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SPREAD SPECTRUM MULTIPLE ACCESS WITH PACKET COMBINING

by

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**A thesis submitted to
the Faculty of Graduate Studies and Research
in partial fulfillment of
the requirements for the degree of
Doctor of Philosophy**

Ottawa-Carleton Institute for Electrical Engineering

Faculty of Engineering

Department of Systems and Computer Engineering

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January, 1996

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COMBINING**

submitted by
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in partial fulfilment of the requirements
for the degree of Doctor of Philosophy

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Abstract

When an uncorrectable error pattern is detected at the receiver of a conventional spread spectrum multiple access (SSMA) ALOHA system, the corresponding packet is discarded and the transmitter is informed about the status of the packet. The transmitter then sends another copy of the packet and decoding is attempted at the receiver without using any information from the previously received copies which were decoded unsuccessfully. This may seem a reasonable approach if a small fraction of the packets are discarded. However, ignoring the previously received packets when the retransmission rate is high does not utilize the potential capacity with SSMA. In this thesis all the received copies of a packet are retained and combined together to improve the performance of the SSMA slotted ALOHA system. This combining has been considered for the following three scenarios.

In the first case, forward error correction with maximum likelihood decoding is used in a direct sequence SSMA ALOHA system. Combining the different copies of a packet corresponds to the technique called code combining. Code combining allows for the combination of multiple copies of the same packet (which are typically discarded), to obtain a lower code rate for that specific packet. This results in a throughput which is an increasing function over a broad range of offered load, and therefore, a more stable system. In addition, code combining provides a higher throughput and a smaller time delay for packet transmission.

Packet combining for a direct sequence SSMA ALOHA system in a Rayleigh fading channel constitutes the second scenario. An exponential multipath intensity profile for all users is considered. When the channel is modeled as a frequency selective channel, combining d copies of a packet, effectively, increases the order of diversity at the RAKE receiver from L , for the conventional case, to $d \times L$. This increase translates into a more stable system, an increase in the throughput, and a decrease in the average number of transmissions necessary to transmit a packet.

The final system considered is a frequency hopped SSMA system with packet combining. Combining the received copies of the same packet results in a system which is equivalent to fast frequency hopping with an adaptive order of diversity and improves the performance of the system.

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List of Abbreviations

AWGN	additive white Gaussian noise
ARQ	automatic-repeat-request
BDD	bounded distance decoding
BER	bit error rate
BPSK	binary phase-shift keying
CDMA	code division multiple access
CRC	cyclic redundancy check
DAMA	demand-assignment multiple access
DS	direct sequence
EC	error correcting
EGC	equal gain combiner
FDMA	time division multiple access
FEC	forward error correction
FFH	fast frequency hopping
FH	frequency hopping
iid	independent identically distributed
MAI	multi access interference
MDS	maximum distance separable
MFSK	M-ary frequency-shift keying
MIP	multi-path intensity profile
MLD	maximum likelihood decoding

List of Abbreviations

MRC maximum ratio combiner

pdf probability density function

PN pseudo-noise

PRN packet radio network

QOS quality of service

RCPC rate compatible punctured convolutional

RS Reed-Solomon

SNR signal-to-noise ratio

SSMA spread spectrum multiple access

TDMA time division multiple access

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To my family.

Chapter 1

Introduction

Fixed assignment multiple access techniques, such as time division multiple access (TDMA) or frequency division multiple access (FDMA), work as circuit-switching networks, where a complete physical path is established from the sender to the receiver that remains for the duration of the connection. Once the circuit is established, data is transferred through the network.

Data traffic generated in many data processing applications has a bursty characteristic (having a large peak-to-average ratio of data rate). That is, if one were to observe a user's transmission for a period of time, one would see that the communication channel is required infrequently, but when it is used the data is on the order of thousands (or even more) of bits and is available instantly for transmission. In such a system, if circuit-switched capacity allocation schemes are employed, each user must be assigned enough capacity to meet its peak transmission rate, with the consequence that the resulting channel utilization is low because of the large peak-to-average ratio of the data rate. Of course one could suggest a TDMA or FDMA system with a demand-assigned multiple access (DAMA) technique, in response to the fluctuating user demands.

In a DAMA system, network capacity is allocated (either in frequency or time

slices) in response to user requests for channel capacity. A special channel, usually called the request channel, is assigned to carry requests for capacity assignments in a DAMA network. Some access method to allow reasonably efficient sharing of the request channel must be found.

A data network can be a collection of a large number of users, each one having a bursty nature. Examples of these types of networks are satellite packet communications and cellular and terrestrial data networks. As the number of users increases the use of fixed-assignment FDMA or TDMA is usually impractical, while the use of DAMA would impose an unreasonable amount of overhead on the network [1].

To efficiently transmit bursty traffic in a data network, the data is formatted into one or more fixed-length packets to form a packet-switched network. In order to provide multiple access to packet data networks, random access techniques, such as ALOHA and spread spectrum multiple access (SSMA) ALOHA, have been used.

An overview of this chapter is as follows. Two random access techniques, narrow-band ALOHA and SSMA ALOHA, are reviewed in Section 1.1. Section 1.2 briefly describes the different forms of a spread spectrum communication system.

1.1 Random Access Techniques

In this section two random access techniques are described, ALOHA and SSMA ALOHA. Each technique can be used in an unslotted or slotted fashion as explained below.

Figure 1.1: Narrow-band slotted ALOHA system block diagram.

1.1.1 ALOHA Protocol

The ALOHA protocol is a random access scheme pioneered at the University of Hawaii for interconnecting terminals and computers via radio and satellites. A channel of capacity R_c bits per second is shared by a large population of n users whose traffic is very bursty.

Unslotted ALOHA

In an unslotted ALOHA system each user station transmits packets randomly at the channel bit rate R_c whenever its buffer contains one packet. Each packet contains parity bits for error detection. If packets from different stations overlap, a packet collision is said to have occurred. When a collision happens, the receivers detect the errors and, by some mechanism, the transmitters are informed about the status of the received packets. The transmitters then retransmit the packets until they are received free of errors. If two packets from two transmit stations collide, they will surely collide again if they are retransmitted after a fixed time delay. To avoid repeated collisions, the interval of packet retransmission is randomized for each station. The unslotted ALOHA protocol is shown schematically in Fig. 1.1.

Figure 1.2: Throughput of the unslotted ALOHA system.

In an ALOHA system, the channel traffic contains newly arrived packets as well as retransmitted packets. Let G denote the average channel traffic (offered load) in packets per packet length and assume this traffic is a Poisson process with mean G . The channel throughput, T , is given by [2]

$$T = G \exp(-2G). \quad (1.1)$$

The throughput of the ALOHA system versus the offered load is given in Fig. 1.2.

From Fig. 1.2 it can be seen that $G = f(T)$ is a double-valued function. For a given value of T there are two values of G , namely G and $G' > 0.5 > G$, such that $T = Ge^{-2G} = G'e^{-2G'}$. This indicates that, as the channel traffic increases past 0.5, the throughput drops because the number of packets that suffer a collision increases, creating more packet retransmissions. Hence, there is a further increase in channel traffic and consequently a decrease in channel throughput, creating a backlog of packets that grows with time (run-away effect). This instability is an inherent characteristic of the ALOHA channel and can only be prevented by operating it well below the maximum throughput with enough margin for peak traffic or with some sort of control.

Slotted ALOHA

The advantage of the pure ALOHA protocol is its simplicity, which can result in low-cost solutions since no synchronization is required between stations in the system. The disadvantage is somewhat inefficient channel utilization; that is, the maximum channel throughput is only 18.4% of the available capacity. One strategy for improving pure ALOHA channel throughput is to coordinate transmissions among the stations by synchronizing the packet transmissions, in effect providing time slots of the same duration as a packet length. A station can transmit a packet only at the start of a slot. Thus, when two packets collide, they will overlap completely, i.e., partial overlaps like those occurring in pure ALOHA channels will never occur. This modified ALOHA scheme is called slotted ALOHA.

If we assume that the user population n is large and the offered load has a Poisson distribution, it can be shown that the slotted ALOHA throughput is given by [2]

Figure 1.3: Throughput of the unslotted ALOHA system.

$$T = G \exp(-G). \quad (1.2)$$

A plot of the throughput, T , versus the channel traffic, G , of slotted ALOHA is given in Fig. 1.3. The slotted ALOHA maximum throughput is 0.368, which is twice the maximum capacity of the pure ALOHA channel.

Figure 1.4: Block diagram of a SSMA ALOHA system.

1.1.2 SSMA ALOHA

A number of narrow-band channels can be combined to form an SSMA channel. An SSMA channel can be used by more than one user at the same time. As will be explained later in this chapter, in such a system, each transmitter-receiver pair is assigned a distinct code sequence. This code sequence is used to extract a particular signal from the received signal, which contains the signals of all other users.

Employing SSMA in a packet radio system is termed SSMA ALOHA. A simple block diagram of a SSMA ALOHA system is given in Fig. 1.4. When an uncorrectable error pattern is detected at the receiver of a DS/SSMA ALOHA system, the corresponding transmitter is informed about the status of the packet. The transmitter then sends another copy of the same packet. The retransmissions continue until the receiver is able to reconstruct the transmitted packet. SSMA ALOHA can be used in an unslotted or slotted fashion. In this thesis only a slotted system is considered.

Figure 1.5: Block diagram of a spread spectrum communication system.

1.2 Spread Spectrum

In a digital communication system a transmission bandwidth much greater than the information rate can be used to form a spread spectrum communication system. The block diagram of a spread spectrum digital communication system is given in Fig. 1.5. There are two pseudo-random or pseudo-noise (PN) sequence generators, one which interfaces with the modulator and a second which interfaces with the demodulator. The generators generate a PN sequence which is impressed on the transmitted signal at the modulator and removed from the received signal at the demodulator.

There are many ways to spread the spectrum of a signal, among them direct-sequence (DS) and frequency hopping (FH) techniques.

1.2.1 DS Spread Spectrum

One method of spreading the spectrum of a data-modulated signal is to modulate the signal a second time using a very wideband spreading signal. This second modulation is usually some form of digital phase modulation. The simplest form of DS spread spectrum employs binary phase-shift keying (BPSK) as the spreading modulation.

A simple block diagram of a DS/SS communication system with BPSK modulation is given in Fig. 1.6. At the receiver, the information-bearing signal, $b(t)$, is first multiplied by the carrier to form a constant-envelope data-modulated carrier given by

$$s_d(t) = \sqrt{2P} \cos[\omega t + \theta_d(t)]. \quad (1.3)$$

where the phase $\theta_d(t)$ can be 0 or 180 (90 or 270) degrees depending on the data bit.

BPSK spreading is accomplished by simply multiplying $s_d(t)$ by a function $a(t)$ which takes on values ± 1 . The transmitted signal $s_t(t)$ is given by

$$s_t(t) = \sqrt{2P}a(t) \cos[\omega t + \theta_d(t)]. \quad (1.4)$$

At the receiver, the demodulation is accomplished by remodulating the received signal by the delayed replica of the spreading code, i.e., $a(t - T_t)$, where T_t is the transmission delay. This remodulation of the received signal with the delayed spreading waveform is called despreading. Despreading is a critical operation in all spread spectrum systems. If the transmission delay is calculated correctly, the signal at the output of the despreader is given by

$$\begin{aligned} \sqrt{2P}a(t - T_t)a(t - T_t) \cos[\omega t + \theta_d(t - T_t) + \phi] \\ = \sqrt{2P} \cos[\omega t + \theta_d(t - T_t) + \phi], \end{aligned} \quad (1.5)$$

since $a(t - T_t)a(t - T_t) = 1$. In this case the signal component of the output of the

Transmitter

Receiver

Figure 1.6: Block diagram of a DS/SS with BPSK.

receiver despreader is equal to $s_d(t)$ except for a random phase ϕ , and $s_d(t)$ can be modulated using a conventional coherent phase demodulator.

1.2.2 FH Spread Spectrum

A second method for widening the spectrum of a data-modulated carrier is to change the frequency of the carrier periodically. Typically, each carrier frequency is chosen from a set of 2^k frequencies which are spaced approximately the width of the data modulation spectrum, although neither condition is absolutely necessary. The spreading code in this case does not directly modulate the data modulated carrier but instead is used to control the sequence of carrier frequencies. Because the transmitted signal appears as a data-modulated carrier which is hopping from one frequency to the next, this type of spread spectrum is called frequency-hop spread spectrum. At the receiver, the frequency hopping is removed by mixing (down-converting) the received signal with the local oscillator. This local oscillator hops synchronously with the received signal. FH spread spectrum systems can be categorized into two categories, slow and fast FH. With slow FH, the frequency band changes more slowly than the symbols come out of the data modulator. Thus, with slow FH one or more data symbols can be transmitted in a frequency band. Conversely, the frequency band changes several times per symbol in a fast FH system.

Most FH spread spectrum systems use either noncoherent or differentially coherent data modulation schemes [3]. The modem block diagram for a FH spread spectrum is given in Fig. 1.7. In contrast to the DS system, where the spreading code was used one bit at a time, the spreading code here is used k bits at a time. Using these k bits, the data modulator output is translated to a new frequency

(among the 2^k frequencies) by the frequency-hop modulator. In the receiver, using a local oscillator and a code generator that is synchronized to the one in the receiver, the transmitted signal is down-converted. The down-converted output can be demodulated using conventional methods, i.e., a bank of bandpass filters with energy detectors at their output for M-ary frequency-shift keying (MFSK).

1.3 Problem Statement and Thesis Contents

The problem considered in this thesis can be stated as follows: *How the uncorrectable received copies of a packet available at the receiver can be combined together to improve the performance of a spread spectrum packet radio network?* The main contributions of the thesis are given in Chapters 2, 3 and 4.

In Chapter 2, maximum likelihood decoding and code combining is applied to a slotted DS/SSMA ALOHA packet radio system with coherent BPSK modulation. Multiple access interference is considered as the only source of errors in this system. Code combining allows for the combination of multiple copies of the same packet (which are typically discarded), to obtain a lower code rate for that specific packet, and therefore an improved probability of successful decoding. Combining multiple copies of the same packet results in a throughput which is an increasing function over a broad range of offered load, so that the system is more reliable from the point of view of stability. In addition, combining provides a higher throughput and a smaller time delay for packet transmission. This is illustrated by means of analytical and simulation results.

Chapter 3 considers a DS/SSMA slotted ALOHA packet radio system using BPSK modulation and a modified RAKE receiver, in a mobile environment. The

Figure 1.7: Block diagram of a FH/SS.

channel is modeled as a frequency-selective Rayleigh fading channel with an exponential multi-path intensity profile (MIP) for all users. Average power control is assumed, which means that on average the same power is received from each user. The RAKE receiver retains and combines all received copies of a packet, instead of discarding those which are detected in error. This effectively results in a multiplication of the diversity order, at the RAKE receiver, by a factor d , where d is the number of packet copies being combined.

In Chapter 4, combining the different received copies of the same packet is applied to a frequency hopped spread spectrum packet radio network. This improves the throughput and reduce the delay necessary to transmit a packet successfully. MFSK modulation with square-law envelope detector is assumed. This system is equivalent to a fast frequency hopped system where the diversity level is adapted according to the channel conditions. Reed-Solomon codes are used for error correction and detection.

It should be mentioned that, in all the above mentioned cases, in order to retain the uncorrectable copies of a packet at the receiver, some extra memory space is needed.

Chapter 2

Maximum Likelihood Decoding and Code Combining for DS/SSMA Slotted ALOHA

2.1 Introduction

Consider a narrow-band ALOHA packet radio network. It is assumed that the multiple access interference (MAI) is the only source of errors in the system. In this case, when only one user is transmitting in a given time period, it can be assumed that the packets are received successfully. However, if two or more users are transmitting simultaneously, mutual interference severely corrupts the packets from these users, rendering the channel useless. This means that there is very little useful information in an erroneous copy of a packet to aid the receiver in recovering the data corresponding to that packet. Thus, when a packet is received in error it is best to simply discard it and request a retransmission of that packet. There is also no benefit in using forward error correction (FEC) for a narrow-band ALOHA system based on the following:

1. When a single user is utilizing the channel, all packets are received correctly.

2. When the channel is unusable, no practical amount of error correction can recover the affected packets.

In a DS/SSMA ALOHA system, (since we assume MAI is the only source of error) the average number of errors introduced into a packet depends on the number of packets that are transmitted simultaneously. The severity of the errors within a packet is a monotonically increasing function of the number of simultaneous users. When the number of these simultaneous users is low, the average number of errors within a packet is small, so the packet contains much useful information about the transmitted data. Therefore, it seems reasonable that combining corrupted copies of the same packet can improve the performance of the system. The use of FEC also seems promising for this system. Previous work by Morrow and Lehnert [4], among others, has confirmed that using FEC can improve the performance.

When an uncorrectable error pattern is detected at the receiver of a conventional DS/SSMA ALOHA system, the corresponding packet is discarded and the transmitter is informed about the status of the packet. The transmitter then sends another copy of the packet and decoding is attempted at the receiver without using any information from the previously received copies which were decoded unsuccessfully. This may seem a reasonable approach if a small fraction of the packets are discarded. However, ignoring the previously received packets when the retransmission rate is high does not utilize the potential capacity with DS/SSMA.

In this chapter we combine all received copies of the same packet. This is done for a slotted DS/SSMA ALOHA system with and without FEC. For the uncoded system, this can be considered as diversity combining, with the same signal transmitted through several different channels (each channel corresponding to one copy). For the coded system, combining the received copies of a packet is termed code

combining, and has been shown to provide effective communications in a very noisy channel [5] [6]. In this case, combining all received noisy copies of the same packet results in a lower code rate for that packet. Thus, if initially a rate r code is used, combining J packets results in a rate r/J code. It is important to note that this lower code rate (and therefore extra redundancy, which results in a lower throughput for a given offered load) applies only to those packets that require more FEC. In effect, redundancy is added only to those packets which are not successfully received. This translates into an overall lower level of interference for the system, which is a key factor in the capacity of a SSMA system (in terms of the number of simultaneous users).

This chapter is organized as follows. In Section 2.2, code combining technique is described. The system description and motivation for this work are given in Section 2.3. The performance analysis of slotted DS/SSMA ALOHA with and without FEC, where different copies of a packet are combined, is given in Section 2.4. Section 2.5 provides some examples and finally, Section 2.6 presents some conclusions based on the results.

2.2 Code Combining

Automatic-repeat-request (ARQ) schemes offer high reliable communications at the expense of throughput degradation at high channel error rates. Forward error correction (FEC) schemes on the other hand provide constant throughput but suffer a decrease in reliability at high channel error rates. A combination of ARQ and FEC, known as hybrid ARQ, can offer a performance superior to either technique alone [7]-[10].

Hybrid ARQ schemes can be classified into two categories, namely type-I and type-II schemes [11]. Type-I hybrid ARQ uses a code which is designed for simultaneous error detection and correction [12]-[13]. When a received word is detected in error, the receiver first attempts to correct the errors. If an uncorrectable error pattern is detected, the receiver rejects the received word and requests a retransmission. The retransmission is the same codeword. When the retransmitted codeword is received, the receiver again attempts to correct the errors (if there are any). If the decoding is not successful, the receiver again rejects the received word and requests another retransmission. This continues until the codeword is successfully decoded. Type-I ARQ schemes are best suited for communication systems in which a fairly constant level of noise and interference is anticipated on the channel.

For a channel with a nonstationary bit error rate, one would like to design an adaptive hybrid ARQ system. When the channel bit error rate is low, the system behaves just like a pure ARQ system, with only parity-check bits for error detection included in each transmission. Therefore, the throughput performance is the same as a pure ARQ system. However, when the channel bit error rate is high, extra parity-check bits are needed. This is the basis of type-II hybrid ARQ schemes. The concept of parity retransmission for error correction was first introduced by Mandelbaum [14]. The first type-II hybrid ARQ using a parity retransmission strategy was proposed by Metzner [15] [16], which was later extended and modified by many others, among them [9] and [10].

In [10] Wang and Lin have suggested using parity retransmission with a rate $1/2$ invertible code to provide FEC capability in a hybrid ARQ system. In the Wang-Lin scheme, an (n, k) block code, denoted C_0 , is used for error detection. Also, a rate $1/2$ invertible code, i.e., a rate $1/2$ memory m $(2, 1, m)$ convolutional code,

C_1 , is used for error correction. Viterbi decoding is assumed at the decoder. Let $(G_1(x), G_2(x))$ be the two generator polynomials of C_1 , with an encoder as shown in Fig 2.1. When a message is ready for transmission, it is first encoded into a codeword $I(X)$ in C_0 . The sequence $I(X)$ is then transformed into two sequences $P_1(X) = I(X)G_1(X)$ and $P_2(X) = I(X)G_2(X)$, each $(n + m)$ bits long. The $2(n + m)$ bit sequence obtained by interleaving $P_1(X)$ and $P_2(X)$ is a code sequence for $I(X)$ based on the rate $1/2$ convolutional code C_1 . The transmitter alternately sends the sequence $P_1(X)$ and $P_2(X)$, upon request. A correct reception occurs if the most recent transmission is received error free, in which case $I(X)$ is recovered by inversion, or if the combination of the last two sequences received contains an error pattern which is correctable by the rate $1/2$ code. Such a system can provide a larger throughput at higher error rates than a pure ARQ system. However, it fails to provide useful throughput at very high channel error rates. Note that the performance of the aforementioned parity retransmission scheme is limited by the performance of the rate $1/2$ code selected.

Chase has suggested a maximum-likelihood approach [5], known as code combining, to provide useful throughput at very noisy channels. With code combining, a receiver keeps all the received copies of a packet and combines them to form a lower rate code for that specific packet. If each packet is a codeword of a rate R code, combining L copies of a packet creates a codeword of a rate R/L code. Furthermore, with code combining, each packet can be weighted according to its reliability. Thus, the less reliable copies have a smaller contribution in the decoding process. Note that maximum likelihood decoding, rather than bounded distance (algebraic) decoding, is a critical feature of code combining.

In [17], Chase incorporated code combining [5] into the Wang-Lin [10] parity

Figure 2.1: Encoder of a memory 2 (4-state) convolutional code.

retransmission scheme. With code combining, the transmission still alternates between $P_1(X)$ and $P_2(X)$ of Fig. 2.1, but the decoder operates on the combination of all previous receptions for a given packet. The most recent reception can be inverted as before, but after the first reception this provides little performance gain. The trellis structure after combining two and three copies of a packet is given in Figs. 2.2 and 2.3, respectively. As can be seen, there are two bits for each branch of the trellis in Fig. 2.2, while each branch of the trellis in Fig. 2.3 maps into three bits. Combining more copies simply means adding more bits to each branch of the trellis, while the trellis structure is preserved. Note that when an invertible rate 1/2 convolutional code is used, inversion is easily carried out by searching the trellis of the rate 1/2 code while treating one of the two digits on each branch as an erasure. Which digit to erase depends on which sequence has been received, $P_1(X)$ or $P_2(X)$. Fig. 2.4 shows the trellis diagram where only one bit on each branch is transmitted (the second bit is erased). As can be seen, two branches depart from each node, an upper branch and lower branch. In each node, the bit corresponding to the upper branch is different from the corresponding bit of the lower branch (if one of them is a 1 the other one is a 0). Therefore, if the starting state of the encoder is known by the decoder, the received sequence is mapped into a single path in the trellis. Assuming that the received packet is error free, the decoder output will be a codeword of C_0 , and the inversion is successfully performed.

In [6], Kallel has analyzed the performance of the Wang-Lin scheme with code combining in more detail. He also has analyzed the scheme where the incremental redundancy bits are provided by the transmitter, if the first transmission of a packet is not successful [18].

In [19], Hagenauer extended the concept of punctured convolutional codes to the

Figure 2.2: Trellis diagram for rate 1/2 code of Fig. 2.1 after receiving the second transmission associated with a packet.

Figure 2.3: Trellis diagram for rate 1/2 code of Fig. 2.1 after receiving the third transmission associated with a packet.

Figure 2.4: Trellis diagram for rate 1/2 code of Fig. 2.1 after receiving the first transmission associated with a packet.

generation of a family of rate compatible punctured convolutional codes (RCPC). A mother rate $1/N$ convolutional code is periodically punctured to obtain a family of rate-compatible codes with decreasing rate. The rate-compatibility restriction to the puncturing rule implies that all the bits of a high rate code in the family are used by the lower rate codes. Based on the RCPC codes, Hagenauer proposed a modified type II hybrid ARQ strategy. In Hagenauer's ARQ strategy, the first transmission of a data packet includes parity bits for error correction according to a high rate code in the family. If the high rate code is not sufficiently powerful to overcome channel errors, then incremental redundancy bits, which were deleted by the puncturing process, are transmitted a few at a time, as needed. This decreases the code rate towards the rate of the mother code.

2.3 Preliminaries

2.3.1 System Description

The DS/SSMA system examined in this chapter is shown in Fig. 2.5 for K users. The modulation used is BPSK. The data and spreading signals are denoted by $b_k(t)$ and $a_k(t)$, respectively. An asynchronous system is assumed where the phase ϕ_k ($k = 0, 1, \dots, K$) is uniformly distributed in the interval $[0, 2\pi)$ and the delay τ_k is uniformly distributed in the interval $[0, T_b)$, where T_b is the data bit duration.

The following assumptions about the user population and packet generation are made. These are essentially the same as those in [4]. It is assumed that the channel is used by a large number of users, n . Each user generates packets of length L at a rate of I packets per time slot. Therefore, the channel receives newly generated

Figure 2.5: DS/SSMA system model.

packets at a rate I (input rate) where

$$I = \sum_{j=1}^n I_j, \quad (2.1)$$

and I has a Poisson distribution. The arrival of previously unsuccessful packets for retransmission also has a Poisson distribution, at a rate of H packets per time slot. Therefore, the composite packet arrival (offered load) distribution is Poisson with rate $G = I + H$. The composite arrival distribution for a given slot, $f_{M_1}(m)$, is given by

$$f_{M_1}(m) = \frac{G^m}{m!} e^{-G}. \quad (2.2)$$

Errors at the output of the demodulator are assumed to be independent, and caused only by MAI (i.e., the effects of thermal noise are negligible). The bit-to-bit error dependence may result in a slightly different throughput [4], [20], but will not significantly affect the comparison between systems. Further, we assume equal received power for all signals and random interfering signature sequences. A cyclic redundancy check (CRC) code is applied to detect erroneous packets after decoding. By properly choosing the parameters of the (n_c, k_c) error detecting code, the undetected error probability will be small (upper bounded by $2^{-(n_c - k_c)}$).

A feedback channel is used to report the status of received packets to the transmitter. In practice feedback channels can be noisy, however, since the objective of the chapter is to compare the slotted DS/SSMA ALOHA using code combining with the conventional system, a perfect return channel, i.e. no errors in the ACK/NAK signals, is considered.

For the case where FEC is not used and several copies of the same packet are available, the matched filter outputs corresponding to each bit are added together

Figure 2.6: Block diagram for the k th user of the DS/SSMA ALOHA system employing code combining.

before the bit decision is made.

When FEC is used, code combining, as explained in Section 2.2 is used to combine the different received copies of the same packet. The system block diagram for the k th user of the DS/SSMA ALOHA system is given in Fig. 2.6.

2.3.2 Motivation

The throughput, T , is defined as the average number of successfully transmitted packets per time slot. To account for the additional redundancy introduced by FEC, the effective throughput is defined as

$$T_{eff} = T \times r, \quad (2.3)$$

where r is the code rate. When FEC is not used, $r = 1$.

To illustrate the motivation behind this work, Figs. 2.7 and 2.8 show simulation and analytical results for the effective throughput of a conventional slotted DS/SSMA ALOHA system when different fixed rate convolutional codes are used¹. The throughput analysis is given in the following sections. These results are based on a rate 1/4, memory 4 convolutional code with generator polynomials (23,35,27,33) [19]. The generator polynomials for the corresponding rate 1/2, 1/4, 1/6, ..., codes are (23,35), (23,35,27,33), (23,35,27,33,23,35), ..., respectively. In all cases, the packets carry 500 information bits. Thus with a rate 1/ i code, each packet contains 500 i bits.

When code combining is used, the 500 information bits are encoded by means of a rate 1/4 convolutional code, which results in 2000 bits. These 2000 bits are divided into four packets, numbered 1 to 4, each consisting of 500 bits corresponding to each

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¹

The reason that the analytical and simulation results differ is that an upper bound was used to determine the error probability for the analytical curve. Therefore the BER is higher than that obtained via simulation. This upper bound becomes very pessimistic when the signal-to-noise ratio (SNR) is low (large number of users). Thus the curves in Fig. 2.8 fall more quickly than in Fig. 2.7. In addition, the analytical results are based on infinite level quantization, compared to the 8-level quantization used for the simulation curves. Therefore, the analytical curves have a higher throughput for high SNR (smaller number of users), since in this case the BER upper bound is tight.

Figure 2.7: Simulated throughput for conventional slotted DS/SSMA ALOHA using convolutional codes with several code rates.

Figure 2.8: Analytical throughput for conventional slotted DS/SSMA ALOHA using convolutional codes with several code rates.

generator polynomial. In the first transmission, only packet number 1 is sent, which is the same as using a rate 1 code. This is the same as was considered in [6]. If this packet is detected in error, packet number 2 is sent. Packets 1 and 2 combined together form a codeword of a rate $1/2$ code. If after decoding this codeword, errors are still detected, packet number 3 is transmitted, and at the receiver it is combined with packets 1 and 2 to form a codeword of a rate $1/3$ code. This process continues until packet 4 is transmitted. These 4 packets are transmitted in turn until decoding is successful. This results in codes of rates $1, 1/2, 1/3, \dots$, after the transmission of the first, second, third, ... copy, respectively. Figs. 2.7 and 2.8 show that a lower rate code has a lower initial throughput, but has a higher throughput at some higher offered load. The use of code combining guarantees that the throughput is at least the envelope of the curves in these figures. In fact, the throughput is actually better than this envelope, and is not bell shaped in terms of the offered load. This result is explained below.

Let us modify the process of code combining to the following inferior system (which is the same as the conventional system with a rate $1/2$ code), with performance as shown in Figs. 2.7 and 2.8. If after receiving the even numbered packets data recovery is still impossible, both previously received packets are discarded and the transmission restarted. Since code combining will perform better than this scheme, the curve corresponding to the rate $1/2$ code in Figs. 2.7 and 2.8 is a lower bound on the throughput with code combining.

Consider a second system which is the same as the conventional system with a rate $1/4$ code, with performance as shown in Figs. 2.7 and 2.8. After receiving packets numbered 1, 2 or 3, no attempt is made to recover the transmitted data. When packet number 4 is received, all 4 packets are combined and decoding at

tempted. If the data is recovered unsuccessfully, all 4 previously received packets are discarded and the transmission restarted. This system is also inferior to that with code combining, since with code combining successful decoding can occur after the reception of any packet. Thus the curve corresponding to the rate 1/4 code in Figs. 2.7 and 2.8 is also a lower bound on the throughput with code combining. Extending this argument to all possible code rates shows that the throughput with code combining is lower bounded by the performance with fixed code rates in Fig. 2.7 and 2.8.

Next, consider one of the bell shaped curves in Figs. 2.7 or 2.8. This curve shows the effective throughput, τ_{eff} , of a conventional slotted DS/SSMA ALOHA system in terms of the offered load, G . It is clear that increasing the offered load increases the throughput until its maximum value, $\max\{T_{eff}\}$, is attained. As long as the input load (or the rate of newly generated packets), I , is smaller than $\max\{T_{eff}\}$, the throughput of the system equals the input load, $I = T_{eff}$. However, when $I > \max\{T_{eff}\}$, on average a higher number of packets per slot are transmitted, which causes a higher symbol error probability so that the resulting throughput T_{eff} is smaller than $\max\{T_{eff}\}$, and therefore smaller than I . This means that the system cannot successfully transmit packets as fast as the rate of newly-generated packets. Let us assume that the system remains in this situation ($I > \max\{T_{eff}\}$) for a long period of time, which creates a backlog of packets in the system. This causes the throughput to become very small. Thus, even if no new packets are generated, the system will be occupied for a long time in clearing the backlog.

With code combining, even if the system throughput does begin to decrease due to the offered load increasing beyond the maximum throughput (in other words, even if a point is reached where any further increase in the offered load decreases

the throughput), restricting the generation of new packets will allow the backlog to be cleared much more rapidly. This is because receiving additional copies of a packet improves the probability of successfully decoding that packet, which is not the case for the conventional system. In addition, these results demonstrate that increasing the offered load does not decrease the throughput over a broad range of the offered load.

2.4 Performance Analysis

2.4.1 DS/SSMA Slotted ALOHA Without FEC

In this section, the performance of the slotted DS/SSMA ALOHA system without FEC is evaluated. BPSK modulation with coherent detection is considered. Two cases are compared. The first is a conventional slotted DS/SSMA ALOHA system which discards all unsuccessfully decoded packets, so that the receiver uses only the most recently received packet in an attempt to recover the transmitted data. The second case combines all received copies of the same packet. This means that when several copies of the same packet are available at the receiver, the decision on a given bit is made based on the sum of the soft decision matched filter outputs corresponding to that bit. Denote this case as slotted DS/SSMA ALOHA with combining.

With SSMA communications, each user has its own PN sequence. This enables many direct-sequence spread spectrum signals (users) to occupy the same channel bandwidth. At the receiver, the signals from the other simultaneous users appear as an additive interference. The level of interference varies depending on the number

of users transmitting at a given time. In this analysis, an asynchronous system with independent errors at the output of the demodulator is considered. A Gaussian approximation is used for the interference [21]. As shown in [20], a Gaussian approximation is valid when the number of simultaneous users is large, and this is the region of interest in this work. The SNR conditioned on the number of simultaneous users, m_i , is then given by [21], [20]

$$SNR(m_i) = \frac{1.5N}{m_i - 1}, \quad (2.4)$$

where N is the number of chips per data bit. Note that (2.4) differs slightly from similar expressions [21], [20] because of a difference in definition. However, they result in the same error probability. The bit error probability, conditioned on m_i , is given by

$$\begin{aligned} p_b(m_i) &= Q(\sqrt{2SNR}) = \frac{1}{2}\text{erfc}(\sqrt{SNR}) \\ &= Q\left(\sqrt{\frac{3N}{m_i - 1}}\right) = \frac{1}{2}\text{erfc}\left(\sqrt{\frac{1.5N}{m_i - 1}}\right), \end{aligned} \quad (2.5)$$

where

$$\begin{aligned} Q(x) &= \frac{1}{\sqrt{2\pi}} \int_x^\infty e^{-t^2/2} dt \\ \text{erfc}(x) &= \frac{2}{\sqrt{\pi}} \int_x^\infty e^{-t^2} dt. \end{aligned} \quad (2.6)$$

Multi-channel signaling, where a signal is sent through more than one channel, is used to evaluate the performance of slotted DS/SSMA ALOHA with combining since the two systems are similar.

Figure 2.9: The multi-channel signaling system.

Consider the multi-channel signaling system shown in Fig. 2.9. In this system, each bit is transmitted via Q different channels with additive white noise having two-sided power spectrum densities $N_{oi}/2$, $i = 1, 2, \dots, Q$. At the receiver, the outputs of all Q matched filters corresponding to the same bit are added together, and this sum is used by the decision device to decide on the transmitted bit. The equivalent SNR, which results from adding the corresponding signals from different channels, is given by

$$SNR_{eqQ} = \frac{Q^2}{\frac{1}{SNR_1} + \frac{1}{SNR_2} + \dots + \frac{1}{SNR_Q}}, \quad (2.7)$$

where SNR_i is the SNR corresponding to the i th channel. Eq. (2.7) is derived in Appendix I.

Slotted DS/SSMA ALOHA with Q copies of the same packet available at the receiver is similar to multi-channel signaling with Q channels. Each copy can be considered as having been received via a unique channel. If a total of m_i packet

were transmitted simultaneously during transmission of the i th packet, the equivalent conditional SNR, $SNR(m_1, m_2, \dots, m_Q)$, which results from combining these Q copies, is given by

$$SNR(m_1, m_2, \dots, m_Q) = \frac{1.5NQ^2}{m_1 + m_2 + \dots + m_Q - Q}. \quad (2.8)$$

Let S_e and S_c denote the events that the received packet is detected in error and the received packet is error free, respectively. Also, let $R_{av}(m_1, m_2, \dots)$ denote the average number of transmissions necessary to transmit a given packet successfully when m_i other users are transmitting simultaneously during the i th, $i = 1, 2, \dots$, transmission of this packet. Then $R_{av}(m_1, m_2, \dots)$ is given by [6]

$$R_{av}(m_1, m_2, \dots) = 1 + p(S_e(m_1 + 1)) + p(S_e(m_1 + 1), S_e(m_1 + 1, m_2 + 1)) \\ + p(S_e(m_1 + 1), S_e(m_1 + 1, m_2 + 1), S_e(m_1 + 1, m_2 + 1, m_3 + 1)) + \dots, \quad (2.9)$$

where $p(S_e(m_1 + 1, m_2 + 1, \dots, m_i + 1))$ is the probability of the event that after combining i copies of the same packet, the receiver has not recovered the transmitted data corresponding to that packet, with the condition that $m_j + 1$ users were simultaneously transmitting during the transmission of the j th, $j = 1, 2, \dots, i$, copy. This probability is given by

$$p(S_e(m_1 + 1, m_2 + 1, \dots, m_i + 1)) \\ = 1 - [1 - \frac{1}{2}\text{erfc}(\sqrt{SNR(m_1 + 1, m_2 + 1, \dots, m_i + 1)})]^L \\ = 1 - [1 - \frac{1}{2}\text{erfc}(\sqrt{\frac{1.5Ni^2}{m_1 + m_2 + \dots + m_i}})]^L. \quad (2.10)$$

where L is the number of bits per packet. Since

$$p(S_e(m_1 + 1), S_e(m_1 + 1, m_2 + 1), \dots, S_e(m_1 + 1, m_2 + 1, \dots, m_i + 1))$$

$$\leq p(S_e(m_1 + 1, m_2 + 1, \dots, m_i + 1)). \quad (2.11)$$

(2.9) can be written as

$$\begin{aligned} R_{av}(m_1, m_2, \dots) \leq & 1 + p(S_e(m_1 + 1)) + p(S_e(m_1 + 1, m_2 + 1)) + \\ & p(S_e(m_1 + 1, m_2 + 1, m_3 + 1)) + \dots, \end{aligned} \quad (2.12)$$

From (2.12), the average number of transmissions for a given packet is

$$\begin{aligned} R_{av} \leq & 1 + \sum_{m_1=1}^{\infty} p(S_e(m_1 + 1))f_M(m_1) + \\ & \sum_{m_1=1}^{\infty} \sum_{m_2=1}^{\infty} p(S_e(m_1 + 1, m_2 + 1))f_M(m_1, m_2) + \\ & \sum_{m_1=1}^{\infty} \sum_{m_2=1}^{\infty} \sum_{m_3=1}^{\infty} p(S_e(m_1 + 1, m_2 + 1, m_3 + 1))f_M(m_1, m_2, m_3) + \dots, \end{aligned} \quad (2.13)$$

where $f_M(m_1, m_2, \dots, m_i)$ is the joint probability density function (pdf) of packet generation (including new packets and packets to be retransmitted). In other words, $f_M(m_1, m_2, \dots, m_i)$ is the probability that m_1 packets are generated in a given time slot, m_2 packets are generated in the next time slot, and so on.

The right hand side of (2.10) is a function of $m_1 + m_2 + \dots + m_i$ only, so that R_{av} can be bounded as

$$\begin{aligned} R_{av} \leq & 1 + \sum_{m_1=1}^{\infty} p(S_e(m_1 + 1))f_{M_1}(m_1) + \\ & \sum_{\substack{m_1=0, m_2=0 \\ m_1+m_2 \geq 2}}^{\infty} p(S_e(m_1 + m_2 + 2))f_M(m_1, m_2) + \end{aligned}$$

$$\begin{aligned}
& \sum_{\substack{m_1 = 0, m_2 = 0, m_3 = 0 \\ m_1 + m_2 + m_3 \geq 3}}^{\infty} p(S_e(m_1 + m_2 + m_3 + 3))f_M(m_1, m_2, m_3) + \dots \\
& \hspace{15em} (2.14)
\end{aligned}$$

It is assumed that the composite packet arrivals (offered load) in different time slots are independent identically distributed Poisson random variables, with pdf given by (2.2). Therefore, the summation of these random variables is also a Poisson distributed random variable. The rate of this new random variable is the sum of the rates of the random variables in the sum. Therefore, for $m = m_1 + m_2 + \dots + m_i$, we have

$$f_{M_i}(m) = \frac{(iG)^m}{m!} e^{-iG}. \quad (2.15)$$

Substituting (2.15) and (2.10) into (2.14) results in

$$R_{av} \leq 1 + \sum_{i=1}^{\infty} \sum_{m=i}^{\infty} \left\{ 1 - \left[1 - \frac{1}{2} \operatorname{erfc} \left(\sqrt{\frac{1.5Ni^2}{m}} \right) \right]^L \right\} \frac{(iG)^m}{m!} e^{-iG}. \quad (2.16)$$

It should be mentioned that for the conventional system, the average number of transmissions to transmit a packet successfully, $R_{av_{con}}$ is given by

$$\begin{aligned}
R_{av_{con}} &= 1 + \sum_{m_1=1}^{\infty} p(S_e(m_1 + 1))f_{M_1}(m_1) + \left[\sum_{m_1=1}^{\infty} p(S_e(m_1 + 1))f_{M_1}(m_1) \right]^2 + \dots \\
&= \frac{1}{1 - \sum_{m_1=1}^{\infty} p(S_e(m_1 + 1))f_{M_1}(m_1)}, \quad (2.17)
\end{aligned}$$

where $p(S_e(m_1 + 1))$ is given by (2.10).

Since $1/R_{av}$ is the average portion of a given packet that is received successfully in each transmission and G is the average number of packets generated during each time slot, the throughput, T , is given by

$$T = \frac{G}{R_{av}}. \quad (2.18)$$

Fig. 2.10 shows the throughput for conventional slotted DS/SSMA ALOHA and slotted DS/SSMA ALOHA with combining. As can be seen, combining results in a higher throughput which is monotonically increasing for a large range of offered load.

2.4.2 DS/SSMA Slotted ALOHA With FEC

When FEC is applied to a slotted DS/SSMA ALOHA system, code combining can be used to improve the performance of the system. This is implemented here with a convolutional code and a Viterbi decoder. After receiving i copies of the same packet, the decoder combines all i copies and attempts to decode this composite packet. Let $D_e^{(i)}$, $D_u^{(i)}$ and $D_c^{(i)}$ denote the events that after receiving i copies, 'the decoded sequence contains detectable errors', 'the decoded sequence contains undetectable errors' and 'the decoded sequence contains no errors', respectively. Since all possible events have been accounted for

$$p(D_e^{(i)}) + p(D_u^{(i)}) + p(D_c^{(i)}) = 1. \quad (2.19)$$

$p(D_u^{(i)})$ can be neglected due to the use of a suitable CRC code. Therefore, we have

$$p(D_e^{(i)}) + p(D_c^{(i)}) \simeq 1. \quad (2.20)$$

Figure 2.10: Throughput for conventional slotted DS/SSMA ALOHA and slotted DS/SSMA ALOHA with combining (no FEC).

An upper bound on the error event probability of the Viterbi decoder for a rate $1/Q$ convolutional code is given by [22]

$$p(E)^{(Q)} \leq \sum_{d=d_{free}Q}^{\infty} a_d^{(Q)} P_d. \quad (2.21)$$

where $d_{free}Q$ is the free distance of the rate $1/Q$ convolutional code and $a_d^{(Q)}$ is the number of codewords of weight d . For the additive white Gaussian noise (AWGN) channel, P_d is given by [22]

$$P_d = \frac{1}{2} \text{erfc}(\sqrt{d \times SNR}). \quad (2.22)$$

Assume that Q copies of the same packet are available at the receiver. Also, assume that m_i simultaneous packets were transmitted in the channel during transmission of the i th, $i = 1, 2, \dots, Q$, copy. When these Q copies are combined together, they form a packet of a rate $1/Q$ convolutional code. For this rate $1/Q$ code, P_d , conditioned on m_i , $i = 1, 2, \dots, Q$, is given by

$$P_d(m_1, m_2, \dots, m_Q) = \frac{1}{2} \text{erfc}\left(\sqrt{d \times \frac{1.5NQ}{m_1 + m_2 + \dots + m_Q - Q}}\right), \quad (2.23)$$

Eq. (2.23) is derived in Appendix I. The packet error probability, conditioned on $m_1, m_2, \dots, m_Q$, is given by

$$\begin{aligned} p(D_e^{(Q)}(m_1, m_2, \dots, m_Q)) &= 1 - (1 - p(E^{(Q)}(m_1, m_2, \dots, m_Q)))^L, \\ &\leq 1 - \left(1 - \frac{1}{2} \sum_{d=d_{free}Q}^{\infty} a_d^{(Q)} \text{erfc}\left(\sqrt{d \frac{1.5NQ}{m_1 + m_2 + \dots + m_Q - Q}}\right)\right)^L, \end{aligned} \quad (2.24)$$

where L is the number of bits per packet. Since the right hand side of (2.24) is only a function of $m_1 + m_2 + \dots + m_Q$, we have

$$p(D_e^{(Q)}(m_1, m_2, \dots, m_Q)) = p(D_e^{(Q)}(m_1 + m_2 + \dots + m_Q)). \quad (2.25)$$

The average number of transmissions, with the condition that m_i other users were transmitting in the i th, $i = 1, 2, \dots$, slot, is then

$$\begin{aligned}
 R_{av}(m_1, m_2, \dots) = & \\
 & 1 + p(S_e(m_1 + 1)) + p(S_e(m_1 + 1), S_e(m_2 + 1), D_e^{(2)}(m_1 + 1, m_2 + 1)) + \\
 & p(S_e(m_1 + 1), S_e(m_2 + 1), D_e^{(2)}(m_1 + 1, m_2 + 1), S_e(m_3 + 1), D_e^{(3)}(\\
 & m_1 + 1, m_2 + 1, m_3 + 1)) + \dots, \tag{2.26}
 \end{aligned}$$

where

$$\begin{aligned}
 & p(S_e(m_1 + 1), S_e(m_2 + 1), D_e^{(2)}(m_1 + 1, m_2 + 1), \dots, S_e(m_Q + 1), D_e^{(Q)}(\\
 & m_1 + 1, m_2 + 1, \dots, m_Q + 1)) \\
 & \leq p(D_e^{(Q)}(m_1 + 1, m_2 + 1, \dots, m_Q + 1)). \tag{2.27}
 \end{aligned}$$

Using (2.25) and (2.27), $R_{av}(m_1, m_2, \dots)$ can be bounded by

$$R_{av}(m_1, m_2, \dots) \leq 1 + p(S_e(m_1 + 1)) + \sum_{i=2}^{\infty} p(D_e^{(i)}(m_1 + m_2 + \dots + m_i + i)). \tag{2.28}$$

Using the fact that $m = m_1 + m_2 + \dots + m_i$, and averaging (2.28), R_{av} can be bounded by

$$R_{av} \leq 1 + \sum_{m_1=1}^{\infty} p(S_e(m_1 + 1)) f_{M_1}(m_1) + \sum_{i=2}^{\infty} \sum_{m=i}^{\infty} p(D_e^{(i)}(m + i)) f_{M_i}(m), \tag{2.29}$$

where $f_{M_i}(m)$ is given by (2.15). It should be mentioned that for the conventional system with a rate $1/Q$ code, the average number of transmissions to transmit a packet successfully, $R_{av_{con}}$, is given by

$$R_{av_{con}} = 1 + \sum_{m_1=1}^{\infty} p(D_e^Q(m_1 + 1)) f_{M_1}(m_1) + [\sum_{m_1=1}^{\infty} p(D_e^Q(m_1 + 1)) f_{M_1}(m_1)]^2 + \dots$$

$$= \frac{1}{1 - \sum_{m_1=1}^{\infty} p(D_c^Q(m_1 + 1))f_{M_1}(m_1)}, \quad (2.30)$$

where

$$p(D_c^Q(m_1 + 1)) = p(D_c^Q(m_1 + 1, \dots, m_1 + 1)), \quad (2.31)$$

which can be calculated using (2.24). Using (2.29) and (2.30) in (2.18) provides the throughput, T .

It should be noted that a CRC code of k_c bits must be added to each packet for error detection. This code is used only to determine if the packet has been decoded correctly. Also, M '0', or known bits, (where M is the memory of the convolutional code) must be added to the end of each packet to terminate the trellis at a known state. Therefore, the throughput is $T \frac{k}{k+k_c+M}$, where k is the number of information bits per packet. Since $k \gg k_c + M$, the factor $\frac{k}{k+k_c+M}$ can be ignored¹.

2.5 Performance Examples

In this section, examples are given to illustrate the performance improvements possible with packet combining. They are based on either analysis alone or on both analytical and simulation results. In the latter case, the packet error probability conditioned on the number of users, (2.24), is evaluated by means of simulation. These simulation results are used in (2.29) and (2.18) to calculate the throughput.

1

The comparison between slotted DS/SSMA ALOHA using code combining and the conventional system is not significantly affected by neglecting this redundancy of $(k_c + M)$ bits, since it must be added to both systems. In addition, typical values are between 16 and 32 bits for k_c , 500 for k , and 6 for M . The overhead due to $k_c + M$ for these values is less than 8 percent.

Simulation results were obtained for a minimum of 100 packets in error and at least 1000 transmitted packets. The packet error probability was evaluated by simulation because for low SNRs, the bound on the error probability, (2.21), and therefore (2.24), is not sufficiently tight. This bound could result in an inaccurate value for the throughput. Note that (2.21) is valid only for infinitely quantized received data, while the simulations employed 3 bit (8-level) receiver quantization, as in [23]. It has been shown that, for an AWGN channel, there is a 0.20-0.25 dB loss associated with 8-level receiver quantization compared to infinite quantization [23].

For the analysis, the weight spectra of the (2,1,4) and (4,1,4) convolutional codes are needed. The weight spectra of the (2,1,4) code with generator polynomials (23,35) are given in [24] and [6]. The weight spectra of the codes based on the (4,1,4) convolutional code with generator polynomials (23,35,27,33) are given in Table 2.1. It should be noted that the rate $1/2$ (2,1,4) code is the same as the rate $1/2$ code based on the (4,1,4) code.

In the following examples, unless otherwise indicated, all packets consist of 500 bits. This was done to provide a fair means of comparison. Thus, for conventional slotted DS/SSMA ALOHA with a rate $1/2$ code, each packet carries only 250 information bits. With code combining, each packet carries 500 information bits, so that the effective code rate is $1, 1/2, 1/3, 1/4, \dots$ depending on the number of retransmissions for a given packet. The number of chips per bit is $N = 63$. It should be noted that for slotted DS/SSMA ALOHA with code combining, the effective throughput is the same as (2.18), while for conventional slotted DS/SSMA ALOHA, (2.18) should be multiplied by the code rate ($1/2$ in our case) to obtain T_{eff} .

Fig. 2.11 shows the throughput for the following cases. The solid and dotted lines are the simulation and analytical results, respectively, for the lower bound on the

rate	generator polynomial	d_{free}	$(a_{d_{free}+l} \quad l = 0, 1, 2, \dots, 17)$ $[c_{d_{free}+l} \quad l = 0, 1, 2, \dots, 17]$
1/3	(23,35,27)	11	$(1, 2, 3, 2, 3, 8, 12, 28, 52, 62, 86, 156, 303,$ $547, 860, 1331, 2114, 3591)$ $[1, 6, 9, 6, 13, 32, 60, 146, 294, 386, 562,$ $1330, 2283, 4326, 7322, 11886, 19878, 35436]$
1/4	(23,35,27,33)	15	$(2, 1, 0, 1, 2, 4, 3, 9, 9, 13, 23, 21, 44, 53, 86,$ $128, 166, 271)$ $[4, 2, 0, 4, 6, 12, 13, 40, 43, 66, 119, 118, 254,$ $336, 566, 874, 1216, 2020]$
1/5	(23,35,27,33,23)	18	$(1, 0, 2, 0, 1, 0, 6, 0, 6, 0, 13, 0, 25, 0, 44, 0,$ $79, 0)$ $[1, 0, 5, 0, 4, 0, 19, 0, 25, 0, 58, 0, 132, 0,$ $229, 0, 487, 0]$
1/6	(23,35,27,33,23,35)	22	$(1, 0, 2, 0, 2, 0, 0, 0, 6, 0, 6, 0, 14, 0, 21, 0,$ $31, 0)$ $[1, 0, 5, 0, 7, 0, 0, 0, 22, 0, 23, 0, 61, 0, 112, 0,$ $175, 0]$
1/7	(23,35,27,33,23,35,27)	26	$(1, 0, 2, 0, 1, 0, 2, 0, 1, 0, 5, 0, 6, 0, 12, 0,$ $16, 0)$ $[1, 0, 5, 0, 4, 0, 6, 0, 2, 0, 17, 0, 27, 0, 54, 0,$ $84, 0]$
1/8	(23, 35, 27, 33, 23, 35, 27, 33)	30	$(2, 0, 1, 0, 0, 0, 1, 0, 2, 0, 4, 0, 3, 0, 9, 0,$ $9, 0)$ $[4, 0, 2, 0, 0, 0, 4, 0, 6, 0, 12, 0, 13, 0, 40, 0,$ $43, 0]$
1/9	(23, 35, 27, 33, 23, 35, 27, 33, 23)	33	$(1, 0, 1, 1, 0, 0, 0, 1, 0, 0, 2, 3, 1, 1, 2, 2,$ $1, 5)$ $[1, 0, 3, 2, 0, 0, 0, 4, 0, 0, 6, 10, 3, 2, 10, 8,$ $5, 22]$
1/10	(23, 35, 27, 33, 23, 35, 27, 33, 23, 35)	37	$(1, 0, 1, 1, 0, 0, 0, 1, 1, 0, 0, 0, 1, 3, 1, 2,$ $2, 2)$ $[1, 0, 3, 2, 0, 0, 0, 4, 3, 0, 0, 0, 3, 8, 5, 10,$ $8, 6]$

Table 2.1: Weight spectra of the repeated convolutional code based on the (4,1,4) code with generator polynomial (23,35,27,33)

throughput of the slotted DS/SSMA ALOHA system using the (4,1,4) convolutional code and code combining. The dashed lines show the throughput of the conventional slotted DS/SSMA ALOHA system employing the (2,1,4) convolutional code. As can be seen, combining the repeated copies of the same packet results in a throughput which continues to increase as the offered load increases. As discussed previously, this is an important factor in maintaining system stability. It should be mentioned that when the error probability is high, the upper bound becomes pessimistic, so simulation results are given in addition to the analytical results.

In Fig. 2.11, the throughput curve for code combining and a small offered load is higher when compared to the conventional system curve. This is due to the fact that for the conventional system, each transmitted packet is a codeword of a rate $1/2$ code, while with code combining, no FEC is provided for error correction in the first transmission. With code combining and a low offered load, most of the packets are delivered to the user without having to transmit any redundancy. For the conventional system, the redundancy due to FEC is always transmitted, even if it is not needed.

Note that there are local minima and maxima in the analytical curves of Fig. 2.11. These can be explained as follows. Recall that to evaluate the throughput, the error probability (2.21) for different rate codes must be evaluated. As mentioned previously, when this error probability is high, (2.21) and therefore (2.24) are not very tight and rather pessimistic. Therefore, the local minima are due to the fact that the theoretical bit error probability, and therefore the theoretical packet error probability, for one of the code rates is evaluated quite pessimistically. In other words, the minima in the theoretical curve, of Fig. 2.11 occur for the same reason that the curves in Fig. 2.8 fall more quickly than the corresponding curves in Fig.

Figure 2.11: Analytical and simulated throughput for conventional slotted DS/SSMA ALOHA using a (2,1,4) convolutional code, and slotted DS/SSMA ALOHA using a (4,1,4) convolutional code with code combining.

2.7. Therefore, these minima can be considered the image of the descent of the curves in Fig. 2.8. The reason that the analytical curves in some portions of Fig. 2.11 are better than the simulation curves, is due to the (8 level) quantization error for simulations, compared to the infinite level quantization for the analytical curves. Thus the superiority of the infinite level quantization over 8 level quantization is evident whenever the bound in (2.21) is tight. This occurs at a different location for each code rate.

Fig. 2.12 compares the analytical throughput of the system using code combining with the (4,1,4) and (2,1,2) convolutional codes. For the latter code, as explained in Section II, the transmitter sends $P_1(x)$ and $P_2(x)$ alternately. With the former code, the transmission sequence cycles through $P_1(x)$, $P_2(x)$, $P_3(x)$ and $P_4(x)$ in order. Fig. 2.12 shows that using the (4,1,4) code results in a higher throughput than using the (2,1,2) code.

The following example considers the case where each packet contains 1000 bits, as in [4]. This allows for a comparison of the above system with the narrow-band slotted ALOHA system. The corresponding throughput is shown in Fig. 2.13. To obtain meaningful results, the two systems should be compared in terms of the same total bandwidth. Since $N = 63$, the throughput of the above system should be compared with 63 narrow-band slotted ALOHA systems.

The network capacity is defined as the maximum achievable effective throughput. As in [4], define the capacity/bandwidth factor, β , as

$$\beta = \frac{T_{effmax}}{T_{nb} \times N} \quad (2.32)$$

where $T_{nb} = 1/\epsilon$ is the throughput with narrow-band slotted ALOHA. From Fig.

Figure 2.12: Comparison of analytical throughput for slotted DS/SSMA ALOHA using (4,1,4) and (2,1,4) convolutional codes, both with code combining.

Figure 2.13: Analytical throughput for conventional slotted DS/SSMA ALOHA using a (2,1,4) convolutional code, and slotted DS/SSMA ALOHA using a (4,1,4) convolutional code and code combining. In both cases, each packet contains 1000 bits.

2.13. T_{effmax} is approximately 43 for an offered load of 180 (although the throughput is still increasing at that point), so that β is almost 1.9. This means that with the same bandwidth, the throughput of slotted DS/SSMA ALOHA with code combining is 1.9 times of that of narrow-band slotted ALOHA. It should be mentioned that the code used here is not a very powerful code. Increasing the constraint length from 5 to 7, or higher, will increase β .

2.6 Conclusions

This chapter considered a slotted DS/SSMA ALOHA system with and without FEC, with a receiver which combines the available copies of a packet. When FEC is used, code combining is applied. Code combining allows for the combination of all available copies of a packet at the receiver, resulting in a lower code rate for that specific packet. Analytical and simulation results have been presented which show that by applying code combining, a higher throughput is achievable than that without code combining. Code combining also causes the effective throughput to be a non-decreasing function of the offered load (at least for a large range of the offered load). This is important from the point of view of system stability.

It was shown that the throughput of slotted DS/SSMA ALOHA with code combining can be 1.9 times that of narrow-band slotted ALOHA, considering the same bandwidth for both cases. This factor can be increased by using a more powerful error correcting code.

Chapter 3

Slotted DS/SSMA ALOHA with Packet Combining in a Rayleigh Fading Channel

3.1 Introduction

A packet radio network (PRN) can be a collection of mobile terminals that communicate with each other through radio links. In such a network, the data is transmitted using a packet format. In this chapter, a DS/SSMA ALOHA PRN is considered in a mobile environment. The channel is modeled as a Rayleigh fading channel and a RAKE receiver with maximum ratio combiner (MRC), as explained in Section 3.2.2, is considered.

If a received copy of a packet contains uncorrectable errors, the receiver of a conventional DS/SSMA ALOHA PRN discards that copy and asks for the retransmission of the same packet. In this chapter, the throughput of the slotted DS/SSMA ALOHA system is enhanced by increasing the order of diversity in the MRC of the receiver. This is achieved by retaining the copies of a packet which were detected in error and combining them with the newly received copy. While using each copy

alone might not enable the receiver to recover the transmitted packet, combining these copies enhances the probability of recovery.

The outline of this chapter is as follows. Section 3.2 characterizes a DS/SS in a Rayleigh fading channel. RAKE-RAKE receiver, which combines several received copies of a packet, is introduced in Section 3.3. The system under consideration is described in Section 3.4. The throughput and the average number of transmissions required to transmit a packet successfully are derived in Section 3.5. Section 3.6 includes some performance examples and finally Section 3.7 presents some conclusions.

3.2 Fading Channels

One of the channel types which frequently occurs in radio communications is the fading channel. One common cause of fading is the multipath nature of propagation media. A multipath propagation for a scattering medium is illustrated in Fig. 3.1. In this case a large number of randomly moving scatterers are located within the propagation path, causing the signal to arrive at the receiver via several time varying paths. These individual paths cannot be distinguished at the receiver and so are all added together. Thus, the signal becomes a replica of the transmitted one with random amplitude and phase. Fading can severely degrade the performance of the communication system. This section characterizes multipath fading channels, and describes the elements necessary for the next sections, which examines the performance of DS/SSMA ALOHA in a Rayleigh fading channel.

Figure 3.1: Multipath fading channel.

3.2.1 Multipath Fading Channel Characterization

A time-varying multipath fading channel can be characterized by its equivalent lowpass impulse response given by [25]

$$h(\tau; t) = \sum_i \alpha_i(t) e^{-j2\pi f_c \tau_i(t)} \delta(\tau - \tau_i(t)), \quad (3.1)$$

where $\alpha_i(t)$ is the attenuation factor for the signal received on the i th path and $\tau_i(t)$ is the propagation delay for the i th path.

For a multipath fading channel, the equivalent low-pass impulse response, $h(\tau; t)$ is a complex-valued zero mean Gaussian random process in the t variable. Assuming that $h(\tau; t)$ is wide-sense-stationary its autocorrelation function is defined by

$$R_h(\tau_1, \tau_2; \Delta t) = \frac{1}{2} E[h(\tau_1; t)h^*(\tau_2; t + \Delta t)], \quad (3.2)$$

where $h(\tau_1; t)$ and $h(\tau_2; t + \Delta t)$ are the impulse responses corresponding to two path delays, τ_1 and τ_2 , respectively. Assume that the scattering at two paths, with corresponding delays τ_1 and τ_2 , is uncorrelated. This is called uncorrelated scattering and is a reasonable assumption for most fading channels [26].

The autocorrelation function for uncorrelated scattering can be expressed as

$$R_h(\tau_1, \tau_2; \Delta t) = R_h(\tau_1; \Delta t)\delta(\tau_1 - \tau_2), \quad (3.3)$$

where $R_h(\tau; \Delta t)$ represents the autocorrelation with time lag Δt for a path delay τ . The value of $R_h(\tau; \Delta t)$ at $\Delta t = 0$ ($R_h(\tau; 0)$), describes the average output power of the channel for a relative path delay τ . The function $R_h(\tau; 0)$ is called the multipath intensity profile or the delay power spectrum of the channel [25] [26].

The multipath spread, T_{max} , is the range of values of the path delay τ for which the multipath intensity profile, $R_h(\tau; 0)$, is non-zero.

3.2.2 DS/SS in a Rayleigh Fading Channel

As was mentioned in Section 1.2.1, at the transmitter of a DS/SS system, the information-bearing signal is multiplied by the spreading code. Thus, each information bit is divided into a number of small increments called chips. Assume a spread spectrum waveform is transmitted over a Rayleigh fading channel with delay spread larger than the chip duration. The received signal is the summation of the replicas of the transmitted waveform arriving from different paths with random delays and phases. It is this summation that causes the signal to fade. It can be assumed

without loss of generality that the first path arrives with zero delay. In a DS/SS system, paths with a delay longer than the chip duration, T_c , will not take part in the summation and therefore do not contribute to the fading of the signal. Thus, when a path delay is greater than the chip duration there is a negligible degradation due to that specific path. In general, however, it is possible to do better. The paths with delay $T_c \leq \tau \leq 2T_c$, effectively form another channel output which is independent of the first one. Furthermore, paths with delay $iT_c \leq \tau \leq (i+1)T_c$, where $i = 2, 3, \dots, L-1$, form another $L-2$ independent channel outputs. Thus, the channel can be modeled as a frequency selective channel with an L tap delay line model as shown in Fig. 3.2. For all practical purposes L is given by [27]

$$L = \lfloor \frac{T_{max}}{T_c} \rfloor + 1. \quad (3.4)$$

The L tap delay line channel model provides L replicas of the same transmitted signal at the receiver. Assume that each channel (among the L channels) can be modeled as a Rayleigh fading channel. It has been shown that a RAKE receiver with diversity combining techniques, such as equal gain combining (EGC) and MRC, can provide better performance than a multipath rejection receiver [28] [29]. A multipath rejection receiver attempts to demodulate the first arriving path, which corresponds to the path with the largest average received SNR. The RAKE receivers with EGC and MRC are shown in Figs. 3.3 and 3.4, respectively, where x^* is the complex-conjugate of x and $a(t)$ is the spreading signal. In the case of MRC, the channel tap weights at the receiver must be determined.

Figure 3.2: Tap delay line channel model of a DS/SS system.

Figure 3.3: The conventional RAKE receiver with an equal-gain combiner.

Figure 3.4: The conventional RAKE receiver with a maximum-ratio combiner.

3.3 RAKE-RAKE receiver in a DS/SSMA ALOHA

In this section a receiver that retains the copies of a packet which were detected in error and combines them with the newly received copy is introduced. For a conventional DS/SSMA ALOHA, regardless of the number of copies of a packet which have previously been transmitted, the receiver always has the structure given in Fig 3.4.

The proposed RAKE-RAKE receiver with d copies of a packet available is shown in Fig 3.5, where x^* is the complex-conjugate of x and $a_j(t)$ is the spreading signal of j th user. Each packet copy can be considered as an additional L paths to be used in the maximum ratio combiner. Therefore, combining d copies of a packet increases the diversity order to $L \times d$, from L for the conventional system.

As mentioned in Section 3.2.2, when the path gains are not available at the receiver, an EGC can be used to form a RAKE receiver in order to demodulate the transmitted signal. Using EGC, the modified RAKE receiver which combines all received copies of a packet is given in Fig. 3.6. Since some of the resolvable paths at the receiver have very small average SNR, one would suspect that considering these paths in the EGC-RAKE receiver would degrade the performance of the receiver (by adding more noise to the received signal). Therefore, it is reasonable to think that there must be an optimum number of paths (depending on the channel specifications) that should be combined at the EGC-RAKE receiver to give the best performance. Analyzing the performance of EGC-RAKE receiver in a DS/SSMA slotted ALOHA system and calculating the optimum number of paths to be combined together is an interesting topic for further research.

Figure 3.5: The slotted DS/SSMA ALOHA receiver with maximum-ratio combiner and packet combining.

Figure 3.6: The slotted DS/SSMA ALOHA receiver with equal gain combiner and packet combining.

3.4 System Description

The slotted DS/SSMA ALOHA system being considered has spreading factor N , chip duration T_c and bit duration $T_b = NT_c$. It is assumed that each packet consists of N_p bits. BPSK modulation is employed with FEC which can correct up to t errors in each packet.

As explained in Chapter 2.3, it is assumed that the channel is used by a large number of users, n . Each user generates packets of length N_p at a rate of I_i packets per time slot [4]. Therefore, the channel receives newly generated packets at a rate I (input rate), where

$$I = \sum_{j=1}^n I_j, \quad (3.5)$$

and I has a Poisson distribution.

We assume that the retransmission of packets received in error is also Poisson distributed, with a rate of H packets per time slot. The composite packet arrival (offered load) distribution is then Poisson with rate $G = I + H$. The composite arrival distribution for a given slot, $f_M(m)$, is given by

$$f_M(m) = \frac{G^m}{m!} e^{-G}. \quad (3.6)$$

It is assumed that packet arrivals in different time slots are independent identically distributed (iid) Poisson random variables.

The channel is characterized as a Rayleigh fading channel with an exponential

multi-path intensity profile (MIP) given by

$$Q(\tau) = \begin{cases} \frac{\sigma^2}{T_m} e^{-\tau/T_m}, & 0 \leq \tau \leq T_{max} \\ 0, & \text{otherwise,} \end{cases} \quad (3.7)$$

where T_{max} is the multipath spread. The MIP gives the average power output of the channel as a function of delay. For all practical purposes, the channel can be modeled as $L = \lfloor T_{max}/T_c \rfloor + 1$ resolvable paths [27], where $\lfloor y \rfloor$ is the largest integer smaller than or equal to y . It is assumed that the path gains are constant during each data bit duration and uncorrelated from bit to bit. In DS/SSMA ALOHA, if the transmitter is unable to recover the transmitted packet, another copy of the packet is retransmitted. Therefore, several copies of a packet might be transmitted before it is received successfully. Define the following parameter $\cdot$ for the z th copy of the j th user signal:

$b_j(t)$: user data signal,

$a_j(t)$: user spreading sequence,

$\beta_{lj}^{(z)}$: path gain for the l th path,

$\tau_{lj}^{(z)}$: time delay for the l th path,

$\phi_{lj}^{(z)}$: phase for the l th path (3.8)

It is assumed that $\beta_{lj}^{(z)}$ is a Rayleigh distributed random variable, $\phi_{lj}^{(z)}$ is a uniform random variable in $[0, 2\pi)$, and $\tau_{lj}^{(z)}$ is a uniform random variable in $[0, T)$ where T is the symbol duration. The multi-path power of each path can be approximated by

$$E\{\beta_{lj}^{(z)^2}\} = \int_{(z-1)T_c}^{zT_c} Q(\tau) = \sigma^2 [e^{-\tau T_c/T_m} - e^{-(z-1)T_c/T_m}] \quad (3.9)$$

It should be mentioned that average power control is assumed, so that, on average, all user signals arrive at the receiver with the same power

The receiver, as illustrated in Fig 3.5, is a modified RAKE receiver, called a RAKE-RAKE receiver, that uses all available copies of a packet at the receiver.

3.5 Performance Analysis

Consider BPSK modulation and define γ_b as half of the instantaneous signal-to-noise plus interference ratio. The conditional bit error rate (BER), conditioned on γ_b , is given by

$$P_b(\gamma_b) = \frac{1}{2} \text{erfc}(\sqrt{\gamma_b}), \quad (3.10)$$

where

$$\text{erfc}(x) = \frac{2}{\sqrt{\pi}} \int_x^{\infty} e^{-t^2} dt. \quad (3.11)$$

A Gaussian approximation is used to calculate the performance. The BER is calculated by first finding the probability density function (pdf) of the signal to noise plus interference ratio, and then averaging it over (3.10). This BER is then used to derive expressions for the throughput and the average number of transmissions to transmit a packet successfully.

3.5.1 BER with one copy

In this section, the BER is derived for the case when a single copy of the received packet is available. The derivation is similar to the analysis in [27]. It should be

mentioned that in this chapter, the superscript is dropped whenever it equals one, i.e., $y^{(1)} = y$. The received signal, $r(t)$, is given by

$$r(t) = A \sum_{l=1}^{\infty} \beta_{l1} a_1(t - \tau_{l1}) b_1(t - \tau_{l1}) \cos(\omega_c t + \phi_{l1}) + A \sum_{l=1}^{\infty} \sum_{k=2}^K \beta_{lk} a_1(t - \tau_{lk}) b_k(t - \tau_{lk}) \cos(\omega_c t + \phi_{lk}) + n(t), \quad (3.12)$$

where $n(t)$ is white Gaussian noise with 2-sided power spectrum density $N_0/2$, and the other parameters are given in (3.8). Using a RAKE receiver, the decision variable for the j th path is given by

$$\xi_j = \int_{\tau_{j1}}^{T+\tau_{j1}} r(t) a_1(t - \tau_{j1}) \beta_{j1} \cos(\omega_c t + \phi_{j1}) dt, \quad (3.13)$$

$$\begin{aligned} \xi_j = & b_0^1 \frac{AT}{2} \beta_{j1}^2 + \frac{A}{2} \beta_{j1} \sum_{\substack{l=1 \\ l \neq j}}^{\infty} \beta_{l1} \cos(\phi_{l1} - \phi_{j1}) \int_{\tau_{j1}}^{T+\tau_{j1}} a_1(t - \tau_{l1}) b_1(t - \tau_{l1}) a_1(t - \tau_{j1}) dt + \\ & \frac{A}{2} \beta_{j1} \sum_{l=1}^{\infty} \sum_{k=2}^K \beta_{lk} \cos(\phi_{lk} - \phi_{j1}) \int_{\tau_{j1}}^{T+\tau_{j1}} a_k(t - \tau_{lk}) b_k(t - \tau_{lk}) a_1(t - \tau_{j1}) dt + \\ & \beta_{j1} \int_{\tau_{j1}}^{T+\tau_{j1}} n(t) a_1(t - \tau_{j1}) \cos(\omega_c t + \phi_{j1}) dt. \end{aligned} \quad (3.14)$$

$$\begin{aligned} \xi_j = & b_0^1 \frac{AT}{2} \beta_{j1}^2 + \frac{A}{2} \beta_{j1} \sum_{\substack{l=1 \\ l \neq j}}^{\infty} \beta_{l1} \cos(\phi_{l1} - \phi_{j1}) [b_{-1}^1 R_{11}(t_{l1}) + b_0^1 \dot{R}_{11}(t_{l1})] + \\ & \frac{A}{2} \beta_{j1} \sum_{l=1}^{\infty} \sum_{k=2}^K \beta_{lk} \cos(\phi_{lk} - \phi_{j1}) [b_{-1}^k R_{k1}(t_{lk}) + b_0^k \dot{R}_{k1}(t_{lk})] + \beta_{j1} \nu. \end{aligned} \quad (3.15)$$

where $t_{lk} = \tau_{lk} - \tau_{j1}$ and ν is a zero mean Gaussian random variable with variance $N_0T/4$, and

$$\begin{aligned} R_{k1}(\tau) &= \int_0^\tau a_k(t - \tau)a_1(t)dt, \\ \hat{R}_{k1}(\tau) &= \int_0^\tau a_k(t - \tau)a_1(t)dt. \end{aligned} \quad (3.16)$$

Pursley [21] has shown that for Gold codes

$$\epsilon^2 = E\left\{\frac{[\alpha_{-1}R(t_1) + \alpha_0\hat{R}(t_1)]^2}{T^2}\right\} = \frac{2}{3N}, \quad (3.17)$$

where $\alpha_i = \pm 1$, $i = -1, 0$, are independent binary random variables. Assuming that the RAKE receiver uses L paths in recovering the data and $LK \gg 1$, the decision variable is given by

$$x = \sum_{j=1}^L \xi_j. \quad (3.18)$$

The mean and variance of the decision variable are given by

$$\mu_x = \frac{AT}{2} \sum_{j=1}^L \beta_{j1}^2, \quad (3.19)$$

$$\sigma_x^2 = \left(\frac{AT}{2}\right)^2 \epsilon^2 K \sum_{j=1}^L \beta_{j1}^2 \sum_{l=1}^{\infty} E\{\beta_{l1}^2\}/2 + \frac{N_0T}{4} \sum_{j=1}^L \beta_{j1}^2, \quad (3.20)$$

$$= \left(\frac{AT}{2}\right)^2 \epsilon^2 K \sum_{j=1}^L \beta_{j1}^2 \sigma^2/2 + \frac{N_0T}{4} \sum_{j=1}^L \beta_{j1}^2, \quad (3.21)$$

where σ^2 is given by (3.7). Using $E_b = A^2T/2$ results in

$$\gamma_b^{(1)} = \frac{1}{2} \frac{S}{N + I} = \frac{E_b \sum_{j=1}^L \beta_{j1}^2}{E_b \epsilon^2 K \sigma^2 + N_0}. \quad (3.22)$$

for $\gamma_b^{(1)}$, which is defined as half of the signal to noise plus interference power.

Equation (3.22) can be written as

$$\gamma_b^{(1)} = \frac{E}{N} \sum_{j=1}^L \beta_{j1}^2 = \sum_{j=1}^L \gamma_j. \quad (3.23)$$

where

$$\frac{E}{N} = \frac{E_b}{E_b \epsilon^2 K \sigma^2 / 2 + N_o}. \quad (3.24)$$

The pdf of γ_{b1} is given by Proakis [25]

$$p(\gamma_b^{(1)}) = \sum_{j=1}^L \frac{\pi_j}{\bar{\gamma}_j} e^{\gamma_b^{(1)}/\bar{\gamma}_j}, \quad (3.25)$$

where

$$\bar{\gamma}_j = \frac{E}{N} E\{\beta_{j1}^2\}. \quad (3.26)$$

$$\pi_j = \prod_{\substack{i=1 \\ i \neq j}}^L \frac{\bar{\gamma}_j}{\bar{\gamma}_j - \bar{\gamma}_i}. \quad (3.27)$$

Averaging (3.10) over (3.25), the bit error probability conditioned on K , the number of simultaneous users during the transmission of the packet, is given by [25]

$$P_b^{(1)}(K) = \frac{1}{2} \sum_{j=1}^L \pi_j \left[1 - \sqrt{\frac{\bar{\gamma}_j}{1 + \bar{\gamma}_j}} \right]. \quad (3.28)$$

3.5.2 BER with two copies of the signal available

In this section, the BER is calculated for the case when two copies of each packet (or bit) are available at the receiver. It is assumed that during the transmission of each packet, K simultaneous users are also transmitting. The decision variable for the j th path and i th copy, $i = 1, 2$ is given by

$$\xi_j^{(i)} = \int_0^T r^{(i)}(t) a_1(t - \tau_{j1}^{(i)}) \beta_{j1}^{(i)} \cos(\omega_c t + \phi_{j1}^{(i)}) dt. \quad (3.29)$$

Using a RAKE-RAKE receiver which combines the two received copies of each bit (with L paths for each copy), the decision variable is given by

$$x = \sum_{i=1}^2 \sum_{j=1}^L \xi_j^{(i)}. \quad (3.30)$$

Following (3.13)-(3.21) and using the fact that the two copies are statistically independent, the mean and variance of the decision variable are given by

$$\mu_r = \frac{AT}{2} \sum_{j=1}^L [\beta_{j1}^{(1)^2} + \beta_{j1}^{(2)^2}]. \quad (3.31)$$

$$\sigma_r^2 = \left(\frac{AT}{2}\right)^2 \epsilon^2 K \sigma^2 / 2 \left[\sum_{j=1}^L \beta_{j1}^{(1)^2} + \sum_{j=1}^L \beta_{j1}^{(2)^2} \right] + \frac{N_o T}{4} \sum_{j=1}^L [\beta_{j1}^{(1)^2} + \beta_{j1}^{(2)^2}], \quad (3.32)$$

respectively. Therefore, half of the signal-to-noise plus interference ratio, $\gamma_b^{(2)}$, is given by

$$\gamma_b^{(2)} = \frac{1}{2} \frac{S}{N + I} = \frac{E}{N} \sum_{j=1}^L [\beta_{j1}^{(1)^2} + \beta_{j1}^{(2)^2}] = \sum_{j=1}^L [\gamma_{j1} + \gamma_{j2}] = [\gamma_{b1} + \gamma_{b2}], \quad (3.33)$$

where E/N is given by (3.24). γ_{b1} and γ_{b2} are two independent identically distributed (iid) random variables with pdfs given by (3.25). Therefore the right side of (3.33) has the following pdf

$$p(\gamma_b^{(2)}) = p(\gamma_b^{(1)}) * p(\gamma_b^{(1)}). \quad (3.34)$$

where $*$ denotes convolution and $p(\gamma_b^{(1)})$ is given by (3.25). After some mathematical manipulations, $p(\gamma_b^{(2)})$ is given by

$$\begin{aligned} p(\gamma_b^{(2)}) &= \sum_{j=1}^L \sum_{\substack{j'=1 \\ j' \neq j}}^L \frac{\pi_j \pi_{j'}}{\tilde{\gamma}_j - \tilde{\gamma}_{j'}} [e^{-\gamma_b/\tilde{\gamma}_j} - e^{-\gamma_b/\tilde{\gamma}_{j'}}] \\ &\quad + \sum_{j=1}^L \left(\frac{\pi_j}{\tilde{\gamma}_j}\right)^2 \gamma_b e^{-\gamma_b/\tilde{\gamma}_j}. \end{aligned} \quad (3.35)$$

Averaging (3.10) over (3.35), the bit error probability conditioned on K , the number of simultaneous users during the transmission of both the first and second packets, is given by

$$P_b^{(2)}(K) = \frac{1}{2} \sum_{j=1}^L \sum_{\substack{j'=1 \\ j' \neq j}}^L \frac{\pi_j \pi_{j'}}{\tilde{\gamma}_j - \tilde{\gamma}_{j'}} \left\{ \tilde{\gamma}_j \left[1 - \sqrt{\frac{\tilde{\gamma}_j}{1 + \tilde{\gamma}_j}} \right] - \tilde{\gamma}_{j'} \left[1 - \sqrt{\frac{\tilde{\gamma}_{j'}}{1 + \tilde{\gamma}_{j'}}} \right] \right\}$$

$$+ \sum_{j=1}^L \pi_j^2 \left(\frac{1 - \mu_j}{2} \right)^2 (2 + \mu_j), \quad (3.36)$$

where

$$\mu_j = \sqrt{\frac{\tilde{\gamma}_j}{1 + \gamma_j}}. \quad (3.37)$$

3.5.3 BER with three copies of the signal available

The BER when three copies of each packet are available at the receiver is now calculated. In this case, the RAKE-RAKE receiver combines the three copies in an attempt to recover the transmitted data. Using the same arguments as for two copies, half the signal-to-noise plus interference ratio, $\gamma_b^{(3)}$, is given by

$$\begin{aligned} \gamma_b^{(3)} &= \frac{1}{2} \frac{S}{N + I} = \frac{E}{N} \sum_{j=1}^L [\beta_{j1}^{(1)2} + \beta_{j1}^{(2)2} + \beta_{j1}^{(3)2}] = \sum_{j=1}^L [\gamma_{j1} + \gamma_{j2} + \gamma_{j3}] \\ &= [\gamma_{b1} + \gamma_{b2} + \gamma_{b3}], \end{aligned} \quad (3.38)$$

where E/N is given by (3.24). γ_{b1}, γ_{b2} and γ_{b3} are iid random variables with pdfs given by (3.25). Therefore we have

$$p(\gamma_b^{(3)}) = p(\gamma_b^{(1)}) * p(\gamma_b^{(1)}) * p(\gamma_b^{(1)}) = p(\gamma_b^{(2)}) * p(\gamma_b^{(1)}), \quad (3.39)$$

where $*$ denotes convolution, $p(\gamma_b^{(1)})$ is given by (3.25) and $p(\gamma_b^{(2)})$ is given by (3.35)

After some mathematical manipulations, $p(\gamma_b^{(3)})$ is given by

$$\begin{aligned} p(\gamma_b^{(3)}) &= 2 \sum_{j_1=1}^L \pi_{j_1} \tilde{\gamma}_{j_1} e^{\gamma_b^{(3)}/\gamma_{j_1}} \sum_{\substack{j_2=1 \\ j_2 \neq j_1}}^L \frac{\pi_{j_2}}{(\gamma_{j_1} - \gamma_{j_2})} \sum_{\substack{j_3=1 \\ j_3 \neq j_1}}^L \left(\frac{\pi_{j_3}}{(\gamma_{j_1} - \gamma_{j_3})} \right) + \\ &\quad 2 \sum_{j_1=1}^L \pi_{j_1} e^{\gamma_b^{(3)}/\gamma_{j_1}} \sum_{\substack{j_2=1 \\ j_2 \neq j_1}}^L \frac{\pi_{j_2} \tilde{\gamma}_{j_2}}{(\gamma_{j_1} - \gamma_{j_2})} \sum_{\substack{j_3=1 \\ j_3 \neq j_2}}^L \left(\frac{\pi_{j_3}}{(\gamma_{j_2} - \gamma_{j_3})} \right) + \end{aligned}$$

$$\begin{aligned}
& 3 \sum_{j_1=1}^L \frac{\pi_{j_1}^2 \bar{\gamma}_b^{(3)}}{\bar{\gamma}_{j_1}} e^{\gamma_b^{(3)}/\bar{\gamma}_{j_1}} \sum_{\substack{j_2=1 \\ j_2 \neq j_1}}^L \frac{\pi_{j_2}}{(\bar{\gamma}_{j_1} - \bar{\gamma}_{j_2})} + \\
& \sum_{j_1=1}^L \pi_{j_1} \bar{\gamma}_{j_1} e^{\gamma_b^{(3)}/\bar{\gamma}_{j_1}} \sum_{\substack{j_2=1 \\ j_2 \neq j_1}}^L \left(\frac{\pi_{j_2}}{(\bar{\gamma}_{j_1} - \bar{\gamma}_{j_2})} \right)^2 - \\
& \sum_{j_1=1}^L \pi_{j_1}^2 e^{\gamma_b^{(3)}/\bar{\gamma}_{j_1}} \sum_{\substack{j_2=1 \\ j_2 \neq j_1}}^L \frac{\pi_{j_2} \bar{\gamma}_{j_2}}{(\bar{\gamma}_{j_1} - \bar{\gamma}_{j_2})^2} + \\
& \frac{\gamma_b^{(3)^2}}{2} \sum_{j=1}^L \left(\frac{\pi_j}{\bar{\gamma}_j} \right)^3 e^{-\gamma_b^{(3)}/\bar{\gamma}_j}. \tag{3.40}
\end{aligned}$$

Averaging (3.10) over (3.25), the BER conditioned on K , the number of simultaneous users during the transmission of the first, second and third packets, is given by

$$\begin{aligned}
P_b^{(3)}(K) = & \sum_{j=1}^L \pi_j \bar{\gamma}_j^2 (1 - \mu_j) O_j^2 + \sum_{j_1=1}^L \pi_{j_1} \bar{\gamma}_{j_1} (1 - \mu_{j_1}) \sum_{\substack{j_2=1 \\ j_2 \neq j_1}}^L \frac{\pi_{j_2} \bar{\gamma}_{j_2}}{\bar{\gamma}_{j_1} - \bar{\gamma}_{j_2}} O_{j_2} + \\
& 3 \sum_{j=1}^L \pi_j^2 \bar{\gamma}_j \left(\frac{1 - \mu_j}{2} \right)^2 (2 + \mu_j) O_j + \frac{1}{2} \sum_{j=1}^L \pi_j \bar{\gamma}_j^2 (1 - \mu_j) Q_j - \\
& \frac{1}{2} \sum_{j=1}^L \pi_j^2 \bar{\gamma}_j (1 - \mu_j) M_j + \sum_{j=1}^L \pi_j^3 \left(\frac{1 - \mu_j}{2} \right)^3 \left(4 + \frac{9}{2} \mu_j + \frac{3}{2} \mu_j^2 \right), \tag{3.41}
\end{aligned}$$

where

$$O_j = \sum_{\substack{j'=1 \\ j' \neq j}}^L \frac{\pi_{j'}}{\bar{\gamma}_j - \bar{\gamma}_{j'}}. \tag{3.42}$$

$$Q_j = \sum_{\substack{j'=1 \\ j' \neq j}}^L \left(\frac{\pi_{j'}}{\hat{\gamma}_j - \hat{\gamma}_{j'}} \right)^2. \quad (3.43)$$

$$M_j = \sum_{\substack{j'=1 \\ j' \neq j}}^L \frac{\pi_{j'} \hat{\gamma}_{j'}}{(\hat{\gamma}_j - \hat{\gamma}_{j'})^2}. \quad (3.44)$$

3.5.4 BER with more than three copies of the signal available

It is obvious that the complexity of the BER expression increases rapidly as the number of combined copies is increased. However, for the range of offered load considered in this chapter, the probability of more than three packet transmissions was determined to be negligible and therefore the case when more than three copies of a packet are available can be ignored.

3.5.5 Throughput

In this section the system throughput is calculated. It is assumed that there are N_p bits in each packet and that FEC is employed which can correct up to t bits in each packet. It is also assumed that the fading statistics are uncorrelated from bit to bit, and remain constant during the transmission of each bit. Let $D^{(i)}(m_1 + 1, m_2 + 1, \dots, m_i + 1)$ be the event that after combining i copies of the same packet, the receiver has not recovered the transmitted data corresponding to that packet, conditioned on $m_j + 1$ simultaneous users during transmission of the j th, $j = 1, 2, \dots, i$, copy. The probability of the event $D^{(i)}(m_1 + 1, m_2 + 1, \dots, m_i + 1)$ is given by

$$p(D^{(i)}(m_1 + 1, m_2 + 1, \dots, m_i + 1)) \leq p(D^{(i)}(m + 1, m + 1, \dots, m + 1))$$

$$= \sum_{j=i+1}^{N_p} \binom{N_p}{j} (P_b^{(i)}(m))^j (1 - P_b^{(i)}(m))^{N_p-j}, \quad (3.45)$$

where $m = \max(m_1, m_2, \dots, m_i)$ and $P_b^{(i)}(m)$ is calculated as in the previous sections. The right hand side of (3.45) is the case when there are $m \geq m_j$, $j = 1, 2, \dots, i$, simultaneous users during the transmission of each copy. Let $R_{av}(m_1, m_2, \dots)$ denote the average number of transmissions necessary to transmit a given packet successfully when m_i other users are transmitting simultaneously during the i th, $i = 1, 2, \dots$, transmission of this packet. $R_{av}(m_1, m_2, \dots)$ is given by [6]

$$\begin{aligned} R_{av}(m_1, m_2, \dots) &= 1 + p(D^{(1)}(m_1 + 1)) \\ &+ p(D^{(1)}(m_1 + 1), D^{(1)}(m_2 + 1), D^{(2)}(m_1 + 1, m_2 + 1)) \\ &+ p(D^{(1)}(m_1 + 1), D^{(1)}(m_2 + 1), D^{(2)}(m_1 + 1, m_2 + 1), \\ &\quad D^{(1)}(m_3 + 1), D^{(3)}(m_1 + 1, m_2 + 1, m_3 + 1)) + \dots \end{aligned} \quad (3.46)$$

where

$$\begin{aligned} &p(D^{(1)}(m_1 + 1), D^{(1)}(m_2 + 1), D^{(2)}(m_1 + 1, m_2 + 1), \dots, D^{(1)}(m_Q + 1), D^{(Q)}(\\ &\quad m_1 + 1, m_2 + 1, \dots, m_Q + 1)) \\ &\leq p(D^{(Q)}(m_1 + 1, m_2 + 1, \dots, m_Q + 1)). \end{aligned} \quad (3.47)$$

Therefore, (3.46) can be written as

$$R_{av}(m_1, m_2, \dots) \leq 1 + \sum_{i=1}^{\infty} p(D^{(i)}(m_1 + 1, m_2 + 1, \dots, m_i + 1)). \quad (3.48)$$

Averaging (3.48) over (3.6), we have

$$R_{av} \leq 1 + \sum_{i=1}^{\infty} \left[\sum_{m_1=0}^{\infty} f_M(m_1) \sum_{m_2=0}^{\infty} f_M(m_2) \cdots \sum_{m_i=0}^{\infty} f_M(m_i) \right] \times \\ p(D^{(i)}(m_1 + 1, m_2 + 1, \dots, m_i + 1)). \quad (3.49)$$

$p(D^{(i)}(m_1 + 1, m_2 + 1, \dots, m_i + 1))$ and $f_M(m)$ are given by (3.45) and (3.6), respectively.

Since $1/R_{av}$ is the average portion of a packet which is received successfully for each transmission, and G is the average number of packets generated during each time slot, the throughput, T , is given by

$$T = \frac{G}{R_{av}}. \quad (3.50)$$

The normalized throughput is defined as throughput divided by the spreading factor, T/N .

3.6 Performance Examples

In this section, examples are given to illustrate the performance improvements possible with packet combining. DS/SS using BPSK modulation with a spreading factor of $N = 64$ is considered. Each packet consists of $N_p = 255$ bits, and the FEC is assumed capable of correcting up to $t = 10$ bits in each packet, i.e., (255,179,10) BCH code. Since the objective is to compare the conventional system with the combining case, the effect of code rate on the throughput for both systems is ignored. It is assumed that $T_m = 0.65 \mu\text{sec}$, σ^2 in (3.7) is 0.0631 (the same as the heavy shadowing case in [29]) and $T_{max}/T_c = 5$. It is also assumed that the number of resolvable paths for each copy of a packet is $L = 6$.

The improvement in BER when multiple copies of a bit are available at the RAKE-RAKE receiver, with $\frac{E_b}{N_0} = 20$ dB and no FEC, is depicted in Fig 3.7. These multiple copies can be due to retransmission (as is the case in this chapter) or due to multiple antennas. As expected, the RAKE-RAKE receiver performance improves as the number of available copies increases.

Fig 3.8 shows the throughput for $\frac{E_b}{N_0} = 20$ dB. It is obvious that combining the copies of a packet results in a higher throughput. The throughput of the conventional system in Fig 3.8 is a bell-shaped curve in terms of the offered load. As the offered load increases above 0, the throughput increases to a maximum value. Increasing the offered load beyond this point forces the system into an unstable region where the transmitter receives packets at a rate higher than can be successfully transmitted. Alternatively, combining the received copies of a packet results in a throughput which is an increasing function of the offered load, and is therefore stable compared to the conventional case (or at least is stable for a greater range of offered load).

Fig 3.9 depicts the average number of transmissions required to transmit a packet successfully vs. the normalized throughput. As expected, this average number is much lower for the combining case. The stability problem can be better explained using this figure. Considering the conventional system, as the average number of transmissions increases above 0, the throughput also increases. This continues until the throughput reaches a maximum. After this maximum, the system enters an unstable state in which the number of transmissions increases rapidly and the throughput decreases. Conversely, the throughput is always an increasing function of the offered load for the combining case (at least for the demonstrated values of offered load).

Figs. 3.10 and 3.11 show the throughput and the average number of transmis-

Figure 3.7: BER for DS/SSMA with multiple copies of a bit available at the RAKE-RAKE receiver.

Figure 3.8: Throughput of slotted DS/SSMA ALOHA with $E_b/N_0 = 20dB$.

Figure 3.9: Average number of transmissions for slotted DS-SSMA M-QPSK with $E_b/N_0 = 20dB$.

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PM-1 3½"x4" PHOTOGRAPHIC MICROCOPY TARGET
NBS 1010a ANSI/ISO #2 EQUIVALENT

PRECISION^{SM} RESOLUTION TARGETS

Figure 3.10: Throughput of slotted DS/SSMA ALOHA with $E_b/N_0 = 17dB$.

Figure 3.11: Average number of transmissions for slotted DS/SSMA ALOHA with $E_b/N_0 = 17dB$.

sions necessary to transmit a packet successfully for $\frac{E_b}{N_0} = 17$ dB. Again, combining the copies of a packet results in a higher throughput which is an increasing function of the offered load. The relative improvement with combining over the conventional system is higher for $\frac{E_b}{N_0} = 17$ (Fig.3.10) than with $\frac{E_b}{N_0} = 20$ (Fig.3.8). This is expected since combining is more effective when the retransmission rate is higher (smaller $\frac{E_b}{N_0}$).

3.7 Conclusions

In this chapter, a slotted DS/SSMA ALOHA PRN in a Rayleigh fading channel has been considered. A frequency selective channel model with an exponential MIP was assumed. A modified RAKE receiver, called a RAKE-RAKE receiver, was used which combines all received copies of a packet. This effectively increases the order of diversity from L , as in a conventional system, to $d \times L$, where L is the order of diversity for each packet copy and d is the number of copies being combined.

The BER with multiple copies of a bit available at the receiver, due to the retransmission of a packet, was calculated. The throughput and the average number of transmissions required to transmit a packet successfully were also derived. It was shown that combining increases the throughput and decreases the average number of transmissions necessary to transmit a packet.

The performance examples given illustrated that a conventional slotted DS/SSMA ALOHA system can become unstable, whereas with combining, the throughput always increases as the offered load increases (or at least the throughput is an increasing function for a larger range of offered load), and is therefore a more stable system.

Chapter 4

A Slotted Frequency-Hopped Multiple Access Network with Packet Combining

4.1 Introduction

In Chapters 2 and 4, DS/SSMA was applied to a packet radio network, with copies of the same packet combined at the receiver. Significant performance improvements were shown over the non-combining and narrow-band cases. In this section, frequency hopped spread spectrum multiple access (FH/SSMA) in a packet radio network is considered. The objective is to improve the throughput of the conventional system by combining the different received copies of the same packet.

A packet radio network is a collection of terminals which can communicate with each other via radio links [30]. Satellite multiple access broadcast channels for packet switching and terrestrial packet radio networks are the primary examples of these types of networks. FH/SSMA packet switching can also be used for any other shared channel, if sufficient bandwidth is available and the data can be packetized. In such systems, each transmitter is a potential source of interference for all terminals

Figure 4.1: Block diagram of a FH packet switching system.

within range of the transmitter. The goal in designing the packet radio network is to provide reliable communication, with a high throughput, in the presence of this multiple access interference. There are several papers in the literature that have considered FH/SSMA packet radio networks, among them [30] and [31]. In all these papers, the affected copies of a packet are discarded.

This chapter is organized as follows. Section 4.2 gives the system description, and the performance analysis is given in Section 4.3. Section 4.4 includes some examples and finally, Section 4.5 presents some conclusions based on the results.

4.2 The Frequency Hopped Packet System

The system considered in this chapter is shown in Fig. 4.1. When k code symbols are ready for transmission, they are encoded into n code symbols (a packet) by an (n, k) Reed-Solomon (RS) encoder. These codes are described below.

4.2.1 Reed-Solomon Codes

(n, k) Q -ary RS block codes are symbol error correcting (EC) codes with symbol size $\log_2 Q$ bits and block length $n = Q - 1$ [32]. RS codes are optimal in the sense that they achieve the maximum possible minimum Hamming distance between codewords, which means they can correct the largest possible number of errors for an EC code with the same code parameters. Since RS codes are maximum distance separable (MDS) [33], with k information symbols per block the minimum Hamming distance is $d_{min} = n - k + 1$. An RS code can correct up to

$$t = \lfloor \frac{d_{min} - 1}{2} \rfloor = \lfloor \frac{n - k}{2} \rfloor$$

symbol errors, where $\lfloor x \rfloor$ is defined as the largest integer less than or equal to x . In addition, up to ϵ errors and s erasures can be corrected where $2\epsilon + s = n - k$. This means that with erasures only correction, $n - k$ errors can be corrected, if the error locations are known.

Shortened, punctured and extended (within limits) RS codes can be obtained which are also MDS [33].

Decoding of an RS code is normally done in one of two ways, maximum likelihood decoding (MLD) or bounded distance decoding (BDD). With MLD, the decoder chooses the codeword closest to the received word, in terms of Hamming distance, as the most likely transmitted codeword. This distance may be greater than t , the error correcting capability of the code. One significant disadvantage of MLD decoding is the associated decoding complexity. MLD decoding is not practical in most RS code applications. With BDD, i.e., the Berlekamp-Massey or Euclidean algorithms [34], decoding is successful only if the received word lies within the decoding sphere of a valid codeword. Otherwise, the decoder reports a decoding failure.

Both algorithms employ an iterative technique to find first the error locations, then the values of the codeword symbols at these locations. Only BDD is considered here.

When a received word arrives at the RS decoder, three events are possible.

1. The number of erroneous symbols in a codeword is smaller than or equal to t and the transmitted packet is correctly recovered (error correction).
2. The number of errors is larger than t and the decoder cannot recover the transmitted packet (error detection).
3. The number of errors is larger than t and the decoder maps the received packet to the wrong codeword (decoding error).

By properly choosing the RS code parameters, (n, k) , the probability of decoding error can be made very small [35]. Therefore the RS code can be used for error detection as well as error correction.

4.2.2 Network Description

At the beginning of a time slot, each user starts packet transmission (if there are waiting to be transmitted) by means of FH. M-ary FSK, with square-law envelope detection at the receiver, is considered. Therefore, the demodulator consists of a bank of M square-law envelope detectors, as shown in Fig. 4.2. Using FH, the total bandwidth is divided into q sub-bands, each of which consists of M bins. We assume that one M-ary (channel) symbol is transmitted in each sub band. This means the channel symbol rate and the hop rate are equal. When a user (reference user) hops into a sub-band, it is possible that another user (or users) hops into the

Figure 4.2: M-ary FSK demodulator.

same channel, creating a hit. Two cases can be considered, hop synchronous, where the hop timing in all receivers is synchronized, and hop asynchronous, where the hop timing is asynchronous. The probability that the reference user is hit by any other particular user is given by [36]

$$p_h = \begin{cases} \frac{1}{q}, & \text{hop synchronous} \\ \frac{2}{q} - \frac{1}{q^2}, & \text{hop asynchronous,} \end{cases} \quad (4.1)$$

In this chapter we assume the system is hop synchronous.

The M-ary FSK demodulator is shown in Fig. 4.2. When a packet is received in the conventional system, F_1 through F_M are calculated for each received symbol. The demodulator chooses the symbol corresponding to the largest $F_j, j = 1, 2, \dots, M$, as the transmitted symbol, and when all the symbols in a packet are estimated, they are passed to the RS decoder as the received word. If the decoder cannot recover the transmitted data, the received packet is discarded and, by some mechanism, the transmitter is informed. Then the transmitter sends another copy of the packet. This process continues until the decoder recovers the transmitted data. As was noted, there is no difference between demodulating the first copy of a packet and demodulating any subsequent copies. Therefore, the receiver does not use any information that might exist in the discarded packets.

In the system that we propose in this chapter, all uncorrectable copies of a packet are kept at the receiver. After receiving the l th copy of a packet, F_{1l} through F_{Ml} are calculated for each symbol. Using the previously received copies of the same packet, the decision variables

$$F_j = \sum_{i=1}^l F_{ij}, \quad j = 1, 2, \dots, M, \quad (4.2)$$

are calculated. The symbol corresponding to the largest F_j is chosen as the transmitted symbol. The success or failure in decoding the transmitted packet determines the need for retransmission of another copy of the same packet.

Having several copies of the same packet can be interpreted as transmitting the same signal through different channels. Therefore, the effect of combining different copies of a packet is the same as a fast frequency-hopped (FFH) system where the diversity level is adapted according to the channel condition. FFH is a well

known technique for combating severe tone jamming (high error probability). In an FH/SS packet radio system, interference from other users appears in the form of tone interference. As the number of simultaneous users increases, the level of tone interference for each user is increased. Combining several copies of the same packet (or effectively using FFH) increases the probability of decoding the packet successfully in the presence of such interference. This increases the overall throughput of the system and reduces the delay necessary to transmit a given packet.

In this chapter we ignore the effects of additive noise and only consider the multiple access interference as a source of error. It should be noted that in a mobile environment, where fading and multipath can dramatically degrade the system performance, diversity is well known as a means of overcoming this severe channel condition [25]. Therefore, it is expected that the scheme proposed in this chapter, which is essentially a form of adaptive diversity, will have a substantial gain in a fading environment as well. This is presently under investigation.

4.3 Performance Analysis

In this section the performance of the slotted FH/SSMA packet radio network using RS codes is evaluated. MFSK with square-law envelope detection is considered. First the channel and code symbol error probability are calculated, followed by the throughput.

4.3.1 The pdf of the output of the square-law envelope detector for FH/SSMA

Assume that K simultaneous users are utilizing the FH/SSMA system. Without loss of generality, assume that the reference user is transmitting the symbol corresponding to the first filter in the bank of M filters in the square-law receiver (Fig. 4.2). In this section, the pdf of the detector output for the reference user's receiver is analyzed. This is required to derive the channel symbol error probability. The pdf of the output of the j th ($j = 1, 2, \dots, M$) square-law detector for a received symbol, $p_{F_j|K}(f_j)$, can be written as

$$p_{F_j|K}(f_j) = \sum_{i=0}^{K-1} p_{F_j}(f_j|i) p_h(K, i), \quad j = 1, 2, \dots, M, \quad (4.3)$$

where $p_{F_j}(f_j|i)$ is the conditional pdf conditioned on i , the number of users that hit the reference user's M -ary sub-band, and $p_h(K, i)$ is the probability that i users among the $K - 1$ other users hit the reference user. $p_h(K, i)$ is given by

$$p_h(K, i) = \binom{K-1}{i} p_h^i (1 - p_h)^{K-1-i}. \quad (4.4)$$

Since it is assumed that the reference user is transmitting the signal corresponding to the first filter in the bank of M filters, we have

$$p_{F_j}(f_j|i) = \begin{cases} \sum_{i_1=0}^i p_{i_1+1}(f_j) p(i_1, i), & j = 1 \\ \sum_{i_1=0}^i p_{i_1}(f_j) p(i_1, i), & \text{otherwise,} \end{cases} \quad (4.5)$$

where $p(i_1, i)$ is the probability that i_1 users, among the i users that hit the reference user's sub-band, are transmitting the j th symbol. This probability is given by

$$p(i_1, i) = \binom{i}{i_1} \left(\frac{1}{M}\right)^{i_1} \left(\frac{M-1}{M}\right)^{i-i_1}. \quad (4.6)$$

$p_g(f_j)$ is the pdf of the squared envelope of a sum of g random phase vectors and is given, using a recursive formula, in Simon [37]. If all signals have the same amplitude (normalized to one), we have

$$\begin{aligned} p_0(f_j) &= \delta(f_j), \\ p_1(f_j) &= \delta(f_j - 1), \\ p_2(f_j) &= \frac{1}{\pi \sqrt{f_j(4 - f_j)}}, \quad 0 \leq f_j \leq 4, \\ &\dots \\ p_l(f_j) &= \frac{1}{\pi} \int_{\zeta_{l-1}}^{\eta_{l-1}} \frac{p_{l-1}(x)}{\sqrt{4x - (x - f_j + 1)^2}} dx, \quad 0 \leq f_j \leq l^2, \end{aligned} \quad (4.7)$$

where

$$\begin{aligned} \eta_{l-1} &= \min((l-1)^2, (1 + \sqrt{f_j})^2), \\ \zeta_{l-1} &= \max(0, (1 - \sqrt{f_j})^2), \end{aligned} \quad (4.8)$$

and

$$\delta(x) = \begin{cases} 1, & x = 0, \\ 0, & \text{otherwise.} \end{cases} \quad (4.9)$$

Combining (4.6), (4.5) and (4.3) we have

$$p_{F_j|K}(f_j) = \begin{cases} \sum_{i=0}^{K-1} \frac{1}{M^i} \sum_{i_1=0}^i \binom{i}{i_1} (M-1)^{i-i_1} p_{i_1+1}(f_i) p_i(K, i), & j = 1, \\ \sum_{i=0}^{K-1} \frac{1}{M^i} \sum_{i_1=0}^i \binom{i}{i_1} (M-1)^{i-i_1} p_{i_1}(f_i) p_i(K, i), & j = 2, 3, \dots, M. \end{cases} \quad (4.10)$$

Note that at least one user (the reference user) is present at the input of the first filter, but none may be present at the input of the other $M-1$ filters.

If all the signals do not have the same amplitude, $p_j(f_j)$ can be calculated using the general recursive formulas given in [37], and the remainder of the analysis follows as above.

4.3.2 Symbol Error Probability

The pdf of the output of each filter of the squared law detector of the reference user, $p_{F_j|K}(f_j)$, was derived in the previous section. Using this pdf the symbol error probability is now calculated. Assume that l copies of a packet are present at the receiver. Also, assume that during transmission of the l th packet, K_l simultaneous users have also utilized the M-ary channel.

Let $p_c^{(l)}(K_1, K_2, \dots, K_l)$ and $p_e^{(l)}(K_1, K_2, \dots, K_l)$ denote, respectively, the probability that after combining the l packets “the reference user recovers the correct symbol” and “the reference user recovers an incorrect symbol”. Using $p_{F_j|K}(f_j)$, it is mathematically convenient to derive $p_e^{(l)}(K_1, K_2, \dots, K_l)$ first. For $l = 1$ the probability is given by

$$\begin{aligned}
p^{(1)}(K) &= P(F_2 < F_1, F_3 < F_1, \dots, F_M < F_1 | K) \\
&= \int_{-\infty}^{\infty} p(F_2 < f_1, F_3 < f_1, \dots, F_M < f_1 | F_1 = f_1, K) p_{F_1|K}(f_1) df_1,
\end{aligned} \tag{4.11}$$

where F_j , $j = 2, 3, \dots, M$ is the output of the j th filter in the detector. Since the number of users that fall into a given bin is random, the F_j , $j = 2, 3, \dots, M$, can be considered statistically independent and identically distributed. Thus,

$$p^{(1)}(K) = \int_{-\infty}^{\infty} (p(F_2 < f_1 | F_1 = f_1, K))^{M-1} p_{F_1|K}(f_1) df_1. \tag{4.12}$$

where

$$p(F_2 < f_1 | F_1 = f_1, K) = \int_{-\infty}^{f_1} p_{F_2|K}(f_2) df_2. \tag{4.13}$$

Substituting (4.13) into (4.12), the probability that the receiver makes a correct decision is given by

$$p^{(1)}(K) = \int_{-\infty}^{\infty} \left(\int_{-\infty}^{f_1} p_{F_2|K}(f_2) df_2 \right)^{M-1} p_{F_1|K}(f_1) df_1. \tag{4.14}$$

Now, consider the case where two copies of the same packet are available at the receiver, $l = 2$. Furthermore, assume that K_1 and K_2 simultaneous users have been transmitting during the transmission of the first and second copies, respectively. This means that for each symbol, two values for the output of the j th filter (in the bank of M filters) are available. At the receiver, these two values are added together and then the symbol corresponding to the largest sum is chosen as the estimate of the transmitted symbol. When these two statistically independent random variables

are added, it results in a random variable with a pdf which is the convolution of the individual pdfs. Therefore, the probability of a correct decision at the receiver after receiving two copies, conditioned on K_1 and K_2 , is given by

$$p_c^{(2)}(K_1, K_2) = \int_{-\infty}^{\infty} \left(\int_{-\infty}^{f_1} p_{F_2|K_1}(f_2) * p_{F_1|K_2}(f_2) df_2 \right)^{M-1} p_{F_1|K_1}(f_1) * p_{F_1|K_2}(f_1) df_1, \quad (4.15)$$

where $*$ denotes convolution. Using the same arguments, the probability of a correct decision after receiving l copies of the same packet is given by

$$p_c^{(l)}(K_1, K_2, \dots, K_l) = \int_{-\infty}^{\infty} \left(\int_{-\infty}^{f_1} p_{F_2|K_1}(f_2) * p_{F_2|K_2}(f_2) * \dots * p_{F_2|K_l}(f_2) df_2 \right)^{M-1} p_{F_1|K_1}(f_1) * p_{F_1|K_2}(f_1) * \dots * p_{F_1|K_l}(f_1) df_1. \quad (4.16)$$

The channel symbol error probability after combining l copies is then given by

$$p_e^{(l)}(K_1, K_2, \dots, K_l) = 1 - p_c^{(l)}(K_1, K_2, \dots, K_l). \quad (4.17)$$

4.3.3 Throughput

When each RS code symbol consists of exactly r channel symbols, the codeword symbol error probability after receiving the l th copy of a packet, $p_e^{(l)}(K)$, is given by

$$p_e^{(l)}(K) = 1 - (p_c^{(l)}(K))^r. \quad (4.18)$$

where $p_c^{(l)}(K)$ is given by (4.16). Note that it is assumed that K simultaneous users are also transmitting during all transmissions. Let $D_e^{(l)}$, $D_u^{(l)}$ and $D_c^{(l)}$ be the events

that after receiving l copies, 'the decoded sequence contains detectable errors', 'the decoded sequence contains undetectable errors' and 'the decoded sequence contains no errors', respectively. Assuming that $p(D_u^{(l)})$ is negligible, we have

$$p(D_e^{(l)}) + p(D_c^{(l)}) = 1. \quad (4.19)$$

Considering an (n, k) RS code, it is assumed that the undetected error probability is negligible. Based on the results in [35], this is true for sufficiently large t , where t is the number of correctable errors. If t is not large enough, a CRC code can be appended to each packet. The detected error probability, conditioned on K , is then given by

$$p(D_e^{(l)}|K) = \sum_{i=\lfloor (n-k)/2 \rfloor + 1}^n \binom{n}{i} p_e^{(l)}(K)^i (1 - p_e^{(l)}(K))^{n-i}, \quad (4.20)$$

where $\lfloor x \rfloor$ denotes the largest integer smaller than x .

The average number of transmissions necessary to transmit a packet successfully conditioned on K , $R(K)$, is given by

$$R(K) = 1 + p(D_e^{(1)}|K) + p(D_e^{(1)}, D_e^{(2)}|K) + p(D_e^{(1)}, D_e^{(2)}, D_e^{(3)}|K) + \dots \quad (4.21)$$

Since

$$p(D_e^{(1)}, D_e^{(2)}, \dots, D_e^{(l)}|K) \leq p(D_e^{(l)}|K), \quad (4.22)$$

$R(K)$ can be bounded as [6]

$$R(K) \leq 1 + p(D_e^{(1)}|K) + p(D_e^{(2)}|K) + p(D_e^{(3)}|K) + \dots \quad (4.23)$$

The throughput, $T(K)$, is defined as the average number of successfully transmitted packets per time slot. Since there are q sub-bands and the code rate is $\frac{k}{n}$, the normalized throughput is given by

$$T(K) = \frac{Kk}{R(K)qn}. \quad (4.24)$$

4.4 Performance Examples

In this section examples are given to illustrate the performance improvements possible with packet combining. In all cases, the number of sub-bands, q , is 100. In the first example a $(63, k)$ RS code is used in conjunction with 8-ary FSK modulation. This means every three and six bits constitutes one channel symbol and one codeword symbol, respectively. Therefore, each codeword symbol consists of two channel symbols. The throughput for various values of k is shown in Fig. 4.3. The dotted line, denote the conventional system, where the uncorrectable packets are discarded. The solid lines show the system considered in this chapter, where the uncorrectable packets are combined together. As can be seen, there is an optimum RS code rate that maximizes the achievable throughput [38]. This figure also shows that the maximum achievable throughput for the combining case is more than twice that of the conventional system. Fig. 4.4 shows the average number of transmissions necessary to transmit a packet successfully for $k = 27$ and $k = 39$, which correspond to the maximum achievable throughput for the conventional and

combining cases, respectively. For the conventional system, the number of transmissions necessary to transmit a packet increases dramatically as the number of users increases. However, combining copies of the same packet substantially reduces this number of transmissions.

The following example demonstrates the case where the channel symbols and RS code symbols are matched. This means that channel symbols and RS code symbols contain the same number of bits. With six bits per channel symbol, a $(63, k)$ RS code is used with 64-ary FSK modulation. The throughput is given in Fig. 4.5. For this case, the improvement with combining is higher than the previous example, and is at least three times the throughput of the conventional system. This higher improvement can be explained as follows. Suppose that the reference users' first copy of a packet is received in error. Using 64-ary FSK (as opposed to 8-ary FSK), the probability that during the transmission of the second copy an interfering user (users) hits the reference user with the same symbol (out of 64 possible tones in a sub-band) as in the first transmission is much smaller than if 8-ary FSK (8 possible tones in a sub-band) were used. Therefore, combining different copies of a packet is more effective with higher M-ary FSK modulation. Fig. 4.6 shows the number of transmissions required to transmit a packet successfully, for $k = 29$ and $k = 53$, which correspond to the maximum achievable throughput for the conventional and combining systems, respectively. In this case, combining copies of a packet substantially reduces the number of transmissions required to successfully receive a packet. It should be mentioned that for a $(63, 53)$ RS code, the decoder error probability cannot be considered negligible, so that a CRC error detecting code may also be necessary.

Finally, in the last example consider a $(255, k)$ RS code with 16-ary FSK. Each

Figure 4.3: Throughput for the conventional and combining systems assuming a $(63, k)$ RS code and 8-ary FSK.

Figure 4.4: Number of transmissions required to transmit a packet successfully for the proposed and conventional systems with 8-ary FSK and a $(63, k)$ RS code.

Figure 4.5: Throughput for the conventional and combining systems assuming a $(63, k)$ RS code and 64-ary FSK.

Figure 4.6: Number of transmissions required to transmit a packet successfully for the proposed and conventional systems with a $(63, k)$ RS code and 64-ary FSK.

code symbol consists of two channel symbols. Fig. 4.7 shows the throughput for this case. For both the conventional and combining systems, there is an optimum code rate which maximizes the throughput. However, the maximum achievable throughput with combining is more than three times that of the conventional case. Fig. 4.8 shows that combining substantially reduces the number of transmissions required to transmit a packet.

4.5 Conclusions

A FH/SSMA packet radio network using RS codes has been considered. Instead of discarding the erroneous copies of a packet, the receiver retains and combines these copies to improve the throughput and reduce the delay necessary to transmit a packet. This is equivalent to a fast frequency hopped system with an adaptive diversity level. It is shown that by combining the received copies of the same packet, the throughput is increased substantially. This combining also dramatically reduces the number of transmissions necessary to transmit a packet successfully.

Figure 4.7: Throughput for the conventional and combining systems assuming a $(255, k)$ RS code and 16-ary FSK.

Figure 4.8: Number of transmissions required to transmit a packet successfully for the proposed and conventional systems with a $(255, k)$ RS code and 16 ary FSK.

Chapter 5

Conclusions

This thesis presented a study of spread spectrum packet radio networks. The key contribution is improved performance by combining the different received copies (due to retransmissions) of a packet. In a conventional system all previously received copies of a packet are discarded. It was shown that combining these copies, results in an increased throughput and a reduction in the delay necessary to transmit a packet successfully. The contributions of this thesis can be divided into three following areas:

1. A slotted DS/SSMA ALOHA system with FEC and maximum likelihood decoding was considered. Multiple access interference was assumed as the only source of error. Code combining was applied to combine the different copies of a packet at the receiver. Code combining allows for the combination of all available copies of a packet at the receiver, resulting in a lower code rate for that specific packet. Analytical and simulation results have been presented which show that by applying code combining, a higher throughput is achievable than that without code combining. Code combining also causes the effective throughput to be a non-decreasing function of the offered load (at least for a

large range of the offered load). This is important from the point of view of system stability.

2. A slotted DS/SSMA ALOHA packet radio network in a Rayleigh fading channel has been considered. The channel was characterized as a frequency selective channel with an L tap delay line model, which provides L replicas of the transmitted signal at the receiver. The multipath intensity profile of the channel was assumed to be exponential. A modified RAKE receiver, called a RAKE-RAKE receiver, which combines all received copies of a packet, was introduced. Using a RAKE-RAKE receiver increases the order of diversity from L , in the conventional case, to $d \times L$, where d is the number of copies being combined.

The BER was calculated for multiple copies of a bit available at the receiver due to the retransmissions of a packet as well as due to frequency selectivity nature of the channel. The throughput and the average number of transmissions required to transmit a packet successfully were also derived. It was shown that combining increases the throughput and decreases the average number of transmissions necessary to transmit a packet successfully.

3. A FH/SSMA packet radio network with MFSK modulation and a square-law envelope detector was considered. A (n, k) RS code was assumed as the forward error correction scheme. The receiver retained and combined the different received copies of a packet, which were normally discarded. This resulted in a system that is equivalent to a fast frequency hopped system, with an adaptive order of diversity. It was shown that combining results in a substantially higher throughput and smaller delay in transmitting a packet

successfully.

In general, combining different available copies of a packet at the receiver causes the system to be more reliable from the stability point of view. Suppose that for conventional slotted SSMA ALOHA, the system remains in the unstable region (where the rate of newly-generated packets exceeds the network throughput) for a significant period of time. Then a backlog of packets to be retransmitted is created at the transmitter. In this case, even if new packet generation is suspended, it will take a long time for the system to clear this backlog. In contrast, suppose that the throughput of the system with combining decreases due to an increase in offered load, so that there exists an unstable region. Further, assume that the system remains sufficiently long in the unstable region to create a backlog of packets for retransmission. By restricting the generation of new packets, the resulting backlog will be cleared much more rapidly compared to the conventional system. This is because receiving additional copies of a packet improves the probability of successfully decoding that packet, which is not the case for the conventional system. In other words, if the conventional system and the system with packet combining start with the same backlog of packets to be retransmitted, the combining system will clear the backlog more rapidly than the conventional system.

It should be mentioned that, in order to retain extra copies of a packet for combining at the receiver, some additional memory space is necessary.

5.1 Future Work

The work in this thesis can be extended in the following directions:

In this thesis a RAKE-RAKE receiver with a maximum ratio combiner was considered for a multipath Rayleigh fading channel. When the path gains are not available at the receiver, EGC can be used to form a RAKE receiver in order to demodulate the transmitted signal. Using EGC, the RAKE-RAKE receiver which combines all the received copies of a packet is given in Fig. 3.6. Since some of the resolvable paths at the receiver have very small average SNR, one would suspect that considering these paths in the EGC-RAKE receiver would only worsen the performance of the receiver, by adding more noise to the received signal. Therefore, it is reasonable to assume that there is an optimum number of paths (depending on the channel specifications) that should be combined in the EGC-RAKE receiver to give the best performance. Analyzing the performance of the EGC-RAKE receiver in a DS/SSMA slotted ALOHA system and calculating the optimum number of paths to be combined together is an interesting topic for further research.

This thesis can also be extended to the area of spread spectrum multimedia communications. The multiple access capability of the CDMA channel can be used to accommodate voice calls, while the data users follow the ALOHA protocol. Since different users require different quality of service (QOS), the channel resources should not be shared equally among the various classes of users. Voice (real-time data) traffic will use the channel resources, i.e., power or bandwidth, to obtain the desired QOS without retransmissions. However, data (non real time) packets will use the remaining channel resources, taking advantage of retransmission and a receiver that combines the different copies of the same packet.

Analyzing the performance of a FH/SSMA packet radio network with packet combining in a multipath fading channel is another potential topic for further research.

Appendix

Derivation of Eqs. (2.7) and (2.23)

In this Appendix, Eq. (2.7) is derived first, followed by Eq. (2.23).

Consider BPSK modulation with coherent detection, where $\pm s(t)$ are transmitted corresponding to ± 1 . The bit error probability, p_b , is given by [25]

$$p_b = Q\left(\sqrt{\frac{2E}{N_o}}\right) = \frac{1}{2}\text{erfc}\left(\sqrt{\frac{E}{N_o}}\right), \quad (.1)$$

where

$$E = \int_0^{T_b} s(t)^2 dt. \quad (.2)$$

and E/N_o is the corresponding SNR.

In the following, we derive the error probability for the multi-channel system of Fig. 2.9 with $Q = 2$. The noise power spectrum density in each branch is $N_{oi}/2, i = 1, 2$. The decision variables, Z_1 and Z_2 , are given by

$$Z_1 = \int_0^{T_b} (s(t) + n_1(t))s(t)dt, \quad (.3)$$

$$Z_2 = \int_0^{T_b} (s(t) + n_2(t))s(t)dt, \quad (.4)$$

where T_b is the symbol duration. Since $n_1(t)$ and $n_2(t)$ are independent additive Gaussian noise random variables, Z_1 and Z_2 are Gaussian random variables with mean E and variances $EN_{o1}/2$ and $EN_{o2}/2$, respectively. In addition,

$$Z = Z_1 + Z_2 = \int_0^{T_b} (2s(t) + n_1(t) + n_2(t))s(t)dt, \quad (.5)$$

is a Gaussian random variable with mean $2E$ the variance $(EN_{o1} + EN_{o2})/2$. Therefore, the error probability is given by

$$p_e = Q\left(\frac{2E}{\sqrt{E\frac{N_{o1}}{2} + E\frac{N_{o2}}{2}}}\right) = Q\left(\sqrt{\frac{4}{\frac{N_{o1}}{2E} + \frac{N_{o2}}{2E}}}\right) = \frac{1}{2}\text{erfc}\left(\sqrt{\frac{4}{\frac{N_{o1}}{E} + \frac{N_{o2}}{E}}}\right). \quad (.6)$$

Thus, the SNR for this multi-channel system, SNR_{eq2} , is

$$SNR_{eq2} = \frac{4}{\frac{N_{o1}}{E} + \frac{N_{o2}}{E}} = \frac{4}{\frac{1}{SNR_1} + \frac{1}{SNR_2}}, \quad (.7)$$

where SNR_i is the SNR for the i th, $i=1,2$, channel.

For the general case of a multi-channel system with Q channels, the equivalent SNR , SNR_{eqQ} , is given by

$$SNR_{eqQ} = \frac{Q^2}{\frac{1}{SNR_1} + \frac{1}{SNR_2} + \cdots + \frac{1}{SNR_Q}}, \quad (.8)$$

which is the same as (2.7).

The derivation of (2.23) is as follows. First consider the case when all of the received bits have the same SNR . Without loss of generality, assume that the all-zero codeword is transmitted. P_d is defined as the probability of error in the pairwise comparison of the transmitted path (all-zero path) and a path of weight d . Let $U^{(0)}$ and $U^{(1)}$ denote the metrics corresponding to the all-zero and the path with weight d , respectively. Since the modulation is BPSK, we have [25]

$$P_d = \Pr[U^{(1)} - U^{(0)} \geq 0] = \Pr\left[\sum_{i=1}^d y_i > 0\right], \quad (.9)$$

where y_i is a Gaussian random variable with mean $-E$ and variance $EN_o/2$. Therefore,

$$P_d = Q\left(\frac{dE}{\sqrt{d\frac{N_oE}{2}}}\right) = \frac{1}{2}\text{erfc}\left(\sqrt{\frac{E}{N_o}}d\right). \quad (.10)$$

Now, consider the situation where two packets are combined together to form a packet of a rate 1/2 convolutional code. It is assumed that each bit has energy E , and the noise power spectrum densities during packet transmission are $N_{o1}/2$ and $N_{o2}/2$. We further assume that when two paths in the trellis differ by d bits, half of these bits belong to the first packet and the remainder to the second packet. Then the y_i 's in (.9) are Gaussian random variables with mean $-E$. Half of the y_i 's have variance $EN_{o1}/2$ while the other half have variance $EN_{o2}/2$. Therefore, P_d is given by

$$P_d = \frac{1}{2}\text{erfc}\left(\sqrt{\frac{2E}{(N_{o1} + N_{o2})}}d\right) = \frac{1}{2}\text{erfc}\left(\sqrt{\frac{2d}{\frac{1}{SNR_1} + \frac{1}{SNR_2}}}\right). \quad (.11)$$

Generalizing this to the case of combining Q packets we have

$$P_d = \frac{1}{2} \operatorname{erfc} \left(\sqrt{\frac{Qd}{\frac{1}{\operatorname{SNR}_1} + \frac{1}{\operatorname{SNR}_2} + \cdots + \frac{1}{\operatorname{SNR}_Q}}} \right). \quad (.12)$$

Substituting (2.4) into (.12) results in

$$P_d = \frac{1}{2} \operatorname{erfc} \left(\sqrt{\frac{1.5dNQ}{m_1 + m_2 + \cdots + m_Q - Q}} \right), \quad (.13)$$

which is the same as (2.23).

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