of Erasure Symbols

$\eta$  
Throughput

FEC  
Forward Error Correction

FOV  
Field-of-View

$f_{3\text{dB}}$  
3-dB Bandwidth

$\Phi_b$  
Branch Field-of-View

GSM  
Global System for Mobile

$G_{rc}(f)$  
Raised-cosine Frequency Response

$g(x), G(x), G(D)$  
Generator Polynomial

$\gamma_s$  
SNR per Symbol

HDD  
Hard-decision Decoding

HEC  
Header Error Control

$H(0)$  
Path Loss

$h(t)$  
Channel Impulse Response

IM  
Intensity Modulation

IM/DD  
Intensity Modulation and Direct Detection

IR  
Infrared

IrDA  
Infrared Data Association

ISI  
Intersymbol Interference
<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
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<tbody>
<tr>
<td>(i_b)</td>
<td>Irradiance of Background Light on the Detector Surface</td>
</tr>
<tr>
<td>(J)</td>
<td>Diversity order</td>
</tr>
<tr>
<td>LD</td>
<td>Laser Diode</td>
</tr>
<tr>
<td>LED</td>
<td>Light-emitting Diode</td>
</tr>
<tr>
<td>LOS</td>
<td>Line-of-Sight</td>
</tr>
<tr>
<td>(L)</td>
<td>PPM Modulation Level</td>
</tr>
<tr>
<td>(L)</td>
<td>Number of Retransmissions</td>
</tr>
<tr>
<td>MDS</td>
<td>Maximum Distance Separable</td>
</tr>
<tr>
<td>ML</td>
<td>Maximum Likelihood</td>
</tr>
<tr>
<td>MLC</td>
<td>Maximum-likelihood Combining</td>
</tr>
<tr>
<td>MLSD</td>
<td>Maximum-likelihood Sequence Detection</td>
</tr>
<tr>
<td>MRC</td>
<td>Maximal-ratio Combining</td>
</tr>
<tr>
<td>MSDC</td>
<td>Multiple-spot Diffusing Configuration</td>
</tr>
<tr>
<td>(M^{(j)})</td>
<td>Maximum-likelihood Metric for (J) Receiving Packets</td>
</tr>
<tr>
<td>(\mu)</td>
<td>Mean of a Random Variable</td>
</tr>
<tr>
<td>ND-NLOS</td>
<td>Nondirected Non-Line-of-Sight</td>
</tr>
<tr>
<td>(N_0)</td>
<td>One-sided Noise Power Spectral Density</td>
</tr>
<tr>
<td>(n(t))</td>
<td>Noise Signal</td>
</tr>
<tr>
<td>OOK</td>
<td>On-off Keying</td>
</tr>
<tr>
<td>PPM</td>
<td>Pulse-position Modulation</td>
</tr>
<tr>
<td>PSD</td>
<td>Power Spectral Density</td>
</tr>
<tr>
<td>(P, p)</td>
<td>Puncturing Period</td>
</tr>
<tr>
<td>(P)</td>
<td>Optical Gain Factor</td>
</tr>
<tr>
<td>(P_\delta)</td>
<td>Puncturing Array</td>
</tr>
<tr>
<td>(P_b)</td>
<td>Bit Error Rate</td>
</tr>
<tr>
<td>(P_{dc})</td>
<td>DC bias</td>
</tr>
</tbody>
</table>
\( P_K \) Probability of PPM Symbol Error
\( P_F \) Outage Probability
\( P_L \) Probability of Discarding a Cell (CLR)
\( P_{\text{packet}} \) Probability of Packet Error
\( P_q \) RS Code Symbol Error Probability
\( P_r \) Average Received Optical Power
\( P_t \) Average Transmitted Optical Power
\( P(\theta) \) Lambertian Angular Distribution
\( P_2(d) \) Pairwise Probability of Error
\( \text{QoS} \) Quality-of-Service
\( \text{QPSK} \) Quadrature Phase Shift Keying
\( q \) Electron Charge
\( \text{RS} \) Reed-Solomon
\( R \) Code Rate
\( R_b \) Information Rate
\( R_s \) Symbol Rate
\( r \) Photodetector Resistivity
\( \text{SDD} \) Soft-decision Decoding
\( \text{SD} \) Selection Diversity
\( \text{SNR} \) Signal-to-Noise Ratio
\( \hat{\text{SNR}} \) Estimate of the SNR
\( S^{\text{shot}}(f) \) Shot Noise Power Spectral Density
\( \sigma \) Standard Deviation
\( \text{TCM} \) Trellis Coded Modulation
\( \text{TDMA} \) Time Division Multiple Access
\( T \) Symbol Period
\( T_c \)  Chip Period
\( T(X) \)  Generating Function
\( T(d_1, d_2) \)  Generalized Transfer Function
\( Tr \)  Expected Number of Transmission Attempts
UEP  Unequal Error Protection
UMTS  Universal Mobile Telecommunication System
VLSI  Very Large Scale Integrated Circuit
\( y(t) \)  Photocurrent at the Photodetector
\( W_j \)  Reliability Factor at the \( j \)th Branch
WATM  Wireless ATM
WMF  Whitened Matched Filter
\( X(t) \)  Instantaneous Optical Power
ACKNOWLEDGMENTS

The work presented here could not have been accomplished without the help and encouragement of my advisor, Professor Mohsen Kavehrad. I am thankful for his teaching, guidance and friendship. His emphasis on making a good balance between theory and practice has been my guideline during my research. I would also like to thank the other committee members, Dr. Lyn Carpenter, Dr. Hao Che, Dr. Shizhou Yin and Dr. Rajeev Sharma for serving on my committee.

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Chapter 1

INTRODUCTION

1.1 Motivation

The reliability of data transmission systems can be increased through the use of error control coding. Coding a packet of information is simply adding redundant information to the packet which can be used by the receiver to correct and/or detect errors induced by the channel. Correction of errors in the received packets is referred to as forward error correction (FEC).

Conventional FEC coding systems transmit packets at a fixed rate regardless of the channel error rate. As the error rate increases, the number of decoder errors increases and hence the reliability of the decoded data decreases. For these types of channels a more complex error control system is required. The code rate at which the data is encoded must change in order to reflect the changing conditions of the channel.

Errors, in general, occur when either the channel attenuation is large or (equivalently) when the noise or interference level at the receiver is high. Although use of variable-rate FEC coding schemes proves to be effective in many cases, there may still be circumstances in which even a very powerful coding scheme is not able to correct all the received errors. If we can supply to the receiver several replicas of the same information signals transmitted over independent channels,
the probability that all the received signal components will fade simultaneously is reduced considerably. This is the essence of diversity.

There are several ways of implementing diversity. One method is to employ time diversity, in which a data frame is transmitted over a number of different time slots. It turns out that FEC coding is a special and implicit (and yet very effective) way of achieving time diversity. Another method is to use space diversity. This can be achieved using multiple antennas at the transmitter and/or the receiver to create a number of independent channels. Diversity schemes, can, in theory, improve the system performance through increasing the capacity and reducing the required transmitted power. It, therefore, becomes a matter of whether a particular diversity technique can respond to all the physical constraints of a system and whether it can satisfy the requirements given by a particular application intended for that system. For instance, delay-sensitive applications may not allow the use of some time diversity techniques, while other considerations such as complexity and cost may prohibit the use of space diversity techniques. Therefore, practical limitations may favor use of one diversity scheme over another. Meanwhile, it is reasonable to expect that, the ideal solution would be to use a combination of diversity schemes of different types and orders.

1.2 Objective

We intend to use rate-adaptive coding (and modulation) schemes along with space and time diversity schemes in an optimal and yet practical way to achieve a high capacity. We will first apply these concepts to some well-known radio channels. We take the example of a wireless ATM network, for which, a guaranteed quality-of-service (QoS) for different applications that run over the network needs be provided. We then propose efficient transmission strategies that would meet those
QoS requirements and at the same time use the channel resources, i.e., available bandwidth, efficiently.

The second part of this work is dedicated to indoor wireless infrared channels. In these channels, it is possible to provide space diversity in a manner somewhat different from the way it is achieved in radio systems. We first investigate the properties of our proposed infrared channel and identify its deterministic as well as statistical characteristics. Based on these characteristics, we propose efficient modulation, coding and combining schemes to achieve the desired goal of QoS provisioning at all possible positions within the coverage area.

A common feature in all the communication systems considered in this thesis is the use of maximum-likelihood criterion, to decode either a single received packet, or a combination of repeated copies of that packet. When convolutional coding is utilized, the maximum-likelihood decoder can be implemented based on Viterbi algorithm. Use of low-complexity adaptive-rate schemes is another common feature in these systems. In all cases, it is assumed that channel conditions are known, apriori, at the receiver and in some cases at the transmitter, as well. Accurate channel estimation is of primary importance in a successful implementation of any rate-adaptive algorithm. We will develop a simple method for estimating the channel using the Viterbi decoder side information. Based on this information, the transmitter chooses the right code in the family of available codes and the receiver adjusts its parameters to allow for maximum likelihood decoding of the received data.

1.3 Thesis Overview

This thesis is divided into eight chapters, including this introduction. In Chapter 2 a family of codes with different rates, known as, rate-compatible punctured convolutional (RCPC) codes [5] is used in a channel with variable error rates,
characterized by lognormal distribution. This model is used for many fixed wireless communication links. Based on the channel state information at the transmitter, the system adaptively and intelligently selects a code in the family to accommodate for protection needs of the transmitting data. The selectable-rate property of the encoder allows for dynamic bandwidth allocation, thereby increasing the overall capacity of the system. This scheme is particularly suitable for ATM networks with wireless links. It provides a means for maintaining QoS guarantees and it remains least demanding on the use of available frequency spectrum.

A different adaptive rate approach would be through retransmitting coded packets with some form of diversity combining. In conventional retransmission systems a receiver requests a retransmission if the received packet cannot be reliably decoded. Original packet is discarded when the retransmitted packet arrives and decoding is performed only on the most recently received packet. In packet combining systems, receiver forms increasingly reliable estimates of the transmitting packet by combining all repeated transmissions of the packet. Chase [6] proposed a method for combining packets via a maximum-likelihood approach. This method allows for reasonable throughput and reliability over most channels. In Chapter 3, maximum-likelihood code combining technique for two important retransmission schemes are developed, tailored to the Viterbi algorithm. Time-varying channel in this chapter is modeled as a slow, flat fading Rayleigh channel. Each packet in this model experiences a constant, independent AWGN channel. Higher orders of diversity are provided, only when needed, by retransmitting repeated copies of a packet.

The technique of maximum likelihood code combining can be further extended to communication systems that operate based on space diversity techniques. A good example of such systems is a wireless infrared communication system which employs
some form of space diversity at the receiver. This is outlined in chapter 4. Some background materials are first provided. Then a novel link design is introduced and thoroughly examined which utilizes multi-spot diffusing configuration (MSDC) in conjunction with multi-branch direction-diversity reception.

The utilization of MSDC along with direction-diversity reception allows for modeling the system as one having a number of separate AWGN channels. Having such a model enables us to apply code combining techniques under various coding and modulation schemes and receiver designs, and evaluate the transmit optical power requirement and the overall system performance, accordingly. Maximum-likelihood code combining in this application represents a technique for combining entire packets, rather than combining just individual symbols as used in conventional diversity schemes. It offers an added dimension to the conventional diversity concepts, thereby improving the overall system performance.

In chapter 5, rate-compatible punctured convolutional codes are used to encode intensity modulated (IM) optical power, to provide an adaptive environment for efficient utilization of channel spectral bandwidth, and to maintain a guaranteed bit-error rate (BER) performance at all receiver positions in a room, being covered. The coding scheme presented in this chapter, offers two attractive features. First, it allows for the design of a low-cost adaptive-rate system since only one convolutional encoder at the transmitter and one Viterbi decoder at the receiver are required. Second, the channel in this case can be accurately estimated using the side information provided by Viterbi decoder. Based on the estimated signal-to-noise ratio, the transmitter and receiver select a code in the RCPC code family with the smallest amount of redundancy that can keep the BER below a target value.

Despite their unique features, RCPC codes do not seem to be the most efficient codes for intensity-modulated (IM) channels. This is because IM channels have some properties that are different from the more typical electrical channels. The
primary concern in these channels is the average optical power which is the average of the transmitted signal amplitude. This suggests that a modulation scheme with a low duty cycle, or equivalently, high peak-to-average ratio be used. An example of such a scheme is $L$-ary pulse-position modulation ($L$-PPM). An immediate question is: What is the achievable rate at some given SNR using $L$-PPM? In chapter 6 we answer to this question by computing the capacity of IM channels subject to $L$-PPM modulation format. The analysis is performed for both hard-decision decoding as well as soft-decision decoding. It is shown that $L$-PPM has a significant capacity at low values of SNR which motivates one to search for efficient codes that can be used along with $L$-PPM modulation.

In chapter 7, we use high-rate Reed-Solomon (RS) codes to further increase the power efficiency of PPM signals. The RS-coded PPM signals are transmitted over several independent channels. Maximum-likelihood code combining is used, once again, at the receiver to recover the PPM symbols. The proposed system can be made rate-adaptive through varying modulation level $L$ and/or code rate $R$ without increasing the complexity, significantly. This provides a dynamic range large enough to allow portable terminals to communicate at their highest permitted data rate, without sacrificing quality-of-service. It is shown that very high data rates, up to hundreds of megabits per second, can be reached with high reliability everywhere within the coverage area, using transmit optical power levels well below one Watt.

1.4 Summary of Contributions

The following is a list of publications as a result of the work described in this dissertation.
Journal Publications


Conference Proceedings


Chapter 2

QoS PROVISIONING FOR WIRELESS ATM USING PUNCTURED CONVOLUTIONAL CODES AND VITERBI DECODER

This chapter presents a solution to the problem of providing Quality-of-Service (QoS) guarantees in an ATM network with wireless links. The suggested resolution to this problem is through selectable-rate channel coding. In particular, rate-compatible punctured convolutional (RCPC) codes are employed due to their several attractive features that make them favorable for ATM-based applications. In this adaptive scheme, the RCPC code rate is intelligently varied in response to the channel’s nonstationary behavior according to the channel state information (CSI). This yields an equivalent stationary channel with a somewhat low and constant bit-error rate (BER) in which QoS can be guaranteed. In order to demonstrate the performance of RCPC codes in maintaining the QoS, a lognormal fading channel model is assumed. A system block diagram for the wireless link is proposed. This system assumes that accurate CSI is available at the transmitter as well as at the receiver. Bandwidth utilization of the proposed system under various channel conditions is then evaluated, plotted, and discussed.

2.1 Introduction

Asynchronous Transfer Mode (ATM) has long been advocated as a leading technology in providing high-quality switching. Recently, there has been a considerable
focus on the topic of Wireless ATM (WATM). As the demand for broadband services in wireline ATM networks increases and as the use of laptops, notebooks and palmtops become more pervasive, it becomes even more sensible that researchers examine the feasibility of extending the ATM virtual connectivity from the wireline to the wireless domain [1].

Like ATM, and unlike other competing technologies, wireless ATM must be capable of supporting a variety of telecommunications services through the ability of providing Quality-of-Service (QoS) guarantees to individual connections. This will require a considerable amount of bandwidth if the rigid-partitioning approach of second-generation digital cellular circuit-switching systems are employed. An example of such an approach is to design an error protection scheme by selecting a fixed channel code with a certain rate and correction capability. The fix code is normally constructed for an expected average or worst-case channel conditions. In many cases, however, the data to be transmitted may not always require the same level of error protection. A new approach is therefore needed to share the entire available spectrum. This approach must not only provide a means for maintaining QoS guarantees, but it must also be least demanding on the use of available frequency spectrum.

Having a family of codes with different rates and correction capabilities, and having the channel state information (CSI) available at the transmitter (as well as at the receiver) will allow one to adaptively and intelligently select a code in the family that can accommodate for the protection needs of the transmitting data while carrying minimum redundancy. For the application in this chapter, a family of codes known as rate-compatible punctured convolutional (RCPC) codes are proposed for their several unique and attractive features. The chapter starts with the analysis of the performance of HEC that leads to the definition of a key QoS parameter,
namely, the cell-loss ratio (CLR). An overview of punctured convolutional codes and their subclass, rate-compatible punctured convolutional codes, is presented and their key features are outlined in section 2.3. In section 2.4, the capability of RCPC codes in providing unequal error protection (UEP) is exploited to overcome channel degradation due to fading. This will be followed by the analysis of RCPC bandwidth utilization over a lognormal fading channel.

2.2 ATM Header Error Control

The performance of ATM Header Error Control (HEC) is evaluated in this section. For this analysis, we recall that an ATM cell header is made up of a 5-octet header and a 48-octet payload, resulting in a total length of 53 bytes. The header is again subdivided into a 4-octet information field and an 8-bit header error control (HEC) field which is used to protect the header from transmission errors. HEC is an 8-bit Cyclic Redundancy Check (CRC) that is calculated based on the remaining 32 bits of the header using a generator polynomial \( g(x) = x^8 + x^2 + x + 1 \), to form a \((40, 32)\) cyclic code. This code is capable of correcting single-bit errors and detecting all double and a large fraction of multiple-bit errors.

At the receiver, the HEC decoder operates in two modes: the correction mode, in which all single errors are corrected and cells with multiple errors are discarded, and the detection mode, in which all cells with detected (single or multiple) errors in the header are discarded. This can be modeled as a Markov chain as depicted in Fig. 2.1[2].

At the initialization, the receiver error-correction algorithm is in the default mode for single error correction. As each cell is received, the decoder attempts to detect all possible errors in the header. As long as no errors are detected, the receiver remains in the error correction mode. When an error is detected, it will correct the error if it is a single-bit error or it will detect that a burst error has
occurred. In either case, the receiver moves to a detection mode. In this mode, no attempt is made to correct errors. The receiver will remain in detection mode as long as erroneous cells are received. This yields a low probability of incorrect cell forwarding under bursty error conditions. When a header is examined and found correct, receiver switches back to its correction mode.

ATM Cells are discarded when errors in the header are detected with no attempts made to correct them. The QoS parameter associated with this event is the cell-loss ratio (CLR) and can be quantified by calculating the probability \( P_L \) that a cell is discarded. This is the probability that the receiver is in correction mode and more than one error has occurred or the receiver is in the detection mode and at least one error has occurred, i.e.

\[
P_L = P_{\text{cor}}P(\geq 2) + P_{\text{det}}(P(1) + P(\geq 2)).
\]  

In order to calculate this probability, we assume that the header at the input to the HEC decoder is corrupted by random bit errors with a probability \( P_b \). Under this condition, the probability that there are \( i \) bit errors in a block of length \( n \) is given by a binomial distribution

\[
P(i) = \binom{n}{i} \cdot P_b^i (1 - P_b)^{n-i}.
\]
Thus, the probability of no error in the cell header is \( P(0) = (1 - P_b)^{40} \), the probability of a single error in the cell header is \( P(1) = 40P_b(1 - P_b)^{39} \) and the probability of more than one error in the cell header is \( P(\geq 2) = 1 - P(0) - P(1) \). To find the correction-mode probability \( P_{\text{cor}} \) and the detection mode probability \( P_{\text{det}} \), respectively, we solve for an appropriate set of state equations in the Markov model in Fig. 2.1. This yields, \( P_{\text{cor}} = P(0) \) and \( P_{\text{det}} = 1 - P_{\text{cor}} = 1 - P(0) \), respectively. The Cell-loss ratio can now be written in terms of \( P_b \) as

\[
P_L = 1 - (1 - P_b)^{40} - 40P_b(1 - P_b)^{40}(1 - P_b)^{39}.
\] (2.3)

### 2.3 Rate Compatible Punctured Convolutional codes

Binary convolutional codes are frequently used in applications that involve transmission of digital data over wireless channels. These codes are popular not just due to their relatively high coding gain, but because of the maximum likelihood sequence decoding (Viterbi decoding) algorithm that can readily be implemented for short memory codes. In bandlimited wireless applications, it is desirable that powerful high-rate \( R = k/n \) convolutional codes be used. For an \((n, k, m)\) convolutional code, there are \(2^k\) paths leaving or merging at each node of the encoder trellis diagram. In the Viterbi decoder, \(2^k\) metric computations must be carried out at each trellis node to select the most likely path. This introduces a practical implementation difficulty for high-rate codes. In order to obtain a simpler Viterbi decoding, punctured convolutional (PC) codes were introduced by Cain et al. [3].

In general, a rate-\(P/(nP-\delta)\) PC code can be obtained from a rate-\(1/n\), \((n, 1, m)\) convolutional code by periodically deleting code symbols from every \(nP\) code symbols generated from encoding of \(P\) information symbols. The low-rate \(1/n\) code is the mother code with a generator polynomial \( G(D) = (g_1^1(D), g_1^2(D), \cdots, g_1^n(D)) \), where \(g_1^i(D)\) is a polynomial of degree \(m\) with coefficients 0 and 1. The deletion of
code symbols is represented by an $n$-by-$P$ puncturing array, $P_\delta$. The elements of the puncturing array are zeros and ones, corresponding to deleting or retaining of the corresponding code symbol at the output of the mother code. The parameter $\delta$ specifies the number of zeros in the array, and can take on values between 0 (mother code) and $(n - 1)P$ (no coding), inclusive. Parameter $P$ is called the puncturing period as the puncturing array is used periodically for every $nP$ code symbols generated in the encoding process.

All PC codes of rate $P/(nP - \delta)$ derived from an $(n, 1, m)$ mother code can be decoded by the Viterbi Algorithm via the same trellis diagram used for decoding the mother code. At the receiver, prior to decoding, dummy symbols are inserted into the received sequence corresponding to the positions of the punctured code symbols at the transmitter. Following this, at the decoding process, the metric computations corresponding to the punctured positions are inhibited.

The upper bound on the bit-error probability of a rate-$P/(nP - \delta)$ PC code with Viterbi decoding can be derived following exactly the same procedure used to derive these bounds for a general rate-$k/n$ convolutional code. This bound requires a full knowledge of the weight distribution of the code, which can be obtained by calculating the code generating function $T(X,Y)$. For coherent BPSK signals with AWGN channel and unquantized received signals, the upper bound on the bit-error probability $P_b$ of a rate-$P/(nP - \delta)$ punctured convolutional code is given by [4]

$$P_b \leq \frac{1}{P} \sum_{d=d_{free}}^{\infty} c_d Q(\sqrt{2dE_s/N_0}), \quad (2.4)$$

where

$$Q(x) = \frac{1}{\sqrt{2\pi}} \int_{x}^{\infty} e^{-t^2/2} dt, \quad (2.5)$$

and $E_s/N_0$ (also denoted by $\gamma_s$) is the ratio of the energy of the received code symbol to the noise power spectral density. This quantity is related to the received SNR.
per information symbol $E_b/N_0$ by $E_s/N_0 = (k/n)E_b/N_0$. In Eq. 2.4, $d_{\text{free}}$ denotes the minimum free Hamming distance of the code and $c_d$ is the total number of information bit errors produced by the incorrect paths of Hamming weight $d > d_{\text{free}}$ that diverge from correct path and remerge with it at some later stage. In order to choose “good” codes, we must maximize $d_{\text{free}}$ and minimize the $c_d$ term.

Rate-compatible punctured convolutional (RCPC) codes [5] are a subclass of PC codes. Rate compatibility constraint in this subclass implies that all the code bits of the high rate punctured code are required by the lower rate codes. This constraint allows for transmission of incremental redundancy in ARQ/FEC schemes and continuous rate variation to change from low to high error protection within a data frame.

PC codes in general, and RCPC codes in particular, are useful in providing unequal error protection (UEP) for bandlimited fading channels. Of course, one way to achieve this goal is to employ a number of channel encoders and decoders, in parallel, to provide the necessary error protection levels when needed. The complexity of this scheme is high when the number of error protection levels becomes large. Punctured convolutional codes, on the other hand, require only one convolutional encoder at the transmitting end and one Viterbi decoder at the receiving end along with a mechanism for puncturing at the encoder and insertion of dummy variables at the decoder.

2.4 QoS Provisioning for WATM

Having reviewed the fundamentals of punctured convolutional codes, we would like to exploit their capability in providing UEP to compensate for signal power degradation due to fading while making minimum use of bandwidth. The selectable-rate property allows for dynamic bandwidth allocation to increase the overall capacity of the system through efficient use of frequency spectrum. Using punctured
convolutional codes for this purpose has the immediate advantage of employing only one convolutional encoder at the transmitter along with a maximum-likelihood decoder using Viterbi Algorithm at the receiver. The basic structure of the encoder is the same as the one designed for the mother code. Only a mechanism for puncturing has to be embedded in this design to allow for selectable-rate transmission. At the receiving end, the same Viterbi decoder for mother code is modified to allow for insertion of dummy symbols and appropriate change of metric computations.

We would also like to apply this concept in the context of Wireless ATM where end users are connected to an ATM network through wireless links. In doing so, we recall how a given service is carried along an ATM network between end-users. ATM is a connection-oriented service, meaning that prior to receiving service, a given user must request a connection to the intended receiver. In order for a logical connection to be established, the admission controller must make sure that the new connection can provide and maintain the QoS requested by the application. If the ATM network is extended to give service to wireless users, then at least one segment of the physical link of the virtual connection (it will be two segments, in case the two end systems have wireless connectivity) will be wireless, with a high BER that is not randomly distributed. In section 2.2, a key QoS measure, namely, CLR was introduced and it was directly related to the BER (Eq. 2.3), a high BER wireless link can act as a bottleneck and may become a major barrier in providing the QoS needs of the intended service unless sufficient amount of bandwidth is provided to reduce the BER (e.g., through coding). Wireless bandwidth resource is very limited and expensive. Therefore, it must be utilized efficiently and only then the capacity of the system can be increased, so more services can be provided to the end-users.

Fig. 2.2 depicts a block diagram of RCPC coding scheme proposed for the wireless link. At the transmitter, ATM cells are encoded and interleaved. We
assume that instantaneous channel state information (CSI) is available at the encoder to allow for proper code rate selection. The purpose of interleaving is to distribute burst errors uniformly in time in order to improve the performance of the Viterbi decoder, since Viterbi decoders tend to propagate burst errors rather than correcting them. The interleaved data is then BPSK modulated before being passed through the channel. The channel is characterized by a known fading distribution and AWGN. At the receiver, the received data is first demodulated and then de-interleaved. Soft decision is employed in the Viterbi decoding algorithm together with the required modifications that allow for insertion of dummy symbols and appropriate change of metric computations, according to the puncturing array, available at the receiver.

\[ \text{Rate Selection} \quad R = P/(nP - \delta) \]

Figure 2.2. Coded data transmission using RCPC codes with interleaving and channel state information (CSI) on fading channels.

The upper bound on the bit-error probability of rate-\(P/(nP - \delta)\) RCPC codes is given in Eq. 2.4. The weight distribution \(c_d\) in this equation can be uniquely obtained for each value of \(\delta\), if the generator polynomial \(G(D)\) of the \((n,1,m)\) mother code and the puncturing array \(P_\delta\) are both known. A search was performed by Hagenauer [5] for the best rate-\(P/(nP - \delta)\) RCPC code family with \(P = 8,\)
and $\delta = 0, 2, \ldots, 22$, among all (4,1,4) mother codes and resulted in a mother code generating polynomial $G(D)$ and a set of puncturing arrays $P_\delta$, $\delta = 0, 2, \ldots, 22$, as presented in [5]. The weight distribution $c_d$ for each $\delta$ was then obtained, accordingly. The result of this search for $m = 4$ will be used in this chapter. Fig. 2.3 depicts a plot of upper bound on the bit-error probability versus $E_s/N_0$ of this code with rates $R_i = 8/(32 - 2i)$, $i = 0, 1, \ldots, 12$.

![Figure 2.3](image)

Figure 2.3. Plot of BER vs. $E_s/N_0$ for RCPC codes of rate $R_i = 8/(32 - 2i)$, $i = 0, 1, \ldots, 12$, given by Hagenauer. The thick lines correspond to an operation mode in which $P_b \leq 10^{-5}$.

As mentioned earlier, there is a one-to-one relationship between cell-loss probability $P_L$ and channel bit-error rate $P_b$, given by Eq. 2.3. Suppose, a new service
demands a CLR less than $2.38 \times 10^{-7}$. This value corresponds to a BER of $10^{-5}$, which is a horizontal line plotted in Fig. 2.3. Suppose the wireless link is a fading channel with a received $\gamma_s$ following a given probability distribution. In order to maintain the pre-assigned CLR, the encoder must continuously adjust its code rate in response to various channel conditions. For instance, suppose at some time during the connection, wireless channel is subject to no fading and the received $\gamma_s$ is above 10 dB. Here, no coding is needed as an uncoded BPSK signal can on its own provide a BER less than $10^{-5}$. Suppose, at a later time, due to rain fading, the received $\gamma_s$ is reduced to 4 dB. In this case, the rate-8/12 code will be the best selected code in the family as all codes of rate more than 8/12 would result in a BER greater than $10^{-5}$, whereas codes of rate less than 8/12, although capable, consume more bandwidth than needed. In general, the code rate is continuously varied based on instantaneous CSI such that QoS is maintained and the available bandwidth is not wasted. In a small fraction of time, however, the received $\gamma_s$ is so small that even the 1/4-rate mother code is not powerful enough to bring the BER to below its prescribed value $10^{-5}$ and at this time an “outage” signal is invoked, advising the user to try again at a later time. We define outage, and denote it by $P_F$, as the probability of failure in providing a guaranteed QoS. This is equivalent to the probability that the received $\gamma_s$ falls below its threshold value $\gamma_{sth}$ at which the BER of the mother code is equal to its prescribed value (i.e., $P_b(\gamma_{sth}) > 10^{-5}$). For a fixed transmitted power, outage will be a function of fading channel probability density $P_{\Gamma_s}(\gamma_s)$ and is given by

$$P_F = \int_{-\infty}^{\gamma_{sth}} P_{\Gamma_s}(\gamma_s) d\gamma_s. \quad (2.6)$$

For a slow-fading lognormal channel, $P_{\Gamma_s}(\gamma_s)$ can be expressed as

$$P_{\Gamma_s}(\gamma_s) = \frac{1}{\sqrt{2\pi}\sigma} e^{-\frac{(\gamma_s-\mu)^2}{2\sigma^2}}, \quad (2.7)$$
where $\mu$ and $\sigma$ are the dB values of mean and standard deviation, respectively. Plots of outage versus guaranteed CLR for a lognormal channel with various means and standard deviations are shown in Fig. 2.4. Note that, the outage probability for any adaptive coding scheme is determined by the lowest-rate code in the family. When RCPC codes are employed, the lowest-rate code is the mother code.

Variable-rate coding schemes save a considerable amount of bandwidth compared to the worst-case scenario in which only a fix-rate code, capable of providing the same outage probability, is used. For the RCPC code used in this work, the fix-rate code is, indeed, the 1/4-rate mother code. A figure of merit can be obtained by calculating the average code rate $R_{\text{ave}}$ defined by

$$R_{\text{ave}} = R_0 \int_{-\infty}^{\gamma_{s1}} P_{\Gamma_s}(\gamma_s)d\gamma_s + \sum_{i=1}^{11} R_i \int_{\gamma_{s_i}}^{\gamma_{s_{i+1}}} P_{\Gamma_s}(\gamma_s)d\gamma_s + R_{12} \int_{\gamma_{s_{12}}}^{\infty} P_{\Gamma_s}(\gamma_s)d\gamma_s, \quad (2.8)$$

under various fading distributions. In this equation, $R_i = 8/(32 - 2i)$ and $\gamma_{s_i}$ is the value of $\gamma_s$ at which a rate transition occurs from $R_{i-1}$ to $R_i$. We can define bandwidth utilization to be the ratio of the average bandwidth used through using RCPC coding scheme to the amount of bandwidth used when only the fix-rate mother code, designed for worst-case scenario is used. This ratio is also equal to $(1/R_{\text{ave}})/(1/R_0) = R_0/R_{\text{ave}}$ and varies between 1/4 and 1. If the wireless channel is clear, no coding will be used and bandwidth utilization will have its lowest and best possible value of 1/4. If, on the other hand, wireless channel is subject to deep fades, the 1/4-rate mother code is always used and bandwidth utilization has its highest value 1.

Plots of bandwidth utilization versus guaranteed CLR in a lognormal channel with various means and standard deviations are provided in Fig. 2.5. As seen, bandwidth utilization is a monotonically decreasing function of guaranteed CLR with an asymptotic value of 1/4. Here, the value of $\sigma$ is determined solely by channel measurements whereas the mean received power $\mu$ is proportional to the transmitted
power with a proportionality factor determined by measuring the mean channel attenuation. Therefore, the transmitted power must be intelligently selected for each application. For example, consider a scenario where symbols with high power are transmitted in a lognormal fading channel to yield received signals with a high mean value ($\mu$). In this scenario, high-rate codes are more frequently activated than the lower-rate ones. Hence, a very low bandwidth utilization is achieved (refer to the plots for $\mu = 16$). During rare fading events, on the other hand, low-rate codes help to provide sufficient protection, as needed to maintain a guaranteed QoS.

Figure 2.4. Outage vs. Guaranteed CLR for a lognormal fading channel with mean values $\mu$ of 3, 6, 9, and standard deviation $\sigma$ of 8, 10 and 12 corresponding to each $\mu$. 
Figure 2.5. Bandwidth utilization vs. guaranteed CLR for a lognormal fading channel with mean values $\mu$ of 3, 6, 9, and standard deviation $\sigma$ of 8, 10 and 12 corresponding to each $\mu$.

2.5 Summary

This chapter was devoted to a case-study of rate-compatible punctured convolutional codes as applied to wireless links of ATM networks. The objective was to maintain QoS guarantees over a fading wireless channel with minimum consumption of the available bandwidth. ATM QoS parameters can generally be classified into two types: (1) BER-oriented parameters which are mainly a function of the channel BER (e.g., CLR) and (2) delay-oriented parameters. For a wireless link with a high BER, the former can be maintained through a powerful FEC and proper interleaving. The latter, on the other hand, strongly depends on the network traffic management. This work primarily dealt with BER-related QoS
issues. We investigated application of RCPC codes in maintaining the cell-loss ratio in a lognormal fading channel. Results show that adaptive coding can deliver a low outage probability corresponding to that of a lowest-rate code in the family when used as a worst-case-scenario code. In return, adaptive scheme provides an efficient way of utilizing the frequency resource, available to the user. This is due to the fact that the adaptive scheme uses the low rate codes only when necessary.
This chapter considers two well-known selective-repeat retransmission schemes, namely, hybrid type-I ARQ and hybrid type-II ARQ, using convolutional coding, in conjunction with maximum-likelihood code combining. Our theoretical analysis, based upon the concept of generalized weight distribution, shows that the use of code combining yields a significant throughput at very high channel error rates not only in constant AWGN channels but also in fading channels. To demonstrate this, we consider a widely-used block-fading Rayleigh channel model, in which the channel is assumed to be constant during each block of data and the fading is assumed to be independent from block to block. A key parameter in designing retransmission protocols for delay-limited applications in such channels is the minimum number of retransmissions needed to achieve error-free decoding at almost all channel conditions (low outage probability). This number can be reduced significantly when code combining is employed.

3.1 Introduction

Data transmission over a slow, flat fading channel is analogous to data transmission over a Gaussian channel with time-varying signal-to-noise ratio (SNR). In periods of deep fades, the decoder-receiver experiences long sequences of very low
signal-to-noise ratio inputs. Consequently, standard coding schemes, effective for classical Gaussian channels of moderate SNRs, cannot be employed directly on fading channels, since the error control capability of these codes is overwhelmed by the long sequences of channel errors. To effectively combat fading, we need to provide the decoder with independent fading statistics. This is the essence of diversity techniques.

One of the simplest diversity techniques is time diversity wherein copies of the same symbol are sent in $L$ different time slots. This scheme amounts to use of repetition coding on two or more channels. The hope is that by some form of combining the received symbols, or perhaps merely selection of the best, the aggregate signal will experience less fading than without diversity. Significant improvement over this scheme can be made possible through the use of code combining. Code combining is a technique for combining entire packets, rather than individual symbols. It introduces an added dimension to conventional diversity schemes that use repetition coding. Code combining techniques are often used in systems that are equipped with some type of retransmission mechanism. Information is transmitted in form of packets encoded with a relatively high-rate code and are repeated only when needed. Receiver forms an increasingly reliable estimate of transmitting packets by combining a minimum number of packets, making a high-throughput end-to-end system.

This chapter considers two well-known retransmission schemes (also known as Automatic Repeat reQuest or ARQ) namely, hybrid type-I ARQ and hybrid type-II ARQ, using convolutional coding in conjunction with maximum-likelihood (ML) code combining [6, 7]. Our approach is somewhat different from [8] in that computer search attempts for finding free distance and weight spectra of the best families of repetition codes of decreasing rate [10] is not required. Our analysis, instead, is
based upon the concept of generalized weight distribution, described in section 3.4. We have also extended the previous results to a block-fading Rayleigh channel, in which the channel is assumed to be constant during a block of \(N\) symbols and the fading is assumed to be independent from block to block. This simplified model for flat, slow-fading Rayleigh channels [11] is useful in characterizing many mobile communications systems including those employing slow frequency hopping with \(N\) symbols per hop [12]. Examples of such systems are the GSM system [13] or the ATDMA proposal for UMTS [14], both using TDMA with slow frequency hopping.

This chapter is organized as follows. The description of the two ARQ protocols is given in section 3.2. In section 3.3 performance bounds for the two ARQ schemes are derived. Section 3.4 introduces the concept of generalized weight distribution, needed for evaluating performance bounds of a type-II ARQ scheme. Numerical results for average throughput efficiency of the two ARQ schemes in an AWGN channel, as well as a block-fading Rayleigh channel are presented in section 3.5.

### 3.2 Description of two ARQ Schemes

This section describes the two retransmission schemes, i.e., hybrid type-I ARQ and hybrid type-II ARQ [15], used in this chapter. Two linear codes are employed in each scheme. One is a high rate \((k,l)\) code \(C_0\) for error detection and the other is a half-rate, \((2,1,m)\) convolutional code \(C_1\) for error correction. Let the generator polynomial for the convolutional code \(C_1\) be denoted by \([g_1(z), g_2(z)]\).

An \(l\) bit message, ready for transmission, is first mapped into a \(k\) bit sequence \(I(z)\) in \(C_0\). The sequence \(I(z)\) is then transformed into two \(k+m\)-bit sequences \(U^{(1)}(z) = I(z)g_1(z)\) and \(U^{(2)}(z) = I(z)g_2(z)\). In the type-I scheme, these two sequences are interleaved at the transmitter to form a half-rate code word \(U(z) = U^{(1)}(z^2) + zU^{(2)}(z^2)\). The sequence \(U(z)\) is then transmitted repeatedly until error-free decoding is declared. In the type-II scheme, on the other hand, interleaving does
not take place at the transmitter. The transmitter alternately sends the sequences $U^{(1)}(z)$ and $U^{(2)}(z)$ until an error-free decoding is declared. In both schemes, the decoder at the receiver operates on the combination of all received copies. Each transmitted packet is assumed to experience a separate Gaussian channel, in which the received signal-to-noise ratio (SNR) for that packet is independent from those of other packets. Knowledge of SNR for each received packet turns out to be necessary when defining the proper metric for maximum-likelihood decoding of the combined code sequence, as explained in the next section.

### 3.3 Performance Analysis

Following the schemes described in the previous section, we consider the case of a BPSK modem in which, prior to transmission, binary packets are mapped into signal packets consisting of -1 (zero bit) and +1 (one bit). The transmitted symbols are assumed to be equally likely with unit energy. We start this section by deriving the performance bounds for the type-II ARQ scheme. The bounds for type-I scheme then follow immediately, as shown later.

#### 3.3.1 Hybrid Type-II ARQ with ML Code Combining

Let’s denote the first transmitted signal packet, corresponding to $U^{(1)}$, by $X^{(1)} = (x_1^{(1)}, x_2^{(1)}, \ldots, x_{k+m}^{(1)})$, and the received signal packet associated with $X^{(1)}$, by $Y^{(1)} = (y_1^{(1)}, y_2^{(1)}, \ldots, y_{k+m}^{(1)})$. Also, denote the energy per symbol in $Y^{(1)}$ by $E^{(1)}$.

For each received symbol in $Y^{(1)}$ we have $y_i^{(1)} = \sqrt{E^{(1)}} x_i^{(1)} + n_i$, $i = 1, 2, \ldots, k + m$, where the noise terms $\{n_i\}$ are statistically independent Gaussian random variables with variance $N_0/2$. The first transmitted packet is not coded and is only detected on a per-symbol basis. The probability of symbol error, $p_e$, after maximum-likelihood detection is:

$$p_e = Q \left( \sqrt{2E^{(1)}/N_0} \right)$$

(3.1)
where
\[ Q(x) = \frac{1}{\sqrt{2\pi}} \int_{x}^{\infty} \exp\left(-\frac{t^2}{2}\right) dt \] (3.2)

The probability that the first packet is in error is:
\[ P_{\text{packet}}^{(1)} = 1 - [1 - p_e]^{k+m} \] (3.3)

When the first packet is detected in error, the receiver requests for the transmission of the second packet \(X^{(2)}\) corresponding to the binary sequence \(U^{(2)}\). The received block \(Y^{(2)}\), with energy-per-symbol \(E^{(2)}\) is then interleaved with \(Y^{(1)}\) at the receiver to form a half-rate convolutionally-coded sequence \(Y = (y_1^{(1)}, y_1^{(2)}, \ldots, y_{k+m}^{(1)}, y_{k+m}^{(2)})\), where \(y_i^{(j)} = \sqrt{E^{(j)}}x_i^{(j)} + n_i^{(j)}, j = 1, 2\). Note that, the symbol energy in this sequence alternates between \(E^{(1)}\) and \(E^{(2)}\). A soft decision maximum-likelihood decoder at the receiver decides in favor of \(\hat{X} = (\hat{x}_1^{(1)}, \hat{x}_1^{(2)}, \ldots, \hat{x}_{(k+m)}^{(1)}, \hat{x}_{(k+m)}^{(2)})\) among all possible \(2^k\) code words \(X\) if and only if \(P(Y|\hat{X}) \geq P(Y|X)\), i.e.,
\[
\prod_{i=1}^{k+m} \prod_{j=1}^{2} \frac{1}{\sqrt{\pi N_0}} \exp \left\{ -\frac{(y_i^{(j)} - \sqrt{E^{(j)}}\hat{x}_i^{(j)})^2}{N_0} \right\} \geq \prod_{i=1}^{k+m} \prod_{j=1}^{2} \frac{1}{\sqrt{\pi N_0}} \exp \left\{ -\frac{(y_i^{(j)} - \sqrt{E^{(j)}}x_i^{(j)})^2}{N_0} \right\}
\] (3.4)

Taking natural logs from both sides and dropping common terms we obtain:
\[
\sum_{i=1}^{k+m} \sum_{j=1}^{2} \sqrt{E^{(j)}} y_i^{(j)} \hat{x}_i^{(j)} \geq \sum_{i=1}^{k+m} \sum_{j=1}^{2} \sqrt{E^{(j)}} y_i^{(j)} x_i^{(j)}
\] (3.5)

Thus, the decision metric \(M^{(2)}\) can be defined as:
\[
M^{(2)} = \sum_{i=1}^{k+m} \sum_{j=1}^{2} \sqrt{E^{(j)}} y_i^{(j)} \hat{x}_i^{(j)}
\]
\[
= \sum_{i=1}^{k+m} \sqrt{E^{(1)}} y_i^{(1)} \hat{x}_i^{(1)} + \sqrt{E^{(2)}} y_i^{(2)} \hat{x}_i^{(2)}
\] (3.6)

Therefore, maximum-likelihood decoding after the reception of two packets requires that each received packet be weighted by its reliability factor (the first packet
by $\sqrt{E(1)}$ and the second packet by $\sqrt{E(2)}$, before they are interleaved. The resulting sequence of length $2(k + m)$ can then be decoded using a conventional soft-decision Viterbi Algorithm for a half-rate convolutional code, in which inner product provides the proper metric.

In general, for $L \geq 2$ received repeated packets, the maximum-likelihood metric $M^{(L)}$ can be written as:

$$M^{(L)} = \sum_{i=1}^{k+m} \sum_{j=1}^{L} \sqrt{E(j)} y_i^{(j)} x_i^{(j)}$$

(3.7)

where, $x_i^{(j)} = x_i^{(1)}$ for odd $j$s, and $x_i^{(j)} = x_i^{(2)}$ for even $j$s. Thus, we get:

$$M^{(L)} = \begin{cases} 
\sum_{i=1}^{k+m} \left\{ \left( \sum_{j=1}^{L/2} \sqrt{E(2j-1)} y_i^{(2j-1)} \right) x_i^{(1)} + \left( \sum_{j=1}^{L/2} \sqrt{E(2j)} y_i^{(2j)} \right) x_i^{(2)} \right\}, & L \text{ even} \\
\sum_{i=1}^{k+m} \left\{ \left( \sum_{j=1}^{(L+1)/2} \sqrt{E(2j-1)} y_i^{(2j-1)} \right) x_i^{(1)} + \left( \sum_{j=1}^{(L-1)/2} \sqrt{E(2j-2)} y_i^{(2j-2)} \right) x_i^{(2)} \right\}, & L \text{ odd}
\end{cases}$$

(3.8)

(3.9)

which is similar in form to Eq. 3.6. This suggests that the same decoding procedure be used when $L$ received copies are available at the receiver: A sequence of length $2(k + m)$ is formed at the decoder input by interleaving the weighted sum of odd-numbered received copies with the weighted sum of even-numbered received copies. The sequence is then decoded via a conventional soft-decision Viterbi algorithm for ideal AWGN channels. Note that, since newly received copies are essentially added (with proper reliability weightings) to the previously received ones, the required buffer length at the receiver will be independent of $L$, number of retransmissions, and therefore finite buffer-length considerations encountered in other ARQ schemes simply do not apply here.

We now derive the pairwise probability of error when $L \geq 2$ copies are available at the receiver. Suppose $L$ is even. We assume, with no loss of generality, that the
all-zero binary message is transmitted. This means the two zero sequences $U_0^{(1)}$ and $U_0^{(2)}$ are alternately transmitted $L/2$ times each. Since all binary symbols in $U_0^{(1)}$ and $U_0^{(2)}$ have zero values (i.e., $u_{i0}^{(1)} = u_{i0}^{(2)} = 0, i = 1, 2, \ldots, k + m$), all symbols in the transmitted signal packets $X_0^{(1)}$ and $X_0^{(2)}$ have a value of -1 (i.e., $x_{i0}^{(1)} = x_{i0}^{(2)} = -1$). The received symbols can thus be written as $y_i^{(j)} = -\sqrt{E^{(j)}} + n_i^{(j)}$, where $j$ varies from 1 to $L$. Consider any error path, $U_e = (u_{1e}^{(1)}, u_{1e}^{(2)}, \ldots, u_{(k+m)e}^{(1)}, u_{(k+m)e}^{(2)})$ of Hamming distance $d$ from the all-zero path. The conditional probability of error in the pairwise comparison of the all-zero path metric $M_0^{(L)}$ and the error path metric $M_e^{(L)}$ is:

$$P_2^{(L)}(E^{(1)}, \ldots, E^{(L)}) = P(M_e^{(L)} \geq M_0^{(L)})$$

$$= P\left\{ \sum_{i=1}^{k+m} \left[ \sum_{j=1}^{L/2} \sqrt{E^{(2j-1)}} y_i^{(2j-1)} (x_{ie}^{(1)} - x_{i0}^{(1)}) + \sum_{j=1}^{L/2} \sqrt{E^{(2j)}} y_i^{(2j)} (x_{ie}^{(2)} - x_{i0}^{(2)}) \right] \geq 0 \right\}$$

$$= P\left\{ \sum_{i=1}^{k+m} \left[ \sum_{j=1}^{L/2} \sqrt{E^{(2j-1)}} y_i^{(2j-1)} u_{ie}^{(1)} + \sum_{j=1}^{L/2} \sqrt{E^{(2j)}} y_i^{(2j)} u_{ie}^{(2)} \right] \geq 0 \right\}$$

(3.10)

Let the Hamming weight of sequences $U_e^{(1)}$ and $U_e^{(2)}$ be $d_1$ and $d_2$, respectively. The conditional pairwise probability in Eq. 3.10 can now be written as:

$$P_2^{(L)}(E^{(1)}, \ldots, E^{(L)}) = P\left\{ \sum_{l_1=1}^{d_1} \sum_{j=1}^{L/2} \sqrt{E^{(2j-1)}} n_{l_1}^{(2j-1)} + \sum_{l_2=1}^{d_2} \sum_{j=1}^{L/2} \sqrt{E^{(2j)}} n_{l_2}^{(2j)} \geq 0 \right\}$$

$$= P\left\{ - \left( d_1 \sum_{j=1}^{L/2} E^{(2j-1)} + d_2 \sum_{j=1}^{L/2} E^{(2j)} \right) + \sum_{j=1}^{L/2} \sqrt{E^{(2j-1)}} n_{l_1}^{(2j-1)} + \sum_{j=1}^{L/2} \sqrt{E^{(2j)}} n_{l_2}^{(2j)} \geq 0 \right\}$$

(3.11)
The argument in Eq. 3.11 is a Gaussian random variable with mean $\mu$ and variance $\sigma^2$ given by:

$$
\mu = - \left( d_1 \sum_{j=1}^{L/2} E^{(2j-1)} + d_2 \sum_{j=1}^{L/2} E^{(2j)} \right) \quad (3.12)
$$

$$
\sigma^2 = \frac{N_0}{2} \left( d_1 \sum_{j=1}^{L/2} E^{(2j-1)} + d_2 \sum_{j=1}^{L/2} E^{(2j)} \right) \quad (3.13)
$$

Thus, for even values of $L$, the conditional pairwise probability of error is given by:

$$
P_{2}^{(L)}(d_1, d_2|E^{(1)}, \ldots, E^{(L)}) = Q \left( \sqrt{\frac{1}{2} \left( \sum_{j=1}^{L/2} d_1 E^{(2j-1)} + d_2 E^{(2j)} \right)} \right) \quad (3.14)
$$

Likewise, for odd values of $L$ we obtain:

$$
P_{2}^{(L)}(d_1, d_2|E^{(1)}, \ldots, E^{(L)}) = Q \left( \sqrt{\frac{1}{2} \left( \sum_{j=1}^{(L+1)/2} d_1 E^{(2j-1)} + \sum_{j=1}^{(L-1)/2} d_2 E^{(2j)} \right)} \right) \quad (3.15)
$$

Note that, in Eqs. 3.14 and 3.15, the Hamming distance $d$ of the error path has been splitted into two distances $d_1$ and $d_2$, where $d_1$ is the Hamming weight of a sequence formed by the first bits in each branch and $d_2$ is the Hamming weight of a sequence formed by the second bits in each branch of the error path.

The conditional first-event error probability can now be upper bounded as:

$$
P_{ev}^{(L)}(E^{(1)}, \ldots, E^{(L)}) \leq \sum_{d_1, d_2} a(d_1, d_2) P_{2}(d_1, d_2|E^{(1)}, \ldots, E^{(L)}) \quad (3.16)
$$

The upper bound given in Eq. 3.16 requires the knowledge of generalized distance spectrum $a(d_1, d_2)$ which can be obtained using the generalized transfer function approach, explained in section 3.4. Finally, the upper bound on the conditional probability of packet error after receiving $L \geq 2$ packets is obtained using:

$$
P_{\text{packet}}^{(L)}(E^{(1)}, \ldots, E^{(L)}) \leq 1 - \left[ 1 - P_{ev}^{(L)}(E^{(1)}, \ldots, E^{(L)}) \right]^{k+m} \quad (3.17)
$$
3.3.2 Hybrid Type-I ARQ with ML Code Combining

The conditional pairwise error probability in a type-I hybrid ARQ scheme with ML code combining can be obtained by remembering that in this case the two sequences $U^{(1)}$ and $U^{(2)}$ are interleaved at the transmitter to form a half-rate code word of length $2(k+m)$. Each code-word packet is then transmitted over a separate AWGN channel. Thus, we let $E^{(2j-1)} = E^{(2j)}$ in Eq. 3.14. Also, we replace $L/2$ by $L$ to allow for $L$ retransmissions. Noting that, $d_1 + d_2 = d$, where $d$ is the total number of nonzero bits in the error path, we obtain:

$$P_2^{(L)}(d|E^{(1)}, \ldots, E^{(L)}) = Q \left( \sqrt{\frac{2d}{L}} \sum_{j=1}^{L} \frac{E^{(j)}}{N_0} \right)$$

(3.18)

The conditional first-event error probability can now be upper bounded as:

$$P_{e_v}^{(L)}(E^{(1)}, \ldots, E^{(L)}) \leq \sum_d a(d)P_2(d|E^{(1)}, \ldots, E^{(L)})$$

(3.19)

where $a(d)$ represents the distance spectrum of the $(2, 1, m)$ code. Finally, the upper bound on the conditional probability of packet error after receiving $L$ packets is given by Eq. 3.17.

As a special case consider the case of a constant AWGN channel in which $E^{(1)} = E^{(2)} = \ldots = E^{(L)} = E$. In this case Eq. 3.18 reduces to:

$$P_2^{(L)}(d) = Q \left( \sqrt{\frac{2dE}{N_0/L}} \right)$$

(3.20)

which is an expected result.

3.4 Generalized Weight Distribution

Let’s consider a $(54,74)$ half-rate convolutional code with the state diagram shown in Fig. 3.1. The labels $d_1$ and $d_2$ in each branch of the state diagram correspond to the existence of a nonzero first bit and a nonzero second bit in that
There are seven non-zero states in this diagram which can be represented by a column matrix $S = [s_1, s_2, \ldots, s_7]^T$. The state equation and the input/output equations are:

\[
S = A \cdot S + B \cdot s_{in} \tag{3.21}
\]
\[
s_{out} = C \cdot S \tag{3.22}
\]

where

\[
A = \begin{bmatrix}
0 & 0 & 0 & 1 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 & d_1 & 0 & 0 \\
0 & d_1 d_2 & 0 & 0 & 0 & 1 & 0 \\
0 & 1 & 0 & 0 & 0 & d_1 d_2 & 0 \\
0 & 0 & d_1 & 0 & 0 & 0 & d_2 \\
0 & 0 & d_2 & 0 & 0 & 0 & d_1
\end{bmatrix},
\]
\[
B = \begin{bmatrix}
d_1 d_2 & 0 & 0 & 0 & 0 & 0 & 0
\end{bmatrix}^T,
\]
\[
C = \begin{bmatrix}
0 & 0 & 0 & d_1 d_2 & 0 & 0 & 0
\end{bmatrix}.
\]

The transfer function $T(d_1, d_2)$ is obtained by solving Eqs. 3.21 and 3.22, simultaneously. This yields:

\[
T(d_1, d_2) = \frac{s_{out}}{s_{in}} = C \cdot (I - A)^{-1} \cdot B = C \cdot (I + A + A^2 + \ldots) \cdot B =
\]
\[
d_1^2 d_2^2 + 3d_1^3 d_2^2 + d_1^6 d_2^2 + 3d_1^4 d_2^2 + d_1^5 d_2^2 + \ldots \tag{3.23}
\]

The first term in the right hand side of Eq. 3.23 shows that there is one path that diverges from and remerges with the all-zero path for the first time, that has a total weight of 6, 4 of which come from the first bits of each branch and the other 2 come from the second bits of each branch, and so $a(4, 2) = 1$. Likewise, we have $a(3, 4) = 3$, $a(6, 2) = 1$, etc. Note that, by letting $d_1 = d_2 = d$, the transfer function $T(d_1, d_2)$ reduces to its conventional form $T(d)$ from which the distance spectra $a(d)$ can be obtained.
3.5 Throughput Analysis

In this section, the ideal selective repeat type-I and type-II ARQ scheme in conjunction with code combining in a block fading Rayleigh channel are analyzed. Let $T_r$ be the expected number of transmission attempts that must be made before a packet is accepted by the receiver. Kallel [9] showed that $T_r$ is tightly upper bounded by:

$$T_r \leq 1 + \sum_{L=1}^{\infty} P_{\text{packet}}^{(L)}(E^{(1)}, \ldots, E^{(L)})$$

(3.24)

where $P_{\text{packet}}^{(L)}(E^{(1)}, \ldots, E^{(L)})$ is the probability of generating retransmission request while decoding the packet formed by combining $L$ received copies of the packet. This probability was derived for type-I and type-II ARQ schemes in section 3.3. The quantity $T_r$ is conditioned on the received energy of each packet. In a block-
fading Rayleigh channel, the received energy values \( E^{(i)} \), \( i = 1, 2, \ldots \) form a set of independent, identically-distributed random variables with a pdf:

\[
f \left( E^{(i)} \right) = \frac{\exp \left\{ \frac{-E^{(i)}}{E_{\text{avg}}} \right\}}{E_{\text{avg}}} , \quad E_{\text{avg}} \geq 0
\]  (3.25)

where \( E_{\text{avg}} \) is the mean received energy for each symbol. The average number of retransmissions over a Rayleigh fading channel can be obtained by integrating Eq. 3.24 over the pdf of each \( E^{(i)} \), i.e.,

\[
T_r \leq 1 + \sum_{L=1}^{\infty} \int_{0}^{\infty} \cdots \int_{0}^{\infty} P_{\text{packet}} \left( E^{(1)}, \ldots, E^{(L)} \right) \times \left( f \left( E^{(1)} \right) \cdots f \left( E^{(L)} \right) \right) dE^{(1)} \cdots dE^{(L)}
\]  (3.26)

which must be evaluated, numerically. The average throughput efficiency \( \eta \) can now be defined as \( 1/T_r \) times an efficiency factor, which is \( l/2(k+m) \) for a type-I ARQ and \( l/k \) for a type-II ARQ. Neglecting the memory length \( m \) of the convolutional code and the \( k-l \) added parity bits for error detection, this factor is approximately equal to 1/2 for a type-I ARQ and unity for a type-II ARQ scheme.

Plot of average throughput efficiency \( \eta \) vs. average received signal-to-noise ratio \( E_{\text{avg}}/N_0 \) for each ARQ scheme in a block-fading Rayleigh channel is shown in Fig. 3.2, where the example code of section 3.4 is used. Here, the value of \( k \) is taken to be 424, i.e., the size of an ATM cell. Fig. 3.2 also includes throughput performance plots for an ideal AWGN channel. It is seen that the throughput curves in a block-fading Rayleigh channel for the two ARQ types follow the same trend as those of an ideal AWGN channel. The average throughput of a type-II ARQ scheme with code combining is always greater than that of a type-I ARQ, especially for high values of average signal-to-noise ratio (SNR). For very low values of average SNR (say, less than 0 dB), where more number of retransmissions is required for error-free decoding, ML code combining can still result in a reasonable average
throughput as high as 0.1. In this case, using a type-I ARQ instead of a type-II ARQ may be more desirable, since it requires, on average, about half the number of retransmissions for error-free decoding, giving rise to a lower average transmission delay.

In real-time applications, such as constant-bit-rate (CBR) applications intended for use in ATM circuit emulation mode [16], received packets must be corrected within a specified time window. In this case, buffers of fixed size are employed to provide a window for error recovery while introducing a fixed delay over the wireless link. A key design issue in this case is to determine the buffer size or equivalently the number of permitted retransmissions in that time window. Alternatively, for a given retransmission scheme, we would like to know what would be the maximum number of retransmissions that would result in an almost error-free communication with, say, a one-percent chance that error-free communication cannot be guaranteed. This is depicted in Fig. 3.3 for a type-I ARQ scheme with ML code combining. Based on this figure, for example, in a block-fading Rayleigh channel characterized by $E_{\text{avg}}/N_0 = 0$ dB, having six copies of a packet at the receiver would be sufficient to achieve an almost error-free decoding ($P_{\text{packet}} \leq 10^{-3}$), more than 99% of the time (%1 outage).

### 3.6 Summary

In this chapter, we have considered both hybrid type-I and type-II ARQ schemes, using convolutional coding in conjunction with maximum-likelihood code combining. It has been shown that the use of code combining allows the system to be robust under severe channel degradation due to fading, while maintaining a high throughput. A high-throughput ARQ system is very important in applications in which received packets must be corrected within a specified time window. Numerical analysis for block-fading Rayleigh channels show that reliable data communication
Figure 3.2. Lower bounds for the average throughput efficiency $\eta$ as a function of average received signal-to-noise ratio $E_{avg}/N_0$ for hybrid type-I and type-II ARQ schemes with code combining in an AWGN channel as well as a block-fading Rayleigh channel using a 1/2-rate (54,74) convolutional code. The points of inflection on the dotted curves correspond to new retransmission events. For a type-II ARQ these points occur at throughput values of 1/2, 1/3, 1/4, 1/5, etc. whereas for a type-I ARQ they occur at throughput values of 1/4, 1/6, 1/8, 1/10, etc.
Figure 3.3. Upper bound in the number of Transmissions required to achieve an almost error-free communication ($P_{packet} \leq 10^{-3}$) with a 99% probability (1% outage) in a block-fading Rayleigh channel, for a hybrid type-I ARQ with code combining.
can be obtained with only a reasonable number of retransmissions making this scheme suitable for delay-limited applications.
Chapter 4

SPACE DIVERSITY FOR WIRELESS INFRARED SYSTEMS USING MULTIPLE-SPOT DIFFUSING CONFIGURATION

The main goal of this chapter and the chapters that follow, is to investigate the communication-theoretic aspects of the physical layer of a wireless infrared network, with emphasis on the effect of code combining [6] on the overall system performance. Code combining can be made possible provided that a proper diversity technique is used. A new link design employing multi-spot diffuse configuration (MSDC) with direction-diversity reception allows for modeling the system as one having a number of separate AWGN channels, creating a system with space diversity. Having such a model enables us to evaluate the optical power requirement and the overall system performance using code-combining techniques. This chapter considers using of infrared radiation as a medium for high-speed in-house wireless digital communication. Some background materials are first provided in section 4.1. MSDC link is then introduced and thoroughly examined in section 4.2. We will use the simulation results obtained in this chapter to evaluate the performance of the systems proposed in the following chapters.
4.1 Introduction

Utilization of infrared (IR) radiation to enable wireless communication has been proposed and widely studied [19, 20]. Infrared\textsuperscript{1} radiation, as a medium for short-range, indoor communication, has several advantages over radio\textsuperscript{2}.

1. The infrared spectral region, offers a virtually unlimited bandwidth, free of regulations. Infrared is close, in wavelength, to visible light and exhibits similar behavior.

2. Infrared light is absorbed by dark objects, diffusely reflected by light-colored objects and directionally reflected from shiny surfaces. It penetrates through transparent objects but not opaque barriers. Therefore, it is confined to the room it originates, preventing interference between links operating in different rooms. This also makes it easy to secure transmissions against casual eavesdropping.

3. When an infrared link exploits intensity modulation with direct detection (see section 4.1.2), the short carrier wavelength and large square-law detector lead to efficient spatial diversity that prevents multipath fading [21], a phenomenon typical of radio links. This greatly simplifies design of infrared links.

Infrared has some drawbacks as well. Some of these drawbacks are enumerated below:

1. Since Infrared cannot penetrate walls, communication from one room to another requires installation of access points that are interconnected via a wired backbone.

\textsuperscript{1}Infrared in this chapter refers to the near-infrared band between about 780 nm and 950 nm.

\textsuperscript{2}The term “radio” is inclusive of the frequency bands that are often referred to as “radio frequency”, “microwave” or “millimeter wave”.
2. Although the transmitting power level can be increased without fear of interfering with other users, transmitter power is limited by concerns of power consumption and eye safety regulations.

3. In many indoor infrared links, intense ambient infrared noise, arising from sunlight, incandescent lighting and fluorescent lighting, induces high noise in an infrared receiver and limits the effective operating range.

4. Typical infrared links are subject to multipath distortion induced by reflections off of walls, floor, and room objects causing intersymbol interference (ISI) further limiting the operating range, for a desired transmission rate.

4.1.1 Infrared Link Designs

Various link designs may be employed in infrared communication systems. The classification can be made based on degree of directionality and existence of a line-of-sight path between transmitter and receiver. Fig 4.1 illustrates several configurations according to this classification.

Directed links employ directional transmitters and receivers, which must be aimed in order to establish a link, while nondirected links employ wide-angle transmitters and receivers. While directed links maximize the power efficiency, nondirected links do not require aiming, which makes it convenient for mobile applications. LOS links rely on an uninterrupted line-of-sight path, while non-LOS links generally rely on reflection of the light from a diffusely reflecting surface like the ceiling. Although LOS paths have a better power efficiency and a lower multipath distortion, non-LOS links are more robust to shadowing.

It is also possible to establish hybrid links, that combine transmitters and receivers, having different degrees of directionality. Of all the possible configurations, non-directed non-LOS link configuration, often referred to as diffuse, is the most
desirable from the user’s point of view, since no alignment is required to use and the system does not require a line-of-sight path. Diffuse systems rely on reflections from ceiling, walls and other reflecting objects, within the room, to provide a good coverage with a high degree of robustness against blockage.

![Diagram of infrared link types](image)

**Figure 4.1.** Classification of simple infrared links according to the degree of directionality of transmitter and receiver and whether the link relies upon existence of a line-of-sight path in between.

### 4.1.2 Intensity Modulation and Direct Detection

Efficient operation of standard modulation techniques, designed for radio links, in wireless infrared systems must employ synchronized local oscillators to enable coherent optical detection. Unfortunately, coherent detection is not feasible in non-directed applications [21]. A practical wireless infrared link uses *intensity modulation* and *direct detection* (IM/DD). Intensity modulation is easily obtained by varying the bias current of a laser diode or LED, whereas direct detection
is obtained at the receiver using a photodiode which produces a photocurrent proportional to the optical power incident upon it. The modeling of infrared channels with IM/DD is illustrated in Fig. 4.2 and can be summarized in a simple form

$$y(t) = r \int_{-\infty}^{\infty} X(\tau) h(t - \tau) d\tau + n(t) \quad (4.1)$$

The received photocurrent $y(t)$ is the convolution of the transmitted optical power $X(t)$ with the channel impulse response $h(t)$, scaled by the photodetector responsivity $r$, plus an additive noise $n(t)$, which is usually modeled as white, Gaussian and independent of $X(t)$. Here, the channel model $h(t)$ is fixed for a given configuration of transmitter, receiver and intervening reflectors.

![Figure 4.2](image_url)

Figure 4.2. (a) Transmission and reception in an infrared link with intensity modulation and direct detection. (b) Modeling link as a baseband linear, time-invariant system having an impulse response $h(t)$, with signal-independent additive noise $N(t)$. Photodetector has a responsivity $r$.

The channel model given by Eq. 4.1, is identical to a conventional linear baseband channel. However, IM/DD infrared links are subject to two additional conditions that are somewhat unusual. First, the input waveform $X(t)$ cannot be negative,
since it describes an optical power. The second condition is the average transmitted power constraint

\[ P_t = \lim_{T \to \infty} \frac{1}{2T} \int_{-T}^{T} X(t)dt. \] (4.2)

which governs the eye safety considerations and electrical power consumption of the transmitter. Eq. 4.2 involves the time integral of \( X(t) \), rather than the usual \( X^2(t) \), which is appropriate when \( X(t) \) represents amplitude. The power constraint given by Eq. 4.2, implies that for a given average optical power \( P_t \), the receiver SNR can be improved by transmitting a waveform having a high peak-to-average ratio. One such waveform is provided by pulse-position modulation which is described later.

The average received optical power is given by

\[ P_r = H(0)P_t \] (4.3)

which is the product of the average transmitted power and the channel d.c. gain, represented by \( H(0) = \int_{-\infty}^{\infty} h(t)dt \).

4.1.3 Multipath Distortion in Nondirected Links

While nondirected propagation alleviates the need for physical alignment between the transmitter and receiver, a major drawback of this approach is signal distortion caused by reflections from ceilings, walls, and other objects. A result of multiple reflections on transmitted digital waveforms is intersymbol interference (ISI).

A useful measure of severity of intersymbol interference induced by a multipath channel \( h(t) \) is the channel root-mean-square delay spread \( D \). The delay spread is a remarkably accurate predictor of ISI-induced SNR penalties, independent of the
particular time dependence of the impulse response of that channel. It is computed using impulse response as a weighting factor:

\[ D = \sqrt{\frac{\int_{-\infty}^{\infty} (t - \mu)^2 h^2(t) dt}{\int_{-\infty}^{\infty} h^2(t) dt}} \]  

(4.4)

where mean delay \( \mu \) is given by

\[ \mu = \frac{\int_{-\infty}^{\infty} th^2(t) dt}{\int_{-\infty}^{\infty} h^2(t) dt} \]  

(4.5)

Note, the impulse response \( h(t) \) and delay spread \( D \) can be considered to be deterministic quantities, in the sense that as long as position of the transmitter, receiver and the intervening reflectors are fixed, \( h(t) \) and \( D \) are fixed. This stands in contrast to the time-varying nature of radio channels, where the r.m.s. delay spread is interpreted as a statistical expectation of a random process [18].

Another measure of severity of intersymbol interference is the channel magnitude response and the corresponding 3-dB bandwidth. The channel magnitude response is defined as \( 10 \log |H(f)| \) with \( H(f) \) being the Fourier transform of \( h(t) \), i.e.,

\[ H(f) = \int_{-\infty}^{\infty} h(t) \exp\{-j2\pi ft\} \, dt \]  

(4.6)

A very distorted channel has a high delay spread and a low 3-dB bandwidth. Throughout this dissertation, we will use the delay spread \( D \) and the 3-dB bandwidth \( f_{3dB} \) interchangeably as channel performance indicators.

4.1.4 Modulation and Demodulation Techniques

The most important criterion for evaluating modulation techniques is the average received power \( P_r \) required to achieve a desired BER. In evaluating the power requirements of modulation techniques at high bit rates, one must consider the impact of multipath ISI as well as any reduction of the ISI that can be achieved through equalization techniques.
The second important attribute of a modulation technique is receiver electrical bandwidth that it requires, since it is difficult to achieve a low noise power level over a wide bandwidth using large-area photodiodes. For all baseband modulation schemes, we define this bandwidth requirement \( BW \) as the span from d.c. to the first null of the PSD of the transmitted waveform \( X(t) \). Usually bandwidth requirement is presented in the form \( BW/R_b \), i.e., it is normalized to the bit rate \( R_b \).

In binary modulation techniques, transmission of binary symbols is of the form

\[
X(t) = A \sum_{n=-\infty}^{\infty} a[n]g_t(t - nT)
\]

where the \( n \)th binary symbol \( a[n] \in \{0, 1\} \) is mapped into the waveform \( a[n]g_t(t) \), \( A \) is a scale factor, chosen such that the average power defined by Eq. 4.2 has the desired value \( P_t \) and \( T \) is a symbol period.

Among all baseband modulation techniques suitable for wireless infrared links, on-off keying (OOK) and pulse position modulation (PPM) are the most popular ones. In what follows, we briefly discuss the performance of these modulation techniques.

**On-Off Keying**

OOK is the simplest to implement. For a distortionless channel, the ideal Maximum Likelihood (ML) receiver for OOK in AWGN consists of a continuous-time filter matched to the transmitted pulse shape followed by a sampler and a threshold detector set midway between “low” and “high” pulse amplitudes (see Fig. 4.3). The performance of OOK in multipath channels without equalization can be computed using the enumeration technique developed in [22]. Results show that the optical power requirement for a given BER essentially depends only on normalized delay spread (the delay spread normalized to a bit duration), regardless of the link configuration or the channel impulse response \( h(t) \). For high bit rates
(100Mb/s and above), unequalized OOK incurs very large power penalties, and yields irreducible BER for normalized delay spreads above about 0.6, implying that on multipath channels with no diversity, unequalized OOK reception is not feasible at high bit rates.

Figure 4.3. Continuous-time block diagram of OOK system.

In the presence of multipath ISI, the optimum receiver for OOK employs a whitened matched filter (WMF) front-end and performs maximum-likelihood sequence detection (MLSD), which can be implemented efficiently using the Viterbi algorithm [17, 18]. MLSD is extremely effective in mitigating ISI, reducing the normalized power requirements. A practical, though sub-optimal, means to mitigate the multipath ISI is using a decision-feedback equalizer (DFE) [17, 18], which can adapt automatically to the channel impulse response. Power requirements of OOK links using DFE have been evaluated in [23]. Results show that power requirement of a DFE is very close to those using MLSD. At high bit rates (100Mb/s) no irreducible BER is observed with a DFE, in contrast to the unequalized case. However, power requirements are still high for shadowed diffuse channels.

**Pulse-Position Modulation**

Block diagram of a $L$-PPM system is shown in Fig. 4.4. Each group of $\log_2 L$ bits is encoded into a $L$-PPM waveform $P_i(t), i = 1, \ldots, L$, which has a duration
Each $P_i(t)$ includes one “chip” of unit amplitude and duration $T/L$, in addition to $L - 1$ chips of zero amplitude. Sequence $P_i(t)$ forms chip waveform $b(t)$, which is scaled by $LP_i$ so that the total average power is equal to $P_t$.

Figure 4.4. Continuous-time block diagram of $L$-PPM system.

PPM is an orthogonal modulation scheme that offers a decrease in average power requirement compared to OOK, at the expense of an increased bandwidth requirement. For a given bit rate, $L$-PPM requires more bandwidth than OOK, by a factor $L/\log_2 L$, e.g., 4-PPM requires two times more bandwidth than OOK. In the absence of multipath distortion, $L$-PPM yields a decrease in average-power requirement that decreases steadily with increasing $L$. Here, the increased noise associated with $L/\log_2 L$-fold wider receiver bandwidth is outweighed by an $L$-fold increase in peak power. In addition to increased bandwidth requirement, two drawbacks of PPM, as compared to OOK, are an increased transmitter peak-power requirement and a need for both chip-level and symbol-level synchronization.

In the absence of multipath distortion, an optimum ML receiver for $L$-PPM employs a continuous-time filter matched to one chip, whose output is sampled at chip rate. Each block of $L$ samples is passed to a block decoder, which makes a symbol decision, yielding $\log_2 L$ information bits. In soft-decision decoding, samples are unquantized and a block decoder chooses the largest of $L$ samples. In hard-decision decoding, each sample is quantized to “low” or “high” using a simple threshold
detector, and the block decoder makes a symbol decision based on which sample is “high”, mediating in cases where no samples, or more than one sample is “high”. Hard decisions result in approximately 1.5-dB optical power penalty with respect to soft decoding [24].

When $L$-PPM is transmitted over a multipath channel, nonzero transmitted chips will induce interference in chips both within the same symbol and chips in adjacent transmitted symbols. These two effects, collectively, are referred to as ISI. Performance of $L$-PPM on non-directed indoor infrared channels (ISI channels), in the absence of equalization is evaluated in [25, 26]. Receiver in this case uses the same receiving filter and a soft-decision decoder as shown in Fig. 4.4 and is optimal only on a distortionless channel. Results show that for small values of normalized delay spread (i.e., small delay spread and/or low bit rates), the power requirement is reduced by using a larger $L$. As normalized delay spread increases, however, power requirement increases more rapidly for PPM than for OOK, and increases more rapidly for a large $L$, because of a short chip duration. Unlike OOK and 2-PPM, $L$-PPM with a large $L$ may incur an irreducible BER.

In presence of multipath ISI, optimum PPM receiver employs a chip-rate WMF, followed by a MLSD, which can be implemented using a symbol rate Viterbi algorithm [21]. Use of MLSD significantly improves the performance of PPM [25, 26], preventing irreducible BERs and restoring the average-power gain obtained by increasing the value of $L$. However, as the normalized delay spread increases, even when this optimum detection technique is employed, power requirement of PPM increases much more rapidly than those of OOK.
4.2 Multi-Spot Diffusing Configuration (MSDC) and Direction Diversity Reception

In section 4.1 various link configurations for Infrared communication systems were discussed (see Fig. 4.1). Link design classification was made based on degree of directionality and existence of a line-of-sight path between transmitter and receiver. Two of these IR links are becoming increasingly popular: (1) directed LOS links, such as those standardized by the Infrared Data Association (IrDA) [27] and (2) non-directed non-LOS links, also known as diffuse links. Directed LOS links offer low path loss and low delay spread, but require aiming of transmitter and are subject to interruption by blockage of beam. Diffuse links, on the other hand, avoid the need for aiming and can tolerate partial obstruction of transmission path and therefore are the most desirable, from user’s point of view. However, they greatly suffer from increased multipath distortion resulting in intersymbol interference.

All these link configurations employ a single-element receiver. A single-element receiver consists of an optical concentrator, whose output is coupled to a single photodetector. In a single-element receiver, the desired signal, delayed multipath components, ambient light noise, and co-channel interference are combined into a single electrical signal.

In this section, we present two modifications which together can result in a significant performance improvement over that of a diffuse link configuration. First, multiple-spot diffusing configuration (MSDC) is used to transmit signal power in form of multiple collimated beams with each beam aiming at a pre-specified direction. Such a transmitting scheme produces multiple line-of-sight (as seen by the receiver) diffusing spots, all of equal power, on an extended reflecting surface, e.g., ceiling of a room. Second, a direction-diversity receiver (also known as angle-diversity receiver) is used, in order to provide a diversity scheme for optimal
rejection of ambient noise power and to substantially reduce the ISI effect. A
direction-diversity receiver utilizes multiple receiving elements, with each element
pointed at a different direction [41, 30]. The photocurrents received in various
elements are amplified separately and resulting electrical signals can be processed
in one of several ways, as described in section 4.2.3. We will see that these
two modifications combine advantages and overcome drawbacks of both direct
line-of-sight and diffuse configurations and result in a significant improvement in
performance.

The remainder of this chapter is organized as follows. In the next section,
we study multiple-spot diffuse configuration (MSDC). A framework for computer
simulations is provided under which properties of room and transmitter are quan-
tified. In section 4.2.2, channel characteristics under MSDC, using a single-element
receiver with a wide FOV is analyzed and compared with that of a pure-diffuse
configuration. Section 4.2.3 is devoted to general concept and rational behind
direction-diversity reception. Various combining techniques for processing the re-
ceived signals at the input of a direction-diversity receiver are also discusses in this
section. In section 4.2.4 we study channel impulse response under MSDC, using a
multiple-branch direction diversity receiver.

4.2.1 Multiple-Spot Diffusing Configuration: Modeling

In this configuration, as opposed to the diffuse configuration, transmitter projects
the light power in form of multiple narrow beams of equal intensity, over a regular
grid of small areas (spots) on a diffusing surface such as a ceiling (see Fig. 4.5). Each
diffusing spot in this arrangement may be considered as a line-of-sight light source
which can be aimed at by a receiver having a wide or narrow field-of-view (FOV).
This way, a good portion of total optical power can be received by each receiving
branch via a finite number of distinct signal paths; a number equal to number of
spots seen by that branch. Each beam should have a divergence angle large enough to make it eye-safe, but small enough that it does not spread excessively when traversing a room. A divergence semi-angle of about $2^\circ$ might be typical. Such a transmitter design, in general, can be used with a single-branch receiver as well as with a multiple-branch receiver as discussed below.

![Diagram](image)

(a)

![Diagram](image)

(b)

Figure 4.5. (a) Diffuse configuration, using a single element receiver. (b) Multiple-spot diffusing configuration using a multi-branch direction-diversity receiver.

There are various ways to implement a multi-beam transmitter. The most straightforward approach is to have several light sources aiming at different di-
rections. For practical reasons, we can not have a large number of beams. A better alternative is to use a holographic optical element mounted on a laser diode. Holograms can be fabricated by conventional optical means, using a multiple-exposure technique [43]. Unfortunately, an exact prescribed ratio between intensity of the beams cannot be achieved with this technique. As a result, optical power cannot be homogeneously distributed within a room. Furthermore, this technique cannot be used for very large spot arrays, nor can it be used for asymmetrical spot arrays. Computer generated holograms (CGH) can be used as a practical alternative to produce wave-fronts with any prescribed amplitude and phase distribution [42]. An ideal wave-front can be computed on the basis of diffraction theory and be encoded into a tangible hologram.

To evaluate the characteristics of the proposed configuration and to perform comparisons with diffuse configuration, a framework for computer simulation must be provided. We consider an empty room with dimensions 6m×6m×3m, as shown in Fig. 4.6. A multi-beam transmitter is assumed to produce 100 beams of equal intensity by means of a CGH [42] to form a 10 × 10 square lattice of equidistant spots of diameter $d = 5\text{cm}$ and spacing $S = 60\text{cm}$, on the ceiling. Transmitter is located at the center of the room at a desktop level (0.9m height) and pointed upward. There is a large window at one side of the room extended from desktop level up to the ceiling. Ceiling, walls, floor, and window are modeled as Lambertian reflectors of first order with reflectivities 0.7, 0.6, 0.1, and 0.1, respectively [31].

To produce realistic simulation results, we assume source of background light in the room is sunlight coming through the window and nine 100W Tungsten lamps, positioned equidistantly on the ceiling. In the spectral range of interest (800-900nm), background spectral radiance of diffuse sky is on the order of $10^{-6}\text{W/cm}^2\text{.nm.sr}$ [45], resulting in a spectral density of 1.3µW/nm for each window.
elementary surface area of 1 cm$^2$. Lamps can be modeled as generalized Lambertian sources of order 2 and optical spectral density of 0.037W/nm [44]. We also assume the total capacitance of receiver is very small. With this assumption, background light due to sunlight and Tungsten lamps is the primary source of noise at the receiver. This noise can be characterized as white and Gaussian with a double-sided power spectral density $S_{\text{shot}}(f) = N_0 = rA_{\text{det}} q i_b$, where $A_{\text{det}}$ is area of photo-detector, $q$ is an electron charge and $i_b$ is irradiance of background light on detector surface [44].

### 4.2.2 MSDC with a Single-Element Receiver

We now look at the characteristics of MSDC using a single-element wide FOV receiver and compare it to that of a pure diffuse configuration. We model the diffuse transmitter as one having an angular distribution of the emitted optical power given by the generalized Lambertian law:

$$P(\theta) = \frac{n + 1}{2\pi} P_t \cos^n \theta$$  \hspace{1cm} (4.8)
with $n = 1$. This model for pure diffuse links has been used in experimental links. In both cases, a total power of 1W is assumed.

Simulation model developed by Barry et al. [22] is used to calculate the channel impulse response. We take the elementary surface elements to have a size of 5cm×5cm for a first bounce. The corresponding elementary time period, (i.e. the time that light needs to travel between the corners of any two neighboring surface elements) is 1.033ns. Reflections up to second order are considered in our simulation. Path loss between transmitter and ceiling was neglected due to the highly collimated nature of beams emerging from the transmitter. The result of simulation is shown in Fig. 4.7.

Impulse response of a multi-spot diffuse link differs considerably compared to that of a diffuse channel. It consists of many peaks corresponding to all the diffusing spots seen by the receiver. A sample of impulse response is shown in Fig. 4.7-b and is compared with that of a pure diffuse link shown in Fig. 4.7-a. Both responses are detected via a 70°-FOV receiver, placed at the position (0.9m, 0.9m, 1.0m) in the room shown in Fig. 4.6.

Figs. 4.7-c and 4.7-d represent the signal power path loss for pure diffuse link and MSDC link, respectively. Due to uniform distribution of spots on ceiling, multispot-diffuse link offers a more uniform distribution of received optical power compared to the case when a Lambertian source is used.

What is not so intuitively clear, and gives rise to main advantage of multi-spot diffuse link over a pure diffuse link is the r.m.s. delay spread distribution shown in Figs. 4.7-e and 4.7-f. Delay spread of a MSDC link decreases as the receiver moves away from the transmitter. It is higher in the center of the room than near the walls and the corners. This is because, in the center of the room, signals with comparable intensity come from all directions, whereas, near walls the number of contributing diffusing spots is smaller. Thus, in MSDC links, increase of path loss with distance
Figure 4.7. The figures on the top represent the channel impulse response at position (0.9m, 0.9m, 1.0m) for (a) pure diffuse link and (b) MSDC link. Two figures on the middle are optical path loss distributions for (c) pure diffuse link and (d) MSDC link. Figures on the bottom represent the channel r.m.s. delay spread distribution for (e) pure diffuse link and (f) MSDC link. In all cases, a single-element receiver with a 90-degree FOV is assumed.
from the center is combined with a decrease in delay spread. The maximum path loss in this case is about 1dB less than that of Lambertian pattern illumination. Delay spread, on the other hand, is nearly less by a factor of two. These simulation results suggest that the overall performance of MSDC link is slightly better than that of a pure diffuse link, in this case.

### 4.2.3 Direction-Diversity Reception

The conventional approach for designing a non-directed receiver with a wide FOV is to use a single-element receiver with one photo-detector, such as the one used in the previous section. This element collects not only the desired signal, but also unwanted ambient light noise. Steady light sources, such as the sun and incandescent lamps, lead to white, nearly Gaussian, shot noise, while periodically modulated sources, such as fluorescent lamps, lead to a cyclostationary noise component [20]. A wide-FOV receiver collects not only the primary beam, but also the ones that have undergone one or more reflections from room surfaces, and are thus delayed. These reflected components, while increasing the collected signal power, lead to multipath distortion.

In direction-diversity (also known as angle-diversity) reception, an array of narrow-FOV receiving elements are oriented along different directions to cover a wide FOV. Each receiving element is equipped with a separate pre-amplifier, and the resulting signal can be processed in various ways as discussed below. Angle diversity receivers offer several advantages over single-element receivers. They can achieve optical gain over a wide field-of-view (FOV). They can significantly reduce the effect of ambient light noise, co-channel interference, and multipath distortion, due to the fact that these unwanted signals are in many cases received from a different direction than the desired signal.
A direction-diversity receiver can be implemented in two ways. One way is to use a composite receiver made of multiple nonimaging elements that are oriented in different directions, as in [30]. Implementation of direction diversity using non-imaging elements requires a separate optical concentrator for each receiving element, which may be excessively bulky and costly. Another way, proposed by Yun and Kavehrad is the fly-eye receiver [41], also known as imaging receiver [29], which consists of a single imaging concentrator (e.g., a lens) that forms an image of the received light on the collection of photo-detectors, thereby separating signals that arrive from different directions. Implementation of an angle-diversity receiver using this scheme offers two advantages over a non-imaging implementation. First, all photo-detectors share a common concentrator, reducing size and cost. Second, all the photodetectors can be laid out in a single planar array, facilitating the use of a large number of receiving elements.

When the signal from a single transmitter is received by the direction-diversity receivers, the overall communication system can be described as a single-input, multiple-output system:

\[ y_j(t) = rX(t) \otimes h_j(t) + n_j(t), \quad j = 1, 2, \ldots, J \quad (4.9) \]

where \( J \) is the total number of receiving elements and the symbol \( \otimes \) denotes the convolution. Here, \( r \) is the detector responsivity, \( X(t) \) is transmitted optical signal, \( h_j(t) \) is the impulse response between the transmitter and the \( j \)th receiving element, \( n_j(t) \) is the noise in the \( j \)th element, and \( y_j(t) \) is the photocurrent in the \( j \)th element.

The advantages achieved by direction-diversity reception depends on how the signals received via the different elements are processed and detected. This is summarized in Table 4.1 [20].

When multipath distortion is significant, the optimum reception technique is maximum-likelihood combining (MLC). In MLC, each \( y_j(t) \) is processed by a sep-
Table 4.1. Combining techniques for direction-diversity receivers.

<table>
<thead>
<tr>
<th>Advantage</th>
<th>MLC</th>
<th>MRC</th>
<th>SD</th>
<th>EGC</th>
</tr>
</thead>
<tbody>
<tr>
<td>Achieves high optical gain and wide FOV simultaneously.</td>
<td>X</td>
<td></td>
<td>X</td>
<td></td>
</tr>
<tr>
<td>Mitigates co-channel interference.</td>
<td>X</td>
<td></td>
<td>X</td>
<td></td>
</tr>
<tr>
<td>Mitigates multipath distortion.</td>
<td>X</td>
<td>Under some circumstances</td>
<td>X</td>
<td></td>
</tr>
<tr>
<td>Can employ a single preamplifier.</td>
<td></td>
<td></td>
<td>X</td>
<td></td>
</tr>
<tr>
<td>Avoids need for channel and noise estimation.</td>
<td></td>
<td></td>
<td></td>
<td>X</td>
</tr>
</tbody>
</table>

arate continuous-time matched filter $h_j(-t)$. The $J$ matched-filter outputs are sampled and combined in a memoryless fashion [30], with the $j$th sample weighted in inverse proportion to the PSD of the noise $n_j(t)$. This weighted sum of the $J$ sample sequences is a sufficient statistic, by which, the receiver performs MLSD which can be implemented using the Viterbi algorithm. Thus, implementation of MLC requires separate estimation of each of the $J$ channel impulse responses and noise PSDs.

The complexity of MLC is likely to be too high for many applications, and a number of simpler approaches are possible. An easily implemented class of techniques involves combining the $y_j(t)$ in a memoryless, linear combiner, filtering the weighted sum using a single continuous-time filter, and sampling the filter output. The resulting sampled process can be processed using MLSD, DFE, or a simple slicer to yield the detected data. This class of techniques includes maximal-ratio combining (MRC), selection diversity (SD), and equal-gain combining (EGC), which differ according to how the combining weights are chosen.

In MRC, the $y_j(t), j = 1, \ldots, J$, are summed together with weights proportional to signal current to noise-PSD ratios, thereby maximizing the SNR of the weighted sum. When multipath distortion is not significant, the optimum MLC reduces to MRC, followed by a simple slicer. MRC can result in a net decrease in multipath distortion, as compared to a single, wide FOV receiver, as long as the ambient light
noise and multipath reflections both arrive from directions sufficiently far away from the strong signal components. If this is not the case, however, an increase in multipath distortion could result [20]. MRC requires estimation of the SNR values in each of the receiving elements, representing an increase in complexity over non-diversity reception.

In SD, only the signal having the best SNR is utilized. This technique can often separate the signal from ambient-light noise, resulting in a SNR improvement, but the gains are not as large as those achieved using MRC. SD can yield a significant reduction in multipath distortion, provided that directional receiving elements are employed, making it a promising technique for high-bit-rate systems. SD is probably not much simpler to implement than MRC, since it still requires SNR estimation.

In EGC, the multiple signals are summed together with equal weights. This technique increases the receiver FOV, but is unable to separate signal from noise or co-channel interference. Moreover, it can result in an increase of multipath distortion making it unsuitable for very high-bit-rate links. It is attractive in its simplicity, as it avoids the need for SNR estimation, and the signals from several photodetectors can be processed by a single preamplifier. EGC is used in some diffuse systems with bit rates up to 4Mb/s, helping them achieve very robust operation, even in the face of shadowing [32].

4.2.4 MSDC with a Multiple-Element Direction-Diversity Receiver

In this study, we consider a composite receiver, consisting of 7 branches of equal FOV, $\Phi_b$ (see Fig. 4.8). Each branch is made of a bare photodetector with a photosensitive area of $1\text{cm}^2$. For simplicity, the responsivity, $r$, of the photodetector is taken to be a constant, equal to 0.6 A/W, over 200nm receiver spectral bandwidth. The central branch is oriented directly towards the ceiling,
while side branches are tilted by $2 \times \Phi_b$ degrees. Two important cases are considered. In the first case, we assume $\Phi_b$ is made large enough so that at least one spot can be sighted by each branch. To calculate $\Phi_b$ we need to consider only the central branch, as its coverage area on the ceiling is encompassed by the coverage area of each one of the side branches. With the aid of Fig. 4.9-a we get $\Phi_b \geq \tan^{-1}(S/h\sqrt{2}) = 11.42^\circ$, where $h$ represents the distance between the receiver and the ceiling. We choose $\Phi_b = 11.5^\circ$. The overall FOV of the receiver thus becomes $3 \times 11.5^\circ = 34.5^\circ$.

![Figure 4.8. Field-of-view of a 7-branch composite receiver.](image)

In the second case, we decrease the value of $\Phi_b$ until no more than one spot is seen by each branch. Such a configuration creates a nearly ideal channel between the transmitter and each active branch (one that sees a spot). To satisfy this condition, we only need to consider one of the side branches as its coverage area on the ceiling encompasses the coverage area of the central branch. With the aid of Fig. 4.9-b, the required FOV $\Phi_b$ is obtained using the relation

$$h [\tan(3\Phi_b) - \tan(\Phi_b)] \leq S - d$$

(4.10)

This yields $\Phi_b \leq 7^\circ$. We choose the case of $\Phi_b = 7^\circ$. The overall FOV of the receiver thus becomes $3 \times 7^\circ = 21^\circ$.

The simulation model developed by Barry et al. [22] is used, once again, to calculate the channel impulse response. This time we take the elementary surface
elements to have a size of 1cm×1cm for the first bounce. The corresponding elementary time period, (i.e., the time that light needs to travel between the corners of any two neighboring surface elements) is 1.033ns. Reflections up to third order are considered in our simulation. To keep computation time within reasonable limits, the elementary surface area for the second and third bounces are taken to be 5cm.

We simulated the impulse response \( h_j(t) \) corresponding to all seven branches \( (j = 1, \ldots, 7) \) for a receiver that is placed (i) near the wall at \((0.60m,3.80m)\) and rotated 47°, (ii) away from walls at \((3.68m,3.86m)\), rotated 36° and (iii) at the room corner at \((0.60m,0.45m)\), rotated 13°. Rotations are counterclockwise, about normal, as seen from the ceiling (see Fig. 4.10). Fig. 4-a shows the channel impulse response and the corresponding frequency response for all 7 branches of an 11.5°-FOV receiver placed near wall at position (i). The results for positions (ii) and (iii) are omitted to avoid redundancy. In general, with 11.5° FOV, channel response on each branch consists of a number, between zero to four, of sharp impulses. This
Figure 4.10. Areas on the ceiling seen by a 7-branch composite receiver with a branch FOV of (a) $\Phi_b = 11.5^\circ$ and (b) $\Phi_b = 7^\circ$, placed (i) near the wall at (0.60m,3.80m) and rotated 47° (ii) away from the walls at (3.68m,3.86m), rotated 36° and (iii) at the room corner at (0.60m,0.45m), rotated 13°.
Figure 4.11. (a) Channel impulse response and frequency response for all 7 branches of an 11.5°-FOV receiver placed near the wall at (0.6m, 3.8m) and rotated by 47°.
Figure 4.11. (b) Channel impulse response and frequency response for all 7 branches of a 7-FOV receiver placed near the wall at (0.6m, 3.8m) and rotated by...
number for each branch is equal to the number of spots that lie within the FOV of that branch. There are also degenerate cases where the number of impulses does not seem to match the number of spots. This is because the length of the signal path, from the transmitter to the receiver, is almost the same for the degenerate spots within the branch. In this case, the impulses lie so close in time to one another that cannot be viewed as separate signals. The minimum 3dB bandwidth, $f_{3dB}$, among all active branches (those that can see at least one spot) in Fig. 4-a is 133 MHz. Fig. 4-b shows the channel impulse response and the corresponding frequency response at all 7 branches of a 7°-FOV receiver that is placed at the same position (i). In this case, the response of an active branch consists only of one impulse. As a result, data rates of hundreds of megabits per second and more can be supported without any need for equalization. In both cases, the first bounce comprises more than 92 percent of the total received power, while the first three bounces comprise more than 98 percent of the total received power. The second order reflection comes into effect only for side branches that cover some portion of a wall. The effect of second and third reflections on the frequency response is seen to be insignificant so that the first bounce can provide sufficient information on the achievable data rate.

The above two designs create nearly ideal channels of high bandwidth and very small ISI. Such an increase in bandwidth is paid for by an increase in the path loss. In the next two chapters, we use power efficient signaling schemes to compensate for this path loss and, based on these schemes, we evaluate the required transmit power that achieves reliable communications at almost all receiver positions within the room.

At this point, it is reasonable to ask if use of a multi-element receiver in a diffuse configuration would be as effective as it is in MSDC. Simulations show that
Figure 4.12. Channel impulse response and magnitude response in a diffuse configuration using a single-element receiver. The receiver is placed at the room corner at (0.5m,0.3m).
Figure 4.13. Channel impulse response and magnitude response under MSDC, using a single-element 12° FOV receiver. The receiver is placed at the room corner at (0.5m, 0.3m).
Figure 4.14. Channel impulse response and magnitude response under MSDC, using a single-element 7° FOV receiver. The receiver is placed at the room corner at (0.5m,0.3m).
the percentage of optical power, received after the first reflection, is substantially higher in the case of Lambertian pattern illumination than in MSDC. Consequently, the decrease in the 3-dB bandwidth due to higher order reflections in the diffuse configuration is much more dramatic than in MSDC. This can be seen by first looking at Fig. 4.12, in which the impulse response and the magnitude response of a 70° FOV receiver in a diffuse configuration is compared to those of a 30° FOV receiver, assuming they are both placed in the same position at the room corner. As seen, the 3-dB bandwidth, after the third bounce, is about the same for both receiver configurations. Therefore, narrowing down the FOV does not result in any significant increase in bandwidth. This is also intuitively true as a decrease in the FOV, must render signal attenuation for all three reflections, keeping the delay spread unchanged. This is, however, not the case for MSDC, as shown in Fig. 4.13. Here, we made the FOV as narrow as 12°, but used MSDC instead. Compared to the previous case, the signal attenuation was maintained the same, but the 3-dB bandwidth increased by an order of magnitude. We can reduce the FOV until we reach the ideal case of 7°, while still receiving a good portion of the optical power (see Fig. 4.14). Note that, the relative size of spots has a secondary effect in determining the 3-dB bandwidth. In fact, it will not cause any significant change up to frequencies above 2 GHz. Therefore, the main source of channel distortion when using MSDC is the existence of more than one spot within the FOV of each branch.

4.2.5 Summary

We intend to evaluate the system performance under MSDC using a multiple-branch direction-diversity receiver. In this chapter, we investigated some of the link characteristics such as the channel impulse response, magnitude response, link coverage range, delay spread distribution and the 3-dB bandwidth, using a single-
branch wide FOV receiver as well as a 7-branch direction-diversity receiver. This study was mainly via computer simulations for a specific room of a given dimension. Most of the results, however, are general in that they do not change significantly by changing the room configuration and dimensions. This model will be used in the subsequent chapters to evaluate the system performance under various coding and modulation schemes and receiver designs proposed in these chapters.

The infrared link made by MSDC has a very high path loss. Therefore, efficient combining schemes, such as MLC may be employed, to reduce the optical power penalty. Significant improvements on the power requirements can be made possible through use of code combining techniques. Code combining represents a technique for combining a minimum number of repeated packets encoded with a code of rate $R$ to obtain a lower rate, and thus more powerful, error-correcting code, capable of allowing communications when channel error rates are less than 50 percent. Two critical features of code combining are as follows:

1. Maximum-likelihood decoding, rather than bounded minimum distance (algebraic) decoding, of $J$ noisy packets as a single low-rate code of rate $R/J$.

2. weighting of each packet by an estimate of its reliability (soft decisions on packets).

The code combiner treats the $J$ received packets as a code of rate $R/J$ and attempts to pick the most probable information packet. The decoder is a soft decoder with the capability of weighting the reliability of each received packet. It is important to note that code combining is designed to work in a very noisy (e.g., infrared links with shadowing and intense ambient light) environment, where conventional diversity schemes, such as maximal ratio combining, etc., can easily break down. This is discussed in more detail in the next chapter.
Chapter 5

CODE COMBINING BASED ON VITERBI DECODER FOR WIRELESS INFRARED SYSTEMS EMPLOYING DIRECTION DIVERSITY

In this chapter, we evaluate the performance of an infrared link composed of a multi-beam transmitter in conjunction with a direction-diversity receiver, employing code combining. Code combining represents an added dimension to the conventional diversity concepts which are limited to combining the individual received symbols. Rate compatible punctured convolutional codes are used to encode intensity modulated OOK optical power, to provide an adaptive environment for efficient utilization of channel spectral bandwidth, and to maintain a guaranteed bit-error rate (BER) performance at all receiver positions. It is shown that a BER not exceeding $10^{-9}$ with 99% probability can be achieved at bit rates up to a few hundreds of Megabits per second, at low transmitted power levels.

5.1 Introduction

The overall performance of an indoor wireless infrared link is closely related to (a) degree of directionality of transmitters and/or receivers, and (b) orientation of transmitters and receivers with respect to one another. Various link designs have been proposed, and analyzed [20]. Directed line-of-sight (D-LOS) links are usually very power-efficient and relatively free from multipath distortion. Such links, however, are very sensitive to blockage. Besides, the desired performance
cannot be achieved unless transmitter and receiver are in line-of-sight. Nondirected non-line-of-sight (ND-NLOS) links, also known as diffuse links, have been under investigation by researchers in the past decade [23]. In contrast to D-LOS links, diffuse links are highly robust against blockage and do not require aiming of transmitter or receiver. Such links, however, are subject to a high path loss and multipath distortion. Furthermore, with a conventional transmitter design, optical power in such links cannot be uniformly distributed over the coverage area. In fact, power efficiency is reduced as the receiver moves away from the transmitter [42]. More efficient utilization of such links thus requires new designs for the transmitter.

The multi-spot diffusing configuration (MSDC), introduced in section 4.2.1, takes advantage of benefits of the previously proposed link designs, while overcoming their drawbacks [41, 42]. In this design, the light power from a transmitter is projected, in form of narrow beams of equal intensity, over several small areas on a diffusing surface. This can be achieved using computer generated holograms (CGH) as discussed in Sec. 4.2.1 and in [42].

MSD configuration has been compared with pure diffuse configuration, using a conventional single element wide field-of-view receiver, and is shown to yield an extended coverage area [42]. Nevertheless, efficient utilization of this transmitter design is only possible when the configuration is used in conjunction with multiple-element narrow field-of-view (FOV) direction-diversity receivers (also known as angle-diversity receivers). Each diffusing spot, in this arrangement, may be considered as a line-of-sight light source, which can be covered by a very narrow FOV receiving branch, thereby, eliminating a great amount of ambient noise power while accepting the same amount of signal power. The narrow FOV also relaxes the problem with spectral response shift of the interference filters used for rejecting the ambient light. Finally, there are only a finite number of distinct signal paths
between a transmitter and a receiving branch, resulting in an undistorted channel, and making the system theoretically easily tractable.

Conventional diversity combining techniques, used for direction-diversity infrared systems include, maximum-likelihood combining (MLC), maximal ratio combining (MRC), selection diversity (SD) and equal gain combining (EGC). Significant improvement on the performance of these combining schemes can be made possible through code combining [6]. Two key features of code combining are (1) maximum-likelihood decoding, rather than bounded minimum distance (algebraic) decoding, of $J$ noisy packets as a single low-rate code of rate $R/J$ and (2) weighting of each packet by an estimate of its reliability (soft decisions on packets). Note, with diversity combining a single error due to an unexpected change on the channel can cause the entire weighted sum of the $J$ received symbols be detected incorrectly, while with code combining, a single error can be typically corrected by the more powerful rate-$R/J$ code. Code combining is a technique for combining entire packets, representing an added dimension to the conventional diversity concepts limited to combining only individual symbols.

In this chapter, we evaluate the performance of an adaptive-rate direction-diversity communication system that is designed for MSD link configuration and uses code combining at the receiver. In the next section, design of the proposed link is briefly reviewed. In section 5.3, punctured convolutional codes are used to provide an adaptive environment for efficient utilization of channel spectral bandwidth while maintaining bit-error rate below a certain level at all receiver positions. We have used a maximum-likelihood decoding algorithm for all $J$ received blocks at the receiver to obtain an estimate of the transmitted information block. In section 5.5, we demonstrate how we can obtain a very accurate estimate of the channel by taking advantage of Viterbi decoder side information. This chapter is ended by
5.2 Link Design

In this section, we evaluate the channel characteristics composed of a multi-beam transmitter and a multi-branch direction-diversity receiver. For our study we consider the empty room of section 4.2.1. A multi-beam transmitter is assumed to produce 100 beams of equal intensity by means of a CGH to form a $10 \times 10$ square lattice of equidistant spots on the ceiling. The transmitter (base station) is placed at the center of the room and pointed upward.

We consider a composite receiver, consisting of 7 branches of equal FOV, $\Phi_b$ sitted on the same level as the transmitter. Each branch is made of a bare photodetector. The central branch is oriented directly towards the ceiling, while the side branches are tilted $2 \times \Phi_b$ degrees. In order to achieve a high optical gain, the FOV for each receiving element is assumed to be wide enough so that at least one spot is sighted by each branch at every possible receiver position in the room. Note that, it is sufficient to consider only the central branch. This is because the area on the ceiling covered by each side branch is an ellipse of semiaxes $a$ and $b$, both greater than the radius of the circular area covered by the central branch. The branch field-of-view was calculated to be $\Phi_b = 11.4^\circ$ (refer to section 4.2.4 for details). We take $\Phi_b = 11.5^\circ$ for our case study. The overall FOV of the receiver thus becomes $3 \times 11.5^\circ = 34.5^\circ$.

The simulation model developed by Barry et al. [22] is used to calculate the characteristics of the channel. To produce more realistic simulation results, we assume there are nine 100 W Tungsten lamps, positioned equidistantly on the ceiling. We assume the total capacitance of the receiver is very small. With this assumption, the background light due to the sun light coming through window, and the Tungsten lamps is the primary source of noise at the receiver. This noise can
be characterized as white and Gaussian with a double-sided power spectral density
\[ S_j^{\text{shot}}(f) = N_0 j = r A_{det} q i_{b_j}, \]
where \( q \) is the electron charge in Coulomb and \( i_{b_j} \) is the irradiance of the background light on the detector surface at the \( j \)th branch [44].

Simulation results for the channel impulse response, \( h_j(t) \), at all 7 branches \((j = 1, \ldots, 7)\) are depicted in Fig. 4.11-b. Reflections up to the third order were considered. For each branch, the impulse response consists of a number of impulses, equal to the number of spots that lie within the FOV of that branch. Note that, the second order reflection comes in to effect only for side branches that cover part of the wall. Both second and third order reflected signals were seen to be too small to produce any significant change in the received power (less than 5 percent). Note, this would not be the case had we used a diffuse transmitter. Plot of magnitude response showed that the 3-dB cut off frequency is on the order of one hundred Megahertz and more, making the assumption of ideal channel valid for symbol rates of one hundred Megabits per second and more\(^1\).

5.3 Communications System Design

Wireless infrared systems are essentially considered as power-limited systems. Therefore, efficient coding schemes may be employed to reduce the optical power requirement at the cost of a reasonable increase in the transmission bandwidth. One way of doing this, which is particularly suitable for space diversity systems such as the one considered in this work, is to send multiple replicas of a coded information block over \( J \) independent channels and to use maximum-likelihood decoding algorithm for all \( J \) received blocks at the receiver to obtain an estimate of the transmitted information block. This way, information is coded not only in time,

\(^1\)Such a symbol rate is comparable to (if not exceeding) the speed of commercially available Viterbi decoders.
but also in space, resulting in a spectrally efficient transmission strategy with a high degree of reliability. In this section we use punctured convolutional codes to provide an adaptive environment for efficient utilization of channel spectral bandwidth. We use maximum-likelihood code combining for it is optimum and matches very well with the structure of convolutional codes. The latter is true since decoding can be achieved with a simple Viterbi algorithm, as shown later in this section.

A simplified block diagram of the proposed infrared communication system is shown in Fig. 5.3. This system can be used for the link proposed in Section 5.2. A binary data source generates equally probable binary symbols at a rate $R_b$ bits/s. Each information block of $K$ bits is sent into a rate-$1/n$ convolutional encoder of memory length $M$. Through the coding process, a rate-$p/(np - \delta)$ punctured convolutional code of puncturing period $p$ is formed by periodically deleting $\delta$ code symbols from every $np$ code symbols created by the original $1/n$-rate mother code [5]. The resulting coded block has a length $N = (np - \delta)(K + M)/p$, provided that $p$, $K$ and $M$ are chosen such that $(K + M)/p$ is an integer. The code symbols “0” and “1” are then mapped into -1 and +1, respectively, prior to entering the pulse shaping filter. The electric signal at the output of this filter in general takes both positive and negative values. A DC bias must therefore be added before the signal is converted to optical power so that no data is lost. Note, in order to keep the information rate equal to $R_b$ bits/s, code symbols must be transmitted at a higher rate $R_s = 1/T = R_b/R$ symbols/s, where $R = K/N \approx p/(np - \delta)$ is the code rate. Let $g_t(t)$ represent the impulse response of the pulse shaping filter. The transmitted optical power $X(t)$ can be written as:

$$X(t) = P_{dc} + P\sqrt{T}\sum_n c[n]g_t(t - nT)$$

(5.1)

where the sequence $c[n] \in \{-1,+1\}$ represents a sequence of coded symbols, $P$ is an optical gain factor and $P_{dc} = \lambda P$ is an added DC bias which ensures that $X(t)$
is everywhere positive. When square shaped pulses are used, the required DC bias is $P_{dc} = P$, i.e., $\lambda = 1$. In this paper we use a square-root raised cosine pulse shape which has a Fourier transform equal to the square root of the Fourier transform of a raised cosine pulse, $G_t(f) = \sqrt{G_{rc}(f)}$. The required DC bias, in this case, can be estimated by looking at the transmit signal eye diagram. Fig. 5.1 shows the eye diagram of an unbiased and uncoded signal $X(t)$, normalized to the value of the gain factor $P$, using square-root raised cosine pulses of $\alpha = 0.45$. The minimum signal value in this case is -1.46, and so $P_{dc}$ must be chosen equal to 1.46$P$. Hence, $\lambda = P_{dc}/P = 1.46$. The coefficient $\lambda$ (or equivalently the required DC bias $P_{dc}$) can be minimized by properly choosing $\alpha$, the rolloff factor. Fig. 5.2 shows the dependence of $\lambda$ on $\alpha$ for an uncoded OOK signal. As $\alpha$ increases, $\lambda$ decreases until it reaches its minimum value near $\alpha = 0.45$. We will use this value of $\alpha$ in our later studies throughout this chapter.

![Eye diagram](image)

Figure 5.1. Eye diagram of an unbiased signal $X(t)$, normalized to the value of the gain factor $P$, using square-root raised cosine pulses of $\alpha = 0.45$. 
Figure 5.2. $\lambda$ as a function of roll-off factor, $\alpha$.

The average optical transmit power $P_t$ is obtained by integrating $X(t)$ over a large interval $\tau$, dividing the result by $\tau$ and letting $\tau$ go to infinity.

\[
P_t = \lim_{\tau \to \infty} \frac{1}{\tau} \int_{\tau} X(t) dt
\]

\[
= P_{dc} + P \sqrt{T} \lim_{N \to \infty} \frac{1}{2NT} \int_{-NT}^{NT} \left( \sum_{n=-N}^{N} c[n] g_t(t - nT) \right) dt
\]

\[
= P_{dc} + \frac{P}{\sqrt{T}} \lim_{N \to \infty} \frac{1}{2N} \sum_{n=-N}^{N} c[n] \int_{-NT}^{NT} g_t(t - nT) dt
\]

Note, for large values of $N$ we have

\[
\int_{-NT}^{NT} g_t(t - nT) dt \approx \int_{-\infty}^{\infty} g_t(t) dt = G_t(0) = \sqrt{G_{rc}(0)} = \sqrt{T}
\]

Hence,

\[
P_t = P_{dc} + P \lim_{N \to \infty} \frac{1}{2N} \sum_{n=-N}^{N} c[n] = P_{dc} + P \cdot E\{c[n]\} = P_{dc} = \lambda P
\]
where it is assumed that code symbols $-1$ and $+1$ occur with the same frequency. This is a valid assumption for the case of convolutional coding.$^2$

Returning to the block diagram of Fig. 5.3, there are $J$ independent diversity channels, characterized by impulse responses $h_j(t), j = 1, \ldots, J$ between transmitter and receiver. These channels are formed as a result of using a $J$-branch receiver. Each channel is assumed to be ideal, with an impulse response $h_j(t) = H_j(0)\delta(t), j = 1, \ldots, J$ where $H_j(0)$ represents path loss. This will be a valid assumption in our case study for bit rates as high as few hundred Megabits per second, as discussed in section 5.2. At the $j$th receiving branch, ambient-induced shot noise $n_j(t)$ with

---

$^2$Though code symbols $-1$ and $+1$ are no longer independent, they occur with the same probability, which is 0.5. This was observed and justified for each code used in this paper by creating a very long code sequence and counting for $+1$ symbols and $-1$ symbols in the sequence.
a two-sided power spectral density $S_{j}^{\text{shot}}(f) = N_{0,j}$ is added to the detected signal. The total signal plus noise is then filtered using a bank of $J$ identical receiving filters $G_r(f)$ and sampled. Time recovery at this stage ensures that the $J$ transmitted code symbols arrive at the same time.

We let each receiving filter also be a square-root raised cosine filter to eliminate ISI, and optimize the received signal-to-noise ratio (SNR). More specifically, we let $G_r(f) = \sqrt{G_{rc}(f)}$ which has a unity energy. Let $y_j(t)$ be the output of the receiving filter at the $j$th branch. After removing the DC term, $y_j(t)$ can be written as:

$$y_j(t) = PrH_j(0)\sqrt{T}\sum_n c[n]g(t-nT) + \nu_j(t)$$  \hspace{1cm} (5.4)

where $g(t) = g_{t}(t) \otimes g_{r}(t) = g_{rc}(t)$ and $\nu_j(t) = n_j(t) \otimes g_{r}(t)$ represents the filtered noise process at the $j$th branch. The equivalent discrete-time sequence for the $j$th branch at the sampling time $kT$ is given by:

$$y_j[k] = \alpha_j c[k] + \nu_j[k]$$ \hspace{1cm} (5.5)

where $\alpha_j = PrH_j(0)\sqrt{T}$. The noise sequence $\nu_j[k]$ is a discrete-time Gaussian random process with an autocorrelation:

$$E\{\nu_j[k]|\nu_j[k']\} = N_{0,j} \int_{-\infty}^{\infty} g_r(t + [k - k']T)g_r(t)dt$$

$$= N_{0,j}g_{rc}([k - k']T) = N_{0,j}\delta_{kk'}$$ \hspace{1cm} (5.6)

Hence, the noise samples $\nu_j[k], k = 1, 2, \ldots$ are uncorrelated (and thus independent), each with a variance $\sigma_j^2 = N_{0,j}$.

The diversity system of Fig. 5.3 is designed based upon the concept of maximum-likelihood code combining [6]. We now show how the corresponding decoder is implemented. Let $Y_j$ be a received block at the $j$th input of the decoder, i.e., $Y_j = (y_j[1], y_j[2], \ldots, y_j[N])$, where $N$ is the block length. With $J$ received coded blocks of length $N$, a soft decision ML decoder decides in favor of $\hat{c} = (\hat{c}[1], \hat{c}[2], \ldots, \hat{c}[N])$
among all possible transmitted coded blocks \( c \) if and only if
\[ P(Y_1, Y_2, \ldots, Y_J|\hat{c}) > P(Y_1, Y_2, \ldots, Y_J|c), \]
i.e.,
\[
\prod_{j=1}^{J} \prod_{k=1}^{N} \frac{1}{\sqrt{2\pi\sigma_j^2}} \exp \left\{ \frac{-(y_j[k] - \alpha_j\hat{c}[k])^2}{2\sigma_j^2} \right\} \geq \prod_{j=1}^{J} \prod_{k=1}^{N} \frac{1}{\sqrt{2\pi\sigma_j^2}} \exp \left\{ \frac{-(y_j[k] - \alpha_jc[k])^2}{2\sigma_j^2} \right\} \quad (5.7)
\]
Taking natural logs from both sides and dropping common terms we obtain:
\[
\sum_{j=1}^{J} \sum_{k=1}^{N} \frac{\alpha_j}{\sigma_j^2} y_j[k]\hat{c}[k] > \sum_{j=1}^{J} \sum_{k=1}^{N} \frac{\alpha_j}{\sigma_j^2} y_j[k]c[k] \quad (5.8)
\]
Therefore, the decision metric \( M^{(J)} \) can be defined as:
\[
M^{(J)} = \sum_{j=1}^{J} \sum_{k=1}^{N} \frac{\alpha_j}{\sigma_j^2} c[k]y_j[k] = \sum_{k=1}^{N} c[k]y[k] \quad (5.9)
\]
where
\[
y[k] = \sum_{j=1}^{J} (\alpha_j/\sigma_j^2) y_j[k] \quad (5.10)
\]
Thus, maximum-likelihood decoding is indeed easy to implement. At the \( j \)th branch, each received symbol \( y_j[k] \) is weighted by its reliability factor \( W_j = \alpha_j/\sigma_j^2 \).
A new received symbol \( y[k] \) at time \( kT \) is then formed by adding up all \( J \) weighted symbols at that time. A new block \( Y \) of length \( N \) consisting of these new symbols is then formed. This block can be decoded using a conventional soft-decision Viterbi decoder, in which inner product provides the proper metric.

Using the metric of Eq. 5.9, the pairwise probability of error can be easily obtained:
\[
P_2(d) = Q \left( \sqrt{d \cdot R \cdot SNR} \right) \quad (5.11)
\]
where \( R \) is the code rate,
\[
SNR = \sum_{j=1}^{J} SNR_j \quad (5.12)
\]
and
\[
\text{SNR}_j = \frac{\alpha_j^2}{\sigma_j^2} = \frac{P^2 r^2 H_j^2(0)}{N_0 R_b} \tag{5.13}
\]

Here, \(d\) is the Hamming distance between the error path and the correct path. The upper bound on the bit-error probability, \(P_b\) of a rate \(p/(np - \delta)\) punctured convolutional code is given by [5]:
\[
P_b \leq \frac{1}{P} \sum_{d=d_{\text{free}}}^{\infty} c_d P_2(d) \tag{5.14}
\]

where \(d_{\text{free}}\) denotes the minimum free Hamming distance of the code and \(c_d\) is the total number of information bit errors produced by the incorrect paths of Hamming weight \(d > d_{\text{free}}\) that diverge from correct path and remerge with it at some later stage.

### 5.4 Performance

Performance evaluation of the proposed coding scheme for the link configuration of section 5.2 requires knowledge of signal-to-noise ratio, defined in Eqs. 5.12 and 5.13, at all possible positions in the room. Note, the amount of received signal and noise power depends strongly on the position of the receiver with respect to the diffusing spot grid and the ambient light sources. Even simple rotation of the receiver about the normal may cause a change of more than 3dB in the signal-to-noise ratio. The only way to properly describe such a model is to use a statistical approach. A total of 4000 random receiver positions and orientations were selected and the probability distribution of SNR\(_n\), i.e., signal-to-noise ratio, normalized to \(P^2/R_b\), is obtained. The shape of this distribution began to converge after 1000 sample points. Fig. 5.4 shows that the normalized signal-to-noise ratio, SNR\(_n\), spans a dynamic range of 14 dB, with a more than 99% probability.

We use two powerful rate-compatible punctured convolutional codes of memory lengths \(M = 4\) and \(M = 6\). Table 5.1 contains the key parameters associated
Figure 5.4. Distribution of normalized signal-to-noise ratio SNR$_n$ obtained using 4000 random receiver positions.

with these two codes. Distance properties of these codes are documented in [5]. Code rates of 8/9 and 8/16, corresponding to $\delta = 7$ and $\delta = 0$, respectively, were used along with the uncoded scheme. Using the distribution given in Fig. 5.4, the optical power $P_t$ required to achieve a given bit-error rate with 99% probability (1% outage) was evaluated as a function of information bit rate $R_b$. Figs. 5.5(a) and (b) show the results using $M = 4$ code for BERs of $10^{-9}$ and $10^{-6}$, respectively. Similar results using $M = 6$ code is shown in Fig 5.6. As expected, the code with $M = 6$ has a slightly better performance than the code with $M = 4$.

The overall system throughput can be increased by having it operate in an adaptive-rate mode. Using RCPC codes, this can be done at minimal cost as only one convolutional encoder at the transmitter and one Viterbi decoder at the receiver are required. The extra cost accounts for implementation of a simple mechanism for puncturing at the encoder and insertion of dummy variables at the decoder. The adaptive-rate system used in this paper takes advantage of the uncoded OOK
Figure 5.5. Plot of average power $P_t$ required to achieve a BER of (a) $10^{-9}$ and (b) $10^{-6}$ with 1% outage as a function of bit rate $R_b$ using the proposed RCPC code of memory $M = 4$.

Figure 5.6. Plot of average power $P_t$ required to achieve a BER of (a) $10^{-9}$ and (b) $10^{-6}$ with 1% outage as a function of bit rate $R_b$ using the proposed RCPC code of memory $M = 6$. 
scheme, a rate 8/9 code derived from a rate 8/16 mother code as well as the mother code itself. To evaluate the performance of the adaptive-rate scheme and compare it with that of the fixed-rate scheme, we assume a fixed bandwidth, BW, is being used and assume that the system is initially set to operate in fixed-rate mode (only the 1/2-rate mother code is used). We calculate the bit-rate according to the relation \( R_b = \frac{2R}{1+\alpha} \cdot \text{BW} \) with \( R = 1/2 \) and \( \alpha = 0.45 \). The optical power \( P_t \) can now be calculated with the aid of Fig. 5.5 or Fig. 5.6, depending on the code used. Using this value of \( P_t \), we now let the system operate in an adaptive-rate mode. In this mode, unlike the fixed-rate mode, a 1/2-rate code is used only when the required BER with 1% outage cannot be achieved by any other code in the family. When the estimated SNR is high, the system switches to a higher-rate code. Using the distribution of Fig. 5.4, the achievable bit rate can be evaluated. Performance results for the adaptive-rate scheme using the \( M = 4 \) and \( M = 6 \) code families are included in Fig. 5.5 and Fig. 5.6, respectively. It is seen that the overall throughput can be increased by nearly 50%.

<table>
<thead>
<tr>
<th>mother code</th>
<th>memory length</th>
<th>period</th>
</tr>
</thead>
<tbody>
<tr>
<td>(23, 35)</td>
<td>4</td>
<td>8</td>
</tr>
<tr>
<td>(133, 171)</td>
<td>6</td>
<td>8</td>
</tr>
</tbody>
</table>

5.5 Channel Estimation

In order to implement the proposed adaptive-rate scheme successfully, an accurate estimate of the channel needs to be provided at the transmitter as well as at the receiver. Based on this estimate, the transmitter will decide on the number of
symbols that need be punctured at the output of the encoder and the receiver will know where to insert dummy variables prior to decoding.

An estimate for the value of SNR can be obtained using decoder side information. More specifically, we define:

$$\hat{\text{SNR}} = \frac{M^{(J)}}{N}$$

(5.15)

where $M^{(J)}$ is the decision metric after receiving $N$ symbols and is given by Eq. 5.9. This estimate has a Gaussian distribution, with a mean of SNR and a variance of $\text{SNR}/N$, provided that all $N$ received symbols are decoded correctly. The probability $P_f$ that $\hat{\text{SNR}}$ falls within $\psi$ percent of its actual value is given by:

$$P_f = 1 - 2Q(\sqrt{\text{SNR}} \psi/100)$$

(5.16)

This estimate is very good, given that $N$ and SNR are sufficiently large. Note again that, this result is accurate if the decoded symbols are error-free. Nevertheless, even if a decoding error does occur, only a relatively small number of trellis branches are likely to be affected. Therefore, the metric value will not change significantly and the estimate will remain reliable.

Based on Fig. 5.4 and Eqs. 5.11 through 5.14 it can be shown that a SNR of more than 10dB is required to achieve a BER of $10^{-6}$, or less with 1% outage. This holds true even when the most powerful code in the family, i.e., 1/2-rate code is used. In Fig. 5.7, we have plotted $N$ as a function of SNR for $\psi$ equal to 1, 2, 5 and 10. The probability $P_f$ in all these plots is taken to be 0.95. A SNR of 10 dB results in a block length $N$ of 154 symbols, assuming $\psi = 5$. This means, using a block length of 154 symbols, the estimate $\hat{\text{SNR}}$ defined by Eq. 5.15, falls in the range between 9.5 and 10.5 (within 5% of its actual value) with a high probability ($P_f = 0.95$).

In order to get the best result, prior to transmission, the system must be set up to use the most powerful code available, i.e., the unpunctured mother code. This
Figure 5.7. A coded block of length $N$ is required to estimate the SNR with a high accuracy determined by $\psi$.

has a twofold advantage. First, since the mother code is the most powerful code in the family, the probability of decoding error is minimized. Second, since the mother code has the largest length $N = n(K + m)$ in the family, the probability $P_f$, given by Eq. 5.16 is maximized. Based on the estimated signal-to-noise ratio, the transmitter and receiver select a code in the family with smallest amount of redundancy that can keep the BER below the desired value.

### 5.6 Summary

The key feature offered by multiple-spot diffusing configuration is that it creates a large number of distinct signal paths between the transmitter and the receiver. One good way of exploiting this feature is to use a direction diversity receiver consisting of a number $J$ of narrow FOV receiving elements to cover a good portion of the diffusing surface. The communication between the two ends can be made robust against blockage by increasing the overall FOV which is achieved by increasing $J$. 
This creates $J$ nearly ideal channels between transmitter and receiver, $J$ being the diversity order.

Code combining at the receiver, using RCPC codes, allows the system to be robust under severe received signal degradation due to ambient light and blockage. It also provides an adaptive environment for efficient utilization of available bandwidth. It allows portable terminals to communicate at their highest permitted bit rate while maintaining a BER below a certain level. To make the communication even more reliable, packets can be retransmitted if they are detected in error. In this case, a block of coded information is transmitted each time through $J$ space diversity channels and is repeated only when needed. With code combining, the receiver forms an increasingly reliable estimate of the transmitted block, forming a high-throughput end-to-end system.
In Chapter 5 we used binary convolutional codes to improve the average power requirement of intensity modulated optical signals. Binary convolutional codes are frequently used in applications that involve transmission of digital data over wired and/or wireless channels. These codes are popular not just due to their relatively high coding gain, but because of the maximum likelihood sequence decoding (Viterbi decoding) algorithm that can readily be implemented for short memory codes. Binary convolutional codes are well-suited for modulation schemes which are also binary, and so, they are widely used in radio systems which use BPSK or QPSK\(^1\). The design goal here is to reduce the transmit power, or equivalently, \(E_b/N_0\).

In wireless optical system, on the other hand, it is desired to reduce the average optical power, i.e., the average optical signal amplitude. This suggests that a modulation scheme with a low duty cycle, or equivalently, high peak-to-average ratio be used. An example of such a scheme is \(L\)-ary pulse-position modulation, discussed in section 4.1.4.

Analytical results show that binary convolutional codes are not suitable coding schemes for \(L\)-PPM [40]. The problem arises from lack of proper matching between

\(^1\)Although QPSK is not a binary modulation scheme, it can be decomposed into two orthogonal BPSK schemes and treated accordingly.
binary coding and non-binary modulation. Specifically, since the signal set in $L$-PPM consists of orthogonal waveforms, a symbol error results in $1$ to $\log_2 L$ bit errors all with the same probability. As a result, bit errors tend to be bunched, and the capacity of binary codes can be easily exceeded. Furthermore, while the Viterbi algorithm can be implemented either for hard-decision or soft-decision decoding according to the metric selected, it is not straight forward to combine binary convolutional codes and $L$-PPM and still use soft-decision decoding. All these problems can be avoided if one resorts to non-binary coding schemes.

Among various non-binary coding schemes, Reed-Solomon codes are one of the most effective, owing to their optimality among all other block coding schemes. In particular, Reed Solomon codes are maximum-distance separable, which means they achieve maximum possible minimum distance (can correct largest possible number of errors). We will use these codes in conjunction with $L$-PPM in the next chapter. Before doing so, however, it is advantageous to answer the following question. What is an achievable rate at some given SNR using $L$-PPM? This question can be answered by computing the channel capacity of $L$-PPM which is the goal of this chapter.

6.1 Introduction

One key to lowering the power requirements of wireless indoor infrared systems is proper use of forward error correction. Performance of FEC coding techniques can vary widely depending upon the code rate and block or constraint length of the code used. However, the limit of performance was found years ago by Shannon when he derived the channel capacity theorem. Shannon found there is a maximum, called channel capacity, to the transmission rate at which any communication system can operate when constrained in average power and bandwidth. Operating at a rate greater than capacity must always result in a high probability of error. The
The channel capacity theorem states that there exists signaling schemes with optimum receivers such that, for information rates less than channel capacity, arbitrarily small error rates may be achieved. Conversely, for information rates greater than channel capacity, probability of error will always be large for every possible signaling scheme.

The channel capacity derived by Shannon can be used to find the limit on minimum required $E_b/N_0$ in order to achieve error free communication. This limit is called the Shannon limit, which is the value of $E_b/N_0$ required to operate at the capacity limit as band spreading gets very large (code rate $R \to 0$). This limit on $E_b/N_0$ is

$$\left( \frac{E_b}{N_0} \right)_{\text{Shannon Limit}} = \log_e 2 = -1.6 \text{ dB}$$

We note that, the Shannon limit as given by Eq. 6.1 is a result that holds when there is no constraint on the modulation type. In practical systems, practical considerations may suggest use of a particular modulation scheme. This signal constraint establishes a higher limit on the average power.

It is important to note that the Shannon capacity formula and its limiting-case result of Eq. 6.1 are derived subject to average transmit power constraint. This is one of the most common input constraints in wired and/or wireless communication systems. In this chapter, we will focus on evaluating the channel capacity under a different constraint. Here, the input signal amplitude needs to be positive and its average must be kept minimal. This is the situation in all indoor wireless infrared channels, in which the signal amplitude represents optical power.

Channel capacity provides an upper bound for what one can achieve with coding prior to modulation. In this chapter, we compute this bound for $L$-PPM modulation. In the next section, we present expressions for achievable information rate on a memoryless channel when the input codewords are independent and identi-
cally distributed with a uniform distribution. We evaluate capacity for $L$-PPM, using both hard-decision decoding (HDD) and soft-decision decoding (SDD) at the receiver and present the numerical results.

6.2 SDD vs. HDD

We now consider transmission of information over an infrared channel using $L$-PPM modulation. A complete end-to-end system is shown in Fig. 6.1-a. The input to the system is a binary information sequence $a[k]$. This sequence enters a channel encoder, which produces a binary coded sequence $b[k]$ at its output. Every block of $\log_2 L$ bits in $b[k]$ is then mapped into a $L$-PPM code vector $c$. This $L$ bit code vector is then sent through the channel of Fig. 6.1-b, in a serial manner. We denote by $c[k]$ a coded binary sequence at the input of pulse shaping filter. The sequence $c[k]$ derives a transmitter filter $g_t(t)$ with a rectangular pulse shape and a unity height. For $L$-PPM, each pulse represents a chip with a duration $T_c = T/L$, $T$ being a symbol duration. To set the average transmit power equal to $P$, prior to transmission, filter output must be multiplied by $LP$. The channel is assumed to be ideal with an impulse response $h(t) = rH(0)\delta(t)$ and corrupted by background white Gaussian shot noise $\eta(t)$. The receiver uses a unit-energy matched filter $g_r(t)$ and samples the output at a chip rate $1/T_c$. The channel of Fig. 6.1-b forms a discrete-input, continuous output (DICO) scalar channel. It is equivalent to the discrete-time model of Fig. 6.1-c, with $\alpha = rH(0)P\sqrt{T}$. Here, $\{n[k]\}$ represents a set of independent zero-mean Gaussian random variables with a variance $N_0$ equal to that of $\eta(t)$. Returning to Fig. 6.1-a, at the receiver is formed, a length $L$ vector $y$ corresponding to $c$. This vector is then used to estimate $a[k]$ using either SDD or HDD. For SDD, it is proper to define an equivalent discrete-input, continuous-output vector channel which takes an $L$-dimensional binary vector $c$ as input and produces an $L$-dimensional continuous output $y$. 
\[ y = \alpha c + n \] (6.2)

where \( n \) represents noise.

![Diagram of the channel model and system block diagram](attachment:figure6.png)

Figure 6.1. (a) System block diagram for transmission of information over an IR link using \( L \)-PPM modulation. (b) Full representation of the scalar discrete-input continuous-output channel of Fig. (a) with input \( c[k] \) and output \( y[k] \). (c) Discrete model for the channel of Fig. (b)

When performing HDD, a detector estimates the transmitted binary sequence \( c[k] \). If we view the decision process at the detector as a form of quantization, we observe that a hard decision in this case corresponds to binary quantization of output \( y[k] \). Once again, it is proper to define an equivalent discrete-input discrete-output vector channel which takes an \( L \)-dimensional binary vector \( c \) as the input and produces an \( L \)-dimensional discrete output \( \hat{c} \) as an estimate of \( c \). This
defines a discrete memoryless channel as shown in Fig. 6.2 with an input alphabet $c = \{c^{(0)}, c^{(1)}, \ldots, c^{(L-1)}\}$ and output alphabet $\hat{c} = \{\hat{c}^{(0)}, \hat{c}^{(1)}, \ldots, \hat{c}^{(L-1)}\}$ and a transition probability

$$P(\hat{c}^{(i)}|c^{(j)}) = \begin{cases} 1 - p, & i = j \\ p/(L-1), & i \neq j \end{cases}$$  \hspace{1cm} (6.3)$$

where $p$ is the probability of symbol error given by

$$p = 1 - \frac{1}{\sqrt{2\pi}} \int_{-\infty}^{\infty} [1 - Q(x)]^{L-1} \exp\left\{-\frac{(x - \sqrt{L \text{SNR}})^2}{2}\right\} dx$$  \hspace{1cm} (6.4)$$

Figure 6.2. Equivalent discrete memoryless channel for $L$-PPM signaling with HDD.

In the next section, we evaluate the channel capacity of $L$-PPM for different values of $L$. We consider both HDD as well as SDD. To perform comparison we define the signal-to-noise ratio SNR, which is directly related to the average incoming optical power, to be

$$\text{SNR} = \frac{r^2 H^2(0) P^2 T}{N_0}$$  \hspace{1cm} (6.5)$$

6.3 $L$-PPM Channel Capacity

We start with finding the channel capacity using HDD at the receiver. The DMC channel in this configuration can be represented as shown by Fig. 6.2. The capacity
of a DMC channel having an input alphabet $c = \{c(0), c(1), \ldots, c^{(q-1)}\}$ and output alphabet $\hat{c} = \{\hat{c}(0), \hat{c}(1), \ldots, \hat{c}^{(Q-1)}\}$ is defined as

$$C = \max_{P(c^{(j)})} I(c; \hat{c})$$

(6.6)

where $I(c; \hat{c})$ is the average mutual information provided by output $\hat{c}$ about the input $c$:

$$I(c; \hat{c}) = \sum_{j=0}^{q-1} P(c^{(j)}) I(c^{(j)}; \hat{c})$$

(6.7)

and

$$I(c^{(j)}; \hat{c}) = \sum_{i=0}^{L-1} P(\hat{c}^{(i)}|c^{(j)}) \log_2 \frac{P(\hat{c}^{(i)}|c^{(j)})}{\sum_{k=0}^{L-1} P(\hat{c}^{(k)}|c^{(j)})}$$

(6.8)

which requires knowledge of the channel transition probabilities $P(\hat{c}^{(i)}|c^{(j)})$. Though, nothing can be said in general about the input probability assignment that maximizes the average mutual information, in many cases, channel transition probabilities exhibit a form of symmetry that results in the maximum of $I(c; \hat{c})$ being obtained when the input symbols are equally probable. In general, necessary and sufficient conditions for the set of input probabilities $P(\hat{c}^{(i)}|c^{(j)})$ to maximize $I(c; \hat{c})$, and thus to achieve capacity on a DMC are that

$$I(c^{(j)}; \hat{c}) = C \text{ for all } j \text{ with } P(c^{(j)}) > 0$$

(6.9)

$$I(c^{(j)}; \hat{c}) \leq C \text{ for all } j \text{ with } P(c^{(j)}) = 0$$

(6.10)

where $C$ is the capacity of the channel.

Following this argument, we note that for all $j$ with $P(c^{(j)}) = 1/L$ we have

$$I(c^{(j)}; \hat{c}) = \log_2 L + p \log_2 \frac{p}{L-1} + (1-p) \log_2 (1-p)$$

(6.11)

Hence, the right-hand side of Eq. 6.11 is indeed the capacity of the channel.

Note that when $L = 2$, Fig. 6.2 becomes a binary-symmetric channel (BSC), and Eq. 6.11 reduces to the well-known formula for capacity of a BSC.
Next, let us consider the discrete-time AWGN channel of Eq. 6.2. The receiver, in this case performs SDD on the received vector $\mathbf{y}$ to obtain an estimate $\hat{a}[k]$ of the information sequence $a[k]$. This channel can be characterized by the discrete input $\mathbf{c}$, the continuous output $\mathbf{y}$ and the set of conditional transition probability density functions

$$P(\mathbf{y}|\mathbf{c}^{(j)}) = \left( \frac{1}{2\pi N_0} \right)^{L/2} \exp \left\{ -\frac{\|\mathbf{y} - \alpha \mathbf{c}^{(j)}\|^2}{2N_0} \right\}$$ (6.12)

The mutual information between the discrete input $\mathbf{c}^{(j)}$ and the output $\mathbf{y}$ in bits/channel use is

$$I(\mathbf{c}^{(j)}; \mathbf{y}) = \int_{-\infty}^{\infty} P(\mathbf{y}|\mathbf{c}^{(j)}) \log_2 \frac{P(\mathbf{y}|\mathbf{c}^{(j)})}{\sum_{k=0}^{L-1} P(\mathbf{c}^{(k)}) P(\mathbf{y}|\mathbf{c}^{(k)})} d\mathbf{y}$$ (6.13)

It is easy to show that for all $j$ with $P(\mathbf{c}^{(j)}) = 1/L$, $I(\mathbf{c}^{(j)}; \mathbf{y}) = C$, where $C$, is indeed the channel capacity given by

$$C = \log_2 L - E \left\{ \log_2 \left[ 1 + \exp\left\{ -\frac{\alpha^2}{N_0} \right\} \exp\left\{ -\frac{\alpha}{\sqrt{N_0}} x_L \right\} \sum_{k=1}^{L-1} \exp\left\{ \frac{\alpha}{\sqrt{N_0}} x_k \right\} \right] \right\}$$ (6.14)

Here, $x_i, i = 1, 2, \ldots, L$ are independent, identically distributed zero-mean Gaussian random variables with unity variance. The above equation contains an $L$-dimensional integral and has no simple closed-form solution. As a consequence, the Monte Carlo method was used to estimate $C$.

Fig. 6.3 shows the channel capacity for $L$-PPM, $L = 2, 4, 8$, employing both HDD and SDD at the receiver. Note, the 2-dB difference between HDD and SDD seen in this case, is in fact a fundamental result and a characteristic that applies in general to coded digital modulation over the AWGN channel.

Fig. 6.4 shows the channel capacity described in terms of bit/sec/Hz. Here we have taken the bandwidth to be the inverse of the duration of the shortest pulse. The Plots suggest that by slightly reducing the information rate through coding
L-PPM using a high-rate code, a gain of more than 6 dB can be attained. A good candidate here is a high-rate Reed Solomon code. This will be discussed in the next chapter.

Figure 6.3. Capacity of 2-PPM, 4-PPM and 8-PPM as a function of SNR using both SDD and HDD.
Figure 6.4. Capacity of $L$-PPM in bit/sec/Hz as a function of SNR using both SDD and HDD.
Chapter 7

REED-SOLOMON CODED PULSE-POSITION MODULATION FOR HIGH-SPEED POWER-EFFICIENT WIRELESS INFRARED LANS

While use of power-efficient signaling schemes appears to be effective in compensating for the inherent high path-loss associated with pure diffuse infrared links, it begins to lose its effectiveness as data rate is increased. At very high data rates, intersymbol interference (ISI) can result in a very high and sometimes irreducible power penalty, preventing the system to operate at a low bit-error probability. In this chapter, we investigate a link design employing a multi-beam transmitter in conjunction with a narrow field-of-view (FOV) direction diversity receiver. The design goal is to eliminate the effect of ISI so that power-efficient signaling schemes such as pulse-position modulation (PPM) can be employed at very high data rates. We also use high-rate Reed-Solomon codes to further increase the power efficiency of PPM signals. The proposed system can be made rate-adaptive through varying modulation level $L$ and/or code rate $R$ without increasing the complexity, significantly. This provides a dynamic range large enough to allow efficient utilization of available bandwidth, i.e., to allow portable terminals to communicate at their highest permitted data rate, without sacrificing the quality-of-service. It is shown that a bit-error rate (BER) not exceeding $10^{-9}$ can be achieved within the link coverage area with 99% probability at bit rates up to a few hundreds of megabits per second, using transmitted power levels well below one Watt.
7.1 Introduction

Non-directed infrared (IR) radiation has been considered as a viable alternative to radio for indoor wireless local area networks [19, 28]. Non-directed transmission alleviates the problem of aligning transmitters and/or receivers and makes the system robust against blockage. The highest degree of robustness in non-directed IR channels can be achieved by projecting a wide beam on a reflecting surface, e.g., ceiling, and having a receiver with a wide FOV face toward that surface. This configuration establishes a non-directed non-line-of-sight link, often referred to as diffuse.

Diffuse link configuration, has been under investigation by researchers in the past decade [20, 22, 35, 36, 23, 37, 38, 39]. An immediate disadvantage of a diffuse IR channel is its increased path loss due to lack of a direct link between transmitter and receiver. One option to compensate for path loss is to increase the transmit power which is not very desirable due to power consumption considerations and eye safety regulations. This motivates use of power-efficient signaling schemes such as pulse-position modulation (PPM).

While use of power-efficient signaling schemes appears to be effective in compensating for the path loss, it begins to lose its effectiveness, as data rate is increased. This is due to a second major disadvantage of a diffuse channel, i.e., its limited bandwidth. The band-limited behavior of the channel is due to multipath dispersion caused by reflections of light from walls and room objects. It creates intersymbol interference (ISI) when signals at high data rates are transmitted over the channel. The effect is particularly pronounced when signals with a poor bandwidth efficiency are used. For example, at high data rates, ISI can bring the power requirements of pulse-position modulation (PPM) to a level exceeding that of a simple on-off keying (OOK) [38].
The effect of ISI can be eliminated when a multi-beam transmitter is used in conjunction with a multiple element narrow field-of-view (FOV) direction-diversity (also known as angle-diversity) receiver [41, 29]. The transmitter in this configuration projects the light power in form of multiple narrow beams of equal intensity, over several small areas (spots) on a diffusing surface such as a ceiling. Each diffusing spot, in this arrangement, may be considered a line-of-sight light source, which can be aimed at by a narrow FOV receiving branch. With this arrangement, a good portion of the total optical power can be received via a finite number of distinct signal paths; a number equal to the number of spots seen by the receiving branch. As the branch FOV is decreased, a smaller number of spots can be sighted and in the limiting case where only one spot is seen, channel behaves as nearly ideal.

Narrowing the FOV offers other advantages, as well. It eliminates a great amount of ambient noise power and relaxes the problem with spectral response shift of the interference filters used for rejecting the ambient light. To make the system robust against blockage, the overall FOV of the receiver must be made wide via increasing the number of branches. This, in turn, increases the effective area of the photo-detector, which allows for receiving more power. These benefits are obviously gained when one trades these for some additional complexity in hardware and a bulkier receiver.

In this chapter, first, we investigate the characteristics of an infrared link composed of a multi-beam transmitter and a multiple-branch narrow FOV receiver. In section 7.2, a framework for computer simulation is provided under which properties of room, transmitter and receiver are quantified. It is shown, via simulation, that channel impulse and magnitude responses can be estimated with a high accuracy using only the first order reflection. Based on this observation, we deliberately
decrease the branch FOV at the receiver until a nearly ideal channel with a very high bandwidth is attained. The multiple-branch receiver used in this chapter creates a direction diversity system, composed of $J$ nearly ideal channels, $J$ being the diversity order. We use power-efficient PPM signals in section 7.3 to transmit data over these $J$ diversity channels. We then use high-rate Reed-Solomon (RS) codes, in section 7.4, to further decrease the power requirements. The proposed system can be made rate-adaptive through varying PPM modulation level $L$ and/or RS code rate $R$ without increasing the complexity, significantly. This is also discussed in section 7.4, and followed by our concluding remarks in section 7.5.

7.2 Link Configuration

In this section, we evaluate the characteristics of the channel formed by a multi-beam transmitter and a multi-branch direction-diversity receiver. For our study we consider the room model of section 4.2.1 (see Fig. 4.6). A multi-beam transmitter is assumed to produce 100 beams of equal intensity by means of a CGH [42] to form a $10 \times 10$ square lattice of equidistant spots on the ceiling. The transmitter is located at the center of the room at a desktop level and pointed upward. There is a large window at one side of the room extended from desktop level up to the ceiling. The ceiling, the walls, the floor, and the window are modeled as Lambertian reflectors of the first order. Both FOVs of $11.5^\circ$ and $7^\circ$ were considered for each receiver branch. For more details about the link configuration refer to sections 4.2.1 and 4.2.3.

We obtained the probability distribution of the path loss by selecting 4000 random positions and orientations for the receiver. After using about 1000 random samples, the distribution began to converge. For reasons that will become clear later, we define the path loss to be $-10 \log H(0)$ where $H(0)$ is

$$H^2(0) = \sum_{j=1}^{J} H^2_j(0)$$

(7.1)
Figure 7.1 shows this distribution at a given point within the room. As seen, for a 7° FOV receiver, path loss falls within the range 69dB to 72dB with a probability more than 0.95. For an 11.5° FOV receiver this range goes from 65dB to 68dB, which compared to the other case, is shifted by 4dB.

Figure 7.1. Probability distribution of the channel path loss between the transmitter and a 7-branch composite receiver that is placed at a random position in the room.

In the following sections, we use power-efficient signaling schemes to compensate for this path loss and, based on these schemes, we evaluate the required transmit power that achieves reliable communication at almost all receiver positions within the room.
7.3 \(L\)-PPM with Diversity

An IR link composed of a multi-beam transmitter and a narrow FOV direction-diversity receiver creates \(J\) communication channels between transmitter and receiver, where \(J\) is the diversity order. In section 7.2 we studied two link designs, one using an \(11.5^\circ\)-FOV receiver and the other using a \(7^\circ\)-FOV receiver. The former creates communication channels with 3-dB bandwidth on the order of 100MHz and more, while the latter creates nearly ideal channels having more than 2GHz of effective bandwidth. The increased bandwidth in this case is accompanied with an increase in the path loss. This motivates use of power-efficient modulation schemes such as \(L\)-PPM which yield a high power-efficiency. The poor bandwidth efficiency of PPM will not be an issue in this case since ISI does not exist.

\(L\)-PPM can be viewed as a nonlinear block code; each block of \(K\) input bits is mapped into a code vector, \(c\), of length \(L = 2^K\) bits with unity Hamming distance. This vector can be transmitted through dividing a time interval \(T\) into \(L\) time slots or “chips” of duration \(T_c = T/L\), to which either a constant power \(LP\), corresponding to the transmission of “one”, or a zero power, corresponding to the transmission of “zero”, is associated. Such signaling requires square shape pulses, hard to realize and occupy a very large bandwidth. In this work, we use a square-root raised cosine pulse shape \(g_t(t)\) which is easy to implement. This pulse has a Fourier transform equal to the square-root of the Fourier transform of the raised cosine pulse, i.e., \(G_t(f) = \sqrt{G_{rc}(f)}\). It has a finite bandwidth equal to \((1 + \alpha)/2T_c\), where \(\alpha \in (0, 1]\) is the roll-off factor. The time-domain representation of this pulse is given by [17]

\[
     g_t(t) = \frac{4\alpha}{\pi \sqrt{T_c}} \cos[(1 + \alpha)\pi(t/T_c)] + \frac{\sin[(1 - \alpha)\pi(t/T_c)]/[4\alpha(t/T_c)]}{1 - [4\alpha(t/T_c)]^2} 
\]  

Note that, the waveforms \(g_t(t - nT_c), n = \ldots, -1, 0, 1, \ldots\) form an orthonormal set of basis functions. This can be proven using Parseval’s relation as follows;
Thus, square-root raised cosine pulse shaping does not affect the orthogonality of L-PPM signaling. As a result, any L-PPM code vector \( \mathbf{c} \) is indeed a vector representation of the corresponding signal waveform in an L-dimensional signal space.

The main problem with having square-root raised cosine pulses is that, they are not always positive, a requirement for transmission of optical power. This can be resolved by adding a DC bias (offset). The instantaneous optical power \( X(t) \) can therefore be written as

\[
X(t) = P_{dc} + PL\sqrt{T_c} \sum_{n} c[n]g(t - nT_c)
\]

(7.4)

where \( c[n] \in \{0, 1\} \) represents the bit sequence (every \( L \) consecutive bits makes an L-PPM code vector \( \mathbf{c} \)) at the input of the pulse shaping filter, \( P \) is a gain factor at the transmitter and \( P_{dc} = \lambda P \) is a minimum DC bias that is added to ensure that the optical power \( X(t) \) is always positive. The coefficient \( \lambda \) is generally a function of \( \alpha \) and \( L \). It can be estimated from the transmitting signal eye diagram. Fig. 7.2 shows the eye diagram of an unbiased signal \( X(t) \), normalized to the value of gain factor \( P \), using 4-PPM \( (L = 4) \) with square-root raised cosine pulses of \( \alpha = 0.45 \). The minimum occurs at -0.86, and so \( P_{dc} \) must be chosen equal to 0.86\( P \). Hence, \( \lambda = P_{dc}/P = 0.86 \).

The total average transmit power can be obtained by integrating \( X(t) \) over a time interval \( \tau \), dividing the result by \( \tau \), and letting \( \tau \) go to infinity.

\[
P_t = \lim_{\tau \to \infty} \frac{1}{\tau} \int_{\tau} X(t) dt
\]
Figure 7.2. Eye diagram of an unbiased signal \( X(t) \), using 4-PPM \((L = 4)\) with square-root raised cosine pulses of roll-off factor \( \alpha = 0.45 \), normalized to the value of \( P \). The pattern is periodic with a period of \( 4T_c \). The minimum occurs at -0.86, and so \( \lambda = P_{dc}/P \) must be chosen equal to 0.86.

\[
\begin{align*}
\lambda &= P_{dc} + PL\sqrt{T_c} \times \lim_{\tau \to \infty} \frac{1}{\tau} \sum_{n} c[n] \int_{\tau} g(t - nT_c)dt \\
&= P_{dc} + P = (1 + \lambda)P \\
&= P_{dc} + P = (1 + \lambda)P
\end{align*}
\]

Note, when square pulses are used, no DC bias is required \((\lambda = 0)\) and the average power reduces to \( P \). Fig. 7.3 shows the transmit power \( X(t) \) using 4-PPM when both square-shape pulses and square-root raised cosine pulses of \( \alpha = 0.45 \) are used. Assuming \( P = 10\text{mW} \), the total average power for the above two cases are \( P_t = P = 10\text{mW} \) and \( P_t = (1 + \lambda)P = 18.6\text{mW} \), respectively.

The coefficient \( \lambda \) can be minimized by properly choosing the roll-off factor. Fig. 7.4 shows the dependence of \( \lambda \) on \( \alpha \) for \( L=4, 8 \) and \( 16 \). A local minimum in all three curves occurs near \( \alpha = 0.45 \). We will use this value for our later studies.
Figure 7.3. Transmitted optical power using 4-PPM with square-shape pulses and square-root raised cosine pulses of $\alpha = 0.45$. Here, we have assumed $P = 10 \text{mW}$ and $P_{dc} = \lambda P = 8.6 \text{mW}$. Throughout this paper.

Fig. 7.5-a illustrates a block diagram of the proposed direction diversity system using $L$-PPM signaling. The optical signal $X(t)$ is transmitted over $J$ diversity channels and is detected at the receiver according to a maximum-likelihood criteria. The channel between the transmitter and each receiving branch is assumed to be ideal with an impulse response $h_j(t) = H_j(0)\delta(t)$, $j = 1, \ldots, J$ where $H_j(0)$ represents path loss of the $j$th channel. This is a valid assumption in our case study for bit rates as high as tens of megabits per second (case 1), and hundreds of megabits per second (case 2), as discussed in section 7.2. A bank of $J$ identical receiving filters $G_r(f)$ are used on the receiver side, at the output of each photodetector. Since the channel is ideal over the bandwidth of interest, we let each receiving filter also be a square-root raised cosine filter in order to eliminate ISI, and to optimize...
Figure 7.4. Variation of $\lambda$ with the roll-off factor, for $L=4, 8$ and 16. For all values of $L$, a local minimum occurs near $\alpha = 0.45$.

the received signal-to-noise ratio (SNR). More specifically, we let $G_r(f)$ be:

$$G_r(f) = \sqrt{G_{rc}(f)}$$

which has a unity energy. Let $y_j(t)$ be the output of the receiving filter on the $j$th branch. After removing the term associated with the input DC bias, we can write $y_j(t)$ as:

$$y_j(t) = rPH_j(0)L\sqrt{T_c} \sum_n c[n]g(t - nT_c) + \nu_j(t)$$

where $g(t) = g_i(t) \otimes g_r(t) = g_{rc}(t)$ and $\nu_j(t) = n_j(t) \otimes g_r(t)$ represents the filtered noise process on the $j$th branch. The equivalent discrete-time output sequence on the $j$th branch and at the sampling time $kT_c$ is given by:

$$y_j[k] = \frac{PrH_j(0)\sqrt{LT}}{1 + \lambda}c[k] + \nu_j[k]$$
\[ \alpha_j c[k] + \nu_j[k] \] (7.8)

where, \( \alpha_j = rH_j(0)P_t\sqrt{LT/(1 + \lambda)} \). The noise sequence \( \nu_j[k] \) is a discrete-time Gaussian random process with an autocorrelation

\[
E \{ \nu_j[k]\nu_j[k'] \} = N_{0j} \int_{-\infty}^{\infty} g_r(t + [k - k']T_c)g_r(t)dt
\]

\[
= N_{0j}g_{rc}([k - k']T_c)
\]

\[
= N_{0j}\delta_{kk'}
\] (7.9)

Hence, the Gaussian noise samples \( \nu_j[k], k = 1, 2, \ldots \) are uncorrelated (and thus independent), each with a variance \( \sigma_j^2 = N_{0j} \).

An optimum \( L \)-PPM maximum-likelihood (ML) detector on each branch \( j \) must wait \( T \) seconds until it receives all the \( L \) sampled outputs associated with the transmission of one \( L \)-PPM code vector (symbol). Note, the ML detector can be viewed as a soft decision decoder for a received \( L \)-bit PPM “code word”. When ML decoding is based on receiving multiple copies of a single code word, from separate channels, the term code combining is often used [6]. For a memoryless channel (such as the one considered here) it is sufficient to consider only one symbol interval, e.g., \( (0, T] \). Let’s denote the received signal vector in this interval, and at the \( j \)th branch, by \( Y_j \), i.e., \( Y_j = (y_j[1], \ldots, y_j[L]) \), where \( y_j[k] \) is given by Eq. 7.8. The elements of this vector are the sufficient statistics for reaching a decision on which of the \( L \) signals were transmitted. When only one channel is used (say, channel \( j \)), the decision rule based on the ML criterion is to find the code vector \( c = (c[1], c[2], \ldots, c[L]) \) that has the largest cross-correlation with the received signal vector \( Y_j \). This is equivalent to choosing a code vector whose nonzero element is at the same position as the largest elements in \( Y_j \). When, on the other hand, \( J > 1 \) diversity channels are used, the receiver must look at all \( J \) received signal vectors \( Y_1, Y_2, \ldots, Y_J \) to make a decision. A maximum-likelihood detector, in this case,
Figure 7.5. (a) System block diagram using pulse-position modulation in an IR link comprised of a multi-beam transmitter and a multiple branch composite receiver. (b) RS coded communication system model.
decides in favor of code vector $\hat{c}$ among all $L$ possible PPM code vectors (denote them in general by $c$), if and only if
\[
Pr(Y_1, Y_2, \ldots, Y_J | \hat{c}) \geq Pr(Y_1, Y_2, \ldots, Y_J | c),
\]
i.e.,
\[
\prod_{j=1}^{J} \prod_{k=1}^{L} \frac{1}{\sqrt{2\pi\sigma_j^2}} \exp \left\{ \frac{-(y_j[k] - \alpha_j \hat{c}[k])^2}{2\sigma_j^2} \right\} \geq \prod_{j=1}^{J} \prod_{k=1}^{L} \frac{1}{\sqrt{2\pi\sigma_j^2}} \exp \left\{ \frac{-(y_j[k] - \alpha_j c[k])^2}{2\sigma_j^2} \right\}.
\]
Taking natural logs from both sides and dropping common terms we obtain:
\[
\sum_{j=1}^{J} \sum_{k=1}^{L} \left[ \frac{\alpha_j^2}{\sigma_j^2} y_j[k] \hat{c}[k] - \frac{\alpha_j^2 c^2[k]}{2\sigma_j^2} \right] \geq \sum_{j=1}^{J} \sum_{k=1}^{L} \left[ \frac{\alpha_j^2}{\sigma_j^2} y_j[k] c[k] - \frac{\alpha_j^2 c^2[k]}{2\sigma_j^2} \right] (7.11)
\]
Note that, $\sum_{k=1}^{L} c^2[k] = 1$ for all $L$ possible code vectors. Hence, the above inequality can be further simplified, as the sum over $k$ of the second term on both sides are equal. The decision metric $M^{(J)}$ can thus be defined as:
\[
M^{(J)} = \sum_{j=1}^{J} \sum_{k=1}^{L} \frac{\alpha_j}{\sigma_j^2} y_j[k] c[k] = \sum_{k=1}^{L} c[k] \left( \sum_{j=1}^{J} \frac{\alpha_j}{\sigma_j^2} y_j[k] \right) = \sum_{k=1}^{L} c[k] y[k] (7.12)
\]
where $y[k] = \sum_{j=1}^{J} (\alpha_j/\sigma_j^2) y_j[k]$. This metric suggests that the same decision rule may be applied when $J > 1$ independent channels are used. Each received vector $Y_j$ is first weighted by its reliability factor $W_j = \alpha_j/\sigma_j^2$. A new vector $Y$ is then formed by adding up all $J$ weighted received vectors. The elements of this vector are the sufficient statistics for reaching a decision. The decision rule, once again, is to find the code vector $c$ that has the largest cross-correlation with the received vector $Y$. This is equivalent to choosing the code vector $c$ whose nonzero element is at the same position as the largest elements in $Y$. 

Based on the metric of Eq. 7.12, it is straightforward to evaluate $P_K$, the probability of symbol error (see [21], for example). The result is

$$P_K = 1 - E \left\{ [1 - Q(x)]^{L-1} \right\}$$  \hspace{1cm} (7.13)

where the expectation is taken over the random variable $x$, which is Gaussian with a variance of $\sigma_x^2 = 1$ and a mean $\mu_x$ of

$$\mu_x = \sum_{j=1}^{J} \frac{\alpha_j^2}{\sigma_j^2}$$

$$= R \cdot L \log_2 L \sum_{j=1}^{J} \frac{P_t^2 r^2 H^2_j(0)}{(1 + \lambda)^2 N_0j R_b}$$

$$= R \cdot L \log_2 L \cdot \text{SNR}$$  \hspace{1cm} (7.14)

where it is assumed that the information bit stream is first encoded into a code of rate $R$ prior to entering the modulator and where SNR is the total signal-to-noise ratio defined by

$$\text{SNR} = \sum_{j=1}^{J} \frac{P_t^2 r^2 H^2_j(0)}{(1 + \lambda)^2 N_0j R_b}$$  \hspace{1cm} (7.15)

As the last remark, note that, at high values of $\mu_x$, the $Q(\cdot)$ term in Eq. 7.13 becomes very small so that $[1 - Q(\cdot)]^{L-1}$ can be approximated by $1 - (L - 1)Q(\cdot)$. After evaluating the expectation in Eq. 7.13 we reach the following simple relation for the symbol error probability

$$P_K = (L - 1)Q \left( \sqrt{\frac{R \cdot L \cdot \log_2 L}{2} \cdot \text{SNR}} \right), \text{ \ \ SNR \gg 1}$$  \hspace{1cm} (7.16)

### 7.4 Punctured Reed-Solomon Codes for $L$-PPM Signaling

$L$-PPM is a multi-level signaling scheme in which every $K$ bits defines one of the $L = 2^K$ symbols. Since the signal set in PPM consists of orthogonal waveforms, a symbol error results in 1 to $K$ bit errors, all with the same probability. This may create error bursts that are not within the error correcting capability of binary
codes. Interleaving can be used to overcome this problem at the expense of extra complexity. Other alternative is to use non-binary codes such as Reed-Solomon (RS) codes, which is used in this paper.

Reed Solomon codes have been widely used for high-speed optical communications systems that use PPM. Recent advances in VLSI technologies and development of efficient decoding algorithms have made it possible to implement RS encoders/decoders on a single chip. Today, synthesizable RS “cores for ASIC implementation” that can process data in excess of 400 Mb/s are being fabricated. These cores can be configured and optimized for a given application. Code parameters such as code word length, number of symbols to be corrected, generating polynomial, erasure support, etc. can be selected by a customer, based on application requirements.

An \((n, k)\) \(Q\)-ary Reed-Solomon code is a symbol error correcting code with a symbol size \(q = \log_2 Q\) bits, and a block length \(n = Q - 1\) that represents \(k\) information symbols. This code has a minimum distance \(d_{\text{min}} = n - k + 1\), and can correct up to \(t = \lfloor(n - k)/2\rfloor\) code word symbol errors. RS codes satisfy the singleton bound \((d_{\text{min}} \leq n - k + 1)\) with equality. Codes that have such a property are referred to as maximum distance separable (MDS) codes.

MDS codes have several interesting properties. In particular, puncturing \(l\) code symbols changes an \((n, k)\) MDS mother code into a \((n - l, k)\) code which is still MDS [49], i.e., \(d_{\text{min}} = n - l - k + 1\). Puncturing can be used to create a family of codes with different rates and error correcting capabilities. The code family can then be utilized in a rate-adaptive system. This does not increase the complexity significantly, as all members of the code family can be decoded using only one standard \(\text{errors and erasure}\) decoder [49]. Such a decoder can correct, in general, all received words containing \(t\) symbol errors and \(e\) symbol erasures with a constraint \((2t + e) \leq d_{\text{min}}\). When decoding a punctured RS code, it fills in the position of \(l\)
punctured symbols with erasures, and treats the resulting word of length \( n \) as if it were from the mother code. The increased minimum distance gained from this treatment is exactly sufficient to decode the appended erasures.

Upper bounds for post-decoding symbol error probability, \( P_s \), and bit error probability \( P_b \) of a \( t \) error correcting \((n,k)\) MDS code can be obtained if \( P_q \), the average code symbol error probability prior to decoding, is known. Assuming that undetected errors probability can be neglected, the probability \( P_s \) is given by

\[
P_s = \sum_{i=t+1}^{n} \binom{n-1}{i-1} P_q^i (1 - P_q)^{n-i}
\]

(7.17)

An upper bound for bit error probability is obtained by assuming that decoded symbol errors make all the associated data bits incorrect. This yields

\[
P_b \leq P_s
\]

(7.18)

The adaptive-rate system considered in this paper has two degrees of freedom: modulation level \( L \) and code rate \( R = k/n \) (or equivalently code word length \( n \) since \( k \) is fixed). These two parameters must be chosen, according to measured channel condition, to satisfy the BER requirements of the system, while utilizing the available bandwidth, efficiently. This means, transmitting data at the highest permitted rate. Efficient bandwidth utilization is possible only when the adaptive rate scheme provides a dynamic range comparable to that of SNR variations at receiver. The combined adaptive \( L \)-PPM modulation and punctured RS coding is aimed to provide such a range and is thus pursued in this paper.

A block diagram of proposed system is shown in Fig. 7.5-b. This system can operate in fixed-rate mode as well as adaptive-rate mode. From a binary data source, \( kq \) bits enter the RS encoder as \( kQ \)-ary symbols \((Q = 2^q)\). The systematic RS encoder appends \( n - k \) parity symbols to these \( k \) symbols and punctures \( l \) of them \((l \text{ can be controlled by user})\) to obtain a code length of \((n - l)\) symbols,
capable of correcting \((n-l-k)/2\) symbol errors. These \(q\)-bit code symbols are then transmitted over the diversity channel of Fig. 7.5-a using \(K\)-bit \(L\)-PPM modulation symbols \((L = 2^K\) and can be controlled by the user). At the receiver, demodulator attempts to determine the transmitted \(L\)-ary symbols. These symbols are then formed into \(q\)-bit RS code symbols. Finally, each block of \(n\) code symbols is RS decoded to obtain the original \(kq\) bits of data, which are passed to a data sink.

Reed-Solomon decoder complexity, in general, increases considerably as code length \(n\) and the error correction capability \(t\) of code increases. Typical values for \(t\) range from 1 to 16, while value of \(n\) used in most applications is 255. The RS mother code, used throughout this chapter, has a symbol size \(q = 8\) bits (one byte per symbol), a length \(n = 2^q - 1 = 255\) symbols (which can be punctured to yield shorter lengths), and an uncoded message length \(k = 223\) symbols. Thus, it can correct up to \(t = (n-k)/2 = 16\) symbols of errors per code word. The implemented RS codec has a generator \(G(x) = (x-a^0)(x-a^1)\ldots(x-a^{31})\), where \(a\) is a root of the binary primitive polynomial \(x^8 + x^4 + x^3 + x^2 + 1\). The code length is programmable by the user and up to 31 code symbols can be punctured.

The modulation level \(L\) can take on values\(^1\) 4, 8 or 16 corresponding to modulation symbol size \(K\) of 2, 3 and 4. In general, one \(q\)-bit code symbol consists of approximately \(m\) \(K\)-bit modulation symbols where \(m\) can be a non-integer

\[
m \approx \log_2 Q / \log_2 L = q/K
\]

When \(K\) divides \(q\), the \(q\)-bit RS symbol error probability \(P_q\) is related to \(P_K\), the \(K\)-bit modulation error probability by

\[
P_q = 1 - (1 - P_K)^m = mP_K - \frac{m(m-1)}{2}P_K^2 + \ldots
\]

\(^1\)2-PPM is not used as it has the same bandwidth efficiency as 4-PPM, while requiring about 3 dB more optical power for a given BER
where $P_k$ is given in Eq. 7.13. When, on the other hand, $K$ does not divide $q$, the analysis becomes more complicated and use of Eq. 7.20, may lead to incorrect results [50]. The exact expression for $P_q$ in this case can be derived by first calculating the probability $p_i$ that $i$ Q-ary symbols out of $K$ Q-ary symbols are in error, for $i = 1$ to $K$. The average RS code symbol error probability $P_q$ is then obtained by evaluating a weighted sum of $p_i$ values and dividing it by $K$, i.e.,

$$P_q = \frac{1}{K} \sum_{i=0}^{K} i p_i$$

(7.21)

Note that, we considered $K$ $q$-bit RS symbols since it includes an integral number of modulation symbols. Note also that, in calculating $p_i$ we must look at the way the modulation symbols overlap the code symbols. Clearly, when a modulation symbol is incorrect, it could cause two code symbols be incorrect when it spans over them both. An exact calculation of $p_i$ for a particular modulation scheme thus requires an analysis of the particular error events, including an enumeration of them all, which is rigorous. This analysis was performed for the case of 3-bit 8-PPM symbols, which are to be matched with 8-bit 256-ary RS codes. The result is given in Table 7.1.

Fig. 7.6, shows the BER performance of $t$-error correcting RS codes with $L$-ary PPM modulation as a function of SNR, defined in Eq. 7.15. In this Figure, $L$ takes on values of 4, 8 and 16, while $t$ is either 3 or 16. As seen, at a BER=$10^{-9}$, a $t = 3$ RS code can result in more than 3dB gain, while a $t = 16$ RS code can result in more than 5dB gain.

Performance evaluation of the proposed system, whether it be used in fixed-rate mode, or in adaptive-rate mode, requires probability distribution of received SNR in a room, normalized to transmit power $P_t$. Based on the assumptions made in section 7.2, we evaluated received SNR (defined in Eq. 7.15) for 4000 random
Figure 7.6. BER Performance of \( t \)-error correcting RS codes with \( L \)-ary PPM modulation, for \( t=3 \) and \( 16 \), and \( L=4 \), 8, and 16, as a function of SNR. The BER performance of uncoded \( L \)-ary PPM is also plotted as a reference.
Table 7.1. Values of $p_i$, $i = 1, 2, 3$ for matching 3-bit 8-PPM symbols with 8-bit 256-ary RS codes.

<table>
<thead>
<tr>
<th>$i$</th>
<th>$p_i$</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>$(1 - P_K)^8$</td>
</tr>
<tr>
<td>1</td>
<td>$\frac{50}{7} P_K - \frac{2190}{49} P_K^2 + \frac{5848}{49} P_K^3 - \frac{8639}{49} P_K^4 + \frac{7626}{49} P_K^5 - \frac{4024}{49} P_K^6 + 24 P_K^7 - 3 P_K^8$</td>
</tr>
<tr>
<td>2</td>
<td>$\frac{6}{7} P_K + \frac{101}{7} P_K^2 - \frac{3636}{49} P_K^3 + \frac{7004}{49} P_K^4 - \frac{7029}{49} P_K^5 + \frac{3032}{49} P_K^6 - 24 P_K^7 + 3 P_K^8$</td>
</tr>
<tr>
<td>3</td>
<td>$\frac{111}{49} P_K^2 + \frac{76}{7} P_K^3 - \frac{1795}{49} P_K^4 + \frac{2138}{49} P_K^5 - \frac{1280}{49} P_K^6 + 8 P_K^7 - P_K^8$</td>
</tr>
</tbody>
</table>

$$P_q = \frac{1}{K} \sum_{i=0}^{K} i p_i = \frac{434}{147} P_K - \frac{443}{147} P_K^2 + \frac{172}{147} P_K^3 - \frac{16}{147} P_K^4$$

positions and orientations of the receiver inside the room. The SNR distribution began to converge after 1000 random samples. The result is given in Fig. 7.7. For a 7°-FOV receiver, SNR is above -5dB with more than 99% probability and spans a dynamic range of nearly 14dB. For an 11.5°-FOV receiver, the distribution is shifted to the right, as expected. It follows almost the same trend as that for the 7°-FOV receiver. The received SNR in this case is above -3dB with a probability more than 99% and spans a dynamic range of nearly 14dB. Note that, although path loss distribution spans a range of only a few dB, SNR distribution spans a much wider range. This accounts for the non-uniform distribution of ambient noise power inside the room, due to window light and Tungsten lamps.

Using the distribution given in Fig. 7.7, the optical power $P_t$, needed to achieve a BER of $10^{-9}$ with a 99% probability (1% outage), was evaluated as a function of
Figure 7.7. Probability distribution of signal-to-noise ratio (SNR), normalized to transmit power $P_t$, at the receiver, placed at a random position in the room.

information bit rate $R_b$. This was also done using a BER of $10^{-6}$. The results, using fixed-rate mode, for 7°-FOV and 11.5°-FOV receivers are shown in Figs. 7.8-a and 7.8-b, respectively. The most powerful scheme, i.e., the one with $t=16$ and $L=16$, is able to bring the power requirement to well below one Watt even at very high bit rates. Use of an 11.5°-FOV receiver yields a more power-efficient system as long as the bit rate is below a certain level. As bit rate increases and goes above, say, 100Mb/s, ISI will start to influence the results shown in Fig. 7.8(b). In this case, use of a 7°-FOV receiver may be a better option.

The system overall throughput can be increased by having it operate in adaptive-rate mode. In Fig. 7.9, we compare the performance of adaptive-rate mode with
Figure 7.8. Plot of optical power $P_t$, that yields a BER of $10^{-9}$ with 1% outage versus information bit rate $R_b$. Results for BER=$10^{-6}$ are shown by dashed lines. The system is assumed to be operating in fixed-rate mode. Plots are for $L=4,8,16$ and $t=3,16$. (a) 7°-FOV (b) 11.5°-FOV.
that of a fixed-rate mode, using \((L = 16, t = 16)\) scheme for transmission. For a fair comparison, we assign equal bandwidth (BW) to both systems. Then we calculate bit-rate for fixed-rate mode according to:

\[
R_b = \frac{2R}{1 + \alpha} \cdot \log_2 L \cdot \text{BW}
\]  

(7.22)

with \(L = 16, t = 16, R = (255 - 2t)/255\), and \(\alpha = 0.45\). Having the bit rate \(R_b\), optical power \(P_t\) can be calculated using Fig. 7.8. Using this value for the transmit power, we now let the system operate in an adaptive-rate mode. In this mode, unlike the fixed-rate mode, the \((L = 16, t = 16)\) scheme is not always used. It is used only when the BER of \(10^{-9}\) with 1\% outage cannot be achieved by other schemes in the family. When estimated SNR is high, system switches to a different scheme achieving a higher throughput. Using the distribution of Fig. 7.7, it is possible to find the average value of achievable bit rate inside the room. This is shown in Fig. 7.9-a and 7.9-b for a 7\(^\circ\)-FOV and an 11.5\(^\circ\)-FOV receiver, respectively. It is seen that using the adaptive-rate mode, overall throughput of the system can on the average be increased by 50%.

### 7.5 Summary

In this chapter, we designed and thoroughly examined a new infrared link composed of a multi-beam transmitter and a multiple-branch narrow FOV composite receiver. The multi-beam transmitter can be implemented using a computer-generated hologram. Our design goal was to narrow down the receiver branch FOV until the effect of ISI is totally eliminated. In order to compensate for path loss, we used high-rate Q-ary Reed-Solomon codes in conjunction with power-efficient \(L\)-PPM signals to transmit data over several independent channels. We then used maximum-likelihood criterion at receiver to recover PPM symbols, and decoded them at RS decoder. We also proposed a low-complexity rate-adaptive scheme
Figure 7.9. Plot of optical power, that yields a BER of $10^{-9}$ with 1% outage versus average bit rate $R_b$, using adaptive-rate scheme. Results for BER=$10^{-6}$ are shown by dashed lines. When the system is operating in adaptive-rate mode, the achievable bit rate is, on the average, more than 50% higher than that of the fixed-rate mode. (a) 7°-FOV (b) 11.5°-FOV.
to increase the overall throughput of system. It was shown that very high data rates, up to hundreds of megabits per second, can be reached with high reliability everywhere within the coverage area, using transmit optical power levels well below one Watt.
Chapter 8

CONCLUSIONS AND FUTURE WORK

8.1 Summary of Results and Contributions

This thesis has concentrated on system design, with emphasis on rate-adaptive coding and combining for high-speed wireless radio and infrared local-area networks. In the case of radio systems, two well-known channel models representing fixed and mobile channels were used to demonstrate effectiveness of some novel rate-adaptive transmission schemes. First, rate-compatible punctured convolutional (RCPC) codes were used in a channel characterized by lognormal distribution. Based on channel state information at the transmitter, system was able to adaptively select a code in the family to accommodate for protection needs of transmitting data. The scheme is particularly suitable for ATM networks with wireless links. It provides a means for maintaining QoS guarantees and uses the available frequency spectrum, efficiently.

Next, we developed a maximum-likelihood code combining technique, tailored to Viterbi algorithm, for two important retransmission schemes, namely, hybrid type-I and hybrid type-II. The time-varying channel in this chapter was modeled as a slow, flat fading Rayleigh channel. A comprehensive analysis, based upon concept of generalized weight distribution, for average throughput efficiency and outage of these two important ARQ schemes was performed. The average throughput of a type-II ARQ scheme with code combining was shown to be always greater than that
of a type-I ARQ especially for high values of average signal-to-noise ratio (SNR). The throughput curves for the two ARQ types followed the same trend as those of an ideal AWGN channel. It was shown that use of code combining allows system to be robust under severe channel degradation due to fading while maintaining a high throughput. The calculated metric in this case suggested that newly received copies of a packet be added (with proper reliability weightings) to the previously received ones. We showed that required receiver buffer length was independent of the number of retransmissions, and therefore finite buffer-length considerations encountered in other ARQ schemes simply did not apply to our proposed algorithm. Also, numerical analysis for block-fading Rayleigh channels showed reliable data communications can be obtained with only a reasonable number of retransmissions, making the scheme suitable for delay-limited applications.

In the case of wireless infrared channels, first we studied, via computer simulation, some of the MSDC IR link characteristics such as channel impulse response, magnitude response, link coverage range, delay spread distribution and 3-dB bandwidth, using a single-branch wide FOV receiver as well as a 7-branch direction-diversity receiver. Simulations showed that percentage of optical power, received after the first reflection, was substantially smaller for MSDC than for Lambertian pattern illumination. Consequently, the decrease in 3-dB bandwidth due to higher order reflections in MSDC was much less significant than that of diffuse configuration. Although results were based on a specific room of a given dimension, most conclusions reached were general in that they did not change significantly by changing the room configuration and dimensions. For receiver design, two important cases were studied. In the first case, receiver branch FOV was small enough so that at most one spot could lie within its coverage area. In the second case, branch FOV was large enough so that at least one spot could be
covered by each branch. Both designs were able to create nearly ideal channels of high bandwidth. The increase in bandwidth was paid for by an increase in path loss. As a result, design of power efficient signaling schemes to compensate for the high path loss was found to be a necessity.

In an effort to design a practical system, RCPC codes were used as a candidate to encode intensity modulated (IM) optical power. The objective here, was not only to provide a coding gain, but also to provide an adaptive environment for efficient utilization of channel spectral bandwidth. Use of RCPC codes here offered two distinct and attractive advantages. It allowed for low-cost implementation and it provided a means for accurate channel estimation. Based on the estimated signal-to-noise ratio, transmitter and receiver select a code in the RCPC code family with the smallest amount of redundancy that can keep the BER bellow a desired value. We developed a simple decoding scheme by evaluating the ML metric. Performance evaluation required the knowledge of SNR at all possible positions in the room, and so we proceeded with a statistical assessment of the MSDC channel. We obtained a probability distribution function for SNR and based on that evaluated average optical power required to achieve a desired BER with a low outage probability. We also proposed a novel method for estimating the SNR using same decoder side information. The estimate was shown to be very accurate given that SNR or packet length are sufficiently large.

In spite of their unique features, RCPC codes do not seem to be the most power efficient codes for intensity-modulated channels. This is because the average power here is defined to be the average optical power which is the average of transmitted signal amplitude. This suggests that a modulation scheme with a low duty cycle, or equivalently, a high peak-to-average ratio be used to reduced the average power. An example of such a scheme is \(L\)-ary pulse-position modulation (\(L\)-PPM). To study
the potential improvement gained through coding a $L$-PPM signal, we calculated achievable information rate for $L$-PPM as a theoretical performance limit. Analysis was performed for both hard-decision decoding as well as soft-decision decoding. It was shown that $L$-PPM has a significant capacity at low values of SNR which motivates one to search for efficient codes that can be used along with $L$-PPM modulation.

Based on this observation, we used high-rate Reed-Solomon codes in conjunction with power efficient $L$-PPM signals to transmit data over several independent channels. We developed a new maximum-likelihood metric for optimal detection of PPM symbols. We then investigated key properties of RS codes and their suitability for use in a concatenated coding scheme, as well as in a rate-adaptive environment. We then analytically evaluated performance of RS codes when used in conjunction with PPM signals. The analysis was particularly involved and required exhaustive search in situations where the modulation symbol length did not divide the code symbol length. We then took advantage of “maximum-distance separable” property of RS codes and proposed a low-complexity rate-adaptive scheme to increase the overall throughput of the system. It was shown that very high data rates, up to hundreds of megabits per second, can be reached with high reliability everywhere within the coverage area, using transmit optical power levels well below one Watt.

8.2 Future Work

The burst error problem with matching a binary convolutional code with a non-binary modulation scheme, i.e., $L$-PPM was outlined in chapter 6. we also mentioned that it is not possible to combine binary convolutional codes with $L$-PPM and still use soft decision decoding. It is, however, possible to perform soft-decision decoding using Viterbi algorithm if one uses non-binary convolutional codes [51] with an alphabet size equal to $L$, the order of PPM. With such an approach,
one gets free from the burst error problem and at the same time gets the benefit of soft-decision decoding. This seems a promising approach of coding in infrared communication systems since it allows the best features of convolutional codes with the power efficiency provided by $L$-PPM.

In chapter 7, Reed-Solomon codes were used in conjunction with $L$-PPM. At the receiver, $L$-PPM symbols were detected and then decoded. We used an algebraic method for decoding which corresponds to hard-decision decoding of Reed-Solomon codes. Soft decision decoding of Reed-Solomon codes has been implemented by using trellis decoding methods. Trellis decoding schemes make it possible to incorporate both hard and soft decision methods, easily. To establish maximum likelihood performance, Viterbi decoding algorithm can be used. The main reason why trellis decoding is not often used for Reed-Solomon codes is the complexity of decoder, especially in the case of long length and high redundancy codes. However, it is still feasible to use trellis decoding if a reduced search algorithm is applied [52]. Reduced-state trellis decoding of Reed-Solomon codes for $L$-PPM is another topic for research.

In 1993, Berrou et al. [53] reported a new class of codes called turbo codes, whose performance in terms of BER was a few tenth of a dB from Shannon capacity. A turbo encoder is based on parallel concatenation of two recursive systematic convolutional codes and an interleaver. While TCM is a bandwidth efficient scheme, turbo coding is a power efficient scheme, and so is a good candidate for wireless infrared channel, where power efficiency is the most important parameter.
8.3 Publications

The following is a list of contributions.

Journal Publications


Conference Proceedings


REFERENCES


Koorosh Akhavan was born in November 20, 1968 in Tehran, Iran. He received his B.S. degree in Electrical Engineering from Sharif (Aryamehr) University of Technology in 1991. He served in the military from 1991 to 1993. He then worked as a System Manager for VAX/VMS computer systems at the Computations Center of Sharif University of Technology.

He began his graduate study in 1994 in Electrical Engineering at the Pennsylvania State University. During his studies towards M.S. degree, he worked as a graduate research assistant at the Nuclear Resonance Control Laboratory of the Penn State University. His Master’s thesis, under the supervision of Professor Jeffrey L. Schiano, involved assisting in the development of a testbed spectrometer for Pulsed Nuclear Quadrupole Resonance (NQR) spectroscopy and modeling of Nitrogen-14 quantum mechanical dynamics. He received his M.S. degree in 1997.

In Fall 1997 Koorosh became a Ph.D. candidate. He started his research work in May 1998 under the supervision of Professor Mohsen Kavehrad at the Center for Information and Communications Technology Research (CICTR) laboratory at the EE department of Penn State. His current research interests include communication systems theory and channel coding for wireless radio and infrared communications systems.