Open Loop Power Control

Solution to the

Direct Sequence Spread Spectrum

Near-Far Problem

A Thesis Submitted to the College of
Graduate Studies and Research
in Partial Fulfillment of the Requirements
for the Degree of Master of Science
in the Department of Electrical Engineering
University of Saskatchewan
Saskatoon, Saskatchewan

by

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1995

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Today's highly mobile, information based society is demanding personal wireless communications. Present solutions, including pagers, cordless telephones, and cellular systems have deficiencies that do not permit roaming telephone communication throughout an office building. A truly mobile, universal telephone service in an office building environment requires a different technology. Many of the systems proposed for the task are based on wireless radio with a modulation format known as Direct Sequence Spread Spectrum (DS-SS). Such systems have the advantages of being spectrum efficient and operating at very low power levels. Unfortunately, the spectral efficiency is dependent on accurate control of the portable telephone's transmission power. This power control is necessary to combat the inherent near-far problem.

Numerous studies have been made on power control issues relevant to Frequency Division Duplex (FDD) systems. The portables and the base station transmit simultaneously on independent frequency bands when they use this protocol. In contrast to FDD, Time Division Duplex (TDD) systems operate using the ping-pong principle where the base station transmits, then the portables transmit—on the same frequency band. This work concentrates on portable transmission power control for a TDD system.

This thesis examines a control method based on fitting Minimum Square Error
(MSE) curves to the forward link power received at the portable. These MSE curves are extrapolated to provide portable transmission power estimates for the next burst. The effect of low order antenna diversity and burst length on this form of power control are also examined. It is shown that MSE control increases the DS-SS system capacity, especially in the presence of diversity and moderate burst lengths.
Acknowledgments

The author would like to express his appreciation to Professor J. E. Salt for the assistance and supervision of this thesis. Thanks also goes to the University of Saskatchewan for support in the form of a Graduate Scholarship.
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List of Symbols

- Channel attenuation.
- Estimated channel attenuation.
- Sinusoid fit amplitude parameter.
- Attenuation factor for a particular bit period.
- RF bandwidth of the spread signal.
- Doppler Spread of a multipath channel.
- Butterworth filter bandwidth.
- Lowpass equivalent impulse response of the channel.
- Fourier transform of the channel impulse response.
- Control gain limiting factor.
- Distance between transmitter and receiver.
- Doppler bandwidth.
- Power measurement data i.
- Expected value operator.
- Energy per bit.
- Frequency in Hz.
- Carrier frequency.
- Coherence bandwidth.
- Doppler fading rate in Hz.
- Doppler shift.
- Noise band center frequency.
- Fourier transform operator.
- Power control factor.
- Power measurement index i.
- Information bearing signal.
- Base transmission power for a particular bit period.
\( \sqrt{-1} \)

- Bessel function of zeroth degree.

\( J_0 \)

- Butterworth filter order.

\( k \)

- Number of portables in a cell.

\( \ln() \)

- Natural logarithm operator.

\( M\text{in}() \)

- Minimization operator.

\( M_{pr} \)

- Measurement of received power at the portable.

\( n \)

- Received symbol index.

\( N \)

- Diversity order.

- Number of antennas in a system using antenna diversity.

- Number of carriers in a system using frequency diversity.

- Number of paths in multipath channel.

- Number of time slots in a system using time diversity.

\( N_b \)

- Generalized broadband noise power per Hz.

\( N_0 \)

- Broadband noise power per Hz.

\( N_m \)

- Power measurement noise.

\( N(t) \)

- Noise that corrupts the transmitted signal.

\( N_s \)

- Noise power per Hz at the receiver.

\( N_{th} \)

- Thermal noise.

\( O_{s} \)

- Sinusoid fit offset parameter.

\( P_0 \)

- A co-user's average power.

\( P_{\text{ave}} \)

- Average co-user noise power received at the base station.

\( P_b \)

- Bit error rate.

\( P_i \)

- The average power the base station receives from co-user \( i \).

\( P_{PN} \)

- Despreading code's average power.

\( P_{\text{rec}} \)

- Signal power received at the base station.

\( PN \)

- Pseudo noise code sequence.

\( \text{rec} \)

- Lowpass equivalent received signal.
$\hat{R}_P$ - Estimated signal power that would be received at the portable in the next burst.

$R_{AA}$ - Autocorrelation of a Rayleigh fading carrier.

$R_B$ - Signal power received at the base station.

Re[] - Real operator.

$R_{env}$ - Rayleigh fading carrier envelope.

$R_P$ - Signal power received at the portable.

$s$ - Transmitted signal.

$S_{despread}$ - Power spectrum at the output of the despreader.

$S_{des\ cod}$ - Despreading code spectrum.

$S_{des\ cu}$ - Post-despreader co-user noise spectrum.

$S_{II}$ - Spectrum of the information bearing signal.

$S_N$ - Signal noise power spectrum.

$S_{PP}$ - Spectrum of the pseudo noise code sequence.

$S_{code\ cu}$ - Pre-despreader co-user noise spectrum.

$S_{rec}$ - Received signal after demodulation and before despreading.

$t$ - Time.

$(\Delta t)_c$ - Coherence time.

$T$ - Symbol (bit) period.

$T_B$ - Base Station transmission power.

$T_c$ - Pseudo noise chip period.

$T_m$ - Delay spread.

$T_P$ - Portable transmission power.

$u$ - Lowpass equivalent transmitted signal.

$V_r$ - Radial velocity.

$w$ - General signal.
<table>
<thead>
<tr>
<th>Symbol</th>
<th>Definition</th>
</tr>
</thead>
<tbody>
<tr>
<td>$W$</td>
<td>Signal bandwidth.</td>
</tr>
<tr>
<td>$\hat{W}_b$</td>
<td>Base station's estimated weighting factor for optimal combining.</td>
</tr>
<tr>
<td>$\hat{W}_p$</td>
<td>Portable's estimated weighting factor for optimal combining.</td>
</tr>
<tr>
<td>$W_i$</td>
<td>Sinusoid weighting function value at index $i$.</td>
</tr>
<tr>
<td>$x$</td>
<td>Received signal.</td>
</tr>
<tr>
<td>$X_1, X_2$</td>
<td>Low pass Gaussian noise sources.</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Definition</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\alpha$</td>
<td>Exponent for distance losses.</td>
</tr>
<tr>
<td>$\alpha_n$</td>
<td>Attenuation factor for path $n$.</td>
</tr>
<tr>
<td>$\gamma_b$</td>
<td>SNR per bit.</td>
</tr>
<tr>
<td>$\delta$</td>
<td>Dirac delta function.</td>
</tr>
<tr>
<td>$\eta$</td>
<td>Mean of the co-user noise on the power measurement.</td>
</tr>
<tr>
<td>$\theta_n$</td>
<td>Phase associated with path $n$.</td>
</tr>
<tr>
<td>$\lambda_0$</td>
<td>Pure tone wavelength.</td>
</tr>
<tr>
<td>$\pi$</td>
<td>Pi.</td>
</tr>
<tr>
<td>$\sigma^2$</td>
<td>Variance of the co-user noise on the power measurement.</td>
</tr>
<tr>
<td>$\tau$</td>
<td>Autocorrelation time index.</td>
</tr>
<tr>
<td>$\tau_n$</td>
<td>Time delay.</td>
</tr>
<tr>
<td>$\phi$</td>
<td>Delay for path $n$.</td>
</tr>
<tr>
<td>$\phi_c$</td>
<td>Rayleigh fading carrier phase.</td>
</tr>
<tr>
<td>$\phi_s$</td>
<td>Time-variant channel transfer function's autocorrelation.</td>
</tr>
<tr>
<td>$\phi_0$</td>
<td>Sinusoid fit phase parameter.</td>
</tr>
<tr>
<td>$\omega$</td>
<td>Frequency in radians per second.</td>
</tr>
<tr>
<td>$\omega_d$</td>
<td>Doppler fading rate in radians per second.</td>
</tr>
<tr>
<td>$\omega_s$</td>
<td>Sinusoid fit frequency parameter.</td>
</tr>
<tr>
<td>$\zeta$</td>
<td>Denotes an outcome of a stochastic process.</td>
</tr>
</tbody>
</table>
- Superscript implies complex conjugate.

* - Denotes convolution.

Λ - Denotes a triangle function.

\( \hat{Z} \) - Symbol denotes an estimate of value \( Z \).
### List of Acronyms

<table>
<thead>
<tr>
<th>Acronym</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>AWGN</td>
<td>Additive Gaussian White Noise.</td>
</tr>
<tr>
<td>BER</td>
<td>Bit Error Rate.</td>
</tr>
<tr>
<td>BPSK</td>
<td>Binary Phase Shift Keying.</td>
</tr>
<tr>
<td>CDMA</td>
<td>Code Division Multiple Access.</td>
</tr>
<tr>
<td>DBPSK</td>
<td>DifferentiallyEncoded Binary Phase Shift Keying.</td>
</tr>
<tr>
<td>DC</td>
<td>Direct Current / has zero hertz frequency.</td>
</tr>
<tr>
<td>DS-SS</td>
<td>Direct Sequence Spread Spectrum.</td>
</tr>
<tr>
<td>FDD</td>
<td>Frequency Division Duplex.</td>
</tr>
<tr>
<td>FDMA</td>
<td>Frequency Division Multiple Access.</td>
</tr>
<tr>
<td>GDV</td>
<td>Gaussian White Noise.</td>
</tr>
<tr>
<td>IF</td>
<td>Intermediate Frequency.</td>
</tr>
<tr>
<td>PN</td>
<td>Pseudo noise, refers to the pseudo noise sequence used by DS-SS systems.</td>
</tr>
<tr>
<td>MSE</td>
<td>Minimum Square Error.</td>
</tr>
<tr>
<td>PSK</td>
<td>Phase Shift Keying.</td>
</tr>
<tr>
<td>RF</td>
<td>Radio Frequency.</td>
</tr>
<tr>
<td>RMS</td>
<td>Root Mean Square.</td>
</tr>
<tr>
<td>SNR</td>
<td>Signal to Noise Ratio.</td>
</tr>
<tr>
<td>TDD</td>
<td>Time Division Duplex.</td>
</tr>
<tr>
<td>TDMA</td>
<td>Time Division Multiple Access.</td>
</tr>
<tr>
<td>WSS</td>
<td>Wide Sense Stationary.</td>
</tr>
</tbody>
</table>
1. Introduction

1.1 Wireless Telephones

There has been an increasing demand for telephone access to people outside their home or office [1]. This has spawned immense interest in mobile communication systems. The demand for such services as cellular telephones and pagers, as well as for cordless telephones, has increased to the point where alternate solutions are being sought. In the move toward universal access, attention is shifting from the mobile cellular “car phone” system to portable office telephones. The indoor systems under investigation are cellular based with restricted cell size. This allows portables within each cell to be linked to the telephone system via low power radio frequency transmissions.

The following desirable features of an indoor mobile communications system have been presented by Cox [2], Pahlavan [3], Kavehrad/McLane [4] and Cox et al. [5]:

- Each subscriber should be accessible at any indoor location at any time of day.
- The system should replace the existing wire-based telephone system for each user, or as many as possible, yet should use the existing Public Switched Telephone Network as much as possible.
- Quality of service must be comparable to wireline service regardless of channel noise and fades.
- Portables should be usable by a person in motion, walking at a reasonable speed while twisting and turning the head, i.e. changing the orientation of the portable.
- The portable must be small in size and lightweight for convenience.
- The battery life of the portable should be at least one month with normal use.
• The system must be secure to eliminate fear of eavesdropping.

• Radio Frequency radiation must be low enough so that the system is not a health risk.

Unfortunately, present systems can only provide some of the desired features. Cellular telephones were originally designed for vehicular use. Use of these within buildings requires high power to penetrate building walls and to have coverage within the large cells. This leads to unacceptably short battery life. The most troublesome problem, however, is the limit on the number of users that can be simultaneously supported in present cellular systems—the system capacity is not sufficient to allow universal access.

The cordless telephone is a very simple form of wireless communication. Unfortunately, this simplicity inspires limitations that prohibit its use for large scale business application. True roaming is not available, as the telephone’s range is restricted by its dedicated base station. Each base station is connected to the network via the standard two wire voice circuit and therefore the “wiring” in an office building is not reduced. Capacity is severely limited due to inefficient use of the radio spectrum. The system is also prone to crosstalk and eavesdropping.

A number of ideas have been put forward for a low power wireless in-building telephone network. Time Division Multiple Access (TDMA) proposals have been suggested by Cox [2] and Cox et al. [5], but many believe that spread spectrum approaches will provide better access in terms of capacity and coverage area. In particular, Pahlavan [3] and Kavehrad/McLane [4] have suggested Direct Sequence Spread Spectrum (DS-SS) systems.

The advantages of spread spectrum have been noted by Marshall et al. [6], Simon et al. [7], Pahlavan [3] and Kavehrad/McLane [4] as well as others, and include:

• Noise resistance and anti-jam capability allows the system to coexist with other radio frequency systems already in place.

• The use of code division for the access mechanism makes the system administration friendly.
• A low probability of intercept makes the system very secure.

• Inherent frequency diversity augments traditional antenna and time (coding) diversity to mitigate the effects of fading.

• The signal quality of a spread spectrum system is a function of the number of users. Thus, when the system is not operating at capacity, which is most of the time, the signal quality will be very good.

• Spread spectrum systems can take advantage of the talking to listening ratio by transmitting only when a user is speaking. This can essentially double the number of users that can be sustained by the system.

These gains do not come without a price. In order to obtain the maximum possible capacity in a DS-SS system, the near-far problem—detailed in Chapter 3—must be addressed. In general terms, the near-far problem can be stated thus: for optimal capacity, all transmissions from the portable telephones must be received at the base station with the same average power. Since the communications channel is time varying, some form of power control is necessary.

1.2 Problem Statement

The problem is to develop and assess the accuracy of a power control algorithm for a DS-SS system in an in-building environment. The power control algorithm must satisfy two criteria: it must be applicable to a system based on time division duplex in a 2 MHz contiguous band of the RF spectrum near 1 GHz, and it must be based on power measurements of the forward link transmissions made at the portable.

The capacity degradation due to imperfect control will be used as a performance measure. The burst length and the order of antenna diversity will be considered as system parameters.

1.3 Thesis Outline

The next chapter describes the indoor channel and introduces the channel model utilized for the research.
The DS-SS communications system under study is detailed in Chapter 3. The treatment there examines system capacity and shows that reverse link power control accuracy has a major effect on the number of users the system can support. Diversity techniques useful in combating multipath fading are discussed as well. The antenna diversity techniques considered in the study, optimal combining and selection diversity, are also dealt with in some detail.

Chapter 4 details the power control mechanism proposed to combat the near-far problem. The simulations constructed to assess the power control capability are also described.

Performance of the power control system is assessed in Chapter 5. The technique used for this purpose is shown and the results of the simulations conducted are discussed.

1.4 Assumptions

The quality of the proposed control algorithms were evaluated using computer simulation. Simulation was chosen over prototyping primarily for simplicity. Prototypes are more difficult to debug and change than computer programs. Further, the focus of this thesis is on the proof of concept and the analysis of the concepts rather than on the construction of the actual system. Unfortunately, the simplicity of simulation comes at a price. The simulation is based on models which approximate the real system. This, and limiting the choice of system parameters, requires a host of assumptions. The following list itemizes the assumptions made in this research work:

- Base station antennas are omni-directional.

- Co-user and environmental noise results in receiver noise that is white, independent, and stationary. The power density of this noise is constant and its average power is proportional to the power sum of the co-user interference.

- Signals received by the base station from all users experience frequency non-selective (flat) slow Rayleigh fading (no line of sight paths).
• Modulation is Differentially Encoded Binary Phase Shift Keying (DBPSK) at a carrier of 900 MHz.

• The data rate is 16 kb/s, and the spreading code rate is 63 chips per bit or 1.008 MHz.

• Portables are in motion such that the maximum Doppler shift experienced is 6 Hz.

• The system is operating at full capacity.

• Demodulation is accomplished with a perfectly coherent carrier.

• Synchronization at the chip and data bit levels is perfect.

• The slow fading due to shadowing and spatial losses are neglected as they introduce second order effects compared to rapid multipath fading. The inherent assumption here is that if the fast fading effects can be followed by the power control, the slow varying effects could be followed as well.
2. Indoor Channel Characterization

2.1 Introduction

The channel model and underlying concepts are presented in this chapter. This includes descriptions of a channel model based on those described by Proakis [8] and Jakes [9]. Terminology and notation is likewise provided. Model parameters used in this thesis are based on data collected by other researchers.

2.2 Channel Characteristics and Terminology

The indoor environment is particularly hostile to radio frequency signals. There are numerous scatterers, reflectors and defractors which produce fading and shadowing. Fading is inherent in a multi-path channel. A multi-path channel is one where a transmitted signal reaches the receiver via two or more separate paths. These paths or routes can be of varying lengths and the signal may experience significantly different phase shifts and attenuations as it traverses each path. The receiver inherently adds these various signals constructively or destructively depending on the phases of the received signals. A fade is the result of destructive addition. Shadowing causes slower power variations and is due to physical structures blocking the transmission path of a signal. Since the transmitters and/or receivers may be moving, fading and shadowing change with time and the position of the portable thereby causing variations in the signal envelope and the channel frequency response [10]. In addition to fading and shadowing, there is a spatial (or distance) loss. Spatial losses are due to the energy density decrease as the distance between the transmitter and receiver increases. In free space the power loss is proportional to \( \frac{1}{D^\alpha} \) where \( D \) is the distance between the transmitter and receiver and \( \alpha \) is equal to 2. As the indoor setting is not free space, the exponent \( \alpha \) varies between 1.2 and 5 depending on such things as building construction, furnishings, and proximity to hallways [11]. Exponents less than 2 are unusual but can occur when a hallway acts as a waveguide.
2.2.1 Channel Impulse Response

When an impulse is applied to a multi-path channel with $N$ paths, an $N$ element impulse train will be received, one impulse for each path. If a signal $s(t)$ is transmitted, the received signal will have the form

$$x(t) = \sum_{n=1}^{N} \alpha_n(t)s[t - \tau_n(t)]$$

(2.1)

where $\alpha_n(t)$ is the attenuation factor for the $n$th path and $\tau_n(t)$ is the delay experienced in traveling the $n$th path. In channels where the transmitter, receiver, and/or scattering objects are moving, the channel impulse response will be time varying, hence the path attenuations and delays are a function of time $t$.

The equivalent low pass received signal at carrier frequency $f_c$ is

$$x(t) = \sum_{n=1}^{N} \alpha_n(t)e^{-j2\pi f_c \tau_n(t)}u[t - \tau_n(t)]$$

(2.2)

where $u(t)$ is the low pass equivalent of the transmitted signal; $u(t)$ is related to $s(t)$ by

$$s(t) = \text{Re}[u(t)e^{j2\pi f_c t}]$$

(2.3)

If the input signal is a unit impulse applied at time $\tau$, the impulse response of the channel becomes

$$c(\tau; t) = \sum_{n=1}^{N} \alpha_n(t)e^{-j2\pi f_c \tau_n(t)}\delta[\tau - \tau_n(t)]$$

(2.4)

If an unmodulated carrier at frequency $f_c$ is transmitted, $u(t) = 1$ and the low pass equivalent received signal $\text{rec}(t)$ will be

$$\text{rec}(t) = \sum_{n=1}^{N} \alpha_n(t)e^{-j2\pi f_c \tau_n(t)}$$

$$= \sum_{n=1}^{N} \alpha_n(t)e^{-j2\pi \tau_n(t)}$$

(2.5)

where $\theta_n(t) = 2\pi f_c \tau_n(t)$. The received signal consists of a sum of time variant vectors (phasors) having amplitudes $\alpha_n(t)$ and phases $\theta_n(t)$. Small changes in the channel,
can cause large variation in phase, amplitude, and delays $\tau_n(t)$ of an individual path. These channel changes are random in nature. This random nature implies that the received carrier can be modeled as a random process composed of the sum of a large number of complex random variables representing each path. Thus, the central limit theorem can be used to model $rec(t)$ as a complex valued random process.

In channels where the low pass equivalent impulse response, $c(\tau; t)$, is modeled with zero mean Gaussian random variables, the envelope, $|c(\tau; t)|$, at any instant in time, $t$, is Rayleigh distributed. This is called a Rayleigh fading channel and occurs where there are no dominant paths. A channel with a significant number of non-moving scatters may result in $c(\tau; t)$ with non-zero mean. The resulting magnitude, $|c(\tau; t)|$, has a Rice distribution. This type of channel is referred to as Ricean fading.

### 2.2.2 Delay Spread

If the channel impulse response, $c(\tau; t)$, is assumed Wide-Sense-Stationary (WSS), its autocorrelation can be defined as

$$\phi_c(\tau_1, \tau_2; \Delta t) = \frac{1}{2} E[c^*(\tau_1; t)c(\tau_2; t + \Delta t)].$$  \hspace{1cm} (2.6)

In uncorrelated scattering, the attenuation and phase shift of the channel at delays $\tau_1$ and $\tau_2$ are uncorrelated. With this assumption, the following equation can be obtained:

$$\frac{1}{2} E[c^*(\tau_1; t)c(\tau_2; t + \Delta t)] = \begin{cases} \phi_c(\tau_1, \tau_2; \Delta t); & \tau_1 = \tau_2 \\ 0 & \text{otherwise.} \end{cases}$$  \hspace{1cm} (2.7)

At $\Delta t = 0$ the resulting autocorrelation $\phi_c(\tau; 0) \equiv \phi_c(\tau)$ yields the average power output as a function of time delay $\tau$, the delay power spectrum. The delay spread of the channel, $T_m$, is the range of values over which the delay power spectrum is non-zero.

The delay power spectrum can be determined by transmitting a narrow pulse over the channel. The average power obtained at the receiver—as a function of time delay—is representative of the delay power spectrum of the channel. The delay spread
of the channel is defined by the time difference between the first and last arrival with significant power.

Since the delay spread is a subjective measure dependent on the judgment of "significant power", Root Mean Square (RMS) delay spread was proposed. The delay power spectrum can be normalized to have an area of 1, corresponding to a probability density function for power with respect to delay

$$\phi_{c\,\text{norm}}(\tau) = \frac{\phi_c(\tau)}{\int_{-\infty}^{\infty} \phi_c(\tau) d\tau}. \quad (2.8)$$

The function's variance can then be computed and the square root taken, leaving the RMS delay spread:

$$\text{RMS delay spread} = \sqrt{E[\phi_{c\,\text{norm}}(\tau)^2] - E[\phi_{c\,\text{norm}}(\tau)]^2}. \quad (2.9)$$

### 2.2.3 Coherence Bandwidth

The channel in the frequency domain may be studied by taking the Fourier Transform of the impulse response to get the time-variant transfer function

$$C(f; t) = \int_{-\infty}^{+\infty} c(\tau; t) e^{-j2\pi f \tau} d\tau. \quad (2.10)$$

An important measure of the channel in the frequency domain is the coherence bandwidth, $(\Delta f)_c$, which is approximately the reciprocal of the multi-path delay spread

$$(\Delta f)_c \approx \frac{1}{T_m}. \quad (2.11)$$

The coherence bandwidth of a channel can be examined by transmitting a flat, wideband signal and examining the frequency content of the received signal. In the presence of multipath, the received signal will no longer be spectrally flat. The width of frequencies that tend to experience the same fading is the coherence bandwidth of the channel.

A signal with bandwidth $W > (\Delta f)_c$ will have frequency components which are
affected differently by the channel—this type of channel is frequency selective. Channels which have coherence bandwidth greater than the signal bandwidth \((\Delta f)c > W\) are frequency non-selective.

2.2.4 Doppler Spread

Time-invariant multi-path channels have a fixed delay spread and a non-varying coherence bandwidth. Time-varying channels have increasing and/or decreasing path lengths which result in a Doppler spreading of the signal spectrum.

If a pure tone is transmitted, a shift in frequency or Doppler Shift, can be caused by the compression or extension of the electro-magnetic wave. The change in the path length brought about by movement of the transmitter, receiver, and/or scatterers induces this Doppler shift. Doppler shift, \(f_d\), defined in terms of the radial velocity, \(V_r\), and the tone’s wavelength, \(\lambda_0\), is

\[
 f_d = \frac{V_r}{\lambda_0}. \tag{2.12}
\]

When multiple paths are present in the channel, there are many opportunities for path lengths to be increasing and decreasing. This results in many tones being received. The width of frequencies the received tones occupy is known as the Doppler spread or fading rate of the channel, \(B_d\). The coherence time of the channel is defined as the reciprocal of the Doppler spread,

\[
 (\Delta t)c \approx \frac{1}{B_d}, \tag{2.13}
\]

and represents the amount of time that the channel remains approximately constant.

The maximum Doppler shift or Doppler (fading) rate can be estimated from a time plot of the normalized received power. Jakes [9] develops a relationship between the maximum Doppler shift and the number of positive slope level crossings. From this relation, the rate of crossing the -3 dB level is approximately equal to the maximum Doppler shift of the channel.

A channel is slowly fading if the signaling period is less than the channel's co-
herence time and the signal bandwidth is less than the coherence bandwidth. A slowly fading channel means that the attenuation and phase induced by the channel is approximately fixed over a symbol interval.

2.3 Fading Channel Model

The nature of an indoor channel is such that channel gain and phase are constantly changing with time. An important consideration is the rate of this change relative to the transmitted bit rate. The ratio of the signal bandwidth to the coherence bandwidth of the channel controls the amount of detail in the frequency domain that the model is required to have. An additional criterion to consider is inter-symbol interference. This factor can be discounted if the delay spread is much less than a symbol period.

Inter-symbol interference can be a limiting factor introduced by some channels. The bit rate of the system presented here is to be 16 kb/s which translates to a bit period of 62.5 μs. This period is much larger than the in-building RMS delay spreads reported by Howard and Pahlavan [12] which are all less than 1 μs. Inter-symbol interference will not be a significant problem and is not considered in the model.

The signal bandwidth for a spread spectrum system is dependent on the spreading introduced by the system. The data signal is effectively spread over a wider spectrum using some pseudo-noise source. In Direct Sequence Spread Spectrum, the transmitted signal bandwidth is dependent on the rate of this pseudo-noise sequence, in this case about 1 MHz. For the channel to be frequency-nonselective, the coherence bandwidth must be greater than the signal bandwidth of 2 MHz (approximated here by the null to null PN sequence bandwidth). Measurements of indoor channels indicate that the 3 dB coherence bandwidth (i.e. the frequency band over which correlation is greater than half the peak) is often greater than 3 MHz [13]. This bandwidth suggests that a frequency non-selective model is appropriate.

Doppler spread is a consequence of a time varying channel. The Doppler spread, $B_d$, of an indoor channel has been measured at 6.1 Hz or less [14] at carrier frequencies in the 0.9 to 1 GHz range. The thesis problem, however, assumes a maximum Doppler
shift of 6 Hz indicating a Doppler spread of about 12 Hz. Regardless, the symbol period of interest, 62.5 $\mu$s, is much less than either channel's coherence time (i.e. $\frac{1}{\Delta f}$). This coupled with the frequency non-selective nature already discussed means the channel is slowly fading\(^1\). A slowly fading channel can be modeled as time invariant over a data bit interval.

Flat Rayleigh fading is simulated by using the model shown in Figure 2.1. This model, found in Jakes [9], allows uniform phase modulation and Rayleigh envelope fading for a particular carrier frequency. Further, the model produces the theoretically derived and empirically observed autocorrelation, $R_{AA}(\tau) = J_0(\omega_d \tau)$; $J_0$ is a Bessel function of zeroth degree and $\omega_d$ is the Doppler fading rate in radians/second [9].

The analysis in this work requires only the magnitude of the envelope provided by Jakes' model. The last mixing and filtering in Jakes' model acts as an envelope detector. Thus, the envelope magnitude may be determined theoretically from the

\(^1\)Shadowing also causes the channel to vary. This effect is, however, slower than multipath fading thus slower than a bit period.
output of the adder in terms of the low pass Gaussian noise sources.

The output of the adder in Figure 2.1 can be represented in terms of this envelope $R_{env}(t)$ and a phase induced $\phi(t)$

$$X_1(t)\sin \omega t + X_2(t)\cos \omega t = R_{env}(t)\cos(\omega t + \phi(t))$$

$$= R_{env}(t)(\cos \omega t \cos \phi(t) - \sin \omega t \sin \phi(t)) \tag{2.14}$$

Observing like terms, two equations can then be formed

$$-R_{env}(t)\sin \phi(t) = X_1(t), \tag{2.15}$$

$$R_{env}(t)\cos \phi(t) = X_2(t). \tag{2.16}$$

Combining Equations 2.15 and 2.16, the envelope $R_{env}(t)$ can be defined in terms of the two independent low-pass Gaussian noise sources $X_1(t)$ and $X_2(t)$

$$R_{env}(t) = \left[X_1(t)^2 + X_2(t)^2\right]^{\frac{1}{2}}. \tag{2.17}$$

The resulting model is shown in Figure 2.2 where GDV is Gaussian White noise.

The process is generated by driving two low pass filters with independent white Gaussian processes. These low pass filters have responses equivalent to the character-
istic Doppler spread of that particular channel [15]. The remainder of the operations create the envelope for the Rayleigh random variable which represents the channel's gain.

The slow fading of the channel relative to the system's bit rate allows the use of a single time index for each transmitted symbol. Any analysis that deals with signal amplitude can thereby be performed symbol by symbol.

The characteristic Doppler rate of interest is the maximum expected in the indoor channel (i.e., 6 Hz). It is necessary to generate a Rayleigh Fading channel model with a 6 Hz Doppler rate in mind. One additional consideration is that the autocorrelation of the channel model should take the form of the Bessel function [9].

Lutz and Plochinger [16] have studied the generation of Rice and Rayleigh processes. In their paper, Butterworth filters were considered as candidates for the low pass filters required by the model shown in Figure 2.2. An approximate equation was derived, which related the Doppler bandwidth, \( Db \), to the Butterworth filter order, \( k \), and bandwidth, \( Bw \),

\[
\frac{Db}{Bw} \approx \frac{1}{1 - \frac{1}{2k}}. \tag{2.18}
\]

Bree [1] showed through extensive simulation that a pair of third order, 5 Hz Butterworth filters generate a 6 Hz Doppler shift with a very similar Doppler spectrum to the in-building channel. Bree's study corroborates Lutz and Plöchinger's approximation.

### 2.4 Simulation of the Indoor Channel

To facilitate writing the simulation program a number of approximations were made regarding the channel. First, the Doppler rate was fixed at 6 Hz. A real channel exhibits variations in the Doppler rate depending on rate of movement, reorientation of the portable, and/or movement of objects near the portable or base station.

The channel's mean attenuation was fixed at one for all of the models for ease of analysis. The channel's mean is typically affected by shadowing and changing distance losses which have a much slower variation than the multipath fading in the channel, thus it is assumed that if a control system can track the multipath fades, the system can track the slower effects as well.
The indoor communications channel under study can be modeled as flat Rayleigh fading. This means that the transmission spectrum fades uniformly, implying that the received signal will be spectrally similar to the transmitted signal. The received signal may, however, be delayed and/or attenuated. If carrier synchronization is attained, perfect demodulation can be assumed, thereby removing the need for simulation of the carrier modulation and demodulation in this power control study.

A slowly fading channel implies that the attenuation in the channel does not change significantly during a bit period. The simulation takes advantage of this by generating the channel attenuation changes at the bit rate.

Simulation testing was carried out to determine the validity of the channel model. The Doppler fading rate was verified by counting the number of significant fades per second and comparing this to the desired Doppler fading rate of 6 Hz. Figure 2.3 shows a typical simulated channel sounding and it can be seen that the trace has about six positive slope crossings of the -3 dB limit. The cumulative distribution function of the simulated Rayleigh fading model is shown in Figure C.1. This distribution is similar to the one reported in [18] where a Rayleigh channel model with a 4 Hz Doppler rate is described.

The channel models used by this thesis are in the form of a channel sounding. This means that the model represents the received power given a constant 1 W transmission power.

2.5 Chapter Summary

In this chapter a slowly fading, flat Rayleigh channel model was developed and parameterized. This model was used in the analysis of the communications system which is found in Chapter 3. Analysis of the proposed control systems in Chapter 4 also utilizes the information presented here.
Figure 2.3 Normalized Channel Model Sounding Given a 1 Watt Transmission Power
3. System Description

3.1 Introduction

It is desirable to examine a system which incorporates Direct Sequence Spread Spectrum (DS-SS) for indoor wireless telephone. The proposed DS-SS system consists of a number of mobile portables which communicate with a central base station, the standard STAR configuration [17]. This system uses the Centralized Multi-Point to Point protocol to handle call processing. Base stations are interconnected to the existing telephone network through which all calls are relayed. Shared use of the spectrum is handled by giving each user a unique, uncorrelated spreading code. Frequency re-use is accomplished by limiting the portable and base station transmission power such that minimum leakage outside the cell is attained. Thus, neighboring cells can utilize the same frequency allocation with minimal adjacent cell interference.

This chapter considers one particular cell which contains a base station and the portables in its servicing area. Direct sequence spread spectrum is discussed as well as transmission schemes and methods used to combat multipath fading.

3.2 Direct Sequence Spread Spectrum

The analytical development of this section is based on the tutorial written by Salt, Kumar, and Dodds [15].

3.2.1 System Realization

Direct Sequence Spread Spectrum is a concept which allows the frequency spreading of a communication signal. The information bearing signal, $I(t)$, is modulated by a broadband Pseudo-Noise code sequence, $PN(t)$. The spread signal is then modulated by a carrier and transmitted. This modulation is illustrated in Figure 3.1.

The studied system is based on Binary Phase Shift Keying (BPSK) where the information is represented by instantaneous phase changes of the carrier by 180 degrees
corresponding to ±1 binary valued data symbols. The data stream is differentially encoded/decoded using differential PSK (DPSK). Assuming the information sequence is random and uncorrelated, the spectrum of the information signal is

$$S_{II}(f) = T \left[ \frac{\sin(\pi f T)}{\pi f T} \right]^2$$  \hspace{1cm} (3.1)$$

where $T$ is the bit time. Thus, the first null bandwidth of the data stream is $\frac{1}{T}$. The spreading function $PN(t)$ has the same statistics as a higher rate data signal giving the PN signal spectrum

$$S_{PP}(f) = T_c \left[ \frac{\sin(\pi f T_c)}{\pi f T_c} \right]^2$$  \hspace{1cm} (3.2)$$

where $T_c$ is the chip duration. In this case, the first null bandwidth of the spreading function is $\frac{1}{T_c}$.

To simplify the analysis, an ideal spreading function is considered. Real spreading functions are not perfectly random and are filtered to limit the bandwidth of the resulting spread signal. Filtering can be done at $\frac{1}{T_c}$, limiting the spreading code bandwidth to $\frac{2}{T_c}$. This spectrum will be approximated by a box of the same bandwidth—the real spectrum would be similar but it would have rounded corners. Since the data symbol rate $\frac{1}{B}$ is much less than the chip rate $\frac{1}{T_c}$, the data spectrum appears essentially as an impulse. The data and the PN sequence are uncorrelated, allowing the spectrum of the spread signal to be calculated as the convolution of the two component spectrums. This approximation results in the spectrum of the spreading code which has a bandwidth of $\frac{2}{T_c}$.

Figure 3.1 Direct Sequence Spread Spectrum.
The optimal way to recover the information is to utilize a matched filter receiver. This operation effectively becomes the integration of an in phase replica of the spreading code multiplied by the received signal. The integration must be symbol synchronized and take place over the whole symbol period.

The signal spectrum should be considered for the different stages of the entire process, Figure 3.2. Initially, the signal consists of only the data stream. After spreading, the power of the signal is spread over a larger bandwidth thereby yielding less power at a particular frequency. The signal is then modulated to RF and transmitted over the channel. At the receiver, the baseband signal is recovered and despreading is performed to recover the information.

### 3.2.2 Effect of Broadband Noise

Broadband noise has a detrimental effect on DS-SS systems and must be considered. For this analysis, noise with a frequency spectrum of width $\Delta f$ and magnitude $N_b$ centered at $f_n$ Hz is studied. The PN code time function used for despreading has a spectrum which is approximated by its null to null bandwidth $\frac{f_c}{2}$ centered at the carrier frequency, $f_c$. The noise and spreading signals are denoted by $N(t)$ and $PN(t)$. The despreading process consists of multiplication in the time domain therefore, the power spectrum at the output of the despread is

$$S_{\text{despread}}(f) = F\{E[N(t)N(t+\tau)PN(t)PN(t+\tau)]\}$$  \hspace{1cm} (3.3)$$

where $F$ denotes the Fourier Transform. Since the noise and PN code are uncorrelated, this simplifies to

$$S_{\text{despread}}(f) = F\{E[N(t)N(t+\tau)]E[PN(t)PN(t+\tau)]\}$$  \hspace{1cm} (3.4)$$

$$= F\{E[N(t)N(t+\tau)]\} * F\{E[PN(t)PN(t+\tau)]\}$$  \hspace{1cm} (3.5)$$

$$= S_N(f) * S_{PP}(f)$$  \hspace{1cm} (3.6)$$

which is the convolution of the noise and spreading code power spectrums.

In analyzing the effects of broadband noise, two frequency bands are considered.
Figure 3.2 Spectrum of a Signal Throughout the DS-SS Process.
The first band lies outside the spread signal bandwidth. Here, the noise has no effect on the reception of the information signal. The signal is only affected by noise within the bandwidth of the spread signal. This means that the broadband noise has the same effect on the despread signal as if the spreading/despreading had not taken place. This simplifies analysis as the Bit Error Rate (BER), for example, can be computed by analyzing the DBPSK system in the presence of noise sources without having to take into account spreading and despreading.

3.2.3 Effect of Co-Users on a Code Division Multiple Access (CDMA) System

The despreading process shrinks the user's signal while expanding the bandwidth of all co-user signals. The effect of despreading on the co-user is analogous to spreading the co-user's signal further since the co-user's and the user's PN codes are different. This spread co-user signal will appear as noise to the user. Fortunately, only the low-frequency portion of the post despread co-user noise will interfere with the information signal, similar to the effect of broadband noise discussed previously.

The approach taken here is to approximate the co-users in a CDMA system as broadband noise with some density constant $N_0$. Since the integrator which assesses the information stream is a low pass device, an estimate of the density constant is the DC value (0 Hz) of the post despreader co-user noise (one sided spectrum).

If the code-phases between users is assumed random, the received co-user signal is independent of the despreading code. This means the DC value of the two-sided post despreader co-user noise spectrum $S_{\text{des cu}}$ is

$$\frac{N_0}{2} = S_{\text{des cu}}(DC) = \int_{-\infty}^{+\infty} S_{\text{rec cu}}(f)S_{\text{des cu}}(f)df$$  \hspace{1cm} (3.7)$$

where $S_{\text{rec cu}}$ is the two sided spectrum of the received pre-despreader co-user noise and $S_{\text{des cu}}$ is the two sided spectrum of the despreading code. The result of the integration yields $\frac{N_0}{2} = \frac{P_{PN}R_0}{2B}$ where $P_{PN}$ is the despreading code power, $P_0$ is the co-user's power, and $2B$ is the RF bandwidth of the spread signal. Assuming $k$ simultaneous users on the system with received co-user powers of $P_i$, the total interference
can be modeled with broadband noise where

$$N_0 = \frac{1}{B} \sum_{i=2}^{k} P_i$$  \hspace{1cm} (3.8)

If the number of users $k$ is large and the variances of $P_i$ are small, this can be estimated as

$$N_0 = \frac{(k - 1)P_{ave}}{B}$$  \hspace{1cm} (3.9)

where $P_{ave}$ is the average of the received power from all the co-users.

### 3.2.4 Capacity of Direct Sequence Spread Spectrum CDMA System

The capacity of a system is dictated by signal characteristics and receiver structure. A simplified receiver diagram is shown in Figure 3.3. Initially, the noisy RF signal is demodulated to provide the spread DBPSK signal. The DBPSK component is then obtained by despreading. The despreading operation does not affect the spectral characteristics of any Additive White Gaussian Noise (AWGN) noise associated with the signal as discussed in Section 3.2.2. Thus AWGN noise present in the spread signal is unaffected by the despreading process. DBPSK receiver performance is well documented when the signal is corrupted with AWGN noise [8]. The BER of such a DBPSK receiver, $P_b$, is

$$P_b = \frac{1}{2} e^{-\gamma_b}$$  \hspace{1cm} (3.10)

where $\gamma_b$ is the Signal to Noise Ratio (SNR) per bit. The SNR per bit may be defined as

$$\gamma_b = \frac{E_b}{N_s},$$  \hspace{1cm} (3.11)

where $E_b$ is energy per bit and $N_s$ is the noise power per Hz at the receiver. Further, the energy per bit may be defined in terms of the received signal power at the receiver, $P_{rec}$, and the data bit period, $T$:

$$E_b = P_{rec} T.$$  \hspace{1cm} (3.12)
RF to IF

Spread

Despread

DBPSK Signal

+ Co-User Noise

+ Thermal Noise

DPSK Receiver

RF Signal

DBPSK Signals

Thermal Noise

Figure 3.3 Receiver Block Diagram.

Specification for signal quality is often made in terms of the maximum tolerable BER. This in turn influences the worst case SNR per bit, $\frac{E_b}{N_0}$. Given this type of specification one can determine the maximum number of users which the system can support. The co-user noise was estimated in terms of broadband white Gaussian noise, Section 3.2.3, to be $N_0 = \frac{(k-1)P_{\text{ave}}}{B}$. In this case the bandwidth, $B$, is approximated by $\frac{1}{T}$. Given this equation and the definition of energy per bit in Equation 3.12, the co-user noise per Hz can be defined by

$$N_0 = (k - 1)P_{\text{ave}}T_c \frac{E_b}{P_{\text{rec}}T},$$

where the difference between $P_{\text{rec}}$ and $P_{\text{ave}}$ is the power control error. Equation 3.13, in terms of the number of users, yields

$$k = \frac{P_{\text{rec}} T}{P_{\text{ave}} T_c} \left( \frac{E_b}{N_0} \right)^{-1} + 1. \quad (3.14)$$

Ideally, all the portable’s transmissions will be received at the base station with the same average power. This would imply that the received signal power under consideration, $P_{\text{rec}}$, would be equal to the average co-user received signal power, $P_{\text{ave}}$. Power control is, however, never perfect. Some variation in the received signal power is to be expected. This has the effect of lowering the SNR and decreasing the capacity of the system. Imperfect power control can be approximated by introducing a power controlled.
control factor, \( F_{pc} \), into the capacity equation. This factor is computed by comparing the minimum received power, \( P_{rec} \), to the average received power, \( P_{ave} \),

\[
F_{pc} = \frac{P_{ave}}{P_{rec}}.
\]  

(3.15)

Most often, the thermal noise, \( N_{th} \), is much less than the co-user noise, \( N_0 \). The receiver noise, \( N_s = N_0 + N_{th} \), can thus be approximated by the co-user noise. This allows Equation 3.14 to be written in terms of the SNR per bit and the power control factor,

\[
k = F_{pc}^{-1} T \left( \frac{E_b}{N_s} \right)^{-1} + 1.
\]  

(3.16)

### 3.2.5 Power Control Justification and the Near-Far Problem

Given the presence of multiple users in a system, co-user noise could mask the desired information bearing signal. The interference on the forward link (base station to portable) is well behaved. All user messages are generated with the same power and travel the same channel to each individual portable. This means that the signal intended for each particular user has the same average energy guaranteeing a constant Signal to Noise ratio after despreading. The portable to base (reverse) link is not so well behaved. Each portable sends its message over a different channel so that each signal experiences different fading, shadowing and distance losses—the near-far problem. Power control must be used to compensate for these different attenuations in order to obtain acceptable capacity.

In addition to compensating for fading, power control is advantageous as the portable’s total energy output can be minimized. At present, the majority of a portable’s size and weight is due to the batteries; reducing the total energy output allows smaller, lighter batteries and/or longer battery life.

### 3.3 Duplexing Schemes

The two most common types of transmission schemes are Time Division Duplex (TDD) and Frequency Division Duplex (FDD). Frequency division duplex is a full
duplex protocol. Separate frequency bands are used by the transmitters at the base station and portable yielding two way communication. These frequency bands must be separated by a significant frequency guard band. This guard band represents a potential loss in signal carrying capacity [18]. In FDD the channel attenuation in the forward and reverse channels are not related. Power control for this scheme is based on the signal power received at the base station. Control information is then transmitted to the portable which adjusts its power output. This system has the advantage in synchronization tracking, however the delay in transmitting control information from the base station to the portable may make the system more sluggish, and therefore less accurate.

Time division duplex is a half duplex protocol. The base station and portable transmit one after the other separated by an appropriate Guard Time. The guard time is used to allow for channel transition and is dependent on the cell size or area that the base station covers as well as synchronization and preamble overhead [18]. Another parameter important to the operation of TDD is the burst length. This is the length of time that the portable spends transmitting a burst and may be measured in terms of the number of symbols that the portable transmits.

The advantage of this method is that both transmitters utilize the same frequency band. Provided the time period over which the portable transmits, or Burst Length is sufficiently short, the attenuation in the forward channel is similar to the attenuation in the reverse channel [18]. Thus, the portable can control its transmission power based on a measurement of the signal power received from the base station in the base station's transmission burst. This is a much less complicated control mechanism.

The system studied in this thesis is TDD where fast open loop power control is being assessed for its ability to track multipath fading more accurately than the slower average power control FDD generally supplies.

3.4 Diversity Techniques to Combat Multipath Fading

Diversity can be used to mitigate the effects of a fading channel. There are three basic types of diversity: frequency diversity, time diversity, and antenna diversity.
Frequency diversity occurs when the signal appears to be received over N distinct carriers. These carriers are separated from each other by more than the coherence bandwidth of the channel so that each carrier experiences independent fading. In this case, for the received signal power to drop significantly, all N carriers must be in a fade. Since the fading is independent, the probability of receiving a weak signal is much reduced. Direct Sequence Spread Spectrum typically uses frequency diversity by transmitting a signal with greater bandwidth $W$ than the channel's coherence bandwidth, $(\Delta f)_c$. It can be shown that $N = \frac{W}{(\Delta f)_c}$ paths may be resolved by such a spread spectrum system [4].

Time diversity is achieved by transmitting the desired signal at N different slots in time, separated by the coherence time of the channel. This diversity technique encompasses areas such as interleaving, block coding and convolutional coding schemes.

Antenna diversity operates by using multiple antennas separated by some distance and/or with different polarity. The spacing and/or polarity differences allow the multiple antennas to each receive independently fading signals. Two principle methods of antenna diversity are utilized by this thesis: optimum combining where the signals are combined, and selection diversity where the strongest received signal is utilized.

### 3.4.1 Selection Diversity

Selection diversity combats multipath fading by using multiple receiver structures to collect the data signal from a number of paths. If a number of independently fading paths are examined, Figure 3.4(a), the likelihood of all the paths experiencing a deep fade decreases as the number of paths (antennas) increases. Thus, choosing the path with the least attenuation, the middle antenna path as shown by Figure 3.4(b), will tend to decrease the number of fades which the system will experience. This in turn will produce less overall variation in the received signal strength. The complexity of the base station is somewhat increased with the additional receiver circuitry and antennas, therefore this thesis examines only second and third order antenna diversity.

Some of the selection diversity system variables are not easy to observe, hence this type of system may be more difficult to control accurately than expected. As the portable receives, it must compute a return transmission power curve. This
Figure 3.4 Selection Diversity Illustration: (a) Portable Transmits, (b) Base Station Transmits.

curve can only be based on the single path that the portable received over. As the portable transmits using this curve, the base station may or may not utilize the path that the portable fit its curve to. In the cases where the base station switches to another antenna, the power curve accuracy is in doubt as the various paths are completely uncorrelated; information about one path is not usable to predict another. It is postulated that this decreased variation in the diverse channel will more than compensate for the small errors induced by the unpredictable antenna switching.

**Operation of the Selection Diversity System**

The base station reception phase requires the switches in Figure 3.5(a) to be in position R. The portable is transmitting so its switch, Figure 3.5(b), is in position T at this time. The portable’s signal is received over each antenna at the base station. Each of these independent signals is demodulated and despread. The maximum power signal is used for each bit in order to get the best signal to noise ratio. The antenna with the greatest received power for the last bit of the base station reception period will be used to transmit during the portable reception stage.

During the portable reception stage, the portable switch is set at R and the base station switches are set to T in Figures 3.5(a) and 3.5(b), respectively. The base station transmits on one antenna selected during the base station reception phase. The portable receives this signal, despreading and demodulating it. The received
Figure 3.5 Selection Diversity System.
power at the portable is then measured for each bit; an algorithm uses this information to control the transmission power during the next data burst.

During the portable transmission stage, the portable switch is set in position T. The algorithm controls the portable's transmitted power for each symbol such that the channel fading is compensated for and a controlled, ideally constant, average signal power is available at the base station.

3.4.2 Optimum Combining

Optimum Combining, as proposed by Winters [19], operates on the theory that appropriately weighting and combining multipath signals will suppress interfering signals and enhance the information bearing signal. Optimal combining results in a signal which is the weighted sum of the multipath components. Examination of this signal shows that the fades are reduced as in selection diversity, yet the signal is more predictable because it appears as a smooth curve, rather than the discontinuous one selection diversity produces.

The advantage of this method over selection diversity is better predictability hence controllability of the system. An increase in complexity of the system, especially in the portable, tends to offset this gain to some degree. The phase component of the optimal weight is also difficult to estimate and inaccuracy here results in less than optimal system system performance.

Operation of the Optimum Combining System

When the base station is receiving, the switches in Figure 3.6(a) are all set to R, whereas the Figure 3.6(b) switch is set to T. The base station receives the portable's signal over the antennas. The signal from each antenna is despread and the gain and phase estimation, $\hat{W}_t$, is made. Each signal is then weighted by an estimate of the complex conjugate of itself, $\hat{W}_t^*$. This optimal combining process is analogous to shifting the signals in time so that the signals add constructively and result in a sum of powers signal with improved SNR characteristics [19].

When the portable is receiving, the switch in Figure 3.6(b) is set to R and base station switches in Figure 3.6(a) are set to T. The base station transmits the desired
Figure 3.6 Optimum Combining System.
signal over each of its antennas such that the signals are out of sync or delayed by 1 chip width between antennas. This delay allows each antenna path's attenuation to be estimated by the portable since the other signals (transmitted on other antennas) will appear as co-user noise.

The portable receives over a single antenna. This information is then taken through a series of 1 chip delays. In this way, antenna 1's path attenuation is tracked on the no delay section, antenna 2's path is tracked by delaying the received signal once, etc. The various antenna path attenuations can now be tracked and an optimal signal can be received. Similar to the base receiver, the delayed signals are despread and the relative gains and phases estimations, $\hat{W}_p$, are made. The delayed signals are then multiplied by their complex conjugates, $\hat{W}_p^*$, and summed. A power control algorithm is used to collect the sum for each received bit and to fit a curve to this data to estimate what power the portable should transmit in the next burst.

The portable uses the gains provided by the algorithm in the portable reception stage to control the transmitted power for each symbol.

3.4.3 Simulation of the Communications System with Antenna Diversity

Antenna diversity involves $N$ independent paths over which a radio frequency transmission is received. As discussed earlier in Section 2.3, the indoor radio channel for this system is flat Rayleigh fading. The system with antenna diversity utilizes $N$ of these paths and therefore $N$ channel models can be used to simulate this effect.

Selection diversity is simulated by creating $N$ of the channel models that are described in Section 2.4. Each channel model represents a channel sounding for a particular antenna. During the base station transmission burst, the power in the first symbol of each channel model is compared. The maximum power indicates which channel model the prediction curve will be fit to. Extrapolation of the prediction curve simulates the portable's transmission power for its return burst. This extrapolation is used in conjunction with the maximum of the $N$ channel models to determine the effective base station received power.

Optimal combining diversity weights the received signals from the $N$ paths ac-
cording to their strengths and phases (i.e. squares the magnitude of the signal) and sums this result. This is simulated by obtaining $N$ independent channel models from Section 2.4 and adding them on a power basis. The portable receives the base station transmission and fits the prediction curve to this sum. The curve is then extrapolated to determine appropriate portable transmit power. The base station received power is determined using the curve extrapolation and the power sum of the channel models for the portable transmission burst.

3.5 Chapter Summary

The overall system architecture has been examined in this chapter. Direct sequence spread spectrum was discussed and the capacity of the system was shown to depend on portable power control. Selection and optimal combining diversity were introduced to combat multipath fading and the simulation strategy for each was outlined.

Spread spectrum and diversity concepts outlined in this chapter are used as the basis for power control analysis in Chapter 4.
4. Power Control Theory and Simulation

4.1 Introduction

Power control is necessary to attain reasonable performance in a DS-SS system. In order to control any system a number of things must be considered. To control system outputs, measurements that indicate the system’s present condition, or state, have to be made. Once the measurements have been obtained, prediction or estimation of system dynamics can be done to allow control of the system.

The system under consideration consists primarily of a dynamic indoor communications channel. The output of this channel (power received at the base station) should be kept constant. In order to do this, the portable’s transmitted power must be varied by an effective control scheme. Since the base station’s transmission power is known, a channel estimate is possible by measuring the power received at the portable. This measurement is corrupted by noise due primarily to the co-users.

4.2 Power Control Analysis

In order to study the effect of reverse link power control on this communications system, a general framework needs to be established. The measurement available is, in general terms, the average signal power received at the portable. The base station transmit power, $T_B$, is constant and, without loss of generality, normalized to 1 watt. The channel attenuation, $A(t;\zeta)$, is some random process with known statistics as discussed in Chapter 2. The portable received power $R_P(t;\zeta)$ can thus be expressed as

\[
R_P(t;\zeta) = A^2(t;\zeta)T_B = A^2(t;\zeta). \quad (4.1)
\]
A curve can be fit to $R_P(t; \zeta)$ and extrapolated to estimate $\hat{R}_P(t; \zeta)$ for the portable's transmit burst. From Equation 4.1, estimating the received power in the simulation is equivalent to estimating the channel attenuation $\hat{A}^2(t; \zeta)$.

The base station received power, $R_B(t; \zeta)$, is based on the portable's transmission power, $T_P(t)$, and the channel attenuation:

$$R_B(t; \zeta) = A^2(t; \zeta)T_P(t). \quad (4.2)$$

For ideal reception of 1 watt at the base station, the portable should transmit $\frac{1}{A^2(t; \zeta)}$. Unfortunately, only the estimate computed above, $\hat{A}^2(t; \zeta)$, is available. This leaves us with the base station received power equation,

$$R_B(t; \zeta) = \frac{A^2(t; \zeta)}{A^2(t; \zeta)}, \quad (4.3)$$

which is used in the control system simulations.

### 4.2.1 Measurement Variable Choice and Derivations

Esmailzadeh and Nakagawa [18] proposed a method of measuring power levels by tracking the peak of the matched filter system used for synchronization and despreading. A derivation of this measurement variable is provided in this section.

The *Average Power* in a signal is defined as

$$P_{\text{ave}}(t) = E[w(t)^2] \quad (4.4)$$

where $w(t)$ is the signal process. The received signal after demodulation is given by

$$S_{\text{rec}}(t) = A(t) I(t) PN(t) + N(t) \quad (4.5)$$

where $I(t)$ is the information signal, $PN(t)$ is the $\pm 1$ binary pseudo random noise spreading function, $A(t)$ is the time varying channel attenuation, and $N(t)$ is estimated by independent zero mean Additive White Gaussian Noise (AWGN). The
average power of the signal received at the portable is

\[ P_{\text{ave}}(t) = E[(A(t)I(t)PN(t) + N(t))^2] \]

\[ = E[(A(t)I(t)PN(t))^2 + N^2(t) + 2A(t)I(t)PN(t)N(t)] \]

\[ = E[(A(t)I(t)PN(t))^2] + E[N^2(t)] \]

\[ = E[A^2(t)]E[I^2(t)] + E[N^2(t)]. \]  

(4.6)

The optimum receiver, a matched filter system, is mathematically equivalent to an integrate and dump of the spread signal multiplied by an in phase copy of the spreading function over the course of a bit period. Assuming bit and PN code synchronization, the resulting processed measurement, \( M_{\text{pr}} \), is

\[ M_{\text{pr}} = \frac{1}{T} \int_{0}^{T} A(t)I(t)PN(t)PN(t) + N(t)PN(t)dt \]

\[ = \frac{1}{T} \int_{0}^{T} A(t)I(t)dt + \frac{1}{T} \int_{0}^{T} N(t)PN(t)dt, \]  

(4.7)

where \( T \) is the bit period. This can be pictured as a noisy measurement where the second term is the noise component.

The indoor channel is considered slowly fading, thus the attenuation in the channel does not change appreciably over one bit period. In addition, the base station transmission power must be known for TDD to operate. This thesis assumes that the base station transmission power is constant, thus the transmitted signal magnitude of a single bit will be constant over a bit period. These assumptions allows Equation 4.7 to be written as

\[ M_{\text{pr}} = \frac{1}{T} A_T I_T \int_{0}^{T} dt + \frac{1}{T} \int_{0}^{T} N(t)PN(t)dt \]

\[ = A_T I_T + \frac{1}{T} \int_{0}^{T} N(t)PN(t)dt \]  

(4.8)

where \( A_T \) is the attenuation of the bit, and \( I_T \) is the known transmitted bit amplitude. As the transmitted power is known, the attenuation factor \( A_T \) can be estimated from
the matched filter peak output, $M_{pr}$. This attenuation factor can in turn be used to
determine the proper portable transmission power.

4.2.2 Effect of Noise on Control

Every electronically controlled system is complicated by the presence of thermal
noise. In addition to this, the CDMA system introduces co-user noise. The presence
of noise limits the measurement accuracy of the control variable. This noise must
therefore be studied to determine its effect on the control system performance.

In order to evaluate the measurement noise, its variance, $\sigma^2$, and mean, $\eta$, will be
determined. Measurement noise, $N_m$, obtained from Equation 4.8, is

$$N_m = \frac{1}{T} \int_0^T N(t)PN(t)dt. \quad (4.9)$$

The noise term, $N(t)$, has been approximated as Zero Mean Gaussian White noise,
Section 3.2.4, which by definition has a delta shaped autocorrelation function. The
PN code, $PN(t)$, is assumed to have a zero mean as well and was approximated in
Section 3.2.1 by a sinc squared function. The sinc squared spectrum implies a triangular
shaped autocorrelation as shown in Figure 4.1. The variance of the measurement
noise can therefore be given by

$$\sigma^2 = \mathbb{E}[N_m(t_1)N_m(t_2)]$$
$$= \mathbb{E} \left[ \frac{1}{T} \int_0^T N(t_1)PN(t_1)dt_1 \frac{1}{T} \int_0^T N(t_2)PN(t_2)dt_2 \right]$$
$$= \frac{1}{T^2} \int_0^T \int_0^T \mathbb{E}[N(t_1)N(t_2)PN(t_1)PN(t_2)]dt_1dt_2$$
$$= \frac{1}{T^2} \int_0^T \int_0^T \mathbb{E}[N(t_1)N(t_2)]\mathbb{E}[PN(t_1)PN(t_2)]dt_1dt_2$$
$$= \frac{1}{T^2} \int_0^T \int_0^T N_0\delta(t_2-t_1) \wedge (t_2 - t_1, \frac{1}{T_c})dt_1dt_2$$
$$= \frac{N_0}{T}, \quad (4.10)$$

where $\wedge(t_2 - t_1, \frac{1}{T_c})$ is the triangular function shown in Figure 4.1. The mean of the
Figure 4.1  Autocorrelation of a PN Spreading Code

measurement noise is zero which follows from

\[
\eta = E[N_m(t)] = E \left[ \frac{1}{T} \int_0^T N(t)PN(t) \, dt \right] \\
= \frac{1}{T} \int_0^T E[N(t)PN(t)] \, dt \\
= \frac{1}{T} \int_0^T E[N(t)]E[PN(t)] \, dt \\
= 0. \tag{4.11}
\]

The transmitted signals are similar on the forward and reverse links, thus the equations derived in Section 3.2 may be re-used. As the base station transmits to the portables, it ensures that the average power of each signal in its transmission is kept the same. On the path to a particular portable, the signal of interest in this transmission will experience the same attenuation as the others which form the co-user noise. This means that each of the co-user signals, including the desired signal, will have the same average power at the portable of interest. This allows the substitution of Equation 3.9 into 4.10 and the measurement noise variance becomes

\[
\sigma^2 = (k - 1)P_{\text{ave}} \frac{T_c}{T} . \tag{4.12}
\]

The SNR per bit \( \gamma_b \) was previously defined in Equation 3.11. With negligible thermal
noise, \( N_{th} = 0 \), and perfect power control, \( P_{rec} = P_{ave} \), this equation becomes

\[
\gamma_b = \frac{T}{T_c (k - 1)}.
\]  
(4.13)

Measurement noise variance can be defined in terms of SNR per bit by combining Equations 4.12 and 4.13 which results in

\[
\sigma^2 = P_{ave} (\gamma_b)^{-1} = (A_T I_T)^2 (\gamma_b)^{-1}.
\]  
(4.14)

The greatest permissible BER is typically specified as a performance parameter. Equation 3.10 can be rearranged and combined into Equation 4.14 to give us the largest permissible noise variance

\[
\sigma^2 = \frac{(A_T I_T)^2}{\ln \left( \frac{1}{3P_b} \right)}
\]  
(4.15)

in terms of the system BER, \( P_b \).

The measurement noise is minimized by the averaging effect of the matched filter itself [18] so long as the noise has no bias. Equation 4.11 illustrates that the mean noise should be zero, leaving no bias to deal with. In addition, Equation 4.14 clearly shows that the co-user induced measurement noise power is dependent on the measured variable itself, \( A_T \). The chief implication is that the SNR of the measurement variable will be constant if the thermal circuit noise, which should be insignificant, are neglected.

### 4.3 Power Control Algorithms and Simulations

It must be recognized that the power measurement is corrupted by noise. This indicates that some type of averaging technique will be required in order to track the time varying channel parameter. The time varying parameter should be predicted to give the best performance possible. Another problem is that the statistics of the channel are not typically known and often vary with time. Computation complexity
is also an issue; simpler algorithms are more desirable as they require less costly (slower) hardware to execute in the finite time available. The data available for control is sampled allowing a digital control system to be easily implemented. 

Three power control schemes are considered in this thesis; each was simulated within its own computer program. Average power control is used as a baseline for comparison of the various control approaches. A second algorithm fit a Minimum Square Error (MSE) polynomial to the measurement data whereas the last method fit an MSE sinusoid.

4.3.1 Power Control Limiting

One important piece of information that can be utilized by the control system is the quantifiable limits on the range of variation in the channel. This means that one can limit the power transmitted by the portable and still be assured of good reception. This is fortunate since in practice a portable's transmission power has a limited range of operation.

4.3.2 Average Power Control

Average power control represents an attempt to follow and negate the effects of the long term average of the channel. This power control method perfectly follows the slowly varying attenuation induced by shadowing and distance losses, but it is unable to track multipath fading. All of the control algorithms studied inherently negate the long term average of the channel. Average power control therefore gives a basis from which the relative performance of the proposed control systems can be measured.

Simulation of average power control is carried out by setting the channel attenuation estimate, \( \hat{A}(t; \zeta) \), equal to the expected value of the channel, \( E[\hat{A}(t; \zeta)] \). In the single path case, the application of this power control is done by scaling the channel model of interest by the reciprocal of the channel's mean. When diversity is simulated, \( N \) channel models with the same mean are utilized by the diversity simulations as described in Section 3.4.3, and the result is scaled by the reciprocal of this amalgamated channel's mean. This section was tested by ensuring that the post-controlled
received power was equal to 1 W, the set-point of interest.

4.3.3 Least Mean Square Polynomial Fit

The objective of the Least Mean Square Polynomial Fit method is to fit a polynomial to the decision variable sequence over a number of data bits. This polynomial is then extrapolated to estimate the future value of this received power and thereby estimate the portable’s power transmission requirement.

This method was proposed and tested by Esmailzadeh and Nakagawa [18] and has a number of attractions. The operation is intuitive and relatively simple to perform. In addition the effect of zero mean noise is mitigated during the curve fitting operation.

Unfortunately, some problems exist which limit the effectiveness of the algorithm. The curves chosen (line, parabola) are inadequate for modeling the power curve at times of deep fades. Rayleigh variables are never less than zero, however a parabola or linear fit may result in a negative result if the channel is experiencing a fast fade. Esmailzadeh and Nakagawa address this issue by limiting the portable’s control gain, which in turn limits the transmitter’s range of operation. In addition, the parabola was fit to the logarithm of the power curve. This ensures negative estimates, which can be produced by the polynomial-fit operation, have meaning.

Control Limits for Linear and Parabola Fit

Limitation of the transmit power is a must when fitting functions like lines and parabolas. Entering and leaving deep fades, the channel curve is very steep and open functions can produce large estimation errors. This effect can be minimized by placing a floor and ceiling on the power transmitted. For simplicity, the range chosen was the approximate extent of the channel’s variation. These values were found by looking at the minimum and maximum received power in the average power controlled channel. When diversity is added, the effective channel variation is decreased. New variation limits are therefore required. These were found by examining the maximums and minimums of the received power under average power control for the system with diversity. A listing of the control gains for each diversity type can be found in Table 4.2.
Simulation of Linear and Parabola Fit Algorithms

Parabola and linear fit algorithms were used to estimate power control for the portable transmitter. A number of data points – one burst length long – were considered and the MSE polynomial was fit using the built-in MATLAB function `polyfit()`. Extrapolation was performed, using the MATLAB function `polyval()`, to estimate the power level over the next transmission set. The transmission level of the portable was then matched to this curve and the resulting base station received power was recorded.

Extrapolation can produce large magnitude approximations because parabolas and straight lines are open ended functions. The large approximations may be rare but they will cause significant errors in the controlled output. To minimize this effect, a limiting parameter, the maximum control gain $CG$, is introduced. This is the range of power which the transmitter can supply. Since the channel – regardless of diversity – has already been normalized to have unity mean (corresponding to 0 dB), the transmit power is allowed to range between $\pm \frac{CG}{2}$ dB.

The program has a number of parameters: the maximum control gain, burst length in symbols, and the data that needs to be fit. The program output consists of the estimated portable transmission power which may be used to calculate the base station received power.

Simulation testing was based primarily on visual analysis. The fit curve was plotted on the same graph as the data the curve was being fit to and comparison was made.

4.3.4 Least Mean Square Sinusoid Fit

The second control method consists of fitting a Minimum Square Error (MSE) sinusoid to the logarithm of the power curve. The function for this operation is detailed in Equation 4.16. In this equation $D_i$ is the vector of points being fit, $O$, is the offset, $A$ is the amplitude, $\omega$ and $\phi$ are the frequency and phase of the sinusoid.
and $W_i$ is a weighting function,

$$MSE = \min \left( \sum_{i=1}^{\text{DataPoints}} (D_i - (O_s + A_s \sin(\omega_s i + \phi_s)))^2 W_i \right). \quad (4.16)$$

The curve is extrapolated to estimate the transmission power for each symbol in the next interval.

**Control Limits for Sinusoid Fit**

To simplify and thus speed up the curve fit operation, it is desirable to eliminate as many variables from the equation as possible and replace these with constants. The sinusoid amplitude, $A_s$, was fixed for this analysis. This has the effect of limiting the variation possible in one burst length. This has the additional benefit of being an inherent form of power limitation. Since the fit-sinusoid’s amplitude is set, the function is self-limiting.

The presence of an unknown DC offset means the power limitation must be different than that used for parabola and linear fit algorithms. Rather than setting a ceiling and floor for transmission power, the power variation within a burst was limited. The sinusoid amplitude was set to half of the largest variation expected in a burst period. The amplitudes were determined from the received power plots gathered when the average power control method was applied to the channel. The control gains used for various system parameters are listed in Table 4.2.

**Simulation of the Sinusoid Fit Algorithm**

As an alternative to the “open ended functions” a sinusoid was used for the prediction curve. The points were fit to a constant amplitude sinusoid with variable frequency and phase such that Minimum Square Error was achieved. The estimates are therefore limited to within certain boundaries. The portable’s transmit power is then estimated by extrapolating the sinusoid.

It is postulated that the later sample points are more important to the prediction process than the earlier ones. This is based on the channel autocorrelation which has a marked roll-off. Weighting of the measurements was done to reflect this relative importance. Three data weighting schemes were investigated. The simplest consisted
of weighting all the data points equally. Two weighting functions were assessed to see how much effect the weighting function itself might have on the prediction accuracy. The two schemes consist of weighting the points using exponentially increasing (hereafter termed \textit{Weighted 1}) and exponentially decaying (\textit{Weighted 2}) weighting functions. These weighting functions were chosen such that the later data points were weighted more than the earlier ones. The parameters of the exponential were evaluated so that the data points would be weighted from 10 percent to 100 percent; for a 100 data bit burst, the curves become \(0.1e^{0.023i}\) (\textit{Weighted 1}) and \(1.1 - e^{-0.023i}\) (\textit{Weighted 2}) \((i\text{ is the data bit number})\) as depicted in Figure 4.2.

The parameters required by the program are simply the length of each data burst in symbols, the data that needs to be fit, and a weight vector used to weight the data points. The program provides the estimated portable transmission power which in turn can be used to give the received power.

Simulation testing of the sinusoid fit algorithm was carried out in a manner similar to that used for the linear and parabola fit algorithms. The fit curve and the data it was being fit to were plotted and a visual comparison was performed.


Table 4.1  Received Power Correlation for Various Burst Lengths

<table>
<thead>
<tr>
<th>( \tau ) in no. of bits for ( f_d = 6 \text{ Hz} )</th>
<th>( 2\pi f_d \tau )</th>
<th>( J_0(\omega_d \tau) )</th>
<th>( R_{AA}(\tau) )</th>
</tr>
</thead>
<tbody>
<tr>
<td>50</td>
<td>0.1178</td>
<td>0.997</td>
<td></td>
</tr>
<tr>
<td>100</td>
<td>0.2356</td>
<td>0.986</td>
<td></td>
</tr>
<tr>
<td>200</td>
<td>0.4712</td>
<td>0.945</td>
<td></td>
</tr>
</tbody>
</table>

4.3.5 Varying the Burst Length Parameter

As discussed in Section 2.3, investigation of the channel attenuation variable \( A(t) \) indicates it has an autocorrelation of \( R_{AA}(\tau) = J_0(\omega_d \tau) \). This autocorrelation gives an indication of the accuracy with which predictions of the channel can be made. As can be seen in Table 4.1, the autocorrelation changes slightly over the burst rates studied. It should be noted there is little difference between the 50 and 100 symbol bursts, whereas a more significant drop is observed between 100 and 200 symbol bursts. This indicates there should be diminishing returns in reducing the burst rate and suggests that a burst length of 100 bits is reasonable. If the burst length is made too short, overhead between bursts can become important. The system capacity will effectively decrease as the burst length approaches the guard time—in opposition to the capacity gains made possible through decreasing channel variation.

4.4 Power Control Analysis with Antenna Diversity

The portable in a single path system is identical to the one needed for a selection diversity system. This means that the processed measurement, \( M_{pr} \), and the associated noise statistics on this measurement are the same. The curve fit algorithms used are the same as those used for the single path case. As far as the portable is concerned, the system has not changed.

The received signal from an optimum combining system consists of the weighted sum of the signals from \( N \) different paths. Using the previous derivations from Section 4.2.1, a similar power measurement can be derived for the optimum combined system by considering \( N \) of the single path measurements. These \( N \) measurements are each weighted by a factor, \( \hat{W}_i^* \), which is intended to remove the phase differences
and weight the received signal. Using Equation 4.8 as a basis and assuming perfect weights, the relation for the optimally combined processed signal, \( M_{pr} \), is

\[
M_{pr} = \left( \sum_{i=1}^{N} A_{Ti}^2 I_T^2 \right) + \text{measurement noise} \tag{4.17}
\]

where \( I_T \) is the bit amplitude and \( A_{Ti} \) is the attenuation of the \( i \)th path.

Power control for an optimum combining system is quite different from single path and selection diversity. The curve is fit to the weighted sum of the available paths, rather than a particular one. The weighted sum has significantly less variability than any one path.

The algorithms used for control of the optimum combined system were the same as for the single path and selection diversity cases. Using the weighted sum tends to moderate the number and severity of fades similar to a selection diversity system, but it lacks the quick transitions that the selection diversity system introduces as it changes antennas. This lack of antenna transitions should result in some savings in terms of power control error.

### 4.5 Summary of Simulations Conducted

The purpose of the simulations was to analyze the effectiveness of Time Division Duplex power control schemes on the indoor channel. To simplify the analysis, channel parameters were chosen to reflect the situation where worst case power control error would be expected. Intuitively, larger error should occur in a channel which has the greater variability—a faster Doppler rate—and this has been observed [18]. Maximum Doppler rate (6 Hz as discussed in Section 2.4) channel models are therefore used for all simulations. Each simulation was conducted assuming a 16 kb/s symbol rate DBPSK modulated spread spectrum system with a 1 MHz chip rate. In addition, the expected received power at the base station was set at a nominal 1 W for ease of computation and comparison.

Table 4.2 details the control mechanisms that were considered. Lines, parabolas, and sinusoids were fit to predict channel variation and to control the base station received power. The sinusoid fit algorithm contained optional weighting functions,
Weighted 1 and Weighted 2, which were used in an attempt to increase the algorithm's accuracy. The burst length parameters of these algorithms were varied in order to study this effect on control error. Burst lengths of 50, 100, and 200 symbols were investigated.

Low order antenna diversity was introduced as a means of mitigating some of the multipath fading effects. Selection and optimal combining diversity were considered as potential ways of minimizing control error induced by deep fades. Order of diversity was investigated (2 and 3 paths) to evaluate this effect on control as well. The various diversity systems studied are itemized in Table 4.2.

In all cases, a control gain parameter was assigned to limit the estimates made by the control algorithms. These values can be determined by examining the base station's received power given a constant portable transmit power. In terms of this analysis, the control gain parameters for each diversity type are determined by examining the received power under average power control conditions. Control gain parameters must be found for each diversity and control mechanism studied as detailed in Section 4.3.1. The control gain parameters used are tabulated in Table 4.2.

4.6 Chapter Summary

Reverse link power control has been discussed in this chapter. Analysis of forward link power measurements was performed including the effect of co-user noise. This analysis was utilized to construct simulations that indicate the effect of the parameters studied on power control. The results of these simulations are presented in Chapter 5.
Table 4.2  Summary of Simulations Conducted

<table>
<thead>
<tr>
<th>Data Rate</th>
<th>Chip Rate</th>
<th>Spreading Gain</th>
<th>Nominal Power</th>
</tr>
</thead>
<tbody>
<tr>
<td>16 kb/s</td>
<td>1 MHz</td>
<td>18 dB</td>
<td>1 Watt</td>
</tr>
<tr>
<td>Channel Noise</td>
<td>Doppler Rate</td>
<td>Burst Length</td>
<td>Modulation</td>
</tr>
<tr>
<td>None</td>
<td>6 Hz</td>
<td>See Table</td>
<td>DPSK</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Diversity</th>
<th>Control Method</th>
<th>10^log Control Gain for Bursts of</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td>50 Symbols</td>
</tr>
<tr>
<td>No Diversity</td>
<td>Linear Fit</td>
<td>20</td>
</tr>
<tr>
<td></td>
<td>Parabola Fit</td>
<td>20</td>
</tr>
<tr>
<td></td>
<td>Sinusoid</td>
<td>15</td>
</tr>
<tr>
<td></td>
<td>Weighted 1</td>
<td>15</td>
</tr>
<tr>
<td></td>
<td>Weighted 2</td>
<td>15</td>
</tr>
<tr>
<td>N=2 Selection</td>
<td>Linear Fit</td>
<td>20</td>
</tr>
<tr>
<td>Diversity</td>
<td>Parabola Fit</td>
<td>20</td>
</tr>
<tr>
<td></td>
<td>Sinusoid</td>
<td>1</td>
</tr>
<tr>
<td></td>
<td>Weighted 1</td>
<td>1</td>
</tr>
<tr>
<td></td>
<td>Weighted 2</td>
<td>1</td>
</tr>
<tr>
<td>N=3 Selection</td>
<td>Linear Fit</td>
<td>20</td>
</tr>
<tr>
<td>Diversity</td>
<td>Parabola Fit</td>
<td>20</td>
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<tr>
<td></td>
<td>Sinusoid</td>
<td>1</td>
</tr>
<tr>
<td></td>
<td>Weighted 1</td>
<td>1</td>
</tr>
<tr>
<td></td>
<td>Weighted 2</td>
<td>1</td>
</tr>
<tr>
<td>N=2 Optimal Combining</td>
<td>Linear Fit</td>
<td>10</td>
</tr>
<tr>
<td>Diversity</td>
<td>Parabola Fit</td>
<td>10</td>
</tr>
<tr>
<td></td>
<td>Sinusoid</td>
<td>1</td>
</tr>
<tr>
<td></td>
<td>Weighted 1</td>
<td>1</td>
</tr>
<tr>
<td></td>
<td>Weighted 2</td>
<td>1</td>
</tr>
<tr>
<td>N=3 Optimal Combining</td>
<td>Linear Fit</td>
<td>10</td>
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<tr>
<td>Diversity</td>
<td>Parabola Fit</td>
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<tr>
<td></td>
<td>Sinusoid</td>
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<tr>
<td></td>
<td>Weighted 1</td>
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</tr>
<tr>
<td></td>
<td>Weighted 2</td>
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</tr>
</tbody>
</table>
5. Data Analysis

5.1 Introduction

This chapter proposes methods of assessing the control system accuracy given various system configurations. Combinations of the following system variations are examined with a burst length of 100 symbols:

- Control system type: average, linear fit, parabola fit, sinusoid fit, and sinusoid fit with weighting functions.
- Diversity: none, two and three antenna selection diversity, two and three antenna optimal combining diversity.

The effect of burst length on power control was studied by simulating systems with 50, 100, and 200 symbol bursts.

5.2 Measurement of Power Control Effectiveness

The effectiveness of the power control algorithms is measured by examining the RMS deviation from the correct value. A qualitative measure of effectiveness can be gleaned by examining the received power envelopes and the probabilistic distributions of the received power. The received power envelopes for a number of parameters are shown in Appendix A whereas the received power density functions are given in Appendix B. The cumulative distribution functions of received power are shown in Appendix C and Appendix D. Comparison of these plots is often enough to determine the better control algorithms. The mean, variance, and percentiles of the received power are also used as performance indicators. These statistics may be found in Table 5.1, 5.2, and 5.3.

Since the controlled power level is expected to fluctuate in an AC fashion about a set-point, the mean received power should be equal to the set point. If the mean received power is different from the set point value, the control system introduces bias.
Table 5.1 Simulation Statistics of Received Power for 100 Symbol Data Bursts.

<table>
<thead>
<tr>
<th>Data Rate</th>
<th>Chip Rate</th>
<th>Spreading Gain</th>
<th>Nominal Power</th>
</tr>
</thead>
<tbody>
<tr>
<td>16 kb/s</td>
<td>1 MHz</td>
<td>18 dB</td>
<td>1 Watt</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Channel Noise</th>
<th>Doppler Rate</th>
<th>Burst Length</th>
<th>Modulation</th>
</tr>
</thead>
<tbody>
<tr>
<td>None</td>
<td>6 Hz</td>
<td>100 symbols</td>
<td>DPSK</td>
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</table>

<table>
<thead>
<tr>
<th>Diversity</th>
<th>Control Method</th>
<th>Received Power (W)</th>
<th>10 log $\frac{\text{Received Power}}{\text{Watt}}$</th>
</tr>
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<tr>
<td></td>
<td></td>
<td>Mean</td>
<td>Variance</td>
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<tr>
<td>No</td>
<td>Average</td>
<td>1.0000</td>
<td>1.077 $10^{+0}$</td>
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<tr>
<td>Diversity</td>
<td>Linear Fit</td>
<td>0.9924</td>
<td>1.235 $10^{-1}$</td>
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<tr>
<td></td>
<td>Parabola Fit</td>
<td>0.9928</td>
<td>3.958 $10^{-2}$</td>
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<tr>
<td></td>
<td>Sinusoid</td>
<td>1.8329</td>
<td>6.210 $10^{+2}$</td>
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<tr>
<td></td>
<td>Weighted 1</td>
<td>2.1655</td>
<td>1.270 $10^{+3}$</td>
</tr>
<tr>
<td></td>
<td>Weighted 2</td>
<td>2.0279</td>
<td>9.821 $10^{+2}$</td>
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<tr>
<td>N=2</td>
<td>Selection</td>
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<td>5.070 $10^{-1}$</td>
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<tr>
<td>Diversity</td>
<td>Linear Fit</td>
<td>1.0619</td>
<td>1.813 $10^{-1}$</td>
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<tr>
<td></td>
<td>Parabola Fit</td>
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<td>2.247 $10^{-1}$</td>
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<td>Sinusoid</td>
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<td>1.133 $10^{-1}$</td>
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<td>Weighted 1</td>
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<td>Diversity</td>
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<td>Parabola Fit</td>
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<td>Weighted 1</td>
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<td>1.0608</td>
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</tr>
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<td>N=2</td>
<td>Optimal</td>
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<tr>
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<td>Linear Fit</td>
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<td>Parabola Fit</td>
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<td>Weighted 1</td>
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<td>Weighted 2</td>
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<td>Combining</td>
<td>Linear Fit</td>
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<td>Diversity</td>
<td>Parabola Fit</td>
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<td>2.191 $10^{-4}$</td>
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<td>Sinusoid</td>
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<td>3.770 $10^{-4}$</td>
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<td>Weighted 1</td>
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Table 5.2  Simulation Statistics of Received Power for 200 Symbol Data Bursts.

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<tr>
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<th>Chip Rate</th>
<th>Spreading Gain</th>
<th>Nominal Power</th>
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<tbody>
<tr>
<td>16 kb/s</td>
<td>1 MHz</td>
<td>18 dB</td>
<td>1 Watt</td>
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</tbody>
</table>

<table>
<thead>
<tr>
<th>Channel Noise</th>
<th>Doppler Rate</th>
<th>Burst Length</th>
<th>Modulation</th>
</tr>
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<tbody>
<tr>
<td>None</td>
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<table>
<thead>
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<th>Diversity</th>
<th>Control Method</th>
<th>Received Power (W)</th>
<th>10 log $\frac{\text{Received Power}}{1\text{ watt}}$</th>
<th>5 and 95 Percentiles</th>
</tr>
</thead>
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<tr>
<td>No Diversity</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
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<td></td>
<td>Average</td>
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<tr>
<td></td>
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<td>Parabola Fit</td>
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<td>5.620 $10^0$</td>
<td>$-1.294 10^0$</td>
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<td>Sinusoid</td>
<td>403.01</td>
<td>9.367 $10^7$</td>
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<td>Weighted 1</td>
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<td>1.184 $10^0$</td>
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<td>Sinusoid</td>
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<td>1.024 $10^0$</td>
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<td>1.399 $10^0$</td>
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<td>Weighted 2</td>
<td>1.3067</td>
<td>1.123 $10^0$</td>
<td>$-3.236 10^-1$</td>
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<td>N=2 Optimal Combining Diversity</td>
<td>Average</td>
<td>1.0000</td>
<td>4.764 $10^-1$</td>
<td>$-6.671 10^0$</td>
</tr>
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<td></td>
<td>Linear Fit</td>
<td>1.0124</td>
<td>5.131 $10^-2$</td>
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<td>1.659 $10^-2$</td>
<td>$-4.504 10^-1$</td>
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<td>Sinusoid</td>
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<td>1.621 $10^-2$</td>
<td>$-4.688 10^-1$</td>
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<td>9.484 $10^-3$</td>
<td>$-4.106 10^-1$</td>
</tr>
<tr>
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<td>Weighted 2</td>
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<td>1.358 $10^-2$</td>
<td>$-4.578 10^-1$</td>
</tr>
<tr>
<td>N=3 Optimal Combining Diversity</td>
<td>Average</td>
<td>1.0000</td>
<td>3.408 $10^-1$</td>
<td>$-6.671 10^0$</td>
</tr>
<tr>
<td></td>
<td>Linear Fit</td>
<td>1.0123</td>
<td>3.395 $10^-2$</td>
<td>$-6.656 10^-1$</td>
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<td>Parabola Fit</td>
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<td>8.564 $10^-3$</td>
<td>$-3.519 10^-1$</td>
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<td>Sinusoid</td>
<td>1.0100</td>
<td>8.164 $10^-3$</td>
<td>$-2.879 10^-1$</td>
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<tr>
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<td>Weighted 1</td>
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<td>6.870 $10^-3$</td>
<td>$-2.388 10^-1$</td>
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<td>Weighted 2</td>
<td>1.0101</td>
<td>7.640 $10^-3$</td>
<td>$-2.655 10^-1$</td>
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### Table 5.3  Simulation Statistics of Received Power for 50 Symbol Data Bursts.

<table>
<thead>
<tr>
<th>Data Rate</th>
<th>Chip Rate</th>
<th>Spreading Gain</th>
<th>Nominal Power</th>
</tr>
</thead>
<tbody>
<tr>
<td>16 kb/s</td>
<td>1 MHz</td>
<td>18 dB</td>
<td>1 Watt</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Channel Noise</th>
<th>Doppler Rate</th>
<th>Burst Length</th>
<th>Modulation</th>
</tr>
</thead>
<tbody>
<tr>
<td>None</td>
<td>6 Hz</td>
<td>50 symbols</td>
<td>DPSK</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Diversity</th>
<th>Control Method</th>
<th>Received Power (W)</th>
<th>10 log Received Power</th>
<th>5 and 95 Percentiles</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td>Mean</td>
<td>Variance</td>
<td>5 and 95 Percentiles</td>
</tr>
</tbody>
</table>

- **No Diversity**
- **N=2 Selection Diversity**
- **N=3 Selection Diversity**
- **N=2 Optimal Combining Diversity**
- **N=3 Optimal Combining Diversity**
which, whether positive or negative, is undesirable. If the mean received power is too large, the portables will be needlessly drawing upon the limited battery resource. If the mean is too low the ambient channel noise will introduce communication errors.

It is important to examine each control system’s worst case operation. Tolerable communication is desired at this worst case. Previous analysis indicates that capacity is dependent on the minimum SNR at the base station receiver, Equation 3.16. Minimum SNR is approximated in Section 3.2.3 by assuming that the noise power at the receiver is constant and equal to the average co-user received power. Given the definition of SNR and the constant noise power, the minimum received power will indicate the minimum SNR. The determination of the absolute minimum is impossible with limited simulation runs as this occurrence is rare. The rare occurrence of the minimum received power also means that this value will have little effect on the system’s overall performance. A percentile is simpler to estimate from limited simulation runs provided that the percentile deviates significantly from zero and one. In addition, the percentile can be chosen so that the value is useful in estimating system performance. The fifth percentile can be used in place of the minimum received power as a control effectiveness indicator.

Base station receiver noise statistics and battery life are additional factors that need consideration. Unfortunately these are much more difficult to examine. The noise power can be approximated as the sum of the received powers of all the co-users in the system. This co-user noise is typically assumed to be AWGN—a very good model if one assumes no fading and identical received power levels for all users. In a system with fading and power control, noise statistics become complicated and seem to exhibit short term time dependence [20]. One way to account for the short term time dependence is to assume a “worst case” noise where all co-users are received with the maximum received power. Using this approximation, over compensation by the control system would be as disadvantageous as under compensation with regards to SNR. This worst case condition is extremely harsh and would be highly improbable when the number of co-users is large.

A related consideration is battery life. Transmitting too much power dramati-
cally increases the demands placed on the portable supply. Co-user noise concerns and portable power supply requirements combine to place some restriction on the allowable maximum portable received power. Measuring and using the maximum received power as a performance indicator is not appropriate. The reasoning for this is similar to that used to rule out the use of the minimum received power as a performance indicator. The ninety-fifth percentile received power was chosen as an indicator by which the control approaches will be compared.

5.3 Channel Simulations

The indoor channel was simulated for a variety of scenarios: no diversity, two and three path selection and optimal combining diversity. In order to compare these channels, the long term average power was removed using average power control. Plots of the received power envelope after average power control is applied are found in Figure 5.1. These plots clearly indicate the large variation of base station received power in the absence of effective power control. Comparison of these plots shows the benefit of diversity; diversity decreases the number and severity of deep fades.

A decrease in the low power tail of the power density histograms illustrates the reduction in fading with the addition of diversity. This can be observed in Figure 5.2 where the frequency of observations between 0.0 and 0.5 watts decreases as diversity is added. Thus, some of the deepest fades were eliminated or decreased in severity as diversity was introduced.

Integrating over the density histogram gives the cumulative distribution function of the received power. These diagrams are scaled in normalized base station received power (dB) and clearly show the improvement experienced with diversity, Figure 5.3. This improvement is reflected in the increasing slope of the distribution curves as a perfect control system would be represented by a vertical line at 0 dB.

The tabulated statistics show that the introduction of diversity decreases the variance of the channel. Table 5.1 clearly illustrates this for a 100 symbol burst length system. In addition, the fifth percentile received power is increased. Although the improvement is less dramatic, the ninety-fifth percentile is also reduced. These
Figure 5.1  Received Power Envelopes with Average Power Control for: (a) Single Path (b) N=2 Selection (c) N=3 Selection (d) N=2 Optimal Comb. (e) N=3 Optimal Comb. Diversity Schemes.
Figure 5.2 Density Histograms of Received Power with
Average Power Control for: (a) Single Path
(b) N=2 Selection (c) N=3 Selection (d) N=2
Figure 5.3 Cumulative Distributions of Received Power for N=2: (a) Selection and (b) Optimal Comb. Diversity.
improvements support the qualitative findings that diversity decreases the variability of the channel.

5.3.1 Power Control for a Single Path

Power control without diversity is not sufficiently accurate for the application, regardless of the control method. This is due to deep fades that are very difficult to predict and track. Comparing the fifth and ninety-fifth percentiles of the sinusoid and parabola fit algorithms in the no diversity section of Table 5.1, one can see very little difference. However, the sinusoid fit algorithms tend to have much larger variances than the linear and parabola fit algorithms. This is likely due to the significantly large power over-estimates evident in the received power envelopes, Figure 5.4, and in the presence of upper tails (1.5 to 2.0 W) in the density histograms of Figure 5.5. These large variations and tails are not as significant when the parabola fit control algorithm is applied. In addition, the weighting functions used with the sinusoid fit algorithms seem to have the negative effect of increasing the received power variance.

The result of these findings is that the sinusoid fit algorithm with fixed amplitude (regardless of weighting functions) does not operate well without diversity. A possible explanation for this failure is the large amplitude required to cover the range of variation in a non-diverse channel causes a great number of extreme under and over-estimates. These poor estimates then contribute to the variance in the received power. This is best illustrated by examining the plots of received power in Figure 5.4. These plots clearly show extreme over-compensation taking place with up to +30 dB errors occurring.

5.3.2 Power Control with Selection Diversity

When a two path selection diversity scheme is introduced, a marked improvement is evident for all control schemes. The decrease in the number and depth of the channel fades afforded by the second path allows closer estimates to be made. A comparison of the received power envelopes in Figure 5.1(a) and (b) clearly shows the improvement in channel estimation. This improvement is also demonstrated by the large decreases in the variance of the received power for all of the control methods
Figure 5.4  Received Power Envelopes with No Diversity for:
(a) Sinusoid Fit, Unweighted  (b) Sinusoid Fit, Weighted 1 (c) Sinusoid Fit, Weighted 2 Control Schemes.
Figure 5.5 Density Histograms of Received Power given No Diversity for: (a) Sinusoid Fit, Unweighted (b) Sinusoid Fit, Weighted 1 (c) Sinusoid Fit, Weighted 2 Control Schemes.
studied, Table 5.1. The Density functions are also narrower as more diversity is used. This is illustrated in Figure 5.6 for the parabola fit algorithm. The other control methods show improvement similarly; the density histograms for these plots can be found by referring to Appendix B.

After selection diversity is introduced, the sinusoid fit algorithm maintains a smaller variance than the linear and parabola fit algorithms. The fifth percentile of the parabola fit method is slightly, but not significantly, smaller than the sinusoid fit algorithms.

Weighting the sinusoid remains ineffective with the introduction of selection di-
Figure 5.7 Cumulative Distributions of Received Power for \( N=2 \) Selection Diversity for Unweighted, Weighted 1, and Weighted 2 Sinusoid Fit Control Schemes.

versity. Figure 5.7 shows the significant portion of the received power cumulative distribution function. The similarity in the curves indicates weighting the sinusoids is not very effective. This can also be seen in Table 5.1 where the mean, variance, and percentile measures are not significantly different from the unweighted case.

One possible explanation for this phenomenon has to do with the nature of selection diversity itself. When the base station transmits, it does so using a single antenna, the one that received maximum power at the end of the last portable transmission. The portable thus has information about one path and transmits based on this path, Figure 5.8. As the portable transmits, the base station may use diversity to switch to another path. Since the paths are uncorrelated, the portable's transmission does not account for variation in the new path. In cases where the base switches antennas, the path initially followed is probably entering a fade, and the portable is compensating for this. The new path may be coming out of a fade or may be
Figure 5.8 Pictorial of a Possible Event Given N=2 Selection Diversity.
fairly constant. In either case, the portable will over-estimate the power requirement, causing excess power transmission, wasting valuable power and possibly introducing intolerable noise to a co-user's signal. Since it is likely that the portable is accounting for the wrong path anyway, weighting those values and changing the estimation slightly will most likely have no beneficial effect.

As a third antenna is added, improvement is evident. However, this improvement is not nearly as great as when the second antenna was added. The variance in received power, for every control method, is noticeably decreased. In addition the ninety-fifth percentile received power is decreased and the fifth percentile is slightly increased. The relative effectiveness of the control methods mirror those found in the examination of N=2 selection diversity.

Using the simplified assumption that capacity varies as the fifth percentile received power, the diminishing returns of adding antennas can be illustrated. Table 5.1 shows an increase from -0.3202 dB to -0.1408 dB for the sinusoid fit (unweighted) algorithm when a second antenna is used. This translates to about a 4.2 percent increase in capacity. As a third antenna is added, the fifth percentile increases to -0.1150 dB, indicating about a 0.6 percent increase.

Despite the slight disadvantage of the fifth percentile of received power, the lower variance of the sinusoid fit algorithms indicates their suitability for use in selection diversity systems. Weighting of the sinusoids has negligible effect and is therefore unnecessary.

5.3.3 Power Control with Optimal Combining Diversity

Optimal combining generally provides more accurate power control than the selection diversity and single path systems. This is most likely due to the tracked signal having smooth transitions and less variation allowing closer fits to be made using the smooth curves. A comparison of the plots of received power in Figure 5.1 shows that the base station received power plot for N=2 optimal combining diversity is noticeably smoother than the corresponding N=2 selection diversity plot. A particular point of comparison is near 1.75 seconds where selection diversity allowed a 3 dB fade whereas the optimal combining system allowed a fade of about 1 dB.
Figure 5.9 Received Power Envelopes Using N=2 Optimal Combining Diversity with: (a) Linear Fit (b) Parabola Fit (c) Sinusoid Fit, Unweighted Control Schemes.

Table 5.1 clearly indicates the benefits of optimal combining over single path and similar order selection diversity. In most cases the received power variance is significantly less for optimal combining. In addition, the percentiles also indicate the control limits are tighter.

In the optimal combined system, the parabola fit algorithm maintains a smaller variance than the other methods studied. In addition, the fifth and ninety-fifth percentiles are best for the parabola fit method. Comparison of the received power envelopes in Figure 5.9 qualitatively supports this. The received power for the parabola fit is less variable than the other methods, and has fewer large errors. The density histograms in Figure 5.10 show the parabola fit algorithm gives the narrowest spike.
indicating tighter control. The cumulative distribution function of received power further supports this: Figure 5.3(b) clearly shows the parabola control trace to have the greatest slope which indicates tighter control.

Optimal combining diversity benefits slightly from the weighting of the sinusoid curves. That is, the variance and percentiles are more optimal for the weighted curves than the non-weighted curve. This improvement is not dramatic and can be attributed to the fact that the effective signal being tracked is a weighted sum of a number of signals. This yields a much smoother curve than the selection diversity and no diversity cases, Figure 5.1. In addition, there are no induced discontinuities
such as those found in selection diversity.

The addition of a third antenna does not allow as much capacity gain as the addition of the second did. This can be illustrated by examining the fifth percentile base station received power with parabola fit power control. The second antenna increased the fifth percentile received power from -0.3199 dB to -0.0784 dB, this represents a capacity increase of about 5.7 percent. The third antenna yielded a fifth percentile received power of -0.0593 dB, an additional capacity gain of only 0.4 percent.

The parabola fit algorithm performed best according to all tests, and is therefore the most suitable algorithm for tracking an optimally weighted system.

5.4 Effect of Burst Length on Power Control

The indoor channel changes rather slowly relative to the bit rate of the system. However, there can be significant change over a data burst. Burst lengths of 50, 100, and 200 symbols were examined to determine the effect this has on power control.

The change in burst length from 200 to 100 symbols introduces significant savings in capacity. This capacity improvement can be illustrated by considering the parabola fit algorithm with three path optimal combining. The statistics collected in Tables 5.1 and 5.2 show that the fifth percentile changes from -0.3519 dB to -0.05927 dB. Using the simple assumption that capacity varies as the fifth percentile received power, a capacity increase of about six percent is evident. This is the order of the improvement all of the control methods experience as burst length is decreased: linear fit, parabola fit, sinusoid, and weighted sinusoid. The received power variance improved as well; the same control method produced a received power whose variance decreased from 0.00857 to 0.00022, a factor of 39 times.

As the data burst period was decreased from 100 to 50 symbols, further capacity improvement occurs, but the result is not nearly as dramatic. Considering the statistics listed in Tables 5.1 and 5.3, the fifth percentile for the parabola fit algorithm with three path optimal combining changes from -0.05927 dB to -0.009657 dB. This indicates an expected capacity improvement in the order of only one percent; the other
control methods indicate similar gains in capacity. Received power variance decreased from $2.191 \times 10^{-4}$ to $5.813 \times 10^{-6}$, a factor of 37.7.

Changing the burst length from 200 to 100 symbols results in a received power variance decrease of the same order of magnitude as that achieved when the burst length is decreased from 100 to 50 symbols. At the same time, the fifth percentile analysis indicated the capacity increase from 200 to 100 symbols is much greater than the increase achieved if the burst period was further reduced to 50. This indicates diminishing returns as burst length is decreased.

As the burst length is decreased the average guard time per burst becomes more important. If the average guard time is assumed to be one symbol long, the transfer efficiency\(^2\) (information divided by the total data transmitted) of a 200 symbol burst is 99.5 percent. Decreasing the burst length to 100 symbols decreases the efficiency to 99 percent, and a 50 symbol burst length is 98 percent efficient. Comparing these efficiency decreases to the power control capacity gains discussed earlier seems to indicate that a burst period of 100 symbols is near the optimal balance of power controllability and sustained burst length.

5.5 Chapter Summary

This chapter provides the analysis of the results collected from the simulations. It was discovered that open loop power control can be effective in controlling the base station received power so long as antenna diversity is available. In addition, decreasing the burst length and increasing system diversity was observed to provide diminishing returns.

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\(^2\)As the transfer efficiency decreases, the system can carry less information in a given time at that bit rate. In order to maintain voice quality, the bit rate may be increased to offset the loss in efficiency. This in turn decreases the number of users sustainable by the DS-SS system.
6. Conclusions

6.1 Summary

This thesis was undertaken to determine the accuracy of open loop power control for a DS-SS system in an in-building environment. This control system relied on channel measurements obtained by using forward link power measurements made at the portable receiver.

Control accuracy was assessed by estimating the capacity degradation due to imperfect power control. The effect of antenna diversity and burst length on capacity were also considered.

The work done to accomplish the thesis objectives included:

- Modeling the indoor channel as a flat Rayleigh fading channel with a maximum 6 Hz Doppler shift.
- Curve fitting to the received power envelope to mitigate the near-far problem for a TDD scenario.
- Comparing of linear, parabola, and sinusoid curve fit algorithms with respect to maximizing system capacity.
- Determining the effect of optimal combining and selection (antenna) diversity on the various power control schemes.
- Determining the effect of burst length on power control error for bursts of 50, 100, and 200 bits.

6.2 Thesis Findings

The proposed TDD based open loop control schemes significantly decrease the effects of multipath fading with minimal antenna diversity. The chief findings of this thesis are:
A parabola fit algorithm offers slightly better power control than a straight line fit.

The sinusoid fit algorithm offers power control of slightly better quality than the parabola fit algorithm when selection diversity is used. The parabola fit algorithm fares better in optimal combining diversity and no diversity cases.

Weighting the sinusoid fit data points is either ineffective or has minimal positive result for the diversity conditions studied.

Antenna diversity is necessary for power control to be effective.

Two antenna diversity offers significant capacity improvement over a single antenna whether selection or optimal combining diversity is used.

There is diminishing returns in increasing the order of antenna diversity. While three antenna diversity offers some improvement over two antenna diversity, the improvement is not significant.

Optimal combining diversity provides more capacity than selection diversity when parabolic or linear prediction is used.

No definite conclusion on optimum burst length can be reached from the research results obtained in this thesis but it appears this optimum is near 100 bits.

Open loop power control has been shown to be effective in TDD CDMA spread spectrum communication. With this control, Time Division Duplex Direct Sequence Spread Spectrum systems are candidates for indoor wireless telephone.

6.3 Future Directions

A number of assumptions and simplifications were made in order to reduce the complexity of the study. Further study could be undertaken to expand this work and enable a more precise picture of the effectiveness of this power control.
• One could look at the effect of a frequency selective channel model as most DS-SS systems gain some advantage from a wider bandwidth.

• Noise effects on the measured power could be modeled and simulated.

• The effect of interleaving to minimize the effects of the “bursty” nature of the received power could be studied for wireless computer network applications.

• A full analysis including spatial and shadowing effects could be done.

• The interdependence of power control and PN-code synchronization could be studied given this form of power control.
References


A. Plots of Normalized Received Power

This appendix illustrates graphically the normalized base station received power, \(10 \log \frac{\text{received power}}{1 \text{ watt}}\), for the various control methods and diversity schemes. Three seconds of the 3.125 second (50000 symbol) simulation run are depicted in the plots. A burst length of 100 symbols (6.25 ms) was used to show the appearance of the received power envelope; 50 and 200 symbol burst length simulations were conducted but the plots of these envelopes are not included here for brevity. The set-point for the received power is 1 W for all simulations.
Figure A.1 Received Power Envelopes with No Diversity with: (a) Average (b) Parabola Fit Control Schemes
Figure A.2 Received Power Envelopes with No Diversity with: (a) Linear Fit (b) Parabola Fit (c) Sine­
soid Fit, Unweighted (d) Sine­soid Fit, Weighted 1 (e) Sine­soid Fit, Weighted 2 Control Schemes.
Figure A.3  Received Power Envelopes Using N=2 Selection 
Diversity with: (a) Average (b) Parabola Fit 
Control Schemes
Figure A.4 Received Power Envelopes Using N=2 Selection Diversity with: (a) Linear Fit (b) Parabola Fit (c) Sinusoid Fit, Unweighted (d) Sinusoid Fit, Weighted 1 (e) Sinusoid Fit, Weighted 2 Control Schemes.
Figure A.5 Received Power Envelopes Using N=2 Optimal Combining Diversity with: (a) Average (b) Parabola Fit Control Schemes
Figure A.6 Received Power Envelopes Using N=2 Optimal Combining Diversity with: (a) Linear Fit (b) Parabola Fit (c) Sinusoid Fit, Unweighted (d) Sinusoid Fit, Weighted 1 (e) Sinusoid Fit, Weighted 2 Control Schemes.
Figure A.7 Received Power Envelopes Using N=2 Selection Diversity Diversity with: (a) Average (b) Parabola Fit Control Schemes
Figure A.8  Received Power Envelopes Using N=3 Selection Diversity with: (a) Linear Fit (b) Parabola Fit (c) Sinusoid Fit, Unweighted (d) Sinusoid Fit, Weighted 1 (e) Sinusoid Fit, Weighted 2 Control Schemes.
Figure A.9 Received Power Envelopes Using N=3 Optimal Combining Diversity with: (a) Average (b) Parabola Fit Control Schemes
Figure A.10 Received Power Envelopes Using N=3 Optimal Combining Diversity with: (a) Linear Fit (b) Parabola Fit (c) Sinusoid Fit, Unweighted (d) Sinusoid Fit, Weighted 1 (e) Sinusoid Fit, Weighted 2 Control Schemes.
B. Density Histograms of Received Power

The figures in this appendix illustrate the density histograms of the power received at the base station. Fifty thousand symbols (3.125 seconds) were simulated and the resulting densities were computed as a percentage of this total and plotted on a semilog graph. A burst length of 100 symbols (6.25 ms) was used to show the appearance of the received power Density Histograms. Fifty and 200 symbol burst length simulations were conducted but the graphics for these are not included here for brevity. The set-point power for these simulations was 1 watt and the bin size was 10 mW.
Figure B.1 Density Histograms of Received Power given No Diversity for: (a) Average (b) Linear Fit (c) Parabola Fit (d) Sinusoid Fit, Unweighted (e) Sinusoid Fit, Weighted 1 (f) Sinusoid Fit, Weighted 2 Control Schemes.
Figure B.2 Density Histograms of Received Power given N=2 Selection Diversity for: (a) Average (b) Linear Fit (c) Parabola Fit (d) Sinusoid Fit, Unweighted (e) Sinusoid Fit, Weighted 1 (f) Sinusoid Fit, Weighted 2 Control Schemes.
Figure B.3 Density Histograms of Received Power given N=2 Optimal Combining Diversity for: (a) Average (b) Linear Fit (c) Parabola Fit (d) Sinusoid Fit, Unweighted (e) Sinusoid Fit, Weighted 1 (f) Sinusoid Fit, Weighted 2 Control Schemes.
Figure B.4 Density Histograms of Received Power given $N=3$ Selection Diversity for: (a) Average (b) Linear Fit (c) Parabola Fit (d) Sinusoid Fit, Unweighted (e) Sinusoid Fit, Weighted 1 (f) Sinusoid Fit, Weighted 2 Control Schemes.
Figure B.5 Density Histograms of Received Power given $N=3$ Optimal Combining Diversity for: (a) Average (b) Linear Fit (c) Parabola Fit (d) Sinusoid Fit, Unweighted (e) Sinusoid Fit, Weighted 1 (f) Sinusoid Fit, Weighted 2 Control Schemes.
C. Cumulative Distribution Functions of Normalized Received Power for Various Control Algorithms

This appendix contains cumulative distribution functions for normalized power, $10 \log \frac{\text{received power}}{\text{1 watt}}$, received at the base station. The integration of the density diagrams in Appendix B results in the curves presented in these figures. Each control measure and diversity situation is analyzed to evaluate the effectiveness of the various control mechanisms at combating multipath fades.

A burst length of 100 symbols (6.25 ms) was used to show the appearance of the cumulative distribution function. Fifty and 200 symbol burst length simulations were conducted but those plots are not included here for brevity.

The desired received power for all simulations was 1 Watt.
Figure C.1 Cumulative Distributions of Received Power for Average Power Control with: (a) Selection and (b) Optimal Combining Diversity Schemes.
Figure C.2 Cumulative Distributions of Received Power with Selection Diversity for: (a) Linear Fit (b) Parabola Fit (c) Sinusoid Fit, Unweighted (d) Sinusoid Fit, Weighted 1 (e) Sinusoid Fit, Weighted 2 Control Schemes.
Figure C.8 Cumulative Distributions of Received Power with Optimal Combining Diversity for: (a) Linear Fit (b) Parabola Fit (c) Sinusoid Fit, Unweighted (d) Sinusoid Fit, Weighted 1 (e) Sinusoid Fit, Weighted 2 Control Schemes.
D. Cumulative Distributions of Normalized Received Power for Various Diversity Schemes

This appendix contains cumulative distribution functions of normalized power, $10 \log \frac{\text{received power}}{1 \text{ watt}}$, received at the base station. This information is organized such that comparisons may be drawn between average, linear, parabola, and sinusoid control with each type of diversity studied.

A burst length of 100 symbols (6.25 ms) was used to show the appearance of the cumulative distribution function. Fifty and 200 symbol burst length simulations were conducted but these plots are not included here for brevity.

The desired received power for all simulations was 1 Watt.
Figure D.1 Cumulative Distributions of Received Power in the Absence of Diversity: (a) Average, Linear Fit, Parabola Fit and Unweighted Sinusoid Fit (b) Unweighted, Weighted 1, and Weighted 2 Sinusoid Fit Control Schemes.
Figure D.2 Cumulative Distributions of Received Power for N=2 Selection Diversity: (a) Average, Linear Fit, Parabola Fit and Unweighted Sinusoid Fit (b) Unweighted, Weighted 1, and Weighted 2 Sinusoid Fit Control Schemes.
Figure D.3 Cumulative Distributions of Received Power for N=3 Selection Diversity: (a) Average, Linear Fit, Parabola Fit and Unweighted Sinusoid Fit (b) Unweighted, Weighted 1, and Weighted 2 Sinusoid Fit Control Schemes.
Figure D.4 Cumulative Distributions of Received Power for N=2 Optimal Combining Diversity: (a) Average, Linear Fit, Parabola Fit and Unweighted Sinusoid Fit (b) Unweighted, Weighted 1, and Weighted 2 Sinusoid Fit Control Schemes.
Figure D.5 Cumulative Distributions of Received Power, N=3 Optimal Combining Diversity: (a) Average, Linear Fit, Parabola Fit and Unweighted Sinusoid Fit (b) Unweighted, Weighted 1, and Weighted 2 Sinusoid Fit Control Schemes.