Coded Still Image Transmission Over Very Slow Fading Channels

by

William Glenn Zeng

Submitted to the Department of Electrical Engineering and Computer Science
in partial fulfillment of the requirements for the degrees of

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Author ........................................
Department of Electrical Engineering and Computer Science
January 30, 1995

Certified by ......................................
Mitchell D. Trott
Assistant Professor of Electrical Engineering
Thesis Supervisor

Certified by ......................................
Vijitha Weerackody
Member of Technical Staff, AT&T Bell Laboratories
Thesis Supervisor

Accepted by ......................................
Frederic R. Morgenthaler
Chairman, Departmental Committee on Graduate Students
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Abstract

Coded still image transmission over very slow fading channels is difficult because transmission errors on these channels occur in long bursts and the raw bit error rates are unacceptably high. In JPEG, the new international standard for color still image compression, certain markers are extremely sensitive to transmission errors such that a single bit error may cause the entire image to be lost. This thesis develops novel transmit antenna diversity techniques and applies these diversity techniques in two transmission protocols for JPEG coded still images on very slow fading channels. The first protocol uses a feedback channel to request retransmission of the erroneous data and to inform the transmitter to use a different transmit antenna when the current diversity branch is detected in a deep fade. An induced fast fading diversity technique that can effectively provide time diversity in slow fading environments is used in the second protocol. This protocol operates in a forward error correction mode in the absence of a feedback channel. Both protocols are shown to have significant performance improvements over the corresponding protocols without diversity.

Thesis Supervisor: Mitchell D. Trott
Title: Assistant Professor of Electrical Engineering

Thesis Supervisor: Vijitha Weerackody
Title: Member of Technical Staff, AT&T Bell Laboratories

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Chapter 1

Introduction

Recent interests in indoor radio communications are closely related to the exciting developments in today’s telecommunications industry. The industry is in the midst of very active research, development, and deployment of a whole host of wireless communication services and products ranging from paging, cellular telephony, and indoor wireless local area networks to more sophisticated wireless networks collectively known as Personal Communication Networks (PCN). With the full deployment of PCN, a user can initiate and receive phone calls, request and send data, transmit and process images, and perform a variety of other functions from his lightweight wireless communication unit, and be able to do so from anywhere, anytime [1]–[3]. Since the users of these communication services spend a large proportion of their time at numerous indoor facilities, any successful PCN implementation must have an important component which provides voice, image, video, and data to people indoors [4].

The communication services and products envisioned above pose challenging questions. Data communication on indoor radio channels is difficult due to the slow fading characteristics of these channels. The received signal strength on these channels is slowly time-varying with a large dynamic range (17–60 dB) [4][5]. A receiver located in a deep fade will remain there for an extended period of time to cause long bursts of data errors. For data applications such as image and video, the required bit error rates are very low ($10^{-5}$ or less). It will be difficult, if not impractical, to achieve
the required low bit error rates without error control. Given the low bit error rate requirements, the chosen error control scheme must also meet the delay constraints of the intended applications. Sophisticated coding and modulation techniques, advanced error control schemes, novel antenna diversity concepts are required to achieve reliable communication over these very slow fading channels.

1.1 Thesis Overview

This thesis describes two transmission protocols for JPEG coded still images on very slow fading radio channels. Both protocols are extensively investigated with computer simulations and supported by analytical studies and laboratory experiments.

Chapter 2 briefly describes fading channels. A statistical model is selected for mathematical analysis and simulations. Several relevant results from the literature are used to illustrate the deleterious effects of fading on data communications.

Chapter 3 introduces some basic automatic repeat request (ARQ) and forward error correction (FEC) concepts. Several simple error control schemes are implemented in software to highlight some performance issues.

Chapter 4 formulates the problems facing JPEG coded still image transmission over very slow fading channels by presenting the JPEG coding structures and the consequences of transmission errors on the received image quality. The major conclusions are that a small percentage of the JPEG compressed still image data contains the important information that must be received error-free when possible. We label this Type-I information. It suffice to say at this point that the Type-I information is so important that a single transmission error may cause the entire image to be lost. The rest of the JPEG coded still image data, the Type-II information by our definition, should be received with the lowest bit error rate possible but transmission errors on this information are not as disastrous as transmission errors on the important Type-I information. The presentation makes a strong case for the chosen error control schemes to be studied and implemented in subsequent chapters.

Throughout the remaining chapters of this thesis, transmit antenna diversity tech-
niques are extensively studied for reducing the detrimental effects of fading on multi-path fading radio channels. In particular, a novel switched transmit antenna with feedback diversity technique is treated in Chapter 5. This technique is applied to an efficient protocol for JPEG coded still image transmission over very slow fading channels. In this protocol, a feedback channel is provided to request retransmission on erroneous data packets and to request the transmitter to use a different transmit antenna when the current diversity branch is detected in a deep fade. The important Type-I information is transmitted with a Stop-and-Wait ARQ protocol while the Type-II information is transmitted with a FEC code in the time slots left idle by the Stop-and-Wait ARQ protocol. The protocol can be efficiently implemented without allocating extra bandwidth to transmit the switch requests or performing the explicit received signal strength measurement as in a conventional switched transmit antenna with feedback scheme. This transmission protocol is shown to have significant performance gain on very slow fading channels.

For practical applications where a feedback channel is not available, a FEC protocol using an induced fast fading diversity technique that can effectively provide time diversity in the slow fading environments is presented in Chapter 6. This protocol multiplexes the transmissions of the Type-I and Type-II information. The Type-I information is allowed for multiple transmissions as in a repetition code and then appropriately combined at the receiver. The Type-II information is transmitted once with a convolutional code. The fast fading characteristics deliberately induced by this diversity technique increase the correcting capability of the channel code drastically and consequently improve the transmission reliability.

Chapter 7 provides laboratory implementation details of the JPEG coded still image transmission protocol. Experimental results are provided to verify the theoretical analysis and software simulation results in the early chapters. A summary and conclusions are given in Chapter 8.
Chapter 2

The Indoor Radio Channel

This chapter very briefly introduces the indoor radio channel. Multipath interference is explained first. Two models for describing the statistics of the received signal are then discussed. Finally, several consequences of Rayleigh fading are collected here to illustrate the detrimental effects of fading on data communication qualities over these channels.

2.1 Multipath Interference

In a typical indoor radio environment, the portable receiver does not receive just one copy of the transmitted signal. Rather, due to scattering and reflection, the transmitted signal reaches the receiver via multiple paths with different time delays, amplitudes, and phases. The received signal is the sum of these component waves. Mathematically, suppose a unit amplitude continuous wave signal

\[ x(t) = \text{Re}\{e^{j2\pi fc t}\} = \cos 2\pi fc t \]  \hspace{1cm} (2.1)

of frequency \( fc \) is transmitted. The received signal is

\[ y(t) = \text{Re}\{\sum_{i=1}^{N} [a_i(t)e^{j\phi_i(t)}e^{jk_z_i}]e^{j2\pi fc t}\} , \]  \hspace{1cm} (2.2)
where \( a_i(t), z_i, \theta_i(t) = \phi_i(t) + k z_i \) are the random time-varying amplitude, path length, and phase of the \( i^{th} \) component wave, respectively. \( N \) is the number of components. The constant \( k = 2\pi/\lambda \) is the wave number, while \( \phi_i(t) \) depends on the time variation of the channel. For electromagnetic wave propagating at the speed of light \( c \), the path length \( z_i \) is related to the arrival time \( \tau_i \) by \( z_i = c \tau_i \). Much effort in recent measurement and modeling of the indoor radio channel is devoted to obtaining statistical models for these random time-varying quantities [4][5]. Because \( \lambda \) at microwave frequency is very short, a small change in \( z_i \) causes a large phase shift in \( \theta_i(t) \). These rapid time variations in the phases \( \{\theta_i(t)\} \) are responsible for the constructive or destructive interference of the component waves. A typical interference pattern resulting from multipath fading is depicted in Fig. 2-1. This plot was generated in software by using a technique described by Jakes [9, pp. 70–76] with \( v = 1 \) km/hr and \( f_c = 900 \) MHz, where \( v \) is the mobile speed. The meaning of these parameters will become clear in later discussions.

![Figure 2-1: Typical Multipath Fading Signal Strength](image)
2.2 Rayleigh Fading

When considering statistical models for the small scale path amplitude distributions, usually two models apply depending on the presence or absence of a line-of-sight component. For indoor environments, a line-of-sight component may be present. But this need not be the case. In our laboratory experiments to be described in Chapter 7, the transmitter and receiver are located separately in two adjacent rooms with metallic walls. The propagation path is further obstructed by typical laboratory equipment and office furniture in these two rooms. The received signal comes from scattering via the ceiling, floor, walls, and furniture. A direct line-of-sight component does not exist.

The Rayleigh fading model is widely accepted for narrowband multipath fading channel without a strong line-of-sight component. The reason is because the average amplitudes of the component waves are statistically independent and identically distributed. When the received signal in (2.2) is considered, two quadrature components \( I \) and \( Q \) exist,

\[
\begin{align*}
I &= \sum_{i=1}^{N} a_i(t) \cos \theta_i, \\
Q &= \sum_{i=1}^{N} a_i(t) \sin \theta_i
\end{align*}
\]

By applying the central limit theorem, it follows that these components are Gaussian distributed. Hence the amplitude of the received signal, \( r = \sqrt{I^2 + Q^2} \) is described by a Rayleigh distribution, and the phase \( \theta = \tan^{-1}(Q/I) \) is uniformly distributed in \((0, 2\pi]):-

\[
\begin{align*}
PR(r) &= \frac{r}{\sigma^2} \exp\left\{-\frac{r^2}{2\sigma^2}\right\} \quad r \geq 0, \\
PE(\theta) &= \frac{1}{2\pi} \quad 0 \leq \theta < 2\pi,
\end{align*}
\]

where \( 2\sigma^2 \) is the second moment of the Rayleigh distribution.

In a Ricean model, a strong direct component is present among other multipath
components. The severity of fading is reduced in comparison to the Rayleigh fading case. The mathematics, however, becomes more difficult. For example, the probability distribution of the amplitude, appropriately called the Ricean distribution, contains a zero-order modified Bessel function of the first kind,

\[ p_R(r) = \frac{r}{\sigma^2} \exp\left(-\frac{r^2 + u^2}{2\sigma^2}\right) I_0\left(\frac{ru}{\sigma^2}\right) \quad r \geq 0, \tag{2.7} \]

where \( I_0(\cdot) \) is a zero-order modified Bessel function of the first kind, and \( u \) is the magnitude of the strong line-of-sight component.

For mathematical analysis, we use the Rayleigh model because it is analytically more tractable and the results obtained with this model can at least be thought of as the lower bounds on system performance when a line-of-sight component does become available at certain instances. In fact, Rayleigh fading is a special case of the Ricean fading model when the line-of-sight component is removed. The Ricean distribution in (2.7) becomes the Rayleigh distribution in (2.5) when the amplitude of the line-of-sight component goes to zero [4].

Once the Rayleigh model is accepted, several well-known consequences [4]–[10] should be discussed. In the presence of Rayleigh fading, as a receiver moves in a spatial field-strength variation profile such as shown in Fig. 2-1, while unusually strong signals (5 dB above \( rms \)) are extremely rare, very weak signals (10 dB below \( rms \)) are common. It can be easily shown [9] that the received signal is 10 dB below the \( rms \) signal strength for 10% of the time, whereas the signal is 5 dB above the \( rms \) signal strength for less than 1% of the time.

Another way to characterize fading is by the temporal, rather than spatial, correlation properties of the received signal. Referring to Fig. 2-1 once again, given the signal is in a deep fade at some time instant, with very high probability the signal is also in a deep fade a short duration away. How quickly the signal at different time instances are decorrelated can be measured by the coherence time of the channel.
This quantity is approximately given by [12, p123]

\[ \tau_c \approx \frac{1}{2\pi f_D}, \]

(2.8)

where \( f_D = \frac{v}{\lambda} = \frac{v}{(c/f_c)} \) is the Doppler frequency associated with the relative motion \( v \) between the transmitter and the receiver. This apparent frequency shift could also be contributed by the changes in the channel medium such as people or equipment moving about the transmitter and receiver. For indoor radio channels, this time variation is very small. Therefore, \( \tau_c \) for indoor radio channels is very large in comparison to the data signaling intervals. Such channels are best characterized as very slow fading. To give a particular example, suppose the channel variation in the indoor radio channel is represented by an equivalent speed of 1 km/hr. At 900 MHz carrier frequency and a data rate of 10 kbits/sec, \( \tau_c \) is the equivalent of 2000 bit durations.

Finally, as another consequence of Rayleigh fading, the average bit error rate (BER) of an uncoded system falls off only inversely with the signal-to-noise ratio (SNR) for large SNR values. In contrast, the same modulation scheme in the presence of additive white Gaussian noise (AWGN) alone gives a BER that decreases exponentially with increasing SNR.
Chapter 3

On Error Control Protocols

The preceding chapter highlighted the severe limitations imposed on channel quality by Rayleigh fading. This chapter investigates several rudimentary error control protocols on fading channels. It is the purpose of the current chapter to interpret some well-known results and clarify certain relationships between physical assumptions and mathematical derivations.

Several classes of error control protocols are introduced in Section 3.1. The average frame erasure rate, a quantity closely related to the throughput efficiency of ARQ protocols, is computed for the very slow and very fast fading channels in Section 3.2. The fundamental differences between very slow and very fast fading channels are then discussed. A simple ARQ protocol is analyzed in Section 3.3. In Section 3.4, the effects of an imperfect feedback channel on the channel throughput efficiency of ARQ protocols are illustrated. A very simple technique for improving the efficiency of ARQ protocols using code-combine decoding [29]–[32] is presented in Section 3.5. Section 3.6 concludes this chapter.

3.1 Error Control Protocols

In this section we summarize some known error control protocols. Automatic repeat request (ARQ), forward error correction (FEC), and hybrid ARQ error control protocols are commonly used to achieve reliable data communication. ARQ protocols use
feedback channels to request retransmission of erroneous data packets. FEC protocols operate in a broadcast mode without feedback channels and use powerful channel codes to correct transmission errors. ARQ protocols tend to be more reliable, but less efficient than FEC protocols. Hybrid ARQ protocols combine both features of FEC and ARQ.

The simplest ARQ protocol is Stop-and-Wait [13]. The transmitter appends parity check bits to the data stream to enable the receiver to detect errors. Following each transmission, the transmitter waits for an acknowledgment from the receiver. If no error has been detected, the received packet is delivered to the data sink and a positive acknowledgment is sent back to the transmitter. If at least one error is detected in the received packet, the receiver discards the packet and a negative acknowledgment is sent to the transmitter for retransmission of the erroneously received packet. The transmitter sends a new packet when a positive acknowledgment is received; otherwise, it retransmits the same data packet. A timer at the transmitter is activated when a packet is sent. If the acknowledgment is lost in the reverse link or arrives after the time-out period, a time-out mechanism is invoked to retransmit the same packet [14, pp. 64–86], [15, pp. 180–187].

The next two protocols are continuous protocols, whereby data packets are sent from the transmitter to the receiver continuously. In Go-Back-N [13], a negative acknowledgment causes the transmitter to go back to the negatively acknowledged packet and start transmitting from there. All packets following this erroneous packet are retransmitted even if some may already have been correctly received while the negative acknowledgment was being transmitted on the feedback channel. By contrast, in Selective-Repeat [13], only the incorrectly received packets are retransmitted. Because of this selective retransmission of erroneous data packets, packets may be received out of sequence in a Selective-Repeat ARQ protocol. A theoretically infinite buffer is needed at the receiver to restore the proper ordering of the received packets before releasing them to the data sink. A time-out mechanism as described above is also needed for practical implementation of these continuous protocols.

The throughput efficiency of an ARQ protocol, defined as the ratio of the average
number of information bits successfully accepted by the receiver to the total number of bits that could be transmitted per unit time [13], changes with the channel condition. Regardless of the throughput efficiency, ARQ protocols are extremely reliable with the appropriate choice of error detection codes.

On the other hand, FEC protocols do not use feedback channels. The transmitter introduces redundancy into the data stream to allow the receiver to correct errors and then deliver the decoded data block to the data sink, although some of the decoded data blocks may contain uncorrectable errors. Since the probability of uncorrectable error is higher than the probability of undetectable error, pure FEC protocols tend to be less reliable than ARQ protocols.

Finally, hybrid ARQ schemes combine both features of FEC and ARQ [13]. The FEC subsystem is used to increase the throughput efficiency by reducing the number of retransmissions. When the feedback channel and error detection are perfect, all residual errors that cannot be corrected by the FEC subsystem are corrected by retransmissions. Hybrid ARQ schemes can be further classified into type-I and type-II. In a type-I hybrid ARQ scheme, the data blocks are coded for both error detection and correction. The resulting low code rate becomes inefficient when the channel condition is good. In a type-II scheme, the initial transmission is coded for error detection only and parity bits for error correction are transmitted only when necessary [13].

### 3.2 Average Frame Erasure Rate

ARQ protocols are based on the transmission of blocks of $N$ sequential bits. Without error correction, one transmission error within a frame is sufficient to cause the entire frame to be rejected by the receiver. Frames are accepted when no error has been detected. Hence, the average frame erasure rate (FER), defined as the probability of any received frame chosen at random contains one or more errors, is directly related to the throughput efficiency of any ARQ protocol. In fact, if we redefine the throughput efficiency of an ARQ protocol as the ratio of the number of successfully delivered
data frames to the total number of transmitted frames on a channel, the average FER is precisely one minus the throughput. The latter definition is called the channel throughput to distinguish it from the overall throughput [16]. The overall throughput includes factors such as the code rate, stuffed bits for code synchronization, and parity check bits. Except for hybrid ARQ schemes, these are mere constant factors and can be easily incorporated into the final analysis when desired.

It would be desirable to obtain FER expressions in terms of the fading rates, but this turns out to be an extremely difficult problem due to bit error correlations on Rayleigh fading channels. Therefore we only look at two limiting cases: very fast fading and very slow fading.

### 3.2.1 Background

The bit error probabilities of several common binary modulation schemes in the presence of additive white Gaussian noise (AWGN) are given by the formulas [18]

\[
P_b(\gamma) = \begin{cases} 
\frac{1}{2} \exp(-\frac{1}{2} \gamma) & \text{NCFSK,} \\
\frac{1}{2} \exp(-\gamma) & \text{DPSK,} \\
\frac{1}{2} \text{erfc}(\frac{1}{\sqrt{2}} \gamma) & \text{coherent FSK,} \\
\frac{1}{2} \text{erfc}(\sqrt{\gamma}) & \text{coherent PSK,}
\end{cases}
\] (3.1)

where \( \gamma \) is the SNR value and \( \text{erfc}(\cdot) \) is the complementary error function,

\[
\text{erfc}(x) = \frac{2}{\sqrt{\pi}} \int_x^{\infty} e^{-t^2} dt .
\] (3.2)

Because bit errors are statistically independent on AWGN channels, with \( t \) bit error correction, the FER, \( P_w(\gamma) \), is

\[
P_w(\gamma) = 1 - \sum_{i=0}^{t} \binom{N}{i} P_b(\gamma)^i [1 - P_b(\gamma)]^{N-i},
\] (3.3)

where \( N \) is the frame size measured in bits. The notation in (3.3) is such that \( P_w(\gamma) \) does not necessary mean there is only one \( \gamma \) value for the entire packet of size \( N \).
There may be several γ values in a packet duration, as is the case in the fast fading limits.

Data transmission over Rayleigh fading channels presents a completely different picture. One of the fundamental differences between AWGN channels and Rayleigh fading channels is that the former is characterized by a steady signal strength reception and the latter by a fluctuating signal strength reception [20]. The received time-varying signal strength is statistically described by a chi-square distribution with 2 degrees of freedom [18]

\[ Pr(\gamma) = \frac{1}{\Gamma} \exp\left(-\frac{\gamma}{\Gamma}\right), \]

(3.4)

where \( \Gamma \) is the average SNR. In this case, BER and FER are derived in the average sense, where the average is taken over the instantaneous received SNR, \( \gamma \). To calculate the average BER, we assume the effect of fading over one bit duration is effectively constant, although varying over a long succession of such bit signaling intervals [19]. For noncoherent frequency shift keying (NCFSK), the average bit error rate is given by

\[ \langle P_b(\gamma) \rangle = \int_0^\infty P_b(\gamma) Pr(\gamma) d\gamma = \frac{1}{2} \left( \frac{1}{1+\Gamma/2} \right). \]

(3.5)

These ideas are extended in the following subsections to compute the average FER. It is noted here that for mathematical analysis, coherent modulation is difficult because of the complex formulas. For differential phase shift keying (DPSK), the formula appears harmless, but because differential encoding and decoding can lead to double errors which are not independent, unless sufficient interleaving is used to decorrelate these double errors, the analysis developed below does not apply to DPSK [20]. Hence, in order to make the mathematical derivations tractable, we carry out the analysis for NCFSK only.

### 3.2.2 Very Fast Fading

When the fading is very fast in comparison to the data symbol unit, bit errors for each symbol occur independently. This statistical independence between bit errors makes the average FER computation trivial. The result is simply obtained by replacing
\(P_b(\gamma)\) in (3.3) with \(\langle P_b(\gamma)\rangle\):

\[
\langle P_w(\gamma) \rangle = 1 - \sum_{i=0}^{t} \binom{N}{i} \langle P_b(\gamma) \rangle^i [1 - \langle P_b(\gamma) \rangle]^{N-i}
\]

\[
= 1 - \sum_{i=0}^{t} \binom{N}{i} \left[ \frac{1}{2} \left( \frac{1}{1 + \Gamma/2} \right) \right]^i [1 - \frac{1}{2} \left( \frac{1}{1 + \Gamma/2} \right)]^{N-i} . \tag{3.6}
\]

### 3.2.3 Very Slow Fading

In the very slow fading case, we assume the signal strength is constant over a block of \(N\) bits, but it varies from block to block. In essence, the entire block is treated like a single bit in the average bit error rate computations:

\[
\langle P_w(\gamma) \rangle = \int_0^\infty P_w(\gamma) Pr(\gamma) d\gamma
\]

\[
= \int_0^\infty \left\{ 1 - \sum_{i=0}^{t} \binom{N}{i} P_b(\gamma)^i [1 - P_b(\gamma)]^{N-i} \right\} Pr(\gamma) d\gamma
\]

\[
= 1 - \sum_{i=0}^{t} \binom{N}{i} \int_0^\infty P_b(\gamma)^i [1 - P_b(\gamma)]^{N-i} Pr(\gamma) d\gamma
\]

\[
= 1 - \sum_{i=0}^{t} \binom{N}{i} \int_0^\infty \frac{1}{2^i} e^{-\frac{1}{2}iy} \sum_{k=0}^{N-i} \binom{N-i}{k} (-1)^k \frac{1}{\Gamma} e^{-\frac{1}{2}ky} \frac{1}{(i+k)\Gamma + 2} d\gamma
\]

\[
\text{where } \Gamma = \frac{1}{\beta}
\]

\[
= 1 - \sum_{i=0}^{t} \sum_{k=0}^{N-i} \binom{N}{i} \left[ \binom{N-i}{k} (-1)^k \frac{1}{(i+k)\Gamma + 2} \right] . \tag{3.7}
\]

### 3.2.4 Interpretation of Results

Calculated results for packet size \(N = 20\) are shown in Fig. 3-1 and Fig. 3-2 for several values of \(t\) bit correction. Without error correction (Fig. 3-1), the average FER is higher at faster fading rates. This is because for both the very fast and very slow fading cases, the average BER is the same; but errors occur in bursts in the slow fading case. The same average number of errors distributed more uniformly in time causes a higher average FER for the fast fading case. This is a well-known
result and is presented in various forms in [21]–[25]. With error correction (Fig. 3-2), the correcting power of the FEC code is enhanced by interleaving or the inherent randomness on the channel at fast fading rates. Hence, the average BER decreases at higher fading rates to give a correspondingly lower average FER. A plot similar to Fig. 3-2 for DPSK with interleaving is shown in reference [21].

These two figures illustrate clearly the two competing effects on average FER: without error correction, the average FER is higher at faster fading rates; with error correction, burst errors at lower fading rates reduce the error correction capability of the channel code to result in higher average FER. A clear understanding of these issues is critical in designing error control protocols.

### 3.3 A Simple ARQ Protocol

This section analyzes a simple Stop-and-Wait ARQ protocol. This protocol has the unique feature that consecutive transmissions of data packets are separated far enough
Figure 3-2: Average FER Performance with Error Correction

apart in time such that the fading processes among these packet signaling intervals are statistically independent. The channel throughput efficiency is obtained analytically and compared to software simulation results. Error detection and the feedback channel are assumed to be perfect.

3.3.1 Channel Throughput Analysis

Instead of computing the channel throughput directly, we compute its inverse, $\eta$, the expected number of transmissions needed to successfully deliver a packet. We obtain $\eta$ by summing the occurrence probabilities for all successive transmissions [28],

$$\eta = 1 + \langle P_d(\gamma_1) \rangle + \langle P_d(\gamma_1, \gamma_2) \rangle + \langle P_d(\gamma_1, \gamma_2, \gamma_3) \rangle + \cdots,$$

where $\langle P_d(\gamma_1) \rangle$ denotes the average probability of receiving an erroneous packet on the first transmission. The average is taken over $\gamma_1$, the SNR on the first transmission.
of a packet. \( P_d(\gamma_1, \gamma_2) \) denotes the average joint probability of receiving erroneous packets on both the first and second transmissions, with the average taken over the joint distribution of \( \gamma_1 \) and \( \gamma_2 \). Intuitively, the first term in (3.8) is the probability of a sure event, namely, at least one transmission is needed to deliver a packet. The second transmission occurs with probability \( P_d(\gamma_1) \), the probability that the first transmission fails. The remaining terms can be similarly explained.

When the \( \{\gamma_n\} \) are correlated, (3.8) is extremely difficult to solve. However, the ARQ protocol we are currently discussing was designed such that the \( \{\gamma_n\} \) between different transmissions are statistically independent because the packets are transmitted far apart in time. In this case, (3.8) can be simplified to

\[
\eta = 1 + \langle P_d(\gamma) \rangle + \langle P_d(\gamma) \rangle^2 + \langle P_d(\gamma) \rangle^3 + \cdots = \frac{1}{1 - \langle P_d(\gamma) \rangle} = \frac{1}{1 - \langle P_w(\gamma) \rangle} .
\]

(3.9)

Since the channel is effectively memoryless, \( \langle P_d(\gamma) \rangle \) in this case is precisely the average FER, \( \langle P_w(\gamma) \rangle \), that was computed previously.

The average FER for NCFSK on the fast fading and slow fading channels are given by (3.6) and (3.7), respectively. By combining (3.6), (3.7), and (3.9), analytical expressions for NCFSK without error correction \( (t = 0) \) are obtained:

\[
\eta^{-1} = \left\{1 - 1/2(1 + \Gamma/2)\right\}^N
\]

(3.10)

for very fast fading and

\[
\eta^{-1} = \sum_{k=0}^{N} \binom{N}{k} \left(-\frac{1}{2}\right)^k \frac{2}{k\Gamma + 2}
\]

(3.11)

for very slow fading. The theoretical and simulation results are plotted in Fig. 3-3 for \( N = 20 \). In the simulations, coherent PSK over a Rayleigh fading channel with additive white Gaussian noise is used for the channel model. The Rayleigh fading signal is generated with a technique described in [9, pp. 70–76], where a set of low
frequency oscillators are used to provide the Doppler shifts to the carrier frequency $f_c$ assuming uniform arrival angles. Slow and fast fading were simulated using 900 MHz carrier frequency with $v = 1$ km/hr at 8 kbits/sec and $v = 200$ km/hr at 1 kbits/sec, respectively. The theoretical curves are plotted with a 6 dB offset because NCFSK performs 6 dB worse than coherent PSK [18] at large SNR values (e.g., the plotted theoretical result at 5 dB is computed using 11 dB SNR value, 10 dB result is obtained with 16 dB, and so on). The discrepancy between the theoretical result and the simulation result at low SNR values is caused by the fact that at low SNR values, the performance difference between coherent PSK and NCFSK is less than 6 dB.

### 3.4 Imperfect Feedback Channel

In the study of ARQ protocols, it is usually assumed that the feedback channel is perfect. This assumption can not be justified when the feedback channel is also
subject to Rayleigh fading. An imperfect feedback channel can cause unnecessary retransmissions and increases the number of undetected errors. The effect of an imperfect feedback channel on the channel throughput is considered in this section. For the current discussion, perfect error detection is assumed.

The analysis is carried out for the ARQ protocol in the previous section with a Rayleigh fading feedback channel. The feedback channel is assumed to be statistically independent from the forward transmission channel. Again, we use \( P_d(\cdot) \) to denote the probability of receiving an erroneous packet and use \( P_c(\cdot) \) to denote the probability of accepting an error-free packet. The quantity \( \eta \) can be expressed as

\[
\eta = 1 + ((P_d^f(\gamma_1))(P_c^b(\gamma_1)) + (P_d^b(\gamma_1)))[(P_d^f(\gamma_2))(P_c^b(\gamma_2)) + (P_d^b(\gamma_2))] + \cdots . \tag{3.12}
\]

Intuitively, \( [(P_d^f(\gamma_1))(P_c^b(\gamma_1)) + (P_d^b(\gamma_1))] \) represents the occurrence probability of the first retransmission. It occurs when the acknowledgment was detected in error (e.g., \( P_d^f(\gamma_1) \)) regardless of what happened on the forward transmission channel or when the acknowledgment correctly indicated the detection of an erroneous packet (e.g., \( P_d^b(\gamma_1) \)). The superscripts \( f \) and \( b \) are used to denote the forward and back channels, respectively. The subscripts in the \{\gamma_n\} are shown explicitly to indicate how the equation was obtained. Since the \{\gamma_n\} are statistically independent, the subscripts can be dropped to further simplify the equation:

\[
\eta = 1 + [(P_d^f(\gamma))(P_c^b(\gamma)) + (P_d^b(\gamma))][((P_d^f(\gamma))(P_c^b(\gamma)) + (P_d^b(\gamma)))^2 + [(P_d^f(\gamma))(P_c^b(\gamma)) + (P_d^b(\gamma)))^3 + \cdots \\
= \frac{1}{1 - [(P_d^f(\gamma))(P_c^b(\gamma)) + (P_d^b(\gamma))]} \cdot \frac{1}{1 - (P_d^b(\gamma))} \cdot \frac{1}{1 - (P_d^b(\gamma))} \\
= \eta^f \eta^b \tag{3.13}
\]

The second last step uses \( (P_c(\gamma)) + (P_d(\gamma)) = 1 \) in the case of perfect error detection.
The result in (3.13) is intuitively satisfying. Since the forward transmission channel and the feedback channel are statistically independent, the expected number of transmissions needed to deliver a packet is the product of the expected number of transmissions needed to deliver a packet when the feedback channel is perfect ($\eta_D$) and the expected number of transmissions needed to deliver the acknowledgment ($\eta_A$).

![Graph showing channel throughput for imperfect feedback channel](image)

Figure 3-4: Channel Throughput for Imperfect Feedback Channel

We proceed to verify the analysis as follows. First, we choose a packet size $N = 20$ for both the data packet on the forward path and the acknowledgment packet on the feedback path. Both channels are fading at the identical rate of $v = 1$ km/hr, $f_c = 900$ MHz, and 8 kbits/sec transmission rate. We then perform software simulation on the ARQ protocol in the previous section (with a perfect feedback channel). The result is shown in Fig. 3-4 as dashed line. The square of the result is used as an estimate of the channel throughput for the same ARQ protocol when the feedback channel is changed from a perfect one to one that is subject to Rayleigh fading. This estimate is plotted as a solid line in Fig. 3-4. Finally, simulations are performed for
the ARQ protocol from the previous section with a Rayleigh fading feedback channel. The results are also plotted in Fig. 3-4. The close agreement between the simulation results and the estimated results renders credence to the analysis.

3.5 Memory ARQ

In the ARQ protocol discussed above, whenever a data packet is detected in error, that packet is discarded. The receiver waits until an error-free packet is received. Error detection is assumed to be perfect. The basic idea behind memory ARQ schemes is to make use of the erroneously received packet and combine them with their retransmitted copies in decoding the transmitted data packets [29]–[34]. By combining an arbitrary number of noisy packets, a simple variable rate repetition code is obtained. The error correction capability of this code increases with decreasing code rate. The reason we introduce this concept here is that a repetition-type FEC scheme using code-combine is discussed in Chapter 6.

The scheme we implemented in software operates as follows. The first time a packet is received, the receiver operates on the received packet alone to decode the transmitted data packet. If the packet is error-free, the receiver sends a positive acknowledgment to the transmitter; otherwise, the soft decision output of the decoder (\( r_1 \)) is saved in a receiver buffer and a negative acknowledgment is sent. If the second transmission is successfully decoded, the receiver buffer is cleared and a request for a new packet is sent; otherwise, the soft decision output of the decoder (\( r_2 \)) is combined with \( r_1 \). The combined packet, \( r_1 + r_2 \) is then decoded. If \( r_1 + r_2 \) is successfully decoded, the receiver buffer is cleared; otherwise a retransmission request is sent. This procedure continues until the packet is correctly received.

The channel throughput efficiency of the ARQ protocol in Section 3.3 with code-combine decoding is presented here. The quantity \( \eta \) becomes:

\[
\eta = 1 + \left(P_d(\gamma_1) + P_d(\gamma_1, \gamma_2, \gamma_1 + \gamma_2) + \cdots \right) .
\]  

(3.14)
Note that this equation is derived from (3.8) by including the code-combine failure probabilities in each successive terms. The meaning of \( \langle P_d(\gamma_1, \gamma_2, \gamma_1 + \gamma_2) \rangle \) should be obvious: it is the probability that errors were detected at the initial transmission and the first retransmission as well as at the first level of code-combine decoding. The notation \( \gamma_1 + \gamma_2 \) in (3.14) is used to convey the fact that the failure probability at the first level of code-combine decoding is depend on the SNR values on both the first and second receptions. It should also be clear that, \( \langle P_d(\gamma_1, \gamma_2, \gamma_1 + \gamma_2) \rangle \) is less than \( \langle P_d(\gamma_1, \gamma_2) \rangle, \langle P_d(\gamma_1, \gamma_2, \gamma_1 + \gamma_2, \gamma_3, \gamma_1 + \gamma_2 + \gamma_3) \rangle \) is less than \( \langle P_d(\gamma_1, \gamma_2, \gamma_3) \rangle \), and so on. As a result, memory ARQ has a higher channel throughput than its corresponding ARQ scheme without code-combine decoding.

The evaluation of \( \eta \) is difficult due to the statistical dependencies among detected error probabilities on individually received frames and the combined frames. We only seek lower and upper bounds on \( \eta \). Following Sindhu's approach in [29], a simple lower bound on \( \eta \) is obtained by neglecting high order terms in (3.14):

\[
\eta > 1 + \langle P_d(\gamma_1) \rangle = 1 + \langle P_w(\gamma) \rangle .
\]

(3.15)

An upper bound is given by \( \eta \) without code-combine decoding, namely

\[
\eta < \frac{1}{1 - \langle P_w(\gamma) \rangle} .
\]

(3.16)

Since the channel throughput is the inverse of \( \eta \), the lower bound on channel throughput is given by the inverse of the upper bound on \( \eta \); likewise, the upper bound on channel throughput is given by the inverse of the lower bound on \( \eta \). The upper performance bounds are plotted in Fig. 3-5 and Fig. 3-6 for the very fast and very slow fading limit, respectively. The lower performance bounds are obtained by simulations since they are the channel throughputs without code-combine decoding. Again, because the bounds are computed for NCFSK and the simulations use coherent PSK, the calculated results are shown with a 6 dB offset as discussed in Section 3.3. The channel fading parameters for the slow and fast fading limits are identical to the ones given in Fig. 3-3. For the results illustrated here, the improvement in throughput
Figure 3-5: Code-Combine Decoding on Fast Fading Channels

is not very significant. This is because code-combine decoding was not used very frequently due to the fact that a great majority of the packets were delivered with fewer than two transmissions. These figures also illustrate that at faster fading rates, time diversity benefit is much greater and the improvements in throughput efficiency due to combine decoding success is significantly higher than the slow fading case.

3.6 Concluding Remarks

This chapter has been a brief tour through several transmission protocols. The protocols were simple enough such that analysis were feasible. The insights gained from the discussion will be used in the following chapters to design practical data transmission protocols.

As an implementation issue, the acknowledgment frame size should be kept to the smallest practical size because of the multiplicative effect an imperfect feedback
Figure 3-6: Code-Combine Decoding on Slow Fading Channels

channel has on the channel throughput.

With memory ARQ, we showed that substantial gain in channel throughput efficiency is possible when erroneous data packets are used in a code-combine decoding scheme. In essence, code-combine is a simple repetition code. Its performance is greatly influenced by the correlation between the erroneous packets that are combined, with the probability of success being higher when they are decorrelated.
Chapter 4

Concerning JPEG Coded Still Images

The previous chapter dealt with some fundamental aspects of error control protocols. In the next three chapters, the concepts developed there will be used and expanded in the design of practical error control schemes for the transmission of JPEG coded still images over slow fading channels. It is therefore appropriate for the current chapter to describe several relevant features of these JPEG coded still images. An understanding of the image coding structure is an essential step toward understanding the error control scheme design philosophy to be presented later.

4.1 Division of JPEG Coded Image Data

A complete description of the JPEG still image coding scheme is given in [40]. We discuss only briefly some relevant features. For the current discussion, the JPEG compressed data can be categorized by their relative importance. Typically less than 1% of the JPEG coded still image data consists of the extremely important markers. These markers contain the start and end of image information, transformation and quantization tables, and other essential information required to interpret and decode the compressed image data. We label this Type-IA information. Throughout the JPEG coded image data, with our choice of parameters for the coding options, another
set of markers occur at the end of every $16 \times 16$ block known as Minimum Coded Unit (MCU). These markers are very important, albeit not as important as the Type-IA markers described above. They contribute about 5–10% of the entire compressed image data. We label this Type-IB information. Both the Type-IA and Type-IB markers can be uniquely identified by a two-byte code. The first byte of the marker is a byte-aligned 0xff (hexadecimal ff) and the second byte is a code that identifies the function of the marker. Collectively, Type-IA and Type-IB markers are called Type-I information. The rest of the JPEG coded still image data contains the entropy-coded segments. We label these entropy-coded segments the Type-II information.

### 4.2 Effects of Transmission Errors

Transmission errors in the Type-I markers have severe consequences on the received image quality. A single bit error in the Type-IA markers may cause the entire image to be lost when the JPEG still image decoder fails to interpret the image and refuses to provide any output. When errors occur in isolated Type-IB markers, the decoder may be able to decode the image and provide an image with $16 \times 16$ blocks of data severely distorted corresponding to the erroneous Type-IB markers. When a Type-IB marker is accidentally converted to a Type-IA marker, the JPEG still image decoder fails with very high probability. For the Type-II information, although it is desirable to achieve the lowest BER possible, transmission errors are not as disastrous. However, the critical thing to look out for when errors occur in these entropy-coded segments is that transmission errors may convert them into Type-I markers which may crash the JPEG image decoder. Burst errors in these entropy-coded segments corrupting a long string of consecutive MCU can also severely degrade the received image quality.

To illustrate these effects, we start with a JPEG coded still image (Fig. 4-1) and selectively introduce errors throughout the coded data stream. We first replace one of the Type-IA markers, 0xffd8, the start of image marker, with 0xffd9, the end of image marker. Unfortunately, the image cannot be shown because the JPEG still image decoder fails to provide any output as a result of this single bit error. In
Figure 4-1: Color Image: Original Image

Figure 4-2: Color Image: Single Bit Error at Type-IB Markers
Fig. 4-2, one of the Type-IB markers, \texttt{0xffd2}, is replaced by \texttt{0xffd3}, yet another Type-IB marker. Note that this is also just a single bit error. The still image decoder fails to decode the block of $16 \times 16$ pixels where the single bit error occurs, and this entire block is unflatteringly shown with uniform gray scale. This can be seen approximately 1.4 inches from the right and .2 inches from the top of the image. Finally, one of the Type-IB markers is accidentally converted to a Type-IA marker (\texttt{0xffd4} converted to \texttt{0xffc4}, which is the marker that defines the Huffman tables); the resulting image is distorted beyond recognition (Fig. 4-3) due to this single bit error. Note that the JPEG still image decoder correctly decodes the image up to the erroneous Type-IB marker, it then gives up decoding the rest of the image beyond this marker.
4.3 Implications for Transmission Protocol Design

Because of the aforementioned transmission error effects, an overall transmission protocol must apply different levels of error protection for the different kind of data segments in order to fully use the scarce resource of the channel.

Obviously, the Type-IA markers require the highest level of error protection. These markers must be received error-free when possible. Although transmission errors in the Type-IB markers are not as critical as errors in the Type-IA markers, the Type-IB markers are more populous than the Type-IA markers. The majority of the JPEG coded still image data is the entropy-coded Type-II segments. Bit error rates of $10^{-4}$ are satisfactory for the entropy-coded Type-II segments provided that transmission errors do not give rise to unexpected markers. When a feedback channel is available, ARQ protocols may be used to transmit the Type-IA and Type-IB markers. In the absence of a feedback channel, the Type-IA and Type-IB markers may be transmitted with a repetition-type FEC scheme. The Type-IA marker may be transmitted more times than the Type-IB markers to provide more protections for the Type-IA markers. Regardless of which scenario is applied, the Type-II information is transmitted only once.

The remaining chapters of this thesis are devoted to the design of practical JPEG coded still image transmission protocols on very slow fading channels. Chapter 5 is devoted to the study of an unequal error protection scheme which utilizes a novel switched transmit antenna diversity technique in combination with an ARQ protocol to transmit the Type-I information while transmitting the Type-II information in the time slots left idle by the ARQ protocol. A repetition-type FEC scheme is considered in Chapter 6. Chapter 7 treats practical laboratory implementation of error control schemes for the transmission of JPEG coded still images.
Chapter 5

Switched Transmit Antenna With Feedback Protocol

Diversity techniques are extensively used in today’s communication systems to reduce the severity of fading. This chapter builds on the foundations of the previous chapters to design a practical JPEG coded still image transmission protocol using a novel switched transmit antenna with feedback diversity technique.

Section 5.1 introduces certain aspects of the 4-DPSK modem and convolutional codes. In Section 5.2, the JPEG coded still image transmission protocol using the switched transmit antenna diversity with feedback technique is treated. Specifically, a Stop-and-Wait ARQ protocol is used to transmit the Type-I information while the Type-II information is transmitted using the time slots left idle by the ARQ protocol. Simulation results are presented to determine the effect on the average BER of the Type-II information and the ARQ system’s channel throughput efficiency of the number of transmit antennas, feedback response time of the system, frame size, and channel fading rate.

5.1 4-DPSK and Convolutional Codes

In this section we summarize certain relevant features of the modulation and demodulation techniques to be used in the simulation studies and implementations. The
basic concepts of convolutional coding and block interleaving are also introduced.

5.1.1 4-DPSK

In preparation for implementation, we chose 4-DPSK as our modulation scheme. A clear understanding of its operating principles is crucial when transmit antenna diversity technique is applied later in this chapter with this modulation scheme.

Referring to Fig. 5-1, the digital data to be transmitted is pairwise mapped to a complex 4-PSK signal using Gray coding. The complex 4-PSK signal is then differentially encoded by multiplying it with a delayed version of the transmitted signal prior to transmission. At the receiver, demodulation is done by making a decision on the product of the received signal with a delayed version of the previously received signal [18, pp. 266–271].

The 4-DPSK demodulator requires very small phase changes between successive data symbols for it to work reliably, hence it performs better on slow fading channels than on fast fading channels. This is verified by a theoretical result given in [35] and
the computer simulation results depicted in Fig. 5-2. The implications can be seen in the switched transmit antenna diversity scheme to be discussed later, where the transmitter switches from one antenna to another in a controlled fashion, creating abrupt phase changes on the received signal. One solution is to add two known bits to the beginning of each data packet for the demodulator to establish a phase reference once an abrupt change occurs. Bit stuffing (two bits per data packet) of the modulator and demodulator is used in the simulation studies.

5.1.2 Convolutional Coding and Block Interleaving

In practice, the data is rarely just modulated and then transmitted across the channel without any form of protection against multipath fading. Convolutional codes are routinely used to improve the signal transmission performance. However, since most convolutional codes are designed for AWGN channels, interleaving is required to randomize the burst errors on fading channels for the channel code to function
effectively. This subsection briefly illustrates these concepts.

The convolutional encoder loads the message bits into a tapped shift register. Certain bits of the fully loaded shift register are modulo 2 added to generate the encoded data bits. Usually several bits are generated for every input bit. At the receiver, the channel decoder usually employs the Viterbi algorithm with soft decisions to decode the received data stream. An in-depth treatment of convolutional coding with Viterbi decoding is given in reference [12, pp. 358–399].

The rationale for interleaving is that it is much more difficult for a convolutional code to correct clustered errors than it is to correct the same number of errors when they are dispersed. A block interleaver writes the data to be transmitted into a $m \times n$ matrix column-wise. The transmitter reads out the bits row-wise. The deinterleaver performs the inverse operation of the interleaver. It stores the received data row-wise in a similar $m \times n$ matrix. The stored data bits are read out column-wise by the decoder [39, pp346]. Therefore, if a burst error of length $n$ were received, the interleaver effectively converts the burst error into $n$ single bit errors, which is much easier for the convolutional code to correct.

![Block Diagram of the Communication System](image)

Figure 5-3: Block Diagram of the Communication System

When the channel coder and interleaver are added to the 4-DPSK modem, the
resulting communication system is depicted in Fig. 5-3. Computer simulations were performed for this communication system to determine its average BER performance at various channel fading rates and interleaver sizes for a rate 1/2, memory 2 convolutional code. The results are shown in Fig. 5-4. As can be seen in Fig. 5-4, the channel code performs better at faster fading rates and larger interleaver sizes because errors are less bursty under these conditions. Also illustrated is the difficulty in obtaining diversity benefit when the channel is very slowly time-varying. At \( v = 1 \text{ km/hr} \), \( f_c = 900 \text{ MHz} \), and 8 kbits/sec, with the 8 x 12 interleaver, there was virtually no coding gain; even when the interleaver size was increased to 64 x 192, the coding gain was still under 10 dB at \( 10^{-3} \) BER level. Note that by rearranging the data, processing delays are introduced. The delay in the 64 x 192 interleaver deinterleaver pair may be too large for some voice or data applications. We will show in Chapter 6 how to increase the correcting power of the channel code with a transmit antenna diversity technique without incurring long delays.
5.2 Switched Transmit Antenna with Feedback

Switched transmit antenna with feedback is a form of space diversity with the diversity branches implemented at the base station (i.e., forward channel transmitter). An excellent treatment of this technique for the case of two transmit antennas in an analog FM system is given in [9, pp. 399-423]. The treatment for an arbitrary number of transmit antennas used in a data transmission system with binary DPSK is considered in reference [36]. In addition to the three conventional space-diversity combining methods at the receiver: selection diversity, equal gain combining, and maximal ratio combining [9, pp. 313-325], switched diversity at the transmitter represents an additional diversity dimension that can be effectively exploited to achieve more reliable communication in the presence of multipath fading. The switched transmit antenna diversity scheme can be simply implemented since a feedback channel already existed in an ARQ system. We devote this section to study such schemes.

5.2.1 Principles of Operation

In a traditional switched transmit antenna with feedback scheme [9, pp. 399-423], the receiver is constantly comparing the received signal strength to a predetermined threshold. When the received signal falls below this threshold, a signal is sent to inform the transmitter to use a different transmit antenna.

In the scheme to be studied, the switch triggering mechanism is very different. To better appreciate the differences, a block diagram of the JPEG coded still image transmission protocol employing switched transmit antenna with feedback diversity is shown in Fig. 5-5. It consists of an ARQ system for transmitting the Type-I information and another system for transmitting the Type-II information. The Type-II information is protected by a rate 1/2, memory 2 convolutional code with block interleaving. The ARQ system uses the standard CRC-16 for error detection. 4-DPSK is used to modulate and demodulate both the Type-I and Type-II information. The feedback channel is not shown. In this scheme, we do not set a signal threshold level. When a positive acknowledgment is received at the transmitter, the channel is as-
Figure 5-5: Switched Transmit Antenna With Feedback Protocol
sumed to be good and there is no need to switch antenna. Otherwise, the channel is assumed to be in a deep fade and the transmitter switches to a different antenna. Since the switch requests are implicitly conveyed by the retransmission requests, no extra bandwidth is needed to transmit the switch requests.

![Diagram of a typical event sequence on the forward transmission path.](image)

Figure 5-6: Typical Event Sequence on Forward Transmission Path

Fig. 5-6 contains a typical event sequence on the forward transmission path to illustrate how the overall system operates. Assume first that each transmitted packet is delayed by a fixed amount of time before arriving at the receiver. This is a reasonable assumption because for indoor radio channels, the receiver is located approximately the same distance from the transmitter and the channel is slowly time-varying throughout the course of the experiment. This assumption is graphically illustrated in Fig. 5-6, where the last bit of a packet is transmitted at $t_0$ and it arrives at the receiver at $t_1$. We also assume that the transmitter is always ready with packets to be sent. The Type-I packets are multiplexed with the Type-II packets, with the multiplexer inserting one Type-I packet every $L$ Type-II packets. We define $L$ to be the multiplex ratio. At the receiver, a demultiplexer dispatches the Type-I packets to the CRC-16 decoder for error detection. An acknowledgment is sent once every $L$ Type-II packets upon receiving a Type-I packet. Referring to Fig. 5-6, A packet arrives at the receiver at $t_1$. Suppose that after certain processing delay at the receiver, this packet is rejected at $t_2$, and a negative acknowledgment is sent. After the fixed transmission delay from the receiver to the transmitter and processing delay at the transmitter, a
different transmit antenna is used starting at $t_3$. We define the time interval between $t_3$ and $t_1$ to be the switch actuation time. Note that when the multiplex ratio $L$ is larger than the switch actuation time, the shortest time interval between consecutive switch of antennas is defined by $L$ independent of the switch actuation time.

The Type-I packets are not protected with a channel code because they are already protected by a retransmission protocol. It is true that the frequency of retransmissions can be reduced when a channel code is applied. However, the channel code introduces additional overhead. This decrease in the overall throughput efficiency due to a lower code rate may not justify the application of such codes when the forward error correction coding gain is not high enough. To illustrate, suppose we have at our disposal a rate 1/2 convolutional code. If the channel throughput is already greater than 50% before this code is applied, there will be a net loss in the overall throughput efficiency when this channel code is actually applied. We saw in Section 5.1 that there is very limited coding gain on very slow fading channels. For these reasons, we do not use a channel code on the Type-I data packets.

By systematic avoidance of deep fades, there should be an average improvement in the received signal. Just how quickly the switch can be actuated after a switch request is made obviously affects the system performance. Likewise, the number of transmit antennas, packet size, channel fading rate, multiplex ratio, and the quality of the feedback channel can all significantly influence the system performance. We are concerned with both the throughput efficiency of the ARQ system for the Type-I information and the average BER performance of the Type-II information.

In a practical system, actually switching between transmit antennas requires turning on and off the power amplifiers for these different antennas. This should be avoided. We will develop a systematic approach to achieve the same diversity benefits using a different transmit antenna diversity technique to be described in Chapter 6. In that scheme, several antennas are transmitting simultaneously with a predetermined set of phase angles at the transmit antennas. The set of phase angles are simply changed to another predetermined set of values when it is necessary to switch to an independent fading branch.
5.2.2 Performance Comparisons

In this subsection, different combinations of the parameters that are important in determining the performance of the switched transmit antenna with feedback protocol are examined. We simplify the discussion by assuming a perfect feedback channel, so there is no uncertainty as to when the base station should use a different transmit antenna.

(a) Number of Transmit Antennas and Fading Rate

In this particular set of simulations, only the Type-I data packets are transmitted (i.e., \( L = 0 \) in Fig. 5-6). For illustration purposes, we assumed zero switch actuation time, i.e., as soon as a Type-I packet arrives at the receiver, the transmitter knows the acknowledgment immediately and can switch to a different transmit antenna the instant a retransmission on the same packet commences (i.e., \( t_1 \) on Fig. 5-6). The result so obtained should be better than the practical case where nonzero switch actuation time is included.

Fig. 5-7 depicts the channel throughput efficiency of the ARQ system with different numbers of transmit antennas on the very slow time-varying channel \((v = 1 \text{ km/hr}, f_c = 900 \text{ MHz}, 1/T_d = 8 \text{ kbits/sec})\). The channel throughput efficiency improves steadily with the number of transmit antennas. A large number of antennas (more than 12) are required to approach the performance achievable with a theoretically infinite number of diversity branches. This is to be expected since the performance should increase so long as the time it takes to sequentially switch through all the antennas is shorter than the coherence time of the channel [36].

On these slowly fading channels, very large gains can be obtained with this switched transmit antenna with feedback scheme. In contrast, performance gains are rather small on fast fading channels. The counterpart of Fig. 5-7 for \( v = 80 \text{ km/hr} \) with the other channel parameters unchanged is shown in Fig. 5-8. At this high speed, adding more than two transmit antennas provides no extra diversity benefits. This is because the channel changes more quickly at this higher speed. Let us compare the
Figure 5-7: Different Numbers of Transmit Antennas on Slow Fading Channels

difference between a scheme using three transmit antennas with a scheme using two transmit antennas. Both schemes are fading at 80 km/hr. In the scheme with three transmit antennas, suppose after sequentially switching through two transmit antennas, both diversity branches are found in a deep fade. Then switching to the third transmit antenna is no better than simply returning to the first transmit antenna because the fading process on the first diversity branch has become independent from that on the same diversity branch the last time it was used.

(b) Multiplex Ratio and Switch Actuation Time

Switch actuation time is very important in determining the overall system performance. The intuitive reason is because the received signal will continue to fade below a usable signal strength threshold between the instant it is detected in a deep fade and when the transmitter actually switches to a different antenna. Even more importantly, the multiplex ratio $L$ determines how frequently the transmitter polls the
channel state information and ultimately determines the shortest time interval between a switch of antennas. For this reason, we focus on the multiplex ratio for now.

The entire system in Fig. 5-5 is simulated with two transmit antennas at the slow fading limit with two different values of $L$. The exact channel variation parameters are $v = 1$ km/hr, $f_c = 900$ MHz, and 8 kbits/sec. Zero switch actuation time is assumed. The channel throughput efficiency of the Type-I information and the average BER of the Type-II information are depicted in Fig. 5-9 and Fig. 5-10, respectively. The results indicate that the incremental diversity gain decreases with $L$. The reason is that the switched transmit antenna with feedback scheme derives its gain from its ability to switch to an independent diversity branch when the current diversity branch is detected in a deep fade. A retransmission request comes every $L$ Type-II data packets. When $L$ is increased, the transmitter polls the channel state information less frequently, and consequently uses the same diversity branch for an extended
period of time regardless of whether the current diversity branch is in a deep fade or not. Hence the diversity gain observed earlier diminishes.

(c) Collection of Simulation Results

Now that we understand how the number of transmit antennas, multiplex ratio, and channel fading rate affect the overall system performance, several sets of simulation results are collected here for evaluating the transmission protocol. Simulation results are provided for two channel fading rates (1 km/hr and 10 km/hr, both at 900 MHz carrier frequency and 8 kbits/sec transmission rate), two packet sizes (128 bits and 256 bits), a multiplex ratio of 3, and four sets of transmit antenna configurations (1 antenna, 2 antennas, 6 antennas, and a theoretically infinite number of antennas).

The overall system performance is measured by the throughput efficiency of the ARQ system and the average BER for the Type-II information.

Fig. 5-11 and Fig. 5-12 compare the system performance at the two channel fading
rates with $N = 128$ and $L = 3$. It is shown in these two figures that the switched transmit antenna with feedback protocol can be very effectively used on slowly time-varying channels to combat multipath fading. Diversity gain becomes very small at 10 km/hr. This is because the coherence time ($\tau_c$) of the channel at 10 km/hr and 8 kbits/sec is approximately the equivalent of one packet duration of 128 bits. The coherence time of the channel was defined in Chapter 2 to provide a quantitative estimate of the time duration over which the received signal is strongly correlated. With $L = 3$, the fading process on the same diversity branch is essentially uncorrelated with the fading process three frames earlier. Hence switching to an independent diversity branch is no better than to staying on the same diversity branch.

Fig. 5-13 and Fig. 5-14 display the effects of packet size on the overall system performance. Two different packet sizes are used, 128 and 256. Since the same speed (1 km/hr), carrier frequency (900 MHz), and transmission rate (8 kbits/sec) are used, a multiplex ratio of 3 used with a packet size of 256 bits is the equivalent of 6 packets.
Figure 5-11: Channel Throughput at $v = 1 \text{ km/hr}$ & $v = 10 \text{ km/hr}$

at a packet size of 128 bits. The incremental diversity gains for the system using 256 bit packet size are not as large as the corresponding system with a smaller packet size mainly because of this equivalent larger multiplex ratio.

For the results given above, should a switch of antenna be needed, the switch is actuated only at the instant a Type-I packet is retransmitted (i.e., $t_3$ on Fig. 5-6). In that case, the switch actuation time is taken to be precisely $L$ packet durations. We will compare the results obtained above with the results obtained with zero switch actuation time. In a practical system, the switch actuation time is between zero and $L$ packet durations. In that regard, when the switch actuation time is the only variable parameter in the system, we can treat the results obtained with switch actuation time of $L$ packets as the lower performance bound and the results achieved with zero switch actuation time as the upper performance bound. It is obvious that the switch actuation time has no effect on the channel throughput efficiency of the ARQ system because, independent of the switch actuation time, the same diversity branch will be
used to transmit a particular Type-I packet during that particular time slot. The average BER of the Type-II information, however, depends heavily on the switch actuation time. The faster the switch can be actuated, the better (see Fig. 5-15).

Since the delay spread for indoor radio channels is small, intersymbol interference is typically not a serious problem, higher transmission rate than 8 kbits/sec may be used. The transmission protocol is simulated for 1 km/hr at 900 MHz and 64 kbits/sec using $L = 3$ and $N = 128$. Zero switch actuation time is assumed. The results are given in Fig. 5-16 and Fig. 5-17. By comparing curve (2) in 5-10 ($L = 1$) with curve (2) in Fig. 5-17 ($L = 3$), we come to the conclusion that with other parameters unchanged, at a higher transmission rate, the multiplex ratio can be increased correspondingly without degrading the average BER performance on the Type-II information.

Finally, we use the result above to briefly address the delay constraint issue. In this protocol, we transmit two types of information: the Type-II information is
transmitted at a constant bit rate using a FEC code and the Type-I information is delivered (i.e., accepted by the receiver) at a variable rate using ARQ. Since the Type-II information is transmitted with a rate 1/2 convolutional code, and there are two additional overhead bits per Type-II packet to synchronize the modem, therefore the Type-II information bits are accepted by the receiver at a constant rate that is approximately 50% of the transmission rate (for large packet sizes, the two bits per packet overhead has little impact). For given transmission reliability requirements (e.g., average BER) and practical hardware constraints (e.g., transmission rate), we cannot decrease the code rate nor increase the transmission rate to meet the delay constraints. The channel throughput efficiency of the ARQ system, however, can be used to meet the delay constraints of the intended applications by designing efficient schemes. To illustrate, in Fig. 5-11, at 10 dB SNR, the channel throughput is 32% for one transmit antenna for the 1 km/hr case. With two transmit antennas, 45% throughput efficiency is achieved. With six transmit antennas, a throughput of 65%
is obtained. These results are compared to Fig. 5-17, where identical parameters as in Fig. 5-11 are used except the transmission rate was increased 8 times to 64 kbits/sec. With two transmit antennas, 52% throughput is achieved at 10 dB. 72% can be obtained with six transmit antennas. With a higher throughput and a higher transmission rate, the delay constraints of the intended applications can be more easily satisfied.

5.3 Summary

We have developed a novel switched transmit antenna with feedback protocol in this chapter. The implementation of this protocol is simplified by two efficient usages of the channel resources: first, the channel state information is provided to the transmitter by the retransmission requests; second, the explicit received signal strength measurements at the receiver is indirectly done by a error detection module. It is
best to use this protocol on slow fading channels with relatively small packet sizes and multiplex ratios. With other parameters unchanged, the switch actuation time does not affect the throughput efficiency of the ARQ subsystem but it plays an important role in determining the average BER performance on the Type-II information. Significant performance gains can be achieved for the slow fading channels with a large number of transmit antennas.

Figure 5-15: Switch Actuation Time on Average BER
Figure 5-16: Channel Throughput at 64 kbits/sec Transmission Rate
Figure 5-17: Average BER at 64 kbits/sec Transmission Rate
Chapter 6

Repetition-Type FEC Protocol

We saw in early chapters that the throughput efficiencies of ARQ systems change with channel conditions. Furthermore, feedback channels are needed in the implementation of such systems. When it is difficult or impractical to provide a feedback channel, alternative transmission protocols to the one given in Chapter 5 are needed. The current chapter deals with a class of repetition-type FEC schemes for the transmission of JPEG coded still images over very slow fading channels.

The basic scenario to be treated in this chapter is as follows. In the typical JPEG coded still images, there is 10 to 20 times more Type-II information than Type-I information to be transmitted. A feedback channel is unavailable. The relative importance of these two types of data is as discussed in Chapter 4. The simplest approach is to transmit the Type-I information and the Type-II information at nonoverlapping time intervals. Each Type-I packet is transmitted for a predetermined number of times and then code-combined at the receiver, giving rise to a simple repetition code. The Type-II information is transmitted only once. There is very little diversity (e.g., time diversity) benefit for this scheme on the very slow fading channel because all repeated transmissions of a Type-I packet are subject to virtually identical fading. More efficient schemes that use multiple transmit antennas will be treated in this chapter.

The method of combining noisy packets to provide a reliable estimate of the transmitted packet is not new. Reference [32] uses a maximum likelihood decoding
approach for combining an arbitrary number of noisy packets. The proposed method in reference [32] matches the code rate (controlled by the number of repeats) to the dynamic channel conditions by applying an error detection code to the transmitted packets in addition to the repetition code, so that the transmitter can determine when to stop repeating the same packet with the aid of a feedback channel. Similar studies are presented in references [29]-[31], [33], and [34]. In all these studies, a significant improvement in channel throughput is obtained in comparison to the scheme where the receiver waits until an error-free packet is received without code combining.

The schemes above all use a feedback channel. This is different from the schemes to be developed in this chapter where a feedback channel does not exist. In Section 6.1 we provide simulation results for the simple scheme described above to provide a reference for later comparisons. Since there is more Type-II information than Type-I information, a simple way to provide limited time diversity benefits [26] to the Type-I information by multiplexing its transmissions with the transmission of Type-II information is discussed towards the end of Section 6.1. However, as will be shown, time diversity cannot be effectively used on very slow fading channels. Other means for obtaining diversity benefits are needed. In Section 6.2, a transmit antenna diversity technique called induced fast fading [11][38][21] is considered. It will be shown that this technique can effectively provide time diversity on very slow fading channels. The bit error rate performance improvement with the induced fast fading technique is demonstrated with simulations. A systematic approach is then given to provide $M$ uncorrelated copies of the same transmitted Type-I packet for combining at the receiver with this diversity technique using $M$ transmit antennas. Finally in Section 6.3 we develop efficient repetition-type FEC protocols for JPEG coded still image transmission over very slow fading channels.

6.1 Simple Scheme

In this simple scheme there is only one transmit antenna. Each Type-I packet is transmitted $K$ times, where $K$ is a predetermined number. The analog received sig-
Signals (soft decisions) from successive transmissions of the same packet are combined to obtain a rate $1/K$ repetition-type error correction code. The resulting signal is then decoded by the receiver and supplied to the data sink. For analysis, a failure is declared when the decoded packet contains errors. Type-II packets are transmitted only once with a rate $1/2$, memory 2 convolutional code. 4-DPSK is used for both data packets. The transmissions for Type-I and Type-II packets take place at nonoverlapping time intervals.

![Simple-Repetition Code using 1 transmit antenna](image)

Figure 6-1: Average Failure Rate on Slow Fading Channels

For the results in Fig. 6-1, the channel variation parameters are $v = 1$ km/hr, $f_c = 900$ MHz, $1/T_d = 8$ kbits/sec. As expected, the failure rate decreases when the code rate is decreased to result in a more powerful error correction code. However, it is clear from Fig. 6-1 that even with a rate 1/8 repetition code and a SNR of 20 dB, the failure rate is still unacceptably high for JPEG applications ($\approx 2\%$). 2% errors at the important Type-I markers is catastrophic. No recognizable image can be received with this transmission scheme.
In contrast, when the channel is fading must faster, the decrease in failure rate is much drastic with a decrease in the code rate. For the results in Fig. 6-2, the same parameters given above for the slow fading channel were used except the speed was increased 80 times to 80 km/hr. When the channel is varying quickly due to high speeds (Fig. 6-2), the correcting power of the repetition code results in a dramatic improvement in the average failure rate in comparison to the slowly fading channels (Fig. 6-1), where the improvement is gradual. The average failure rate is nearly zero at 20 dB for four repeats at 80 km/hr. In contrast, at 1 km/hr with 20 dB and eight repeats, the average failure rate is still about 2%.

For slowly fading channels, the situation depicted in Fig. 6-1 may be improved slightly by exploiting time diversity. It is inefficient to finish transmitting the Type-I information before transmitting the Type-II information because time diversity is left unexploited. In the JPEG compressed data stream, there is more Type-II information than Type-I information. The transmissions of Type-I information can be
multiplexed with the transmissions of Type-II information, thereby providing limited
time diversity benefits to the Type-I information. Again, we define the multiplex ratio
$L$ as the number of Type-II packets transmitted for every transmission of a Type-I
packet. Note that in such a scheme, extra memory storage is needed to store the
packets prior to combine decoding and this storage requirement increases with $K$.

![Simple-Repetition Code using 1 Transmit Antenna](image)

$N = 128$
$1$ km/hr, $900$ MHz, $8$ kbits/sec

1. $L = 0$
2. $L = 10$
3. $L = 20$
4. $L = 50$

Figure 6-3: Average Failure Rate for Different Multiplex Ratio $L$

Fig. 6-3 illustrates the resulting failure rates when variable amount of time di-
versity is achieved for the Type-I information by changing the multiplex ratio. Each
Type-I packet is repeated twice in this set of simulations, where the two copies of the
packet are separated by $L$ Type-II packets. Time diversity can be quantitatively re-
lated to the coherence time ($\tau_c$) of the channel. At $v = 1$ km/hr, $f_c = 900$ MHz, and
$8$ kbits/sec, $\tau_c$ is the equivalent of $1600$ bit durations, or about $12$ packet durations
for a packet size $N = 128$. When the multiplex ratio is very large in comparison to
the coherence time of the channel ($L = 50$), the fading statistics on the two copies
of the same packet to be combined are independent, thereby achieving the maximum
diversity benefits for two repeats. The question is whether we can afford such wide time separations in a practical transmission scheme. Typically, there is 10 to 20 times as much Type-II information as Type-I information for JPEG coded still images. If we allow each Type-I packet to be repeated twice, \( L \) should be no greater than 10. With such narrow time separations between consecutive transmissions, Fig. 6-3 clearly indicated there is virtually no time diversity gain.

### 6.2 Induced Fast Fading Diversity

The study in the previous section indicated for slow fading channels that time diversity benefits are difficult to achieve. It was also shown in Section 5.1 that the coding gain for a rate 1/2, memory 2 convolutional code was negligible at very slow fading rates with limited interleaver sizes. Various techniques do exist that can achieve a greater interleaving depth [12, pp. 348–354]. We could, for example, interleave several adjacent packets, but it is not very practical to increase the degree of interleaving because it increases the processing delay and requires extra memory. Besides, Fig. 5-4 in Chapter 5 shows very limited gain by interleaving several adjacent packets to allow a larger interleaver size. Other techniques to increase the coding gain are therefore extremely desirable. Induced fast fading diversity is such a technique. This technique was developed independently by Weerackody [11] [38] and by Adachi et al. [21].

We first describe the principles behind this technique. The average BER performance improvement employing this diversity technique is then demonstrated with computer simulations. We also provide a systematic approach for obtaining diversity order \( M \) with \( M \) transmit antennas with this technique.

#### 6.2.1 Principles of Operation

A detailed description of this scheme is given in [11] and [38]. Only the intuitive physical arguments are reproduced here. The simulation results to be given later complement the results given in [11] at faster inherent channel variations.

The basic requirements for this scheme are multiple transmit antennas and a suit-
able channel code. All the transmit antennas send the same signal. But a slow time-varying phase offset is provided to each transmit antenna. Suppose two transmit antennas are used in this scheme. Before the phase offsets are introduced, typical situations where deep fades can occur are illustrated in Fig. 6-4. In Fig. 6-4a, the received signals from the two antennas interfere destructively. Due to the slow time-varying characteristics of these indoor radio channels, it is likely that the receiver will be in this deep fade for a long period of time. In Fig. 6-4b, the received signal is weak because both signals are weak. Deep fades such as Fig. 6-4a can be significantly reduced when the phase offsets are introduced to the transmit antennas because the resulting rotating phasors cancel each other only a small fraction of the time. Although deep fades illustrated in Fig. 6-4b cannot be reduced by rotating the phasors, the probability that all received signals are small decreases with the number of transmit antennas. Hence, significant improvements in transmission reliability can be achieved.

One major concern in the implementation of such diversity schemes is that the induced channel variations should not be so rapid as to cause the demodulator to loses phase coherence, in which case the demodulation error would dominates the overall performance.

Using the same system model as in [11], the signal to be transmitted is denoted by $u(n)$, where $n$ is the discrete time index. The signal $u(n)$ is then amplitude weighted

Figure 6-4: Typical Scenarios for Weak Signal Reception
by $A_i(n)$ and given a phase offset of $\theta_i(n)$ on the $i^{th}$ transmit antenna. The inherent Rayleigh fading multiplicative noise $r_i(n)$ on the $i^{th}$ diversity branch is a complex Gaussian random processes with zero mean. Ignoring the effects of the additive noise, the received signal from the $M$ transmit antennas is

$$s(n) = \sum_{i=1}^{M} A_i(n) e^{j\theta_i(n)} r_i(n) u(n) .$$

(6.1)

We can further denote the complex envelope of the received signal by

$$\beta(n) = \sum_{i=1}^{M} A_i(n) e^{j\theta_i(n)} r_i(n) = \sum_{i=1}^{M} \beta_i(n) .$$

(6.2)

The following combinations of the transmitter induced amplitude and phase offset were used in the simulations in reference [11]:

$$A_i(n) = \frac{1}{\sqrt{M}}$$

$$\theta_i(n) = 2\pi f_i n T_d$$

(6.3)
where \( f_i \) is the induced frequency offset on the \( i^{th} \) transmit antenna. For this particular choice of amplitude weights, the total transmitted power is the same independent of \( M \). Reference [11] selected the frequencies \( f_i \) according to

\[
    f_i = \frac{-(M - 1)}{2} f_\Delta + (i - 1)f_\Delta ,
\]

where \( f_\Delta \) is a small percentage of the transmission rate. For example, \( f_\Delta \) can be chosen to be 2\% of 8 kbits/sec transmission rate, providing \( \pm 80 \) Hz frequency offsets to the two transmit antennas.

### 6.2.2 BER Improvement

In this subsection we provide computer simulation results to illustrate the improvement in the average BER performance when the simulated fast fading diversity technique with a rate 1/2, memory 2 convolutional code is used on a very slow fading channel. The improvements are determined for different frequency offsets and numbers of transmit antennas. The simulations are performed for very slow fading channels (\( v = 1 \) km/hr, \( f_c = 900 \) MHz, and at 8 kbits/sec) to complement the results given in [11]. The communication system used in the following sets of simulations is given by Fig. 5-3 in Chapter 5.

(a) Different Frequency Offsets

Fig. 6-6 illustrates the effects of applying different frequency offsets to two transmit antennas on the average BER performance. The frequency offsets are expressed as a fraction of the data rate. As larger and larger frequency offsets are introduced to the transmit antennas, because of the time diversity gain, the correcting power of the convolutional code increases. Correspondingly, the average BER decreases. However, when the frequency offset becomes too large, the 4-DPSK demodulator functions unreliably to cause the average BER to increase. At 2\% frequency offset, more than 10 dB gain is achieved at the \( 10^{-3} \) BER level.
(b) Different Numbers of Transmit Antennas

Fig. 6-7 illustrates the average BER performance with different numbers of transmit antennas. The frequency offsets are indicated on the figure. The average BER decreases as more and more transmit antennas are used.

6.2.3 Obtaining Diversity Order $M$ with $M$ Transmit Antennas

In this subsection we provide a systematic approach for obtaining diversity order $M$ with $M$ transmit antennas. Perhaps the best way to introduce this method is by starting with a simple configuration with $M = 2$. By using simple geometrical interpretations and a little inductive reasoning, the results can be generalized.

The problem to be considered is as follows. We have at our disposal two transmit antennas. The channel is fading very slowly so that time diversity cannot be effectively
applied. In this transmission protocol, a Type-I packet is transmitted twice and then combined at the receiver to be decoded. If the two transmissions of the same packet are subject to nearly identical fading, the resulting repetition code is not very effective, as was shown in Section 6.1. The question is can we do better with two transmit antennas.

An elegant solution is as follows. The first time a packet is transmitted, we introduce no phase offset to either antenna, i.e., \( \theta_1 = \theta_2 = 0 \) (see Fig. 6-5). The complex envelope of the first reception is \( (\beta_1 + \beta_2) \), where \( \beta_1 \) and \( \beta_2 \) are the complex Rayleigh fading coefficients on the first and second diversity branch, respectively. On the second transmission, phase offsets of \( \theta_1 = 0 \) and \( \theta_2 = \pi \) are introduced to the two transmit antennas. The complex signal envelope is now \( (\beta_1 - \beta_2) \). Since \( (\beta_1 + \beta_2) \) and \( (\beta_1 - \beta_2) \) are statistically independent (i.e., \( \langle (\beta_1 + \beta_2)(\beta_1 - \beta_2)^* \rangle = 0 \)), the received signals at two time slots are completely uncorrelated. In short, with this diversity scheme, one of the transmit antennas maintains a zero phase offset while the other

Figure 6-7: Average BER for Different Numbers of Transmit Antennas
one is given an oscillating $0, \pi, 0, \pi, \ldots$, phase offset. Note that the phase offsets are held constant for the duration of a packet.

This method can be generalized to $M$ transmit antennas. An equation is obtained in [38] for the phase offset to be introduced on the $i^{th}$ transmit antenna during the $l^{th}$ transmission,

$$\theta_i(l) = \frac{2\pi}{M} (i - 1)(l \mod M) \quad i = 1, 2, 3, \ldots, M$$

(6.5)

where $l \mod M$ represents the modulo $M$ value of $l$. With this generalized diversity scheme, $M$ independent copies of the transmitted signal can be received in $M$ signaling periods to achieve diversity order $M$.

It was mentioned in Chapter 5 that switching on and off power amplifiers is problematic. The scheme we just developed here can achieve similar diversity benefits as the switched transmit antenna diversity scheme without switching on and off power amplifiers. In this new scheme, we could transmit using one set of phases for the $M$ transmit antennas and then switch to a different set of phases in response to a switch request. Since the effect of using a different set of phases for the transmit antennas is equivalent to obtaining an independent diversity branch, the diversity benefits of the switched transmit antenna with feedback scheme as discussed in Chapter 5 can be achieved with the new technique.

### 6.3 Improved Repetition-Type FEC Schemes

In this section we consider the application of the diversity technique described in Section 6.2 to the transmission of JPEG coded still images over very slow fading channels. The Type-I information is transmitted using a repetition code while the Type-II information is transmitted only once.

Fig. 6-8 compares the simple scheme described in Section 6.1 to the scheme with two transmit antennas. The phase offsets provided to the transmit antennas are given by (6.5) with $M = 2$. One of the transmit antennas receives zero phase offset while
the other antenna receives an oscillating $0, \pi, 0, \pi, \ldots$, phase offset. The channel fading parameters are given in the figure. Significant improvement in the failure rate with this diversity technique is observed in Fig. 6-8. Note that for 16 dB or higher SNR values, a rate 1/2 repetition code with this diversity technique achieves a lower failure rate than a rate 1/8 repetition code without this diversity technique. We also point out that curve (2) in Fig. 6-8 for combining two independent copies of the same packet using this generalized diversity scheme agrees with curve (4) in Fig. 6-3 with the largest multiplex ratio ($L = 50$). This is an intuitively satisfying result since in both cases, the order of diversity is the same.

The failure ratio can be further improved when four transmit antennas are used. In that case, four independent copies of the same packet can be received in four packet signaling intervals by providing the phase offsets given by (6.5) with $M = 4$ to result in a rate 1/4 repetition code. The resulting repetition code is much more effective than the rate 1/4 code with two transmit antennas. The difference is between a

Figure 6-8: Improved Success Rate with Tow Transmit Antennas
diversity of order 4 and a diversity of order 2 (Fig. 6-9). In fact, very close to 100% success ratio is achieved with SNR values greater than 20 dB using four transmit antennas in a rate 1/4 repetition code.

For completeness, we determine if this scheme may be used on fast fading channels. Fig. 6-10 depicts the results using two transmit antennas and four transmit antennas with rate 1/2 and rate 1/4 repetition codes, respectively, on the fast fading channels ($v = 80 \text{ km/hr}$, $f_c = 900 \text{ MHz}$, at 8 kbits/sec). Diversity gain is still possible with two transmit antennas, but the gain is less impressive with the rate 1/4 repetition code since there is inherent diversity on the channel at this high speed before further diversity is added with the transmit antennas.

Finally, we demonstrate the overall system performance of this repetition-type FEC transmission scheme. In the simulations, each Type-I packet is repeated $K$ times. The Type-II information is transmitted only once. The phase offsets given by (6.5) (e.g., constant phase offset over a packet duration) apply when a Type-I
packet is transmitted while the transmission of a Type-II packet receives the phase offsets given by (6.3) (e.g., continuously time varying even within a packet duration). One Type-I packet is transmitted for every $L$ Type-II data packets in a multiplexed manner as described earlier.

We first consider the effect of different multiplex ratio on the system performance. In Fig. 6-11 and Fig. 6-12, two transmit antennas are used in a rate 1/2 repetition code using the diversity technique developed in Section 6.2 for different values of $L$. Since a diversity of order 2 is achieved regardless of $L$, the average failure rate is the same for all values of $L$. As was mentioned in Chapter 4, suppose our objective is to make sure that nearly 100% of the Type-I information is received error-free while the Type-II information be received with an average BER of no more than $10^{-4}$. From Fig. 6-11 and Fig. 6-12, the objective may be achieved with two transmit antennas using 25 dB transmitted power. In contrast, more than 35 dB transmitted power is required to achieve the same objective without this induced fast fading diversity. 

Figure 6-10: Diversity Gain on Fast Fading Channels
Figure 6-11: Average Failure Rate for Different Multiplex Ratios

As a final evaluation of this transmission protocol, simulations were performed for two different speeds \( (v = 1 \text{ km/hr and } v = 10 \text{ km/hr}) \), both at 900 MHz carrier frequency and 8 kbits/sec transmission rate. Three different transmit antenna configurations were used: two, three, and four transmit antennas used in a rate 1/2, 1/3, and 1/4 repetition code for the Type-I data packets, respectively. The results are illustrated in Fig. 6-13 and Fig. 6-14. With four transmit antennas, the objective of achieving nearly 100% success rate for the Type-I information and an average BER of \( 10^{-4} \) for the Type-II information may be achieved with 20 dB transmitted power, a 5 dB saving in power compared to two transmit antennas.

We have been comparing the protocols in terms of the failure rate for the Type-I information and the average BER for the Type-II information. We now briefly mention the delay constraint issue in the current context. Due to the transmission error effects described in Chapter 4, the protocol in this chapter is of little merit for JPEG
applications when delay constraints are too stringent such that the failure rate for the Type-I information is unacceptably high due to high code rates (small number of repeats) and small SNR values. For example, we can compare the rate 1/3 and rate 1/4 repetition code in Fig. 6-13 in the very slow fading case, curve (4) and curve (6), respectively. Suppose we are limited to 15 dB power. It is true that the rate 1/3 code will meet the delay constraints with a wider margin than the rate 1/4 repetition code, but the rate 1/4 code will obviously provide better image quality because of its lower failure rate on the Type-I markers. Therefore, when the repetition-type FEC protocol is used to transmit JPEG coded still images, delay constraint must be of secondary importance. The most important goal is to achieve the lowest failure rate for the Type-I markers.
Figure 6-13: Average Failure Rate at $v = 1$ km/hr & $v = 10$ km/hr
Figure 6-14: Average BER at $v = 1$ km/hr & $v = 10$ km/hr
Chapter 7

Laboratory Implementation

In this final chapter before the conclusion, we describe laboratory implementations of error control schemes for the transmission of JPEG coded still images. Implementation details are given in Section 7.1. The transmission protocol is described in Section 7.2. Received images and quantitative performance comparisons are presented in Section 7.3.

7.1 Communication System Hardware

A block diagram of the communication system hardware is depicted in Fig. 7-1. It consists of transmitter and receiver pairs for the forward channel (as shown) and the feedback channels (not shown). Following the logical data path, two types of data packets are processed separately as shown in the transmitter section of Fig. 7-1 and then multiplexed prior to transmission. A Type-I packet is transmitted for every three Type-II packets ($L = 3$). All these functions are performed on one side of SURFboard, a DSP module with six AT&T DSP32C processors. An inside view of SURFboard comes shortly. The 4-DPSK encoded discrete time signal is then convolved with a square root Nyquist pulse in the D/A converter to generate a continuous-time signal for transmission over the continuous-time radio channel. In the RF section, the in-phase and quadrature components are modulated with two carriers that are 90 degrees out of phase with one another. The mixer produces the sum and difference frequencies.
Figure 7-1: Block Diagram of Communication System Hardware
The bandpass modulated signal (900 MHz band) is amplified and then transmitted. Although one transmitter antenna is shown in Fig. 7-1, for diversity schemes, three transmit antennas are present. Each antenna transmits about 10 mW power. The received signal is first amplified with a low noise amplifier. It is then filtered and down converted to the baseband frequency for demodulation. The continuous-time signal is sampled to obtain a discrete-time signal. Note that the delay spread for indoor channel is too small to cause significant intersymbol interference. The discrete-time signal is demultiplexed and processed separately as shown in the receiver section of Fig. 7-1. An ACK or NAK is sent to the transmitter via the feedback channel after a Type-I packet is detected for error. The feedback channel uses the 920 MHz band.

Figure 7-2: SURFboard DSP Module

SURFboard is a DSP module with six programmable AT&T DSP32C Digital Signal Processors [37] with an interface that conforms to the industry standard VME bus. The SURFboards are directly interfaced to two SGI workstations. A SURFboard is
divided into receiving and transmitting sections as shown in Fig. 7-2. Since these are programmable devices, the transmission protocol can be very simply modified by updating the software and downloading the DSP. 1 Mbytes of DRAM memory is shared by the receiving and transmitting sections on one SURFboard. The acknowledgments received by the receiving section can be accessed by the transmitting section in this shared memory space. Two SURFboards implement a full duplex communication system nicely.

7.2 Coded Still Image Transmission Protocol

The images to be transmitted are first processed by a module which separates the Type-I information from the Type-II information. It is sufficient to use sequence number modulo 2 to identify the Type-I data packets because this is a Stop-and-Wait ARQ protocol. A discussion on simple ARQ protocol implementations is given in [14, pp. 66–85]. Both the sequence number (SN) and request number (RN) are initialized to zero upon entering the protocol. The forward channel receiver upon receiving an ARQ packet does the CRC-16 error detection. If no error has been detected, SN is extracted from the packet and compared to RN. If RN is different from SN, the packet is rejected; otherwise, the packet is accepted. The forward channel receiver then updates RN in a DRAM location where the feedback channel transmitter checks for RN to be transmitted. When the acknowledgment containing RN arrives at the feedback channel receiver, the receiver performs error detection on the acknowledgment and extracts RN when no error has been detected. If RN is greater than SN, RN is stored in a DRAM location where the forward channel transmitter looks for its SN. If RN is greater than SN, the forward channel transmitter increments SN to RN and sends a new packet; otherwise, it retransmits the same packet.

The same rate 1/2, memory 2 convolutional code is applied to both Type-I and Type-II data packets in this particular implementation. In the discussions earlier, a convolutional code was not used for the Type-I data packets. The reason for this
modification is that when the induced fast fading diversity technique is applied in
the laboratory, the transmit antennas receive their respective frequency offsets from
several signal generators at the RF stage. A channel code such a convolutional code
must be applied to both Type-I and Type-II data packets in order to capitalize on
the time diversity benefit achieved with this induced fast fading technique. To see
this more clearly, suppose the Type-I data packets are transmitted without a channel
code. The continuously time-varying frequency offsets will be in effect throughout the
experiment. The resulting fast fading characteristics will cause the 4-DPSK demod-
ulator to provide unreliable outputs in the absence of the channel code. The ARQ
protocol we implemented is a simple type-I hybrid ARQ protocol. We can, in accor-
dance with the earlier discussions, provide constant phase offsets during the Type-I
data packets in DSP without the signal generators. In that case, a convolutional code
is not required for the Type-I data packets.

Post-processing on the received image data are done to detect whether any of
the Type-II information has been accidentally converted to a Type-I marker due to
transmission errors. These unexpected markers are replaced with random sequences
when found.

7.3 Experimental Results

In the laboratory, the transmitters and receivers are located in two adjacent rooms
with metallic walls. The propagation paths of the transmitted electromagnetic waves
are further obstructed by typical laboratory equipment and office furniture. The
transmitters are stationary. The transmission rate is approximately 8 kbits/sec. The
receivers are moving very slowly at a speed of no greater than 1 km/hr. Each ex-
periment consists of transmitting 10 JPEG coded still images. The total size of
these images is about 1 Mbytes. Each image is about 100 kbytes. Experiments were
performed with and without the induced fast fading diversity scheme. When the
induced fast fading diversity scheme is applied, the Type-I data packets are trans-
mitted with the aforementioned ARQ protocol in conjunction with the induced fast
fading diversity scheme. The performance comparisons are based on the average channel throughput efficiency for the ARQ system, the average BER for the Type-II information, and the number of images distorted beyond recognition (to be clarified later).

Three transmit antennas were used in the experiments. The same average power (30 mW) was transmitted with and without the induced fast fading diversity scheme. When diversity is applied, one of the transmit antennas receives zero phase offset while the other two receive independent time-varying frequency offsets to be specified when the results are presented.

An example of what we called distorted beyond recognition (DBR) image is illustrated in Fig. 7-3. This particular image was received when no diversity was applied. All markers (both Type-IA and Type-IB) were protected with an ARQ system. Because the errors in Fig. 7-3 were very bursty, several consecutive Minimum Coded Units (MCU) were corrupted, and the JPEG image decoder failed to decode the image beyond the first instance of such long burst errors. Fig. 7-4 was also received without
the induced fast fading diversity scheme. The BER of this image (BER = 0.002214) is not very different from that in Fig. 7-3 (BER = 0.003403). However, errors occur more uniformly throughout Fig. 7-4 with the defects shown as blocks with distorted gray scales. These two images clearly illustrate the need to use antenna diversity techniques to improve the BER of the Type-II information and to randomize the errors to avoid corrupting consecutive MCU.

Fig. 7-5 illustrates the experimental results with and without the induced fast fading diversity scheme. From Diversity 1 to Diversity 3, successively larger frequency offsets are introduced. Exceptional image qualities were received in Diversity 2. All the images were virtually error-free and there was no visible defects in any of the 10 images received. Channel throughput was close to 100%. Average BER for the Type-II information was at the $10^{-6}$ level. When the frequency offsets were too large (Diversity 3), the 4-DPSK demodulator was unable to track the phase changes between consecutive data symbols and the overall system performance became worse than no diversity.
<table>
<thead>
<tr>
<th>Frequency Offset given to each antenna in Hz</th>
<th>No Diversity</th>
<th>Diversity 1</th>
<th>Diversity 2</th>
<th>Diversity 3</th>
</tr>
</thead>
<tbody>
<tr>
<td>$f_1 = 0$</td>
<td></td>
<td>$f_1 = 22$</td>
<td></td>
<td>$f_1 = 500$</td>
</tr>
<tr>
<td>$f_0 = 0$</td>
<td></td>
<td>$f_0 = 0$</td>
<td></td>
<td>$f_0 = 0$</td>
</tr>
<tr>
<td>$f_2 = 0$</td>
<td></td>
<td>$f_2 = 46$</td>
<td></td>
<td>$f_2 = 600$</td>
</tr>
</tbody>
</table>

| Channel Throughput Efficiency              | 89.0 %      | 98.5 %      | 99.8 %      | 67.9 %      |

| Overall BER for Type-II Information        | 0.00521     | 0.000851    | 0.000001    | 0.00825     |

| Percentage of DBR Images                   | 50 %        | 10 %        | 0           | 60 %        |

Figure 7-5: Performance Comparisons With and Without Diversity

<table>
<thead>
<tr>
<th>Image No.</th>
<th>Incorrect Bits</th>
<th>% of total incorrect bits</th>
<th>Incorrect Bits</th>
<th>% of total incorrect bits</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>133</td>
<td>0.2 %</td>
<td>339</td>
<td>7 %</td>
</tr>
<tr>
<td>2</td>
<td>0</td>
<td>0 %</td>
<td>178</td>
<td>3.7 %</td>
</tr>
<tr>
<td>3</td>
<td>52</td>
<td>0.079 %</td>
<td>172</td>
<td>3.6 %</td>
</tr>
<tr>
<td>4</td>
<td>374</td>
<td>0.57 %</td>
<td>108</td>
<td>2.2 %</td>
</tr>
<tr>
<td>5</td>
<td>556</td>
<td>0.84 %</td>
<td>865</td>
<td>17.9 %</td>
</tr>
<tr>
<td>6</td>
<td>2987</td>
<td>4.5 %</td>
<td>1002</td>
<td>20.7 %</td>
</tr>
<tr>
<td>7</td>
<td>13983</td>
<td>21.2 %</td>
<td>327</td>
<td>6.8 %</td>
</tr>
<tr>
<td>8</td>
<td>14071</td>
<td>21.4 %</td>
<td>998</td>
<td>20.6 %</td>
</tr>
<tr>
<td>9</td>
<td>16784</td>
<td>25.5 %</td>
<td>703</td>
<td>14.5 %</td>
</tr>
<tr>
<td>10</td>
<td>16890</td>
<td>25.7 %</td>
<td>151</td>
<td>3.1 %</td>
</tr>
</tbody>
</table>

Figure 7-6: Error Distribution With and Without Diversity
Also as an illustration of how the transmission errors are distributed with and without the induced fast fading diversity scheme, we compared the bit errors on the 10 consecutive images for the two experiments identified as No Diversity and Diversity 1 in Fig. 7-6. In Diversity 1, transmission errors were distributed randomly throughout the duration of the experiment. In No Diversity, errors occur in extremely long bursts. This is typical of what one can expect for the indoor radio channels.

Figure 7-7: Color Image: Original Image, BER = 0

Finally, we transmit the same JPEG coded still image using the experimental setting given in Fig. 7-5 under No Diversity, Diversity 1, and Diversity 3. The original image is given in Fig. 7-7. When the induced fast fading diversity technique is not used, the resulting image is shown in Fig. 7-8. This image is distorted beyond recognition due to bursty errors. When this diversity technique is applied, the resulting image is given in Fig. 7-9. The counterproductive nature of the induced fast fading technique when the induced frequency offsets are too large is shown in Fig. 7-10.
Figure 7-8: Color Image: Transmitted Without Diversity, BER = 0.001953

Figure 7-9: Color Image: Transmitted With Diversity, BER = 0.000628
Figure 7-10: Color Image: Frequency Offset Too Large, BER = 0.014467
Chapter 8

Conclusion

This thesis develops two efficient error control schemes for the transmission of JPEG coded still images over very slow fading radio channels. These transmission schemes use novel transmit antenna diversity techniques to improve the reliability of transmission. When a feedback channel is available, a switched transmit antenna with feedback protocol such as in Chapter 5 can be used. This protocol is extremely effective for very slow fading channels because when the channel errors are sufficiently bursty at slow fading rates, the requirement on polling the channel state information is not as stringent as that on fast fading channels. Provided we have a large number of transmit antennas, using reasonable packet sizes (128, 256 bits) and multiplex ratios (3 or less), very large diversity gains can be achieved with this scheme. More and more transmit antennas are needed to achieve additional diversity gains on slower and slower fading channels (e.g., higher transmission rate).

When a feedback channel is lacking, we developed a repetition-type FEC scheme using the induced fast fading diversity technique. Very close to 100% success ratio for the Type-I information and an acceptable average BER ($10^{-4}$ or lower) for the Type-II information may be achieved using 2 to 4 transmit antennas. With appropriate phase offsets introduced to the transmit antennas, the artificially created fast fading enhances the correcting power of the channel codes (repetition code for the Type-I data packets and convolutional code for the Type-II data packets) to improve the transmission performance.
A hybrid of the two transmission schemes is implemented in the laboratory. Specifically, a Stop-and-Wait Type-I Hybrid ARQ scheme is implemented to transmit the Type-I data packets while the Type-II data packets are transmitted in the idle time slots of the ARQ protocol using the simulated fast fading diversity technique. It is shown that with this transmission protocol, excellent image quality and throughput efficiency are achieved when the induced fast fading diversity technique is applied in this protocol.
Bibliography


