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**Direct Sequence Spread Spectrum Multiple Access
System Performance in an Indoor Milieu
Characterized by Rician Channel.**

by

Eng Chai Lim

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

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ABSTRACT

The objective of this thesis is to investigate the performance of a direct sequence spread spectrum multiple access (DS/SS-MA) system in an indoor environment which is expected to undergo a large amount of multipath fading. The indoor wireless communication (IWC) system in our study employs selection diversity (i.e., DS/SS and space diversity) and differential phase shift keying (DPSK) modulation of the information data bit stream. In our work, we simulate the system in the form of discrete models which lends itself to the study of the effects of resolving the multipath and exploiting it in the form of diversity. We determined that DS/SS diversity, in comparison to no diversity, improves the performance of the IWC system in a multipath fading indoor environment. However, we show that the improvement, in system performance, is greater when space diversity reception operates in tandem with DS/SS modulation.

The performance analyses are based on actual indoor radio propagation measurements and Rician channel model reported in [9]. The propagation measurements were conducted in order to determine the characteristics of 910 MHz radio channels operating in several different indoor office environments. The performance of the scheme under various possible scenarios was computed as a result of this reported channel characteristics. We compared the performance of this scheme for two different code sequence sets ($N = 127$ Gold and $N = 255$ Kasami codes) and determined that there is significant advantage in deploying the $N = 255$ Kasami sets. Specifically, the performance of the IWC scheme at information data rates of 32 kb/s, 64 kb/s and 144 kb/s were found. These data rates are compatible with the new cellular access digital network (CADN) architecture recently proposed in [1].

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I. INTRODUCTION

1.1 INDOOR WIRELESS COMMUNICATIONS SYSTEMS

In recent years, indoor wireless communications (IWC) systems have become an active research area [1-9]. This research is spurred on by the increasing proliferation of terminals, workstations, private branch exchanges (PBX's) and telephones in the office environment. However, while there is sufficient technology to transmit high volumes of voice and data information within an integrated network, the interface of user terminals to the network remains costly and restrictive.

The existing communication network which provides connectivity via wire cables, or in recent years fibre optic cables, are not cost efficient or readily amenable to changes and enhancements. The installation of a new terminal or relocation of an existing equipment may necessitate cable pulling or laying of more conduits (if there is no spare or available connections at the chosen site) and often requires a significant lead time for installation. In addition, subscriber loops and building wiring is expensive to install and raises the overall construction costs for new buildings. Wired connections also limit users' mobility. The need for such physical facility can be avoided with wireless communication. The potential cost reductions, flexibility and mobility features provided by wireless communications will make possible a variety of applications not only in residential and building offices, but also in other structures such as stock exchanges, warehouses, manufacturing plants, carriers (land, sea and space), instrumentation environments, etc.

Three different types of signals may be used for indoor wireless communications [S]: ultrasonic, infrared and radio frequency (RF) signals. Among these three, infrared and RF signals are the most promising candidates as the limited ultrasonic band would preclude providing high data rate services. Infrared signalling does not seem as promising as RF signalling due to limitations imposed by interference sources such as ambient light, humidity and smoke in the working environment which reduce the effectiveness of infrared signalling at high data rate transmission.

The application of RF signalling for indoor application requires selection of an RF band. The very high frequency (VHF—30 to 300 MHz) and ultra high frequency (UHF—300 MHz to 1 GHz) bands are both suitable. However, because of the overcrowding and congestion (coupled with the limited total bandwidth allocation) in the VHF and lower end of the UHF spectrum by both commercial and military users, frequencies in the 900 MHz range (formally released for advanced high capacity cellular land mobile radio communication systems in 1981 by FCC and DOC) seem most appealing for low power communication applications in the indoor environment. The 900 MHz frequency band selection is a direct result of the increasing demand for mobile radio communication services such as radiophones, radio paging, radio or emergency dispatch, navigation control, status reporting, etc. The 900 MHz band has sufficient bandwidth (a 40 MHz bandwidth allocation) to accommodate this increased demand. Several cellular mobile radio systems currently operate in this band in various countries. Cellular mobile radio systems are efficient spectrum users due to their high density spatial co-channel frequency reuse.

In the indoor environment, the rapidly growing demand for new communication services along with recent advances in technology have created a strong rationale to provide an overall integrated services digital network (ISDN). The type of services, existing or proposed, run the gamut from telemetry applications (security/alarms or annunciators, telecontrol or supervisory control and data acquisition activities such as in the form of energy management) to high definition television programming with channel capacities requirements ranging from 1 kb/s up to about 1 Gb/s. These indoor communication services can be classified by information type (such as, voice, sound, video and includes text, graphics and data communications) in each of the different types of market (residential and business). The ultimate goal of any indoor communication network is to convey all types of information (voice and data) among machines and people at all desired locations. The emergence of residential cordless telephones, however popular, cannot replace wired phone sets within an office building because they are constrained by regulatory requirements (FCC and DOC), geographical range, susceptibility to interference and lack of privacy. This is a result of the wireless telephones operating in a limited voice-bandwidth allocated in the overcrowded VHF (46 and 49 MHz) range, which is heavily occupied by a plethora of commercial and military users. *Cellular technology*, at the higher 900 MHz carrier frequency band, applied within a building provides a means of overcoming these shortcomings

with its greater set of features such as its ability to page called wireless terminals, regulate channel access and direct transmission to different wired cell sites as a terminal moves through the building.

Recent indoor radio propagation studies, such as in [5], have shown the propagation channel of the office environment has a proclivity for frequency selectivity fading for RF propagation at 900 MHz. To counter the effects of frequency selectivity fading, one method suggested is to employ *direct sequence spread spectrum (DS/SS) modulation* [2-4,9,10]. DS/SS modulation, until recently, has been an esoteric method widely used in military communication systems, mainly to provide resistance to jamming and interception measures by the opposing side. In the same way, the IWC system employing this DS/SS modulation technique, should cause little interference as the signal will be severely attenuated outside an office building. The capability to accommodate frequency reuse by signal overlay, using code division multiple access (CDMA), is the key reason for the relaxation and encouragement of recent rulings by the FCC [11,12] which permit spread spectrum development for commercial applications. However, the major motivation for employing DS/SS modulation in the indoor environment remains that of providing a useful multipath countermeasure.

In the implementation of any communication system the selection of a suitable modem (modulation-demodulation) technique requires some tradeoffs. In general, coherent demodulation techniques offers superior performance especially for stationary or fixed-point to point use. However, attaining synchronous carrier recovery in a multipath fading milieu is difficult in practice. The need to recover the RF carrier in such a system thus leads us to consider incoherent detection as an alternative. Therefore for our purpose we select differential phase shift keying (DPSK) modulation as a viable technique. To counter the deleterious effects of multipath fading, the combination of DS/SS modulation with space diversity is advocated. In practice, space diversity (in the form of antenna selection diversity) is more efficient and is less complex than maximal ratio combining especially when two antennae is configured for diversity reception. We note that selection diversity, in our case, can be realized either by space diversity or by resolving the multipaths or both. We contend that space diversity reception when operating in tandem with DS/SS modulation will have synergistic countereffects against multipath fading typical of an indoor environment. We shall examine this aspect of such a system in detail in this thesis.

In this thesis, we consider a DS/SS-MA scheme, illustrated in Fig. 1.1, for our IWC research. In the IWC system under consideration, we assume that there are K spur links where the users are simultaneously trying to communicate with a base station using DS/SS. Each user is assigned a distinct spread spectrum code sequence of period N from a set of sequences with low cross-correlation properties. This allows the receiver at the central station to separate the different received signals on the basis of code-division multiplexing.

This DS/SS-MA scheme was analysed in [3,4] with switching functions based on digital PBX's (handled using standard PBX features) and shared with wired-based ISDN applications. The performance of this scheme at information data rate of 32 kb/s was investigated assuming the indoor environment could be modelled by a *Rayleigh fading channel*. More recently, measurements of some actual indoor propagation characteristics [7] have suggested that a *Rician fading channel model* is more appropriate. Our intent in this thesis is to evaluate the system performance with actual path data and at higher data rates (up to 144 kb/s). Before we proceed with the systematic study of the scheme let us outline the objectives and general specifications of an IWC system.

1.2 OBJECTIVES AND SPECIFICATIONS OF AN IWC SYSTEM

1.2.1 Integrated System Approach

In designing an IWC system, the engineering objectives should be considered from a wider perspective rather than just simply providing a link between terminals and central stations or providing voice communications between people working in their offices. As a result of the need to provide an overall integrated system approach to IWC with access to advanced integrated services digital network (ISDN), a new cellular access digital network (CADN) was proposed recently [1]. The new CADN architecture (Fig. 1.2) will introduce a systematic approach to IWC by including the capabilities of cellular technology to indoor digital radio transmission technologies. The key components of the architecture are the proposed functional standards; the cellular integrated services digital networks (ISDN) standards (Fig. 1.3) and the cellular digital RF standards (Fig. 1.4). The main features of this system are the all digital transmission, increased bandwidth efficiency, enhanced control over network functions and *wireless access* to many integrated services digital network (ISDN) features to be introduced to the public networks.

Fig. 1.1. IWC-DS/SS-MA scheme

The scheme illustrated in Fig. 1.1 shows a basic indoor cell which may represent an indoor cell site (ICS) in the proposed CADN architecture. The base or central station can in turn be linked to other base stations (indoor or outdoor cell sites) based on the proposed digital cellular standards. As discussed above, the key feature of this system is the provision of wireless accesses to ISDN.

The adoption of a CADN network or similar structure will provide a systematic approach to IWC. Each CADN user will have access to two types of channels: I (information) channels and C (control) channels as illustrated in Fig. 1.5. The I channels will be used for carrying user packet data while the C channels, will be designated for transmitting network control signalling information and user information. The CADN user will be allowed to either (2I+C) channels, similar in some aspects to the ISDN basic access (2B+D) structure, (I+C) channels or a C channel at any given moment and understand that the need maybe allowed to change during a call and that the channel rates may vary. Based on these interface standards, we will investigate the performance of the IWC system for three different source rates, i.e., 32 kb/s, 64 kb/s and 144 kb/s. We note that the indoor CADN switching functions could also be based on digital PBX's (with call set-up handled using standard PBX features) and shared with wired-based ISDN applications.

In summary the overall IWC system should offer the following:

- (a) low cost and ease in implementation,
- (b) high quality and reliability,
- (c) expandability, flexibility, portability, mobility,
- (d) inherent addressing, security, interception resistance and hence traffic privacy,
- (e) multipath resistance capability, i.e., system designed should be robust to withstand varying channel conditions, i.e., entail requirement of a robust modulation technique,
- (f) frequency reuse by signal overlay with respect to other systems and lastly
- (g) be an integral part of the overall CADN architecture.

Fig. 1.2. CADN Network (from [1])

Fig. 1.3. Cellular ISDN Protocol Architecture (from [1])

Fig. 1.4. Cellular Digital RF Protocol Architecture (from [1])

Fig. 1.5. CADN Basic Access (from [1])

1.2.2 IWC System Design Considerations

1.2.2.1 Indoor Radio Propagation Considerations

From a system engineering viewpoint, based on past experience, a considerable amount of time and resources are required to study the radio propagation environment for which the radio systems are meant to be used. To achieve a stable radio propagation path in any environment, stable or almost ideal path conditions have to exist such as ideal climatic or ambient conditions and suitable siting of radio terminals. This requires that no path obstructions with intervening terrain sufficiently “broken” to eliminate ground reflections and that there be no external reflecting bodies (termed *signal reflectors*) or objects or *scatterers* within the vicinity. Obviously such ideal conditions seldom exist, especially in the UHF and microwave radio environment, and the bane of many a radio system engineer is the existence of multipath propagation effects due to lack of some of the ideal or stable factors described above.

To illustrate the multipath fading environment let us consider an IWC link or a single hop radio system as shown in Fig. 1.6. We note that in an IWC system, there are three possible types of RF links: *fixed-point to point link* (i.e., fixed-terminal to fixed-base station), *mobile to fixed-point link* (i.e., portable terminal to fixed-base station) and *mobile to mobile link* (could be direct portable terminal to another portable terminal or in the case of two hop radio link, portable to portable terminals via a fixed-base station). The potential for multipath interference arises when transmitted signals, as a result of signal reflectors and or scatterers in the medium, reach the receiver via two or more paths with different overall lengths and arrivals. Depending on the frequency and relative path lengths, the possibility of the signal over one path arriving in phase with signals over other paths, and remain so, is extremely remote. Consequently, the signals may add destructively producing *deep nulls* or *trough coincidences* in the signal power available at the receiver as graphically demonstrated in Fig. 1.7.

Fig. 1.6 An IWC radio link under multipath fading illustrating discrete multipath channel

Fig. 1.7. Deep nulls or Trough co-incidences

Basically, there are two kinds of channel variations, based on distinctions in time scales for the fluctuations, known as: (a) *short-term fading* and (b) *long-term fading*. Roughly, the short-term variations concern the duration (typically ranging from fractions of seconds to a few seconds) of the temporal fades and affect the details of the received signals structure and thus concern the possibility of errors in reception of the information signal. Long term fading is concerned with the availability of the channel, where the actual structure of the channel and the mean path attenuation changes substantially, with a downtime magnitude in the order of minutes, hours or days for that matter. In this thesis we are primarily concerned with short term fading. The multipath fading phenomenon may also be classified into *slow or fast fading* mode. The slow fading mode pertains to variations of the local mean space over larger distances. It is primarily attributed to a change in the indoor configuration or topography and generally due the presence of large objects or scatterers interposed between the IWC link and usually at greater distances from the intended receiver. Fast fading, on the other hand, is a rapid variation of the received signal over a short length (i.e., at a quarter to three quarters of a wavelength where one wavelength corresponds to about 36 cm. for 1 GHz frequency range) with maximum variation of 40 dB reported in [7] and 60 dB in [13]. This fast fading is due to the interference pattern caused by the local scattering of the signal, i.e., the effects of local scattering results in the combination of multipaths of about the same strength but with different phases.

In the indoor environment, there are two situations where variations in the received signal envelope as a result of multipath fading should be taken into consideration. For a fixed-radio link, *temporal variations* in the received signal envelope, caused by the movement of personnel and machinery, should be taken into consideration. For a mobile link, in addition to this temporal fading, considerations must be given for signal variations due to movement in the *spatially distributed field patterns*. That is, the multipath fading model in the mobile indoor system should take into account the relative movement of the radio terminals, in addition to other changes in the environment.

It has been found [14,15] that when the received signal power is measured, as a result of either the spatial (in a vicinity of tens of wavelength) and or temporal (in tens of seconds) variations of multipath fading channel, fast and deep fluctuations are observed in the received signal envelope. This is due to the large number of waves propagating

randomly in different directions as a result of randomly moving scatterers or presence of reflectors in the propagating medium. In mobile radio propagation studies, in most instances (i.e., when there is no predominant or contributing component), the fluctuated signal or temporal fading envelope observed exhibit *Rayleigh fading distributions*. However, in addition to the randomly moving scatterers or presence of reflectors, when there is a dominant or contributing component (viz a direct line of sight or a specularly reflected component) the envelope fading is said to be characterized by a *Rician distribution* such as in the case of the indoor environment observed in [7].

There are instances where, as a result of a moving or mobile receiver (in the case of mobile to fixed-point or mobile to mobile links), the phenomenon of *Doppler effect* should be considered. This effect shifts the frequency of the signal. This shift is dependent upon the speed of the moving receiver relative to the transmitter and the angle of arrival between the signals and the direction of the heading mobile receiver. As the received signal is a combination of several multipath components from different directions, the Doppler shift is not uniform and thus spreads the signals over a bandwidth wider than originally transmitted. *Doppler spread or time-selective fading* effect is regarded as the temporal decorrelation effect of the multipath fading and may impose a limit on the bit error rate (BER) performance which cannot be improved by increasing the transmitter power. The maximum Doppler frequency or Doppler spread (denoted as B_d) of a channel is the measure of the spectral width of the received carrier when a single sinusoidal carrier is transmitted through the multipath fading channel. In practice, the associated term *coherence time* (which is inversely proportional to the Doppler frequency) is usually defined as the time interval to obtain an envelope correlation of 0.9.¹

Similarly, based on the dual relationship between time and frequency, the term *coherence bandwidth* (denoted as Δf_c) is defined as the frequency spacing to have an envelope correlation of 0.9 or less. It can be shown that this bandwidth is inversely proportional to the *time delay spread*.² The time delay spread of a channel is the measure of the temporal

---

¹ In some references or texts such as [16,17] the precise definitions of coherence time and its exact inverse relationship to the Doppler spread, often differs based on the threshold value of the correlation used to determine coherence time.

² As above, the precise definitions of coherence bandwidth and its exact inverse relationship to the multipath delay spread, are usually based on the the threshold value of the correlation used to determine coherence bandwidth.

width of a received multiple-impulsive carrier transmitted through the multipath fading channel. The effect of time delay spread, which can be regarded as *frequency-selectivity fading*, must be taken into account when the bandwidth (denoted as W) of the transmitted signal is greater than the coherence bandwidth of the channel. The delay spread in our system is actually a random quantity; the precise definition of the time delay spread that we use here is that of the root-mean-square delay spread defined in [6] by

$$T_m = \sqrt{\sum_l E\{[\tau_l - \bar{\tau}]^2\}}$$

where τ_l is the delay of the l th component and $\bar{\tau}$ is the mean delay. Thus, in summary, the multipath fading phenomena can be divided into three categories: (a) Envelope fading (nonselective), (b) Time-selective fading (Doppler Spread) and (c) Frequency selective fading (time delay spread).

Having focused on the parameters of the multipath channel briefly, let us consider some common techniques to overcome the effects of multipath fading.

1.2.2.2 Diversity Techniques (Multipath Fading Countermeasures) Considerations

As a result of deep nulls or large channel attenuation due to the effects of multipath fading, errors occur in reception of the information transmitted. To combat the deleterious effects of multipath fading, diversity techniques are employed. The deployment of diversity techniques is based on the notion that if we can provide to the receiver several (or M) copies of the same information signals transmitted over what are essentially independently fading channels, the probability that all the signal components will fade simultaneously is reduced considerably. To provide this diversity there are several different techniques we can employ such as:

- (a) Frequency Diversity - The information signal is transmitted over several or M carriers where the guardbands or separation between successive carrier frequencies equals or exceeds the coherence bandwidth of the channel.
- (b) Space Diversity - This method involves employing more than one (M) antennae, spaced sufficiently apart so that the multipath components of the signal transmitted arrive with significant different propagation delays at the receiving antennae. The common method is to employ one transmitting antenna and multiple receiving antennae (judiciously spaced apart).

- (c) Time Diversity - The information signal is transmitted in M different time slots where the separation between successive time slots equals or exceeds the coherence time of the channel.
- (d) Wideband Diversity - This method is based on the application of the signal, with a bandwidth (W), transmitted through a channel with multipath delay spread (T_m) where W is much greater than the coherence bandwidth ($\Delta f_c \approx 1/T_m$) of the multipath channel. Such a signal when received will provide resolvable multipath components with a resolution in time delay of $1/W$, i.e., it will provide the receiver with M (where $M \approx T_m W$) separate independently fading signal paths. This technique is thus viewed as an alternate method for providing frequency diversity of order M .

In the indoor environment, multipath fading is a major concern and to ameliorate this problem appropriate countermeasures are required. In practice, space diversity is a simple and effective multipath countermeasure and has been efficiently employed. When combined with or operating in tandem with another diversity technique such as frequency diversity (typically found in microwave radio systems that operate over hostile multipath fading environment) the overall system performance and path reliability has proven to have increased. In our IWC study, we will employ a DS/SS-MA system configured with space diversity. The application of DS/SS modulation technique is based on the notion that spreading the transmitted signal bandwidth (where the resultant bandwidth will be much greater than the coherent bandwidth of the channel) will provide the inherent frequency diversity requirement.

To illustrate this concept, let us consider, hypothetically, transmitting a very narrow pulse or equivalently a wideband signal over a multipath fading link as illustrated in Fig. 1.6. The received signal, as a result of multipath scattering, might appear as discrete pulses as (Fig. 1.8) with a time delay spread T_m . Suppose, in lieu of the narrow pulse signal, we had a DS modulated signal transmitted over the same multipath channel. As a result of the multipath effects we assume that the discrete multipath channel (see [18] for the description of Turin's classical discrete multipath channel with DS/SS signalling) will provide, in this case, $M = 4$ resolvable multipath components as shown by the correlator or matched filter outputs (Fig. 1.9).

Fig. 1.8 Illustration of time delay spread with four discrete multipath components as a result of transmission of narrow pulse signal via discrete multipath channel of Fig. 1.6

Fig. 1.9 Illustration of DS modulated signal with four resolved multipath components for a single transmission via discrete multipath channel of Fig. 1.6

Note the peak of the main lobes of each component have to be separated by at least one chip time, denoted by T_c (which is inversely proportional to the DS/SS signal bandwidth), for it to be resolved. This technique is thus viewed as an alternate method for providing “frequency diversity” of order $M = 4$. This diversity technique actually dates back (circa 1958) to the concept of RAKE correlator or matched filter for a HF system [19].

In summary, the application of direct sequence spread spectrum (DS/SS) is motivated by the following advantages:

- (a) asynchronous multiple access capability via code division multiplexing (CDMA),
- (b) inherent addressing, security, interception resistance and hence traffic privacy features,
- (c) frequency signal overlay with respect to other RF systems
- (d) low cost and ease in implementation.
- (e) compared with infrared or ultrasonic signalling, DS/SS offers higher data rate and range,
- (f) inherent frequency diversity. SS systems are less susceptible to multipath signal variations than conventional systems. In a DS receiver if the reflected signal is delayed, compared to the direct-path signal by more than one code chip, the reflected signal is treated exactly the same as any other uncorrelated signal. The higher the code chip rate for a particular system the smaller the multipath problem, i.e. as the code rate is increased the path length for which the reflected signal interferes is further reduced.

However, there are some drawbacks that has to be addressed such as

- (a) the presence of near-far field problem which requires average power control policies and
- (b) the reduction or elimination of potential interference to other existing systems.

1.3 THESIS ORGANIZATION

The major part of the two chapters of this thesis, following this introductory chapter, provides review material. Chapter II provides relevant background material on multi-channel digital signalling in an AWGN environment. The general concepts of transmitter, channel and receiver models (with the typical DPSK modem as an example) are presented as stepping stones or guides to our DS/SS-MA system modelling in subsequent chapters. A detailed discussion on the multipath fading phenomena is presented and the statistical models (Rayleigh and Rician distributions) for multipath fading are developed. Chapter II also develop an expression for the probability of error of the modulation-demodulation schemes based on the characteristic function approach. Chapter III then discusses the DS/SS-MA scheme in the context of CDMA network architecture. The DS/SS-MA technique is briefly reviewed and the selection of DS codes for CDMA application is detailed. Specifically, two suitable CDMA codes, Gold and Kasami codes, with their low cross-correlation properties, are discussed. A design implementation of the proposed IWC system is included to complete this chapter.

The remaining chapters represent the research contributions of this thesis. Chapter IV deals with the modelling of the IWC system considered in this thesis and derives the system performance equations based on the proposed system models. Chapter V discusses the implementation of the simulator developed for our IWC study. The performance evaluation of the IWC system is derived based on the system performance parameters (given in Chapter IV) and the probability of error expression (developed in Chapter III). The details and results of the simulation, based on the IWC system models and actual Rician path values, are included in this chapter. Lastly, Chapter VI outlines the conclusions reached with some suggestions for further research.

II. MULTICHANNEL DIGITAL SIGNALLING-REVIEW

Before we discussed the general transmitter, channel and receiver models of a multi-channel system let us review the binary DPSK modem scheme. This signalling technique will serve as an example in our multichannel system.

2.1 DPSK MODULATION-DEMODULATION SCHEME

The DPSK signalling scheme makes use of an ingenious technique to avoid the need for a coherent reference signal at the receiver, i.e., the phase reference for demodulation is derived from the phase of the carrier from the previous signalling interval. As a result the receiver is able to decode the information from the differential phase as described below.

At the DPSK transmitter, the information is encoded into phase differences between two successive signalling intervals as shown in the block diagram in Figure 2.1. The differential encoder maps the information signal, $v(t)$ (where $v(t)$ is a NRZ PAM signal), represented by the binary data stream $\dots, g_{-1}, g_0, g_1, \dots$ ($g_n \in \{-1, 1\}$) into the binary sequence ($b_n \in \{-1, 1\}$) through the following logic operation³

$$b_n = g_n b_{n-1}. \quad (2.1)$$

Thus, as a result of the differential encoding process the transmitted signal, $s(t)$, changes its phase by shifting the carrier by π radians relative to the phase in the previous signalling interval when $g_n = -1$. When $g_n = 1$ there is a zero or no phase shift relative to the phase in the previous signalling interval. Hence the transmitted signal can be represented by

$$s_o(t) = \cos(\omega_c t + (b_n \pi/2) + \phi), \quad (2.2)$$

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³ If the input binary sequence is given by the sequence ($g_n \in \{1, 0\}$), the output of the differential encoder is generated by $b_n = b_n \oplus b_{n-1}$. A level shift is then required to remap the encoded bit stream, $b_n \in \{0, 1\}$, into $b_n \in \{1, -1\}$.

for $(n-1)T \leq t \leq nT$, where T is the symbol duration, ω_c is the angular carrier frequency and ϕ the carrier phase. In a bandlimited system the transmitted signal, at the n th signalling interval, is represented as

$$s_o(t) = (\cos(\omega_c t + (b_n \pi/2) + \phi)) * h_T(t), \quad (2.3)$$

where $h_T(t)$ is the impulse response of the transmit filter and $*$ denotes convolution operation.

Fig. 2.1. Block diagram of a DPSK transmitter

Figure 2.2 shows the block diagram of the DPSK receiver. The received signal is filtered through a predetection BPF to limit the noise entering the differential detector. The filtered received signal, at the n th signalling interval, can be represented as

$$r_o(t) = (\cos(\omega_c t + (b_n \pi/2) + \phi)) * h_I(t) + n(t), \quad (2.4)$$

where $n(t)$, the narrow-band Gaussian noise, is given by

$$n(t) = n_c(t) \cos(\omega_c t) + n_s(t) \sin(\omega_c t) \quad (2.5)$$

and $h_I(t)$ is the equivalent IF filter representing the overall transmit and predetection bandpass filtering.⁴ The receiver correlates the received signal in (2.4) with a delayed

⁴ For simplifying this discussion and within the context of this thesis, we will not consider the bandlimiting case and hence assume that there is no intersymbol interference (ISI).

(delay equivalent to 1-bit or symbol duration) version of itself as shown in the figure. The output of this correlator is passed through a low pass filter (LPF) or an integrator to remove the unwanted second harmonics terms and then compared at the threshold detector. It can be shown that the resulting signal is proportional to

$$\cos((b_n\pi/2) - (b_{n-1}\pi/2)). \quad (2.6)$$

Since $((b_n\pi/2) - (b_{n-1}\pi/2))$ equals to 0 or π we have (2.6) equals to 1 or -1. In summary, demodulation of the DPSK signal is accomplished by comparing the phase of the received signal between two successive intervals.

Fig. 2.2. Block diagram of a DPSK receiver

2.2 CHANNEL MODEL - MULTIPATH MODELS

In general, in the transmission of digital signals the modulator (of a hypothetical transmitter) performs the function of impressing the digital information-bearing signal on a carrier either by modulating the frequency, amplitude or the phase of the carrier. As a result, the output of the modulator is a bandpass signal which can be represented in one of the following complex forms as

$$\begin{aligned} s(t) &= \text{Re}[u(t)e^{j2\pi f_c t}] \\ &= \text{Re}[c(t)\exp(j(2\pi f_c t + \phi(t)))] \\ &= c(t) \cos[2\pi f_c t + \phi(t)], \end{aligned} \quad (2.7)$$

where $u(t) = c(t)e^{j\phi(t)}$ is the complex envelope, lowpass or baseband equivalent, of $s(t)$, $c(t)$ represents its amplitude, $\phi(t)$ its phase and f_c is the carrier. The baseband signal $u(t)$ depends on the method by which the binary information sequence is mapped into the set of discrete frequencies, set of discrete amplitudes or the set of discrete phases.

As indicated in Chapter 1, the two fundamental traits of a multipath channel in an indoor environment are the time spread introduced in the signal when it is transmitted through the channel and the time variations in the structure of the channel. To illustrate these channel characteristics let us consider, what we would receive if we transmit a very narrow pulse (ideally an impulse) $p(t - t_o)$ over a time-varying multipath channel. The received signals might appear as trains of varying amplitude pulses, one pulse for each distinct path, as shown in Figure 2.3. Our first observation is that the received signal is broadened out to a time spread, T_m . Secondly, as a result of the time variations of the structure of the medium, the nature of the multipath itself varies with time. From this, we would find changes in the received pulse train depending on the time the pulse is sent (illustrated by the changes in the size of the individual pulses, changes in the relative delays among the pulses and quite often changes in the number of pulses in the received pulse train) as shown in the same figure. In most situations the time variations appear unpredictable to the user of the channel. Therefore one can only characterize the time-varying multipath channel statistically.

From the above hypothesis we can safely assume that there are L (which may vary over time) propagation paths with different delays, $\tau_l(t)$, and attenuations, $\beta_l(t)$, for the l th path. Thus if the bandpass signal, represented by (2.7), is transmitted through the channel the received bandpass signal as a result of the channel modifications can be expressed in the form

$$r(t) = \sum_{l=1}^L \beta_l(t) s[t - \tau_l(t)]. \quad (2.8)$$

Transmitted signal

Received signal

Fig. 2.3 Illustration of the response of a discrete time varying multipath fading channel to a narrow pulse signal.

If we substitute $s(t)$ from (2.7) into (2.8) we get

$$r(t) = \text{Re} \left(\left\{ \sum_{l=1}^L \beta_l(t) e^{-j\gamma_l(t)} u[t - \tau_l(t)] \right\} e^{j2\pi f_c t} \right), \quad (2.9)$$

where $\gamma_l(t) = 2\pi f_c \tau_l(t)$ corresponds to a phase-shift due to the delay $\tau_l(t)$. Thus from the above equation the equivalent low-pass received signal is

$$w(t) = \sum_{l=1}^L \beta_l(t) e^{-j\gamma_l(t)} u[t - \tau_l(t)]. \quad (2.10)$$

Since $w(t)$ is the response of the equivalent low-pass channel to the equivalent low-pass signal $u(t)$, its clear that the equivalent low-pass channel maybe described by the *time-varying* impulse response

$$h(\tau; t) = \sum_{l=1}^L \beta_l(t) e^{-j\gamma_l(t)} \delta[t - \tau_l(t)], \quad (2.11)$$

i.e., as a function of τ -variable to denote the propagation delay characteristic and the time dependence to indicate that the very structure of the impulse response changes with time. The above expression represents a channel that contains discrete multipath components. This discrete approach makes our system modelling mathematically tractable and suitable for analysis later. In short, we can assume that the discrete multipath fading channel can be characterized by the set of three independent random path variables

$$\{\beta_l(t), \tau_l(t), \gamma_l(t)\}_1^L. \quad (2.12)$$

To provide further insight into the nature and characteristics of IWC multipath channel, let us consider the transmission of an unmodulated carrier at frequency f_c , i.e., $u(t) = 1$ for all t . The equivalent lowpass received signal given in (2.10) then reduces to

$$w(t) = \sum_{l=1}^L \beta_l(t) e^{-j\gamma_l(t)}, \quad (2.13)$$

where $\gamma_l(t) = 2\pi f_c \tau_l(t)$. The received baseband signal (2.13) can also be represented by

$$w(t) = x(t) - jy(t), \quad (2.14)$$

where $x(t)$ and $y(t)$ are the inphase and quadrature components of the received baseband signal with

$$x(t) = \sum_{l=1}^L \beta_l(t) \cos \gamma_l(t), \quad y(t) = \sum_{l=1}^L \beta_l(t) \sin \gamma_l(t). \quad (2.15)$$

With reference to the equation (2.13) above, we make the following observation:

- (a) The received signal $w(t)$ consists of a sum of time varying vectors with amplitudes $\beta_l(t)$ and phases $\gamma_l(t)$, where $\beta_l(t)$ and $\gamma_l(t)$ are independent random variables.
- (b) Large dynamic changes in the channel conditions are needed in order for amplitudes $\beta_l(t)$ to impact on the overall received signal strength.
- (c) The phases $\gamma_l(t)$ are critically sensitive to changes in path length (will change by 2π radians whenever the delay parameter $\tau_l(t)$ changes by a wavelength or $1/f_c$) which implies that $\gamma_l(t)$ can change by 2π radians with relatively small motions in the channel.
- (d) The delays $\tau_l(t)$ associated with different signal paths change at different rates and in an unpredictable or random manner.

In phasor terms, the received composite baseband signal is a vector sum of several phasors, represented by $\{\beta_l(t)e^{-j\gamma_l(t)}\}$, with the phase of each varying individually and randomly over a full $(0, 2\pi)$ range. If there is no dominant component among the various phasors, the statistical averages (denoted by $\overline{x(t)}$ and $\overline{y(t)}$) of the quadrature components of the received baseband signal, represented by (2.15) are found to be zero, i.e.,

$$\begin{aligned} \overline{x(t)} &= \sum_{l=1}^L E[\beta_l(t) \cos \gamma_l(t)] = \sum_{l=1}^L E[\beta_l(t)] \underbrace{E[\cos \gamma_l(t)]}_{=0} = 0, \\ \overline{y(t)} &= \sum_{l=1}^L E[\beta_l(t) \sin \gamma_l(t)] = \sum_{l=1}^L E[\beta_l(t)] \underbrace{E[\sin \gamma_l(t)]}_{=0} = 0, \end{aligned} \quad (2.16)$$

and their mean square values

$$\begin{aligned} \overline{x^2(t)} &= \sum_{l=1}^L E[\beta_l^2(t) \cos^2 \gamma_l(t)] = \sum_{l=1}^L E[\beta_l^2(t)] \underbrace{E[\cos^2 \gamma_l(t)]}_{=1/2} = \frac{1}{2} \sum_{l=1}^L E[\beta_l^2(t)], \\ \overline{y^2(t)} &= \sum_{l=1}^L E[\beta_l^2(t) \sin^2 \gamma_l(t)] = \sum_{l=1}^L E[\beta_l^2(t)] \underbrace{E[\sin^2 \gamma_l(t)]}_{=1/2} = \frac{1}{2} \sum_{l=1}^L E[\beta_l^2(t)]. \end{aligned} \quad (2.17)$$

Hence, from the above, it is shown that they have equal variances, denoted by σ^2 , where

$$\sigma^2 = \overline{x(t)^2} - \overline{x(t)}^2 = \overline{y(t)^2} - \overline{y(t)}^2 = \frac{1}{2} \sum_{l=1}^L E[\beta_l^2(t)]. \quad (2.18)$$

As a result of the above, we can deduced that, for large number of paths, the received signal envelope can be modelled as a zero-mean complex-valued Gaussian random process based on the central limit theorem. Thus, the impulse response of the channel, $h(\tau; t)$ given in (2.11), can be described by a zero-mean complex-valued Gaussian random variable in the t -variable. Here, the envelope ($|h(\tau; t)| = \sqrt{x^2(t) + y^2(t)}$), at any instant time t can be found to be *Rayleigh* distributed.⁵ Physically, the modelling of the channel impulse response $h(\tau; t)$ as a zero mean complex-valued Gaussian random process is a result of the randomly moving scatterers or signal reflectors in the channel. However, when there is a *direct line of sight or specularly reflected component*, in addition to these randomly moving scatterers, the impulse response $h(\tau; t)$ cannot be modelled as a zero mean complex-valued Gaussian random process. In such a situation, the envelope $|h(\tau; t)|$ has a *Rician distribution*.

The multipath propagation model shown in the received signal (2.13) represents multipath fading. The fading phenomenon is a direct result of the randomly time variations in the phases $\{\gamma_l(t)\}$. That is, based on vector addition concepts, the sum of the vectors $\{\beta_l(t)e^{-j\gamma_l(t)}\}$ can at times result in the enhancement of the normal signal to a partial or even complete cancellation in the resultant received signal. Thus the amplitude variations, due to the time varying characteristics of the channel, in the received signal gives rise to the *multipath fading* phenomenon.

In an IWC system, the multipath channel is characterized by short term fading or temporal variations which is mainly observed as a result of movement of personnel and equipment in the indoor environment.⁶ The point to be made is that the rate of these variations are very slow compared to the baseband data rates which are in the order of tens of kilobaud. Hence, based on this relative timing, for our analytical purposes the

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⁵ The derivations of these envelope distributions will subsequently be given in the next section.

⁶ In this study we assume that the same fading parameters apply to all the users regardless of their positionings. In fact, some difference is likely to be found although this is not expected to be significant.

parameters in (2.12) can be treated as *time-invariant* random variables, i.e., identical random variables with respect to t . In this instance we may drop the dependence of these variables on ' t ' and just consider the random variables

$$\{\beta_l, \tau_l, \gamma_l\}_1^L. \quad (2.19)$$

We note that the statistical behaviour of the path attenuation random variables, $\{\beta_l\}_1^L$, together with the phase random variables, $\{\gamma_l\}_1^L$, provides us with the signal strength of the received signal. However, to derive the envelope fading distributions models, let us assume that the multipath channel is characterized by short term fading, where the fading is slow compared to the data rate. Thus we can assume that the transmitted baseband signal, within this fading window, has a constant amplitude c_0 and phase ϕ_0 . The bandpass received signal in (2.8) is thus given by

$$r(t) = \text{Re}\left(\left\{\tilde{r}(t) e^{j\phi_0}\right\} e^{j2\pi f_c t}\right), \quad (2.20)$$

where $\tilde{r}(t)$ is the complex envelope of the received signal $r(t)$. With reference to (2.9) the complex envelope can be expressed as

$$\begin{aligned} \tilde{r}(t) &= \sum_{l=1}^L c_0 \beta_l e^{-j\gamma_l} \\ &= c_0 \sum_{l=1}^L [\beta_l \cos \gamma_l - j \beta_l \sin \gamma_l] \\ &= c_0 (x - jy) \\ &= r e^{-j\theta}, \end{aligned} \quad (2.21)$$

where r is the amplitude and θ is the phase of the complex envelope $\tilde{r}(t)$ and x and y are the quadrature components.

2.2.1 Rayleigh Distribution Model

As discussed in the previous section, if there is no dominant component among the various multipath components we can reasonably assumed that the composite received signal can be modelled as a zero-mean complex-valued Gaussian random process. In the case of even a few number of multipath components of roughly the same strength, the received

signal is found to have the characteristics of a very narrow band of stationary Gaussian noise, i.e., its complex envelope consists of zero-mean Gaussian quadrature components x and y as clearly proved in the previous section. The joint probability density function (pdf) of the random variables of this stationary complex zero-mean Gaussian process is described by

$$f(x, y) = \frac{1}{2\pi\sigma^2} \exp\left(-\frac{x^2 + y^2}{2\sigma^2}\right), \quad (2.22)$$

where the quadrature components x and y are Gaussian random variates with zero means and equal variances σ^2 . The complex envelope parameters, at any instant time t , are

$$r = c_0 \sqrt{x^2 + y^2}, \quad 0 \leq r < \infty;$$

and

$$\theta = \arg(x + jy), \quad \theta \in [-\pi, \pi]; \quad (2.23)$$

where r , represented by the independent random variable R , denotes the magnitude and θ , represented by the independent random variable Θ , the phase of the complex envelope. It can be shown [27], via a Jacobian transformation of the rectangular variates x and y , the joint pdf of r and θ is given by

$$f_{R,\Theta}(r, \theta) = \frac{r}{2\pi\sigma^2} \exp\left(-\frac{r^2}{2\sigma^2}\right), \quad r \geq 0, -\pi \leq \theta < \pi. \quad (2.24)$$

By integrating the equation (2.24), over the magnitude r , the density function of the phase represented by the random variable Θ can be shown to be given by the *uniform distribution*

$$f_{\Theta}(\theta) = 1/2\pi, \quad 0 \leq \theta < 2\pi. \quad (2.25)$$

Similarly, we can derive the density function of R , by integrating the joint pdf expressed in (2.24) over the phase θ , which shows that R has a *Rayleigh distribution*

$$f_R(r) = \begin{cases} \frac{r}{\sigma^2} \exp\left(-\frac{r^2}{2\sigma^2}\right) & r \geq 0. \\ 0 & r < 0. \end{cases} \quad (2.26)$$

2.2.2 Rician Distribution Model

The *Rician distribution* is more appropriate for describing multipath fading channels where there is a direct line of sight or specularly reflected component (which contributes

to the received signal) in the medium. In such a situation, the path parameters in the received bandpass signal in (2.20) then consist of two components: a fixed or specular component and a completely random or scatter component, i.e.,

$$\begin{aligned} r(t) &= \text{Re} \left(\left\{ c_0 r e^{-j\theta} e^{j\phi_0} \right\} e^{j2\pi f_c t} \right) \\ &= \text{Re} \left(\left\{ c_0 (s e^{j\varphi} + S e^{j\Psi}) e^{j\phi_0} \right\} e^{j2\pi f_c t} \right) \end{aligned} \quad (2.27)$$

where s and φ are respectively the magnitude and phase of the dominant or specular component, S is an independent random variable with Rayleigh density and average power $2\sigma^2$, and Ψ is an independent random variable uniformly distributed over $(0, 2\pi)$. We note that the discrete multipath fading channel characterized by the set of independent random path variables given in (2.27) can be treated as independent vector quantities. The *joint probability distribution* of the path length and angle of the sum of the fixed vector (s, φ) and the random vector (S, Ψ) is ([28,29])

$$\begin{aligned} f_{R,\Theta}(r, \theta) &= \frac{r}{2\pi\sigma^2} \exp \left(-\frac{r^2 + s^2 - 2rs \cos(\theta - \varphi)}{2\sigma^2} \right), \quad 0 \leq r < \infty, -\pi \leq (\theta - \varphi) \leq \pi \\ &= 0 \quad \text{elsewhere} \end{aligned} \quad (2.28)$$

Integrating over Θ , it can be shown [16] that the density function of the envelope R is a Rician probability density function:

$$f_R(r) = \begin{cases} \frac{r}{\sigma^2} \exp \left(-\frac{r^2 + s^2}{2\sigma^2} \right) I_0 \left(\frac{rs}{\sigma^2} \right), & r \geq 0 \\ 0 & r < 0 \end{cases} \quad (2.29)$$

where I_0 is the modified Bessel function of the first kind and order zero. In our study, to characterise these distributions, we shall make use of

$$\kappa^2 = 2\sigma^2/s^2, \quad (2.30)$$

the *random to specular power ratio*.

In general, there are two different ways we can develop statistical models for multipath fading. The usual norm is to ascribe a physical model of the radio propagation environment and develop statistical equations based upon the assumptions of the physical model. The results obtained from this type of model is then compare, with results based upon the analysis of field data measured subsequently, to validate its usage. If this approach

is considered, care must be exercised to ensure that the underlying physical assumptions upon which the model hinges on are not seriously violated. The second approach involves employing empirical type of model (often derived for other applications) with known probability distributions parameters. Similarly, care must be taken especially when it is used to predict signal behaviour outside the range of operation for which the data were originally measured. Our study is based on the latter approach with the Rician path model parameters values given in [7].

Thus, the general form of the multipath channel model should include a specular component in addition to a purely random fading component. This model can be represented by the deterministic and random filters shown in Figure 2.4. which illustrates the channel model for a m th diversity channel) which we will form part of our multichannel system discussion.

2.3 THE MULTICHANNEL MODEL

Fig. 2.4. General channel model for m th diversity channel

In most radio systems applications it is prudent to transmit over two or more channels when there is a high probability that one or more of the channels will become unreliable from time to time due to adverse channel conditions. In our case, the indoor channels experience

multipath fading which must be mitigated. We can establish multiple channels in effect in these circumstances by employing some well-known “diversity techniques” that were briefly mentioned in Chapter I.

Fig. 2.5. Block diagram of multichannel system under consideration

The effects of the use of general diversity techniques can be described by a *multichannel model* such as is shown in Figure 2.5 where each diversity channel (Figure 2.4) can be represented by a parallel combination of a random (to represent the random or scatter component which is modelled by a complex-valued Gaussian random variable) and a deterministic (to represent the specular component) *time-varying linear*⁷ filter. Figure 2.5

⁷ The concept of representing a radio channel as a linear filter is based on the premise that virtually all radio channels are characterized as linear in their effect upon transmitted signals.

shows the input complex envelope $s(t)$ fed to a parallel combination of a deterministic time-variant filter and a random variant filter. The complex envelopes of these filters' outputs are combined with the additive white complex Gaussian noise term to form $r_m(t)$, the complex envelope of the channel output. We note that $r_m(t)$, is given by the sum of a deterministic component and a complex-valued Gaussian process. In the case of a multi-channel model with *time diversity* based on a resolvable multipath situation, $r(t) = \sum_{l=1}^L r_l(t)$ where $r_l(t)$ represents the signal received over the " l th link" of the L resolvable paths.

The physical realizations of such a multichannel communication model are typified by various examples. In *space diversity* mode, $r_m(t)$ will represent the complex envelope of the signal received via the " m th link" of the M antennae. Similarly, the model may be made to correspond to a *frequency diversity* situation where $r_m(t)$ represents the signal received over the " m th link" of the M nonoverlapping carrier frequency bands. The model may also illustrates the *time diversity* of a resolvable multipath situation where $r_m(t)$ represents the signal received over the " m th link" of the M resolvable paths.

To represent different receiver techniques (that will encompass incoherent, differentially coherent and coherent detectors) we will assume switched diversity combining with a general "quadratic form receiver". Figure 2.6 illustrates one of the M independent diversity branches. For each diversity branch, the receiver performs linear filtering (the processes $X_m(t)$ and $Y_m(t)$ could be matched filter outputs, or $X_m(t)$ could be a filter output and $Y_m(t)$ the reference waveform) on the received signal $r_m(t)$ to establish a reference, $Y_m(t)$, and an information bearing signal, $X_m(t)$. Since the linear operation leaves the Gaussian nature of the process unchanged, it follows that X_m and Y_m each may be described by a complex-valued Gaussian process (with possibly non-zero means). The two signals are then periodically sampled once per signalling interval and weighted in a quadratic combiner. If T is the signalling interval and $X_m \equiv X(mT)$, $Y_m \equiv Y(mT)$, then the decision variable on each branch is given by

$$d_m = A|X_m|^2 + B|Y_m|^2 + C X_m^* Y_m + C^* X_m Y_m^*. \quad (2.31)$$

The output from each diversity branch are added to give

$$D = \sum_{m=1}^M d_m \quad (2.32)$$

the decision variable which is then compared to a zero threshold for the binary decision. As an example, the “quadratic form receiver” will correspond to a DPSK receiver if the information filter is matched to the transmitted signal while the reference filter is a cascade of the information filter and a time delay of one signalling period (i.e. $z_1 = X_m, z_2 = Y_m$). The coefficients of the quadratic combiner in this case are $A = B = 0$ with $C = 1$. Similarly, most binary receivers designed for either coherent or incoherent detection of FSK, DPSK and PSK can be shown to be mathematically equivalent to the receiver described by (2.31) and (2.32) by properly specifying the information and reference filters of the receiver and the coefficients of the quadratic combiner [16].

Fig. 2.6. General receiver model for m th diversity channel

2.3.1 Probability of Error for Multichannel Reception of Binary Signals

Having provided a brief overview of the general concepts of the class of binary receivers let us evaluate the probability of error, P_e of a receiver operating according to the quadratic decision rule $D < 0$ where D is defined in (2.32). Let us begin by denoting $\bar{X}_m, \bar{Y}_m$ as the statistical averages of the complex variables X_m and Y_m , and

$$\mu_{xx} = E[|X_m - \bar{X}_m|^2],$$

$$\mu_{xy} = E[(X_m - \bar{X}_m)^*(Y_m - \bar{Y}_m)],$$

and

$$\mu_{yy} = E[|Y_m - \bar{Y}_m|^2] \quad (2.33)$$

as the various second (central) moments of the complex-valued Gaussian variables X_m and Y_m . We note that the probability of binary error can be given by

$$\begin{aligned} P_e &= \Pr(D < 0 | \text{"1" sent}) \\ &= \Pr(D > 0 | \text{"0" sent}) \end{aligned} \quad (2.34)$$

Thus, we see from the above that to evaluate the desired binary error probabilities, it is sufficient to compute the conditional probability that the probability of error of the quadratic form receiver is the probability that $D < 0$, i.e.,

$$\begin{aligned} P_e &= \Pr(D < 0 | \text{"1" sent}) \\ &= \int_{-\infty}^0 p(D) dD \end{aligned} \quad (2.35)$$

where $p(D)$ ⁸ is the conditional probability density function of D . This conditional density function can be given in terms of the characteristic function, denoted by $\Phi_D(j\omega)$, of the quadratic decision rule $D < 0$ through:

$$p(D) = \frac{1}{2\pi} \int_{-\infty}^{\infty} \Phi_D(jv) e^{-jvD} dv. \quad (2.36)$$

As a result,

$$P_e = \int_{-\infty}^0 \frac{1}{2\pi} \int_{-\infty}^{\infty} \Phi_D(jv) e^{-jvD} dv dD \quad (2.37)$$

To facilitate the analysis we will substitute $f = jv$ and noting that the characteristic function of D is then

$$\Phi_D(f) = \int_{-\infty}^{\infty} p(D) e^{fD} dD. \quad (2.38)$$

Since D is the sum of M statistically independent random variables, d_m , the characteristic function of D factors into a product of the M characteristic functions of the individual random variables d_m :

$$\Phi_D(f) = \prod_{m=1}^M \Phi_{d_m}(f)$$

where

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⁸ For simplicity of notation, we use $p(D)$ to denote $p(D | \text{"1" sent})$. Subsequently the ensuing terms' notations are referred to the quadratic decision rule $D < 0$.

$$\Phi_{d_m}(f) = e^{\overline{f d_m}} \quad (2.39)$$

is the characteristic function of d_m . From (2.31) we see that we can express d_m as Hermitian form

$$d_m = W^{T*} Q W \quad (2.40)$$

where

$$W = [X_m \ Y_m]^T,$$

and

$$Q = \begin{bmatrix} A & C \\ C^* & B \end{bmatrix}$$

is a Hermitian matrix. The characteristic function of the above Hermitian form d_m can be shown [24] to be

$$\Phi_{d_m}(f) = \det [I - f L Q]^{-1} \exp [-\bar{W}^{T*} L^{-1} [I - (I - f L Q)] \bar{W}] \quad (2.41)$$

where I is the identity matrix, $\bar{W} = [\bar{X}_m \ \bar{Y}_m]^T$, L is the complex covariance matrix

$$L = \begin{bmatrix} \mu_{xx} & \mu_{xy} \\ \mu_{xy}^* & \mu_{yy} \end{bmatrix}$$

and

$$\Phi_{d_m}(f) = \frac{\exp \left[\frac{f \lambda_1 |p_{1m}|^2}{(1-f\lambda_1)} \right]}{(1-f\lambda_1)} \frac{\exp \left[\frac{f \lambda_2 |p_{2m}|^2}{(1-f\lambda_2)} \right]}{(1-f\lambda_2)} \quad (2.42)$$

where λ_1, λ_2 are eigenvalues of the product matrices LQ and p_{1m} and p_{2m} are two parameters related to X_m, Y_m . If we let

$$R = LQ = \begin{bmatrix} r_x & r_{xy} \\ r_{yx} & r_y \end{bmatrix} = \begin{bmatrix} \mu_{xx} & \mu_{xy} \\ \mu_{xy}^* & \mu_{yy} \end{bmatrix} \begin{bmatrix} A & C \\ C^* & B \end{bmatrix}, \quad (2.43)$$

then

$$\begin{aligned} r_x &= A\mu_{xx} + C^*\mu_{xy}, & r_{xy} &= C\mu_{xx} + B\mu_{xy}, \\ r_{yx} &= A\mu_{xy}^* + C^*\mu_{yy}, & r_y &= C\mu_{xy}^* + B\mu_{yy} \end{aligned} \quad (2.44)$$

and

$$\Delta_R = \det R = [AB - |C|^2] [\mu_{xx}\mu_{yy} - |\mu_{xy}|^2]. \quad (2.45)$$

We can show that for 2×2 matrices [25] the partial term in the exponential expression in (2.41) can be expressed as

$$\mathbf{L}^{-1} [\mathbf{I} - (\mathbf{I} - f\mathbf{L}\mathbf{Q})^{-1}] = \frac{f^2 \Delta_R \mathbf{L}^{-1} - f\mathbf{Q}}{\Delta} \quad (2.46)$$

where

$$\Delta = \det[\mathbf{I} - f\mathbf{L}\mathbf{Q}] = 1 - f(r_x + r_y) + f^2 \Delta_R \quad (2.47)$$

If we substitute the above term in (2.38) then the characteristic function becomes

$$\Phi_{d_m}(f) = \frac{\exp \left[-\frac{f^2 E_m + f F_m}{1 - f(r_x + r_y) + f^2 \Delta_R} \right]}{1 - f(r_x + r_y) + f^2 \Delta_R} \quad (2.48)$$

where

$$\begin{aligned} E_m &= \Delta_R \bar{\mathbf{W}}^* \mathbf{L}^{-1} \bar{\mathbf{W}} \\ &= (AB - |C|^2) (|\bar{X}_m|^2 \mu_{yy} + |\bar{Y}_m|^2 \mu_{xx} - \bar{X}_m^* \bar{Y}_m \mu_{xy} - \bar{X}_m \bar{Y}_m^* \mu_{xy}^*) \\ F_m &= \bar{\mathbf{W}}^* \mathbf{Q} \bar{\mathbf{W}} \\ &= A|\bar{X}_m|^2 + B|\bar{Y}_m|^2 + C\bar{X}_m^* \bar{Y}_m + C^* \bar{X}_m \bar{Y}_m^*. \end{aligned} \quad (2.49)$$

are the quadratic forms in means $(\bar{X}_m, \bar{Y}_m)$. If we substitute $jv = f$, $\alpha_{1m} = -E_m$, and $\alpha_{2m} = F_m$ then the characteristic function of d_m is

$$\Phi_{d_m}(jv) = \frac{\exp \left[-\frac{(jv)^2 (-\alpha_{1m}) + (jv) \alpha_{2m}}{1 - jv(r_x + r_y) + (jv)^2 \Delta_R} \right]}{1 - jv(r_x + r_y) + (jv)^2 \Delta_R} \quad (2.50)$$

If we let

$$1 - jv(r_x + r_y) + (jv)^2 \Delta_R = \Lambda_1 v^2 + \Lambda_2 v + 1$$

then

$$\Lambda_1 = -\Delta_R = [|C|^2 - AB] [\mu_{xx} \mu_{yy} - |\mu_{xy}|^2]$$

and

$$\Lambda_2 = -j(r_x + r_y) = -j(A\mu_{xx} + B\mu_{yy} + C^* \mu_{xy} + C\mu_{xy}^*) \quad (2.51)$$

If we factorize

$$\Lambda_1 v^2 + \Lambda_2 v + 1 = \Lambda_1 (v - r_1)(v - r_2)$$

where r_1, r_2 are the roots of

$$\Lambda_1 v^2 + \Lambda_2 v + 1$$

The characteristic function

$$\Phi_{d_m}(jv) = \frac{\exp\left[-\frac{v^2 \alpha_{1m} + jv \alpha_{2m}}{\Lambda_1(v-r_1)(v-r_2)}\right]}{\Lambda_1(v-r_1)(v-r_2)} \quad (2.52)$$

The roots are determined to be

$$r_1 = \left(-\frac{\Lambda_2}{2\Lambda_1}\right) - \sqrt{\left(-\frac{\Lambda_2}{2\Lambda_1}\right)^2 + \left(-\frac{1}{\Lambda_1}\right)}$$

and

$$r_2 = \left(-\frac{\Lambda_2}{2\Lambda_1}\right) + \sqrt{\left(-\frac{\Lambda_2}{2\Lambda_1}\right)^2 + \left(-\frac{1}{\Lambda_1}\right)}. \quad (2.53)$$

To show that the equations derived are consistent with that in [16, Appendix 4B] we will substitute

$$\Lambda_1(v-r_1)(v-r_2) = \Lambda_1(v+jv_1)(v-jv_2)$$

where $v_1 = jr_1$ and $v_2 = -jr_2$

The resulting characteristic function of d_m is

$$\Phi_{d_m}(jv) = \frac{v_1 v_2}{(v+jv_1)(v-jv_2)} \exp\left[\frac{v_1 v_2(-v^2 \alpha_{1m} + jv \alpha_{2m})}{(v+jv_1)(v-jv_2)}\right] \quad (2.54)$$

where the parameters $v_1, v_2, \alpha_{1m}, \alpha_{2m}$ are functions of the means $\bar{X}_m$ and $\bar{Y}_m$ and the second moments $\mu_{xx}, \mu_{xy}, \mu_{yy}$ of the complex Gaussian random variables X_m and Y_m . If we define, $w = \left(\frac{j\bar{b}}{2\Lambda_1}\right)$, then, with substitution we can show that

$$v_1 = -w + \sqrt{w^2 + \left(\frac{1}{\Lambda_1}\right)}$$

and

$$v_2 = w + \sqrt{w^2 + \left(\frac{1}{\Lambda_1}\right)}$$

As a result,⁹

$$v_1 = -w + \sqrt{w^2 + \left(\frac{1}{[|C|^2 - AB][\mu_{xx}\mu_{yy} - |\mu_{xy}|^2]}\right)}$$

and

---

⁹ The resulting equations defined above are similar to those of [16, 4.b.6] with the correction of a couple of typographical errors in the text.

$$v_2 = w + \sqrt{w^2 + \left(\frac{1}{[|C|^2 - AB] [\mu_{xx}\mu_{yy} - |\mu_{xy}|^2]} \right)}$$

where

$$\begin{aligned} w &= \frac{A\mu_{xx} + B\mu_{yy} + C^*\mu_{xy} + C\mu_{xy}^*}{2[|C|^2 - AB] [\mu_{xx}\mu_{yy} - |\mu_{xy}|^2]} \\ \alpha_{1m} &= -(AB - |C|^2)(|\bar{X}_m|^2\mu_{yy} + |\bar{Y}_m|^2\mu_{xx} - \bar{X}_m^*\bar{Y}_m\mu_{xy} - \bar{X}_m\bar{Y}_m^*\mu_{xy}^*) \\ \alpha_{2m} &= A|\bar{X}_m|^2 + B|\bar{Y}_m|^2 + C\bar{X}_m^*\bar{Y}_m + C^*\bar{X}_m\bar{Y}_m^*. \end{aligned} \quad (2.55)$$

We note that $\Lambda_1 = (1/v_1v_2)$ in the above. The characteristic function of D represents the sum of M independent random variables d_m , i.e.,

$$\Phi_D(jv) = \prod_{m=1}^M \Phi_{d_m}(jv) \quad (2.56)$$

can be expressed as

$$\Phi_D(jv) = \left[\frac{v_1v_2}{(v+jv_1)(v-jv_2)} \right]^M \exp \left[\frac{v_1v_2(-v^2\alpha_1 + jv\alpha_2)}{(v+jv_1)(v-jv_2)} \right] \quad (2.57)$$

where

$$\alpha_1 = \prod_{m=1}^M \alpha_{1m} \quad (2.58)$$

$$\alpha_2 = \prod_{m=1}^M \alpha_{2m} \quad (2.59)$$

The probability of error in (2.37) is evaluated by substituting the expression for $\Phi_D(jv)$ given above. To evaluate the double integral the order of integrations is interchanged to give

$$P_e = -\frac{(v_1v_2)^M}{2\pi j} \int_{-\infty+j\epsilon}^{\infty+j\epsilon} \frac{dv}{v(v+jv_1)^M(v-jv_2)^M} \exp \left[\frac{v_1v_2(-v^2\alpha_{1m} + jv\alpha_{2m})}{(v+jv_1)(v-jv_2)} \right] \quad (2.60)$$

where to facilitate the integration due to the singularity at $v = 0$ and allow for the smooth interchange in the order of integrations a small positive number ϵ was included. It can be shown¹⁰ that the resulting probability equation is

$$P_e = Q(a, b) - I_0(ab) \exp \left(-\frac{a^2 + b^2}{2} \right)$$

---

¹⁰ Evaluation of the integral can be accomplished in two major steps, for further reference see [16, 4.B.10 to 4.B.20].

$$\begin{aligned}
& + \frac{I_0(ab) \exp [-(a^2 + b^2)/2]}{(1 + v_2/v_1)^{2M-1}} \sum_{m=0}^{M-1} \binom{2M-1}{m} \left(\frac{v_2}{v_1}\right)^m \\
& + \frac{\exp [-(a^2 + b^2)/2]}{(1 + v_2/v_1)^{2M-1}} \sum_{n=1}^{M-1} I_n(ab) \left\{ \sum_{m=0}^{M-1-n} \binom{2M-1}{m} \right. \\
& \times \left[\left(\frac{b}{a}\right)^n \left(\frac{v_2}{v_1}\right)^m - \left(\frac{a}{b}\right)^n \left(\frac{v_2}{v_1}\right)^{2M-1-m} \right] \Big\} \quad M > 1 \quad (2.61)
\end{aligned}$$

and

$$P_e = Q(a, b) - \frac{v_2/v_1}{1 + v_2/v_1} I_0(ab) \exp \left(-\frac{a^2 + b^2}{2} \right) \quad M = 1 \quad (2.62)$$

where

$$a = \left[\frac{2v_1^2 v_2 (\alpha_1 v_2 - \alpha_2)}{(v_1 + v_2)^2} \right]^{1/2}, \quad (2.63)$$

$$b = \left[\frac{2v_1 v_2^2 (\alpha_1 v_2 + \alpha_2)}{(v_1 + v_2)^2} \right]^{1/2}, \quad (2.64)$$

$$Q(a, b) = \exp \left(-\frac{a^2 + b^2}{2} \right) \sum_{n=0}^{\infty} \left(\frac{a}{b} \right)^n I_n(ab) \quad b > a > 0$$

denotes the Marcum Q function and $I_n(\cdot)$ denotes the n th order modified Bessel function of the first kind.

2.4 PROBABILITY OF ERROR FOR SS-MA WITH DPSK MODULATION

Our main interest in this major exercise is to obtain the desired expression for the probability of error in the case of our CDMA indoor wireless system with direct sequence spread spectrum and DPSK signalling technique. The application of this expression will subsequently be shown in Chapter 4.

The decision variable can be expressed as a special case of the general quadratic form as follows:

$$D = d_1 = \frac{1}{2} (X_1 Y_1^* + X_1^* Y_1). \quad (2.65)$$

To explain the above expression, we note that we set M is equal to 1 for we are interested in the one particular channel as we consider all the other channels as interferers and without considering space diversity. Here, the pair of complex Gaussian random variables, X_1 and Y_1 denotes the complex envelope of the matched filter output at the sampling instant

and the complex envelope of the matched filter output at the previous sampling instant respectively, with the constants $A = B = 0$, and $C = C^* = 1/2$. As a result we make the following observations (in conjunction with the probability of error expression of the general quadratic form that forms the decision variable);

In the case of DPSK, with the decision variable $D < 0$, we can ascertain, with reference to (4.34), that $\bar{X}_1 = \bar{Y}_1 = \mu$. Together with the second (central) moments, defined in (2.33), of the complex-valued Gaussian variables and noting that μ_{xy} is real we can show that upon substitution,

$$\begin{aligned}\alpha_1 &= \frac{\mu^2}{4} (\mu_{xx} + \mu_{yy} - 2\mu_{xy}), \\ \alpha_2 &= \mu^2, \\ w &= \frac{2\mu_{xy}}{\mu_{xx}\mu_{yy} - \mu_{xy}^2}, \\ v_1 &= \frac{2}{\sqrt{\mu_{xx}\mu_{yy} + \mu_{xy}^2}}, \\ v_2 &= \frac{2}{\sqrt{\mu_{xx}\mu_{yy} - \mu_{xy}^2}}.\end{aligned}\tag{2.66}$$

Hence the desired expression for the probability of error is:¹¹

$$P_e = Q(a, b) - \left(\frac{1+r}{2}\right) I_0(ab) \exp\left(-\frac{a^2 + b^2}{2}\right)\tag{2.67}$$

where

$$\begin{aligned}r &= \frac{\mu_{xy}}{\sqrt{\mu_{xx}\mu_{yy}}}, \\ a &= \frac{\mu}{\sqrt{2}} \left[\frac{1}{\sqrt{\mu_{xx}}} - \frac{1}{\sqrt{\mu_{yy}}} \right], \\ b &= \frac{\mu}{\sqrt{2}} \left[\frac{1}{\sqrt{\mu_{xx}}} + \frac{1}{\sqrt{\mu_{yy}}} \right],\end{aligned}\tag{2.68}$$

$I_0(\cdot)$ denotes the zeroth order modified Bessel function of the first kind and $Q(a, b)$ as previously described.

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¹¹ The above expression for the probability of error is consistent with that described in [16, Appendix 7A.4] in the case of Rician Fading Channels!

III. DS/SS-MA SCHEME

3.1 OVERVIEW AND INTRODUCTION TO INDOOR DS/SS-MA SYSTEM

In the indoor wireless communication system under consideration, we assume that there are K users simultaneously trying to communicate in a star network configuration (as shown in Figure 1.1), with a base station using DS/SS. This system corresponds to an indoor cell site (ICS) in the proposed cellular access digital network (CADN) architecture of Figures 1.2–5.

At each cell site, the member users communicate on the same carrier frequency (f_0) employing binary DPSK. We assumed that the users transmit at a common baseband data rate, R bits/s, within an allocated channel bandwidth of W Hz. In order to utilize the full channel bandwidth, the phase of the carrier is shifted pseudo-randomly according to a code pattern from the psuedo-noise (PN) sequence generator at a rate of W times/s. This technique, characterized by the use of the high rate codes, has the effect of *spreading* the bandwidth of the information data signal. This direct sequence spread spectrum multiple access (a.k.a. phase-coded SS-MA) technique, shown in Figure 3.1, (i.e., the carrier is phase modulated by the information data bit stream and the code sequence) is the most common form of spread spectrum modulation. In each cell, each user is assigned a distinct PN code sequence (consisting of square pulses referred to as chips) of period N from a set of sequences with low crosscorrelation properties. This allows the receiver at the central station to separate the different received signals with usually acceptably small interference between users signals.¹² The scheme for separating the different signals using the same frequency band is referred to as code-division multiple access (CDMA) scheme.

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¹² The codes may be reused in adjacent cell sites provided the carrier frequencies for each sites are slightly offset by a value equal to the baseband data rate and the receivers are equipped with match filters that have sharp frequency band-selectivity characteristics with carrier frequency offsets.

Fig. 3.1. Typical D/S modulation-demodulation scheme

Fig.a

Fig.b

Fig. 3.2. a. Basic chip function and b. D/S signal spectrum.

3.1.1 Processing gain (G_p) and limitations of code rates (R_c)

The basic square pulse or chip function is shown in Figure 3.2.a, where T_c is the chip interval or time duration of the pulse. The reciprocal of T_c , denoted as R_c , is commonly referred as the chip rate in chips per second. If we denote the time duration of a data bit as T (where $T = 1/R$), then clearly $N = T/T_c = R_c/R$ is the number of chips per information bit. The *bandwidth expansion factor*, B_e , (sometimes known as the *processing gain*) is expressed as $B_e = W/R \geq R_c/R$. The processing gain (G_p) however is a function of the RF bandwidth of the signal transmitted compared with the bit rate of the information. In practice this RF bandwidth is taken as the 3 dB-down bandwidth of the main lobe of the $[(\sin x)/x]^2$ which is about 0.88 times the bandwidth-spreading code clock rate in our case [20]. The waveform $[(\sin x)/x]^2$ is the frequency spectrum of the spread spectrum modulated signal as shown in Figure 3.2.b

Currently, integrated circuit chips are available that allow code rates up to 200 or even to 500 MHz. However, operating high-speed digital circuits entails some major drawbacks such as: higher noise sensitivity (greater susceptibility to errors) and higher consumption of power (and therefore high levels of heat generated). As a result, a practical code rate of 50–100 MHz are the highest recommended for general use now and in the near future. Hence, for our IWC study we will set an optimal or practical code rate of 50 MHz. If an RF bandwidth (null-to-null main lobe bandwidth) of 80 MHz is allocated at 900 MHz, the code rate will be at least 40 MHz. To provide spread spectrum modulation for an information baseband data rate of 144 kb/s we require a code length of not more than 255. Therefore for a system with a 40 MHz code clock rate and a 144 kb/s data rate, the maximum processing gain available is:

$$G_p = (0.88 \times 36.72 \times 10^6) / (144 \times 10^3) = 23.5 \text{ dB}.$$

Another term *jamming margin* (M_j) takes into consideration the capability of the system to perform in an environment with interference due to other interfering signals, the requirement for a useful system output SNR and allows for internal losses. The jamming margin of a system is always less than the processing gain and can be calculated as follows:

$$M_j = G_p - [L_{sys} + \text{SNR}].$$

For example if the maximum processing gain of a spread spectrum system is 23.5 dB, with a minimum SNR of 10 dB required at the information output and allowing 2 dB for system losses we would have a jamming margin of 11.5 dB. Thus we would not expect the system to operate with interference more than 11.5 dB above the desired signal. To mitigate against possible interference preventive measures are required as discussed herein.

Finally, we are cognizant of the difference in the effects of errors by high speed code rates and high information data rates. Errors in a data modulated system may be tolerated up to some correctable limit whereas errors in a code generator would completely disable the code modulated system until resynchronization occurs.

3.1.2 Generation, transmission and reception of DS/SS signals

Generally, generation and transmission of DS/SS signals is an easy task and can be accomplished in various ways. Generation of PN codes can be accomplished by surface acoustic wave (SAW) or charge-coupled devices, shift registers (digital LSI), etc. However, to receive such signals successfully, it is necessary that the DS receiver have a reference signal that is a replica of the transmitted signal and is in time synchronization with it. The establishment of such a communication link and detection of the desired DS/SS signal requires two synchronization phases, namely:

- (i) Initial Synchronization or Acquisition Phase and
- (ii) Tracking or Maintaining Synchronization Phase.

The acquisition phase is the initial phase in which the receiver adjusts its time so that its own reference signal is aligned (within one chip period) in step with the incoming signal. Having accomplished acquisition, the next synchronization phase (or tracking phase) involves a finer adjustment and maintaining the local reference signal in step with the desired signal. A detailed description of the various synchronization techniques available is specific to the particular DS/SS scheme and is beyond the scope of this thesis. However, in the next section we will proposed a DS/SS-MA scheme and outlined the receiver's technique in receiving, despreading and recovery of the desired signal.

A simple and efficient method of initial synchronization is to matched filter detect the IF signal directly. Matched filtering here is based on the notion that if the IF signal is passed through a device whose impulse response is the time inverse of the received coded signal then the received signal is despread directly. Surface acoustic wave (SAW)

devices are ideally suited to matched filter synchronization at IF. Such a matched filter basically generates the autocorrelation function of the code sequence which consists of a narrow large triangular pulse in each period. This pulse can be detected and used to synchronize the receiver code sequence to within one chip. The SAW devices are designed to output a pulse when a specific coded sequence is received. When this pulse is sensed the receiver code sequence generator is started using an initial condition corresponding to the received coded sequence and synchronization is complete. A major drawback of this technique is that SAW devices can be built only for waveforms of short durations which limits substantially the time-bandwidth products and hence the processing gain that can be achieved. However, a side benefit of using SAW devices is that we can reuse the SS codes in adjacent ICS with slightly offset carrier frequencies because of the sharp frequency band-selectivity feature of SAW devices with carrier frequency offset.

One of the most desirable features of a DS/SS-MA system is its asynchronous multiple access capability. However, because of the asynchronous operation of the system, the need for considering aperiodic crosscorrelation effects is required. In order to understand the role of aperiodic crosscorrelation properties in the selection of good CDMA codes we will focus on the key correlation parameters next.

3.2 DIRECT SEQUENCE CODES FOR CDMA APPLICATION

3.2.1 Direct Sequence Codes' Parameters

Before proceeding any further, we note that in our analysis, all PN sequences with elements $\{0, 1\}$ (represented by the sequence α) can be mapped into corresponding PN sequences with elements $\{-1, 1\}$ (represented by the sequence a).¹³ In practice, the spreading is achieved, if the information data bits and the code sequence have elements $\{0, 1\}$, by adding each data bit modulo-2 to every chip in a full PN sequence. Conversely, if the data bits and the PN sequence have elements $\{-1, 1\}$, spreading is carried out by multiplying the chips by the data bit. This distinction will help to clarify the various notations used for denoting the PN sequences in this chapter.

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¹³ This will facilitate in the computation of correlation values and in our system modelling.

The correlation parameters described in this section for the sequences (a_k^j) and (a_i^j) (where $j = 1, \dots, N$; $i, k = 1 \dots K$, $i \neq k$) can generally be defined in terms of the *discrete aperiodic crosscorrelation* function (denoted as $C_{k,i}$) [21] as follows: Let $u = (a_k^j)$ and $v = (a_i^j)$ be $\{-1, 1\}$ valued sequences of period N . The aperiodic crosscorrelation function for u and v is then given by

$$C_{u,v}(x) = \begin{cases} \sum_{j=0}^{N-1-x} u^j v^{j+x}, & 0 \leq x \leq N-1; \\ \sum_{j=0}^{N-1+x} u^{j-x} v^j, & 1-N \leq x \leq 0; \\ 0, & |x| \geq N. \end{cases} \quad (3.1)$$

The *periodic crosscorrelation* function (denoted as $\Phi_{k,i}$ or $\Phi_{u,v}(x)$) following the same notation above, is given by

$$\Phi_{u,v}(x) = \sum_{j=0}^{N-1} u^j v^{j+x} \quad (3.2)$$

for any integer x . It can be shown that $\Phi_{u,v}(x)$ can be expressed as the aperiodic cross-correlation function in (3.1) above, i.e.,

$$\Phi_{u,v}(x) = C_{u,v}(x) + C_{u,v}(x-N), \quad 0 \leq x \leq N-1. \quad (3.3)$$

We note that $\Phi_{u,v}(x) = \Phi_{v,u}(N-x)$ above is typical of a periodic or *even crosscorrelation* function. We also define

$$\hat{\Phi}_{u,v}(x) = C_{u,v}(x) - C_{u,v}(x-N), \quad 0 \leq x \leq N-1; \quad (3.4)$$

where $\hat{\Phi}_{u,v}(x)$ is called the *odd crosscorrelation* function as it has the property $\hat{\Phi}_{u,v}(x) = -\hat{\Phi}_{v,u}(N-x)$. The *even and odd autocorrelation* parameters, i.e., $\Phi_{u,u}(x)$ and $\hat{\Phi}_{u,u}(x)$, are denoted by $\Phi_u(x)$ and $\hat{\Phi}_u(x)$ respectively. Having defined the above, let us focus on the key parameters that are involved in the selection of CDMA codes.

(A) The *peak crosscorrelation* parameters are given by

$$\begin{aligned} \Upsilon_{u,v} &\equiv \max\{|\Phi_{u,v}(x)| : 0 \leq x \leq N-1\}, \\ \hat{\Upsilon}_{u,v} &\equiv \max\{|\hat{\Phi}_{u,v}(x)| : 0 \leq x \leq N-1\}, \end{aligned} \quad (3.5)$$

where $\Phi_{u,v}(x)$ and $\hat{\Phi}_{u,v}(x)$ are as described previously.

(B) Similarly, the *peak autocorrelation* parameters are given by

$$\begin{aligned}\Upsilon_u &\equiv \max \{|\Phi_u(x)| : 1 \leq x \leq N-1\}, \\ \hat{\Upsilon}_u &\equiv \max \{|\hat{\Phi}_u(x)| : 1 \leq x \leq N-1\},\end{aligned}\quad (3.6)$$

where $\Phi_u(x)$ and $\hat{\Phi}_u(x)$ are as defined previously above.

(C) We are also interested in the number of times the maximum is attained when $\Upsilon_u = \Upsilon_v$ and $\hat{\Upsilon}_u = \hat{\Upsilon}_v$ as defined by the following parameters;

$$\begin{aligned}L_u &= ||\{x : |\Phi_u(x)| = \Upsilon_u, \quad 1 \leq x \leq N-1\}||; \\ \hat{L}_u &= ||\{x : |\hat{\Phi}_u(x)| = \hat{\Upsilon}_u, \quad 1 \leq x \leq N-1\}||;\end{aligned}\quad (3.7)$$

where $||W||$ denotes the the number of elements in the set W and the parameter L_u indicates the number of times that $|\Phi_u(x)|$ attains its maximum value.

(d) In addition to the peak correlation parameters we are also interested in the *mean square autocorrelation* parameters which are defined as

$$\sum(u) = \sum_{x=1}^{N-1} \Phi_u^2(x), \quad (3.8)$$

$$\widehat{\sum}(u) = \sum_{x=1}^{N-1} \hat{\Phi}_u^2(x) \quad (3.9)$$

and the *sidelobe energy* parameter

$$S_u = \sum_{x=1}^{N-1} C_u^2(x). \quad (3.10)$$

The parameters cited above are of importance in CDMA application due to synchronization and multipath considerations, of which we are primarily interested in the latter. Specifically we will want to pick codes, in *auto-optimal* phase,¹⁴ where Υ_u attains its maximum

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¹⁴ The phase j of a sequence α is an auto-optimal(AO) phase if $\hat{\Upsilon}(T^j\alpha) = \hat{\Phi}_{AO}(\alpha)$ and if $\hat{L}(T^j\alpha)$ is the minimum of $L(T^i\alpha)$ over all i for which $\hat{\Upsilon}(T^i\alpha) = \hat{\Phi}_{AO}(T^i\alpha)$ where α is a sequence of period N i.e. $(\alpha_0, \alpha_1, \dots, \alpha_{N-1})$ and $T^j\alpha$ denotes the N -tuple sequence $(\alpha_j, \alpha_{j+1}, \dots, \alpha_{j+N-1})$. We refer to $T^j\alpha$ as the j th phase of the sequence α and $T^j\alpha$ is called the j^{th} cyclic shift of the N -tuple α . In certain m -sequences there may be more than one AO phase. To further differentiate these AO phases we will focus on those AO phases with the least sidelobe energy (AO/LSE) which is defined as the phase j of a sequence α which is AO and if $S(T^j\alpha) \leq S(T^i\alpha)$ for all auto-optimal phases i .

value the least possible number of times to reduce the probability of acquiring false synchronization. In CDMA systems the knowledge of the sidelobe energy of a sequence can be exploited or use to bound the value of the average other user interference between codes.

Ideally in CDMA the PN sequences assigned to each user should be mutually exclusive so that the level of interference experience by any one users from transmissions of other users add on a power basis. However in practice the PN sequences exhibit some crosscorrelation. For CDMA applications, among the class of PN sequences, Gold and Kasami sequences have better periodic crosscorrelation properties, i.e., low crosscorrelation between different channels. For completeness, we will review a general and widely known class of binary PN sequences. They are maximum-length shift register sequences or m -sequences for short. The application of m -sequences, arises in connection with, when two m -sequences, with specific properties, are combined to form CDMA codes (such as Gold and Kasami sequences) that have better periodic crosscorrelation properties.

3.2.2 Maximal Length Sequences

A m -sequence, by definition, has length $N = 2^m - 1$ bits or chips and is generated by an m -stage shift register with linear feedback as shown in Figure 3.3. In general binary linear feedback shift registers (SRG) are described by polynomials. A polynomial $f(x)$ of degree m is *irreducible* if it is not divisible by any polynomial of degree less than m but greater than 0. An *irreducible* polynomial, when used to define SRG, will generate a sequence with a length that is a factor of $2^m - 1$. If this sequence, that consists of periodic length of $N = 2^m - 1$, has N as a prime number, the SRG that generate this sequence is said to generate a maximal length sequence. Thus a SRG described by a *primitive* polynomial will generate an m sequence.

Fig. 3.3. An m -stage shift register ($m = 7$)

The illustrated $m = 7$ -stage shift register above correspond to the polynomial $f(x) = x^7 + x^6 + x^3 + x + 1$ which can be represented as 011 001 011 in binary or 313 in octal notation. The properties of m -sequences of relevance to this situation are :

- (a) The number of m -sequences of a given length is an important property. This number can be shown to be

$$\frac{1}{m} \phi(2^m - 1), \quad (3.11)$$

where $\phi(k)$ is the number of positive integers less than k and relatively prime to k . This is frequently referred to as the Euler totient function. Examples of the number of sequences of different lengths are shown in Table III.1.

- (b) The m -sequence is periodic with period N . Each period of the m -sequence contains 2^{m-1} ones and $2^{m-1} - 1$ zeros. This is typical of a balanced code. We note that an unbalanced code will produce spectral emissions containing uneven or unusual large peaks or spikes which may cause interference to other RF systems which may share the same transmission medium. This one-zero balance is also important because, when modulating a carrier, it limits the degree of carrier suppression obtainable. Carrier suppression is dependent on the symmetry of the modulating signal, hence the longer the code sequence the less the effect of the code sequence on the carrier balance.
- (c) For m -sequences, the periodic autocorrelation function is

$$\Phi_u(x) = \begin{cases} N, & x = 0; \\ -1, & 1 \leq x \leq N - 1; \end{cases} \quad (3.12)$$

We note that *autocorrelation* is the degree of correspondence between a code and a phase-shifted replica of itself. For m -sequences with long period N , the off-peak values of $\Phi_u(x)$ relative to the peak value i.e. $\Phi_u(x)/\Phi_u(0) = -1/N$ is small and thus insignificant from a practical viewpoint. The auto-correlation plot for the maximal codes are two-valued with a peak only at the zero shift point as shown in Figure 3.4. This is an important feature as it allows receiver to discriminate between signals on a yes-no basis. Thus m -sequences are ideal when viewed in terms of their autocorrelation function.

Fig. 3.4. Autocorrelation plot

Other properties of m -sequences are described in [21].

In DS/SS-MA application the crosscorrelation properties of PN sequences are as important as the autocorrelation properties. We note that *crosscorrelation* is the measure of agreement between two different codes. In a DS/SS-MA scheme, each user is assigned a distinct PN sequence. Ideally the PN sequences assigned should be mutually orthogonal so that the level of interference experienced by any one user from the transmissions of other users add on a power basis. However, in practice the PN sequences used exhibit some correlation. The crosscorrelation function between pairs of m -sequences of the same period can have relatively large peaks. The peak magnitudes for the periodic crosscorrelation between pairs of m -sequences are listed in Table III.1.

Table III.1. Peak crosscorrelations of m -sequences and Gold sequences*

m	$N = 2^m - 1$ $\Phi_u(0)$	Number of m sequences	Peak Cross Correlation** $\Upsilon_{u,v}^m(m)$	$\frac{\Upsilon_{u,v}^m(m)}{\Phi_u(0)}$	Peak Cross Correlation $\Upsilon_{u,v}^g(m)$	$\frac{\Upsilon_{u,v}^g(m)}{\Phi_u(0)}$
3	7	2	5	0.71	5	0.71
4	15	2	9	0.60	9	0.60
5	31	6	11	0.35	9	0.29
6	63	6	23	0.36	17	0.27
7	127	18	41	0.32	17	0.13
8	255	16	95	0.37	33	0.13
9	511	48	113	0.22	33	0.06
10	1023	60	383	0.37	65	0.06
11	2047	176	287	0.14	65	0.03
12	4095	144	1407	0.34	129	0.03

* Reproduced from [16].

** $\Upsilon_{u,v}^m(m)$ denotes crosscorrelation values of m -sequences while $\Upsilon_{u,v}^g(m)$ denotes cross-correlation values of Gold sequences.

We note that the peak magnitude, for most of the m -sequences, of the crosscorrelation function is a large percentage of the peak value of the autocorrelation function. These high percentages are intolerable in CDMA application. Although it is possible to select a smaller subset of m -sequences that have relatively smaller crosscorrelation peak values, the number of sequences in the set is usually too small for CDMA purposes. However, the application of m -sequences usage arises in connection with DS/SS-MA application when two m -sequences are combined to form CDMA codes (viz Gold sequences which we have indicated earlier have better periodic crosscorrelation properties). A procedure similar to that of generating Gold codes will generate another well-known CDMA code known as Kasami codes.

3.2.3 Gold Codes

Gold codes are generated by modulo-2 addition of a pair of m -sequences (where the pair is judiciously selected for its well-defined correlation properties) as detailed in this section. Gold [22] showed that certain pairs of m -sequences of length $N = 2^m - 1$ exhibit a three-valued crosscorrelation functions with values $\{-1, -\Upsilon_{u,v}^g(m), \Upsilon_{u,v}^g(m) - 2\}$ where

$$\Upsilon_{u,v}^g(m) = \begin{cases} 2^{(m+1)/2} + 1, & m = \text{odd}; \\ 2^{(m+2)/2} + 1, & m = \text{even}. \end{cases} \quad (3.13)$$

His results are summarized by the following theorem:

Gold's Theorem: Let $f(x)$ and $g(x)$ be a preferred pair of primitive polynomials of degree m whose corresponding shift registers generate maximal length sequences of period $2^m - 1$ and whose cross correlation function satisfies the inequality

$$|\Phi_{u,v}(x)| \leq \Upsilon_{u,v}^g(m), \quad (3.14)$$

where $\Phi_{u,v}(x)$ and $\Upsilon_{u,v}^g(m)$ are as defined before. The shift register corresponding to the product polynomial $f(x)g(x)$ will generate $2^m + 1$ different linear sequences each with period $2^m - 1$. The crosscorrelation of any pair of these linear sequences will satisfies the above inequality. Two sequences that satisfy (3.14) are referred to as a pair of *preferred sequences*. Let us consider an example.

Gold Code Design Example

Consider the pair of primitive polynomials

$$\begin{aligned} f(x) &= x^7 + x^3 + 1, \\ g(x) &= x^7 + x^3 + x^2 + x + 1, \end{aligned}$$

from which we form the product polynomial

$$h(x) = f(x)g(x) = x^{14} + x^9 + x^8 + x^6 + x^5 + x^4 + x^2 + x + 1, \quad (3.15)$$

as illustrated in Figure 3.5. According to Gold's theorem, in this design the corresponding shift register will generate 129 (i.e., $2^7 + 1$) different linear sequences of period length 127 (i.e., $2^7 - 1$). The maximum crosscorrelation, $\Upsilon_{u,v}^g$, value for any pair of sequences from the set is

$$\Upsilon_{u,v}^g(7) = 17. \quad (3.16)$$

(The maximum peak autocorrelation in contrast is 127). The generated codes in this example have three-valued autocorrelation function sidelobes and a three-valued crosscorrelation taking on values from the set $(-1, -17, 15)$ as shown in Tables III.2 and III.3. In contrast the largest set of m -sequences of period 127 for which this bound holds contains only 6 sequences. In short the significance of Gold's Theorem is that a preferred pair of

m -sequences can provide large number of non-maximal sequences with a bound on their cross correlation.

Fig. 3.5. Gold code generator

Table III.2. Correlation values for a set of AO/LSE Gold sequences (N=127)*

k	$\mu = \alpha_0^{(k)}$	Υ_μ	L_μ	$\Sigma(\mu)$	$\hat{\Upsilon}_\mu$	$\hat{L}_\mu$	S_μ
1	01450	17	14	10430	21	2	5447
2	15167	17	18	13374	19	2	5563
3	17160	17	18	13374	19	2	5463
4	33112	17	18	13374	19	2	5571
5	35576	17	18	13374	19	2	5651
6	05573	17	18	13374	21	4	6535
7	03665	17	18	13374	19	6	6307
8	15665	17	18	13374	13	8	4815

Table III.3. Peak Correlation values for a set of AO/LSE Gold sequences (N=127)*

k	1	2	3	4	5	6	7	8
1	21	27	27	31	31	43	35	31
2	17	19	27	35	27	31	29	29
3	17	17	19	31	37	31	33	31
4	17	17	17	19	31	27	25	31
5	17	17	17	17	19	27	35	25
6	17	17	17	17	17	21	27	29
7	17	17	17	17	17	17	19	31
8	17	17	17	17	17	17	17	13

* Reproduced from [21]

3.2.4 Kasami Codes

A technique due to Kasami [23], similar to the one used to generate Gold sequences will generate a smaller set of $2^{m/2}$ binary sequences of period $2^m - 1$, where m is even. The procedure to generate the Kasami codes is as follows:

- Beginning with an m -sequence (denoted as c), a binary sequence (denoted as b) is formed by taking every $2^{m/2} + 1$ bit of c , i.e., sequence b is formed by decimating the sequence a by $2^{m/2} + 1$. We note that the resulting sequence b is periodic with period $2^{m/2} - 1$. For example, if $m = 8$, the period of c is 255 and the period of b is 15.
- Next, by taking $N = 2^m - 1$ bits of the sequences c and b we form a new set of sequences (termed Kasami sequences) by adding, modulo-2, the bits from c and the

bits from **b** and all $2^{m/2} - 2$ cyclic shifts of the bits from **b**. By including **c** in the set, a set of $2^{m/2}$ binary sequences of length $N = 2^m - 1$ is derived.

The autocorrelation and crosscorrelation functions of these Kasami sequences take on values from the set $(-1, -(2^{m/2} + 1), 2^{m/2} - 1)$. The maximum crosscorrelation, $\Upsilon_{u,v}^k$, value for any pair of sequences from the set is

$$\Upsilon_{u,v}^k(m) = 2^{m/2} + 1. \quad (3.17)$$

To illustrate these codes, consider the following example.

Kasami Code Design Example

Fig. 3.6. Kasami code generator

The procedure for designing Kasami sequences is quite similar to the one used to generate the Gold codes. If we consider $m = 8$ with the following primitive polynomial

$$f(x) = x^8 + x^4 + x^3 + x^2 + 1$$

and a suitably chosen primitive polynomial with degree $m/2 = 4$, such as

$$g(x) = x^4 + x + 1,$$

we will form the product polynomial

$$h(x) = x^{12} + x^9 + x^7 + x^6 + x^5 + x^4 + x^2 + x + 1 \quad (3.18)$$

as illustrated in Figure 3.6 above.

We note that the above 12 stage linear feedback shift register corresponding to the product polynomial register will generate a set of 16 ($= 2^{8/2}$) different linear sequences of period length 255 ($= 2^8 - 1$). The generated codes have three-valued autocorrelation function sidelobes and a three-valued crosscorrelation taking on values from the set $(-1, -17, 15)$ as shown in Tables III.4 and III.5. Hence the maximum crosscorrelation of any pair of these sequences is

$$\Upsilon_{u,v}^k(8) = 17. \quad (3.19)$$

Table III.4. Correlation values for a set of AO/LSE Kasami sequences ($N=255$)*

k	$\mu = \alpha_0^{(k)}$	Υ_μ	L_μ	$\Sigma(\mu)$	$\hat{\Upsilon}_\mu$	$\hat{L}_\mu$	S_μ
1	2357	17	110	60606	31	2	27807
2	2326	17	112	60734	31	4	27127
3	3551	17	112	60734	29	2	28227
4	7430	17	112	60606	33	2	29179
5	5636	17	112	60734	31	2	28603
6	3710	17	112	60734	29	6	26399
7	5044	17	112	60734	29	4	27903
8	0261	17	112	60734	31	2	27211
9	0617	17	112	60734	31	2	29287
10	4666	17	110	60606	31	4	26543
11	0454	17	110	60606	29	4	29407
12	3134	17	110	60606	31	2	27079
13	2577	17	110	60606	31	2	28755
14	3416	17	110	60606	27	8	28387
15	7577	17	110	60606	31	4	27867
16	205	17	254	254	25	6	9199

Table III.5. Peak correlation values for a set of AO/LSE Kasami sequences ($N=255$)*

k	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16
1	31	51	43	41	41	37	51	47	39	45	45	43	45	45	41	51
2	17	31	51	45	43	49	41	41	39	49	39	43	41	43	41	45
3	17	17	29	45	41	47	47	41	51	45	49	41	47	49	45	43
4	17	17	17	33	47	41	41	39	45	41	45	43	43	41	41	41
5	17	17	17	17	31	43	39	45	43	41	37	39	41	43	51	41
6	17	17	17	17	17	29	49	41	41	39	47	45	39	47	37	49
7	17	17	17	17	17	17	29	49	45	39	47	43	43	49	47	43
8	17	17	17	17	17	17	17	31	41	41	45	43	45	39	41	45
9	17	17	17	17	17	17	17	17	31	39	45	43	41	49	45	43
10	17	17	17	17	17	17	17	17	17	31	39	49	49	49	41	43
11	17	17	17	17	17	17	17	17	17	17	29	47	43	43	43	49
12	17	17	17	17	17	17	17	17	17	17	17	31	37	45	35	51
13	17	17	17	17	17	17	17	17	17	17	17	17	31	45	43	41
14	17	17	17	17	17	17	17	17	17	17	17	17	17	27	39	43
15	17	17	17	17	17	17	17	17	17	17	17	17	17	17	31	39
16	17	17	17	17	17	17	17	17	17	17	17	17	17	17	17	25

* Reproduced from [21]

3.3 PROPOSED DS/SS-MA RADIO SYSTEM DESIGN

In this section we will present the DS/SS transmitter and receiver design as illustrated in Figures 3.7-8.

3.3.1 Transmitter Design

For each user, the transmitter consists of a code sequence generator and a modulo-2 adder prepended to a typical DPSK transmitter as shown in Figure 3.7. In this DS/SS system, the chips of the code sequence are modulo-2 added to each of the incoming differentially encoded data bits. The resultant bit stream is then used to modulate an intermediate frequency (IF) carrier with a BPSK modulator. The IF modulated signal is upconverted to radio frequency (RF) using a mixer and a RF local oscillator. The upper sideband of the signal is then selected using a high pass filter. It is then filtered at the bandpass filter, amplified through a linear power amplifier and finally fed to the antenna.

Fig. 3.7. DS/SS Transmitter

3.3.2 Receiver Design

The signal arriving at the receiver, Figure 3.8, is first converted from RF to IF using a conventional downconverter. Since we used DPSK signalling, precise carrier recovery is not required. The main steps that must be performed on the received IF signal in order to recover the information data are then despreading and data recovery. The

simple and efficient method of initial synchronization by matched filter (SAW devices) detecting the IF signal directly, as described earlier, is used. We assume that perfect chip synchronization is achieved in detecting the signal.

As shown in Figure 3.8, the received signal is received at the antenna connected to a low noise amplifier (LNA) and then downconverted to the IF stage. The IF signal output is bandpass-filtered through a IF filter onto an IF amplifier, with an automatic gain control (AGC), and then to the matched filter. The matched filter is a SAW device (with center frequency matched to the IF carrier and prefabricated or programmed with DS codes assigned specifically for each user) will provide an autocorrelation peak response in the presence of a matching sequence. If there is no match the SAW output will be that of crosscorrelation sidelobes. The output of the matched filter is passed to a broadband amplifier to compensate for the insertion loss of the SAW device.

Fig. 3.8. DS/SS Receiver

The largest outputs among the matched filters, in the case of selection diversity, is selected by the CWU, demodulated and then low-passed filter to eliminate the harmonics. The resulting baseband bipolar signal is then amplified and window detected, with dual comparators, for positive and negative peaks.

3.3.3 Communication Watchdog Unit/Average Power Control

In our system the extent of the interuser interference depends on the correlation properties of the codes as well as the relative strength of the signal from a desired source and the interferer at the receiver. Situations can arise where the levels are very different producing large interference. Such a case can occur when signals (with less path attenuations) arrive from nearby transmitters to swamp the weaker signals arriving from transmitters further afield. This situation is referred to as the *near-far field effect*. To overcome this near-far field effect some form of automatic power control be used, with the aim being to have all signals arrive at the base station receiver with equal strength (in the opposite direction the near-far problem is minimal or non-existent since all the transmitters are colocated at the master station). There are several ways to accomplish this. They are all based on the reciprocity assumption that propagation loss from a user to the base station is the same as the loss from the base station to the user terminal. This assumption should be valid if a common receiving/transmitting antenna is used at the base station. In this case if a known signal is transmitted, the receiver can determine the propagation loss to the base station and adjust its transmitted power level to produce a specified receiver power level. In our study we assume that perfect average power control is attainable.

Two reference signals have been proposed: a common modulated DS signal shared by all receivers and a pilot tone inserted inband. As an example, assume that there are 15 users at a site (ICS) communicating with the master station. Referring to Table III.1. we assigned Kasami codes of length 255 with one distinct code for each user (i.e., $k = 1$ to 15) and a common code ($k = 16$) for average power control and other housekeeping purposes. This average power control feature is provided in what we termed the “communication watchdog unit” (CWU) as illustrated in Figures 3.9 and 3.10. There are two possible space diversity configurations for the case of two antenna diversity arrangement as shown in the figures. Two sets of DS/SS receivers, for each user, are required in the first configuration. In the latter, based on the CWU monitoring, the DS/SS receiver is switched over to the antenna with the largest outputs. Having proposed the DS/SS-MA radio design let us model this system in the next chapter.

Fig. 3.9. Communication Watchdog-Configuration A

Fig. 3.10. Communication Watchdog-Configuration B

IV. DS/SS-MA SYSTEM MODEL & PERFORMANCE EVALUATION

4.1 SYSTEM MODELS

The transmitter, channel and receiver models for the DS/SS-MA IWC system (illustrated in Figure 1.1) is described in this chapter. Figure 4.1 illustrates the corresponding DS/SS receiver for each user. We repeat that the DS/SS receiver employs DPSK demodulation and selection diversity (i.e., resolvable path diversity with direct sequence spread spectrum and space diversity) reception.

Fig. 4.1. DS/SS receiver with selection diversity.

4.1.1 Transmitter Model

The signal transmitted by the k th user is given by

$$s_k(t) = A_k a_k(t) b_k(t) \cos(\omega_c t + \phi_k), \quad (4.1)$$

where $b_k(t)$ is a NRZ PAM signal representing a differentially encoded binary data stream $\dots, b_k^{-1}, b_k^0, b_k^1, \dots$ ($b_k^i \in \{-1, 1\}$), to be transmitted, $a_k(t)$ is a similar signal corresponding to the periodic spreading sequence assigned to the k th user ($a_k^i \in \{-1, 1\}$), ω_c is the common (angular) carrier frequency, ϕ_k is the carrier phase, and A_k is the transmitted signal amplitude. If T_c is the duration of a bit in the spreading sequence and T the duration of a data symbol, then clearly $T = NT_c$. In our analysis we assume that $\omega_c T = 2l\pi$ for some integer l .

To ameliorate the near-far field effect, we assume the A_k s are all adjusted so that all signals from the individual users arrive at the base station with the same average power.

4.1.2 CHANNEL MODEL

In our study we assume that the channel between the k th user and the base station has Rician multipath fading channel characteristics. The lowpass equivalent impulse response of the channel can be modelled [16] as

$$h_k(\tau) = \sum_{l=1}^L \beta_{lk} \delta(\tau - \tau_{lk}) \exp(-j\gamma_{lk}), \quad (4.2)$$

where β_{lk} , τ_{lk} and γ_{lk} are the gain, delay and phase respectively of the l th 'path' identified between the k th user and the base station, and L represents the maximum number of paths that may be discernable. Basically to resolve the number of multipaths we will base it on the time resolution of the direct sequence signals, i.e., two signals has to be separated by at least one chip time T_c . In our analysis we assume that there is no significant intersymbol interference as a result of the multipath fading phenomena, i.e., the maximum multipath delay spread, T_m , is much less than the data symbol duration T . Hence,

$$L = \lfloor T_m/T_c \rfloor + 1, \quad (4.3)$$

where $\lfloor x \rfloor$ denotes the largest integer less than or equal to x . Furthermore, we assume that τ_{lk} may be described by a random variable uniformly distributed over $[0, T_m]$, the

path phase γ_{lk} may be described by a random variable uniformly distributed over $[0, 2\pi]$, β_{lk} may be described by a Rician random variable, and all these random variables are independent for all l and k .

The density function of a Rician random variable β depends on two parameters:

$$f_R(r) = \begin{cases} \frac{r}{\sigma^2} \exp\left(-\frac{r^2 + s^2}{2\sigma^2}\right) I_0\left(\frac{rs}{\sigma^2}\right), & r \geq 0; \\ 0, & r < 0; \end{cases} \quad (4.4)$$

where s is a "noncentrality parameter," $s^2 + 2\sigma^2$ is the mean square value of β , and I_0 is the modified Bessel function of the first kind and order zero. The ratio $\kappa^2 = \sigma^2/s^2$ is particularly significant for our work here for it represents the *random to specular power ratio*.

4.1.3 RECEIVER MODEL

The main receiver at the base station consists of a number of antennas (to provide space diversity) with an identical bank of filters (one filter matched to the code waveform of each user) connected to each. That is, each user has a corresponding DS/SS receiver at the base station as shown in Figure 4.1.

In our study, the receiver model employs selection diversity. The receiver contains a filter matched to the coded waveform of the reference user, which is cascaded to a DPSK demodulator and a threshold detector. We note that the path that provided the largest peak signal is selected and demodulated. That is, the largest signal among the corresponding filter outputs is selected and passed to the DPSK demodulator and threshold detector. This selection diversity rule is based on two fundamental steps. First, the largest S/S correlation peak among the various L discernible S/S signals is selected per RF antenna (path selection). Next, the largest signal among the L_a number of antennae is then chosen (antenna selection). This will give an overall diversity of order M , i.e., $M = L_a L$.

Next, we will assume that the system is corrupted by additive white Gaussian noise (AWGN) and are similar for each of the diversity receiver. This will make our analysis more mathematically tractable as we assumed that the power levels of the corrupted AWGN noise in the system to be much greater than the noise figure of the front end of each diversity receiver. In our analysis, with the A_k s adjusted for constant average power as previously described, the products $A_k \beta_{lk}$ will be assumed to be given by independent

identically distributed Rician random variables G_{lk} . Thus the received signal at the input of the matched filter of the receiver can be easily shown to be

$$r(t) = \sum_{k=1}^K \sum_{l=1}^L G_{lk} a_k(t - \tau_{lk}) b_k(t - \tau_{lk}) \cos(\omega_c t + \theta_{lk}) + n(t), \quad (4.5)$$

where

$$n(t) = n_c(t) \cos(\omega_c t) + n_s(t) \sin(\omega_c t) \quad (4.6)$$

is expressed in terms of its lowpass equivalent components. We note that the carrier phase, ϕ_k , has been absorbed in the random phase θ_{lk} , associated with each channel path, i.e.,

$$\theta_{lk} = \phi_k - \gamma_{lk}. \quad (4.7)$$

4.2 SYSTEM PERFORMANCE EVALUATION

In this section we will derive the system performance equations based on the system models proposed in the previous section and the general probability error expression for multichannel digital signalling in an AWGN environment given in (2.61)–(2.64). To determine the performance of the system we will like to determine the average probability of error or bit error rate (BER) as a function of the average received signal-to-noise ratio, $\gamma_0 = 2\rho E_b/N_0$. We note that in the system we are considering we assume perfect power control, so the performance will be independent of the distance from the receiver to the users in a sense.

4.2.1 Demodulation

To evaluate system performance we may simply choose one user as a reference user and calculate the average probability of error for any particular data symbol of this user. We shall choose user 1 and the data symbol b_1^0 . Let $g_x(t)$ and $g_y(t)$ be the inphase and quadrature components of output of the filter matched to user 1 assuming a sampling instant at time $t = T$. If we denote z_1 as the complex envelope of the matched filter output at the sampling instant then

$$z_1 = g_x(T) + jg_y(T) \quad (4.8)$$

with

$$\begin{aligned} g_z(T) &= \sum_{k=1}^K \sum_{l=1}^L G_{lk} \cos \theta_{lk} [b_k^{-1} R_{1k}(\tau_{lk}) + b_k^0 \hat{R}_{1k}(\tau_{lk})] + \eta, \\ g_y(T) &= \sum_{k=1}^K \sum_{l=1}^L G_{lk} \sin \theta_{lk} [b_k^{-1} R_{1k}(\tau_{lk}) + b_k^0 \hat{R}_{1k}(\tau_{lk})] + \nu, \end{aligned} \quad (4.9)$$

where b_k^{-1} and b_k^0 are the data bits of the k^{th} user in the previous and current bit intervals. The remaining terms in the above equation are the continuous-time partial cross correlations between the regenerated chip sequence $a_1(t)$ and a delayed version of the various interfering codes, $a_k(t)$ which is given by

$$\begin{aligned} R_{1k}(\tau) &= \int_0^\tau a_k(t - \tau) a_1(t) dt, \\ \hat{R}_{1k}(\tau) &= \int_\tau^T a_k(t - \tau) a_1(t) dt, \end{aligned} \quad (4.10)$$

for $0 \leq \tau \leq T$.

Alternately, we can represent [21] the above partial cross correlation parameters as ,

$$\begin{aligned} R_{1k} &= C_{1k}(x - N)T_c + [C_{1k}(x + 1 - N) - C_{1k}(x - N)](\tau - xT_c), \\ \hat{R}_{1k} &= C_{1k}(x)T_c + [C_{1k}(x + 1) - C_{1k}(x)](\tau - xT_c), \end{aligned} \quad (4.11)$$

for $0 \leq xT_c \leq \tau \leq (x+1)T_c \leq T$, where, from (3.1), the discrete aperiodic cross-correlation function C_{1k} for the sequences $\{a_1^i\}$ and $\{a_k^i\}$ is given by

$$C_{1k}(x) = \begin{cases} \sum_{i=0}^{N-1-x} a_k^i a_1^{i+x}, & 0 \leq x \leq N-1; \\ \sum_{i=0}^{N-1+x} a_k^{i-x} a_1^i, & 1-N \leq x \leq 0; \\ 0, & |x| \geq N. \end{cases} \quad (4.12)$$

Lastly, the noise samples η and ν are independent zero mean Gaussian random variables, with variances equal to $N_0 T$, as given by

$$\begin{aligned} \eta &= \int_0^T a_1(s) n_c(s) ds, \\ \nu &= \int_0^T a_1(s) n_s(s) ds. \end{aligned} \quad (4.13)$$

As before, based on user 1 as reference, we will assume that the j^{th} path between the transmitter of user 1 and the corresponding receiver at the base station is the primary path between user 1 and the base station, and all other paths give rise to interference. In this situation, we referenced the time and phase origins so that $\tau_{j1} = 0$, $\theta_{j1} = 0$ (so as to facilitate our computation and thus complete the path characteristics for the dominant path of our reference user). We will denote z_2 as the complex envelope of the matched filter output at the previous sampling instant ($t = -T$). Since the fading is slow compared to the data rate, i.e., we assumed that the channel within this time window is *time-invariant*, it will be reasonable to assume that the path characteristics used in computing z_1 can be identically applied in computing z_2 . Thus, z_2 differs from z_1 only due to a shift in the data bits and in the changed noise samples only. Hence

$$\begin{aligned} z_1 &= G_{j1} T b_1^0 + \sum_{k=1}^K (b_k^{-1} U_k + b_k^0 \hat{U}_k) + j \sum_{k=1}^K (b_k^{-1} V_k + b_k^0 \hat{V}_k) + (\eta_1 + j\nu_1), \\ z_2 &= G_{j1} T b_1^{-1} + \sum_{k=1}^K (b_k^{-2} U_k + b_k^{-1} \hat{U}_k) + j \sum_{k=1}^K (b_k^{-2} V_k + b_k^{-1} \hat{V}_k) + (\eta_2 + j\nu_2), \end{aligned} \quad (4.14)$$

where for ease of analysis purposes, U_k , V_k , $\hat{U}_k$, and $\hat{V}_k$ are random variables defined as follows:

$$\begin{aligned} U_1 &= \sum_{\substack{l=1 \\ l \neq j}}^L R_{1l}(\tau_{1l}) G_{1l} \cos \theta_{1l}, & \hat{U}_1 &= \sum_{\substack{l=1 \\ l \neq j}}^L \hat{R}_{1l}(\tau_{1l}) G_{1l} \cos \theta_{1l}, \\ V_1 &= \sum_{\substack{l=1 \\ l \neq j}}^L R_{1l}(\tau_{1l}) G_{1l} \sin \theta_{1l}, & \hat{V}_1 &= \sum_{\substack{l=1 \\ l \neq j}}^L \hat{R}_{1l}(\tau_{1l}) G_{1l} \sin \theta_{1l}, \end{aligned} \quad (4.15)$$

and, for $K \geq 2$

$$\begin{aligned} U_k &= \sum_{l=1}^L R_{1k}(\tau_{lk}) G_{lk} \cos \theta_{lk}, & \hat{U}_k &= \sum_{\substack{l=1 \\ l \neq j}}^L \hat{R}_{1k}(\tau_{lk}) G_{lk} \cos \theta_{lk}, \\ V_k &= \sum_{l=1}^L R_{1k}(\tau_{lk}) G_{lk} \sin \theta_{lk}, & \hat{V}_k &= \sum_{\substack{l=1 \\ l \neq j}}^L \hat{R}_{1k}(\tau_{lk}) G_{lk} \sin \theta_{lk}. \end{aligned} \quad (4.16)$$

We note that the output of the DPSK demodulator at the sampling instant given by the decision variable, Γ' , relates to the difference in the arguments of the two complex numbers

(i.e., projecting z_1 onto z_2 and then use the argument of the resulting complex number) as indicated below

$$\Gamma' = z_1 z_2^*. \quad (4.17)$$

In binary DPSK case, the two possible transmitted phase difference are 0 and π radians. Thus for probability of error determination purpose, the real part of the decision variable will suffice, i.e., in the case of no diversity

$$\Gamma = \text{Re}[\Gamma'] = [z_1 z_2^* + z_1^* z_2]/2. \quad (4.18)$$

In the case of selection diversity of order M , the decision variable is

$$\Gamma_{\max}^M = \max_{i=1, \dots, M} \Gamma_i. \quad (4.19)$$

We note that $\Gamma_i, i = 1, \dots, M$ are independent random variables statistically identical to Γ . The probability of error P_e can then be shown to be,

$$\begin{aligned} P_e &= P[\Gamma_{\max} < 0 | b_1^0 b_1^{-1} = 1] \\ &= P[\Gamma_{\max} > 0 | b_1^0 b_1^{-1} = -1]. \end{aligned} \quad (4.20)$$

4.2.2 Probability of Error Expression

4.2.2.1 DS/SS Diversity only

For the system configured with DS/SS diversity only, the average probability error expression for selection diversity of order M (where M is the order of selection diversity exactly equal to the number of L paths resolved in this case) can be based on the general expression given in (2.62), (2.63) and (2.64).¹⁵ In our analysis, we assume that the largest value, i.e., G_{j1} obtained from the M demodulated signals in $\Gamma_{\max}^M$ is selected then the error probability for DPSK, in an additive white Gaussian noise channel, with the noise and interference terms treated as Gaussian¹⁶ can be shown to be

$$P_e = Q(a, b) - \frac{1}{2} [1 + r] I_0(ab) \exp\left(-\frac{a^2 + b^2}{2}\right), \quad (4.21)$$

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¹⁵ For completeness, we repeat the BER expression from Chapter 2.

¹⁶ The treatment of other multiusers as interferers in terms of Gaussian noise will simplify our analysis and has been applied in various investigative works

where

$$\begin{aligned} a &= \sqrt{\frac{|m_x/\sqrt{\mu_{xx}} - m_y/\sqrt{\mu_{yy}}|^2}{2}}, \\ b &= \sqrt{\frac{|m_x/\sqrt{\mu_{xx}} + m_y/\sqrt{\mu_{yy}}|^2}{2}}, \end{aligned} \quad (4.22)$$

$$\begin{aligned} m_x &= E[z_1], \\ m_y &= E[z_2], \end{aligned} \quad (4.23)$$

$$\begin{aligned} \mu_{xx} &= \text{var}(z_1), \\ \mu_{yy} &= \text{var}(z_2), \end{aligned} \quad (4.24)$$

$$\begin{aligned} \mu_{xy} &= E[(z_1 - m_x)(z_2 - m_y)^*], \\ r &= \frac{\mu_{xy}}{\sqrt{\mu_{xx}\mu_{yy}}}, \end{aligned} \quad (4.25)$$

$$I_0(ab) = \sum (ab/2)^{2k} / (k!)^2, \quad (4.26)$$

and $Q(a, b)$ is the Marcum Q function, assuming the delays $\{\tau_{lk}\}$, G_{j1} , K and L are all given.

If the parameters $\{\tau_{lk}\}$, G_{j1} , K and L are given we can reduced the computation requirements further. First, we made the following observations in conjunction with the random variables U_k , $\hat{U}_k$, V_k , and $\hat{V}_k$ given in (4.15) and (4.16). We note that the concatenated random variables $G_{lk} \cos \theta_{lk}$ and $G_{lk} \sin \theta_{lk}$ in the cited random variables can be evaluated as follows. Since G_{lk} , θ_{lk} are independent random variables where θ_{lk} is uniformly distributed in $[0, 2\pi]$ and G_{lk} follows a Rician distribution then the statistics of the concatenated random variables are

$$\begin{aligned} E[G_{lk} \cos \theta_{lk}] &= E[G_{lk}] \underbrace{E[\cos \theta_{lk}]}_{=0} = 0, \\ E[G_{lk} \sin \theta_{lk}] &= E[G_{lk}] \underbrace{E[\sin \theta_{lk}]}_{=0} = 0. \end{aligned} \quad (4.27)$$

As a result from the above, the random variables U_k , $\hat{U}_k$, V_k , and $\hat{V}_k$ can be shown to be zero-mean independent random variables. Similarly,

$$E[G_{lk}^2 \cos^2 \theta_{lk}] = E[G_{lk}^2] \underbrace{E[\cos^2 \theta_{lk}]}_{\frac{1}{2}} = \frac{1}{2} E[G_{lk}^2] = \frac{1}{2} A_k^2 [\sigma_{\theta_{lk}}^2 + s_{lk}^2] = A_k^2 \rho_{lk} = P_{lk},$$

$$E[G_{lk}^2 \sin^2 \theta_{lk}] = E[G_{lk}^2] \underbrace{E[\sin^2 \theta_{lk}]}_{\frac{1}{2}} = \frac{1}{2} E[G_{lk}^2] = \frac{1}{2} A_k^2 [\sigma_{\beta_{lk}}^2 + s_{lk}^2] = A_k^2 \rho_{lk} = P_{lk}. \quad (4.28)$$

As a result, we derived the following equations for the second moments:

$$\begin{aligned} E[U_1^2] &= \sum_{\substack{l=1 \\ l \neq j}}^L R_{11}^2(\tau_n) E[G_n^2 \cos^2 \theta_n], \\ E[V_1^2] &= \sum_{\substack{l=1 \\ l \neq j}}^L R_{11}^2(\tau_n) E[G_n^2 \sin^2 \theta_n], \\ E[\hat{U}_1^2] &= \sum_{\substack{l=1 \\ l \neq j}}^L \hat{R}_{11}^2(\tau_n) E[G_n^2 \cos^2 \theta_n], \\ E[\hat{V}_1^2] &= \sum_{\substack{l=1 \\ l \neq j}}^L \hat{R}_{11}^2(\tau_n) E[G_n^2 \sin^2 \theta_n], \\ E[U_1 \hat{U}_1] &= \sum_{\substack{l=1 \\ l \neq j}}^L R_{11}(\tau_n) \hat{R}_{11}(\tau_n) E[G_n^2 \cos^2 \theta_n], \\ E[V_1 \hat{V}_1] &= \sum_{\substack{l=1 \\ l \neq j}}^L R_{11}(\tau_n) \hat{R}_{11}(\tau_n) E[G_n^2 \sin^2 \theta_n]. \end{aligned} \quad (4.29)$$

With the results from (4.28) substituted into (4.29) we obtained

$$\begin{aligned} E[U_1^2] &= E[V_1^2] = \sum_{\substack{l=1 \\ l \neq j}}^L R_{11}^2(\tau_n) A_1^2 \rho_n = \sum_{\substack{l=1 \\ l \neq j}}^L R_{11}^2(\tau_n) P_n, \\ E[\hat{U}_1^2] &= E[\hat{V}_1^2] = \sum_{\substack{l=1 \\ l \neq j}}^L \hat{R}_{11}^2(\tau_n) A_1^2 \rho_n = \sum_{\substack{l=1 \\ l \neq j}}^L \hat{R}_{11}^2(\tau_n) P_n, \\ E[U_1 \hat{U}_1] &= E[V_1 \hat{V}_1] = \sum_{\substack{l=1 \\ l \neq j}}^L R_{11}(\tau_n) \hat{R}_{11}(\tau_n) A_1^2 \rho_n = \sum_{\substack{l=1 \\ l \neq j}}^L R_{11}(\tau_n) \hat{R}_{11}(\tau_n) P_n. \end{aligned} \quad (4.30)$$

For $K \geq 2$,

$$\begin{aligned}
E[U_k^2] &= \sum_{l=1}^L R_{1k}^2(\tau_{lk}) E[G_{lk}^2 \cos^2 \theta_{lk}], \\
E[V_k^2] &= \sum_{l=1}^L R_{1k}^2(\tau_{lk}) E[G_{lk}^2 \sin^2 \theta_{lk}], \\
E[\hat{U}_k^2] &= \sum_{l=1}^L \hat{R}_{1k}^2(\tau_{lk}) E[G_{lk}^2 \cos^2 \theta_{lk}], \\
E[\hat{V}_k^2] &= \sum_{l=1}^L \hat{R}_{1k}^2(\tau_{lk}) E[G_{lk}^2 \sin^2 \theta_{lk}], \\
E[U_k \hat{U}_k] &= \sum_{l=1}^L R_{1k}(\tau_{lk}) \hat{R}_{1k}(\tau_{lk}) E[G_{lk}^2 \cos^2 \theta_{lk}], \\
E[V_k \hat{V}_k] &= \sum_{l=1}^L R_{1k}(\tau_{lk}) \hat{R}_{1k}(\tau_{lk}) E[G_{lk}^2 \sin^2 \theta_{lk}].
\end{aligned} \tag{4.31}$$

Similarly with the result from (4.28) substituted into (4.31) we obtained

$$\begin{aligned}
E[U_k^2] &= E[V_k^2] = \sum_{l=1}^L R_{1k}^2(\tau_{lk}) A_k^2 \rho_{lk} = \sum_{l=1}^L R_{1k}^2(\tau_{lk}) P_{lk}, \\
E[\hat{U}_k^2] &= E[\hat{V}_k^2] = \sum_{l=1}^L \hat{R}_{1k}^2(\tau_{lk}) A_k^2 \rho_{lk} = \sum_{l=1}^L \hat{R}_{1k}^2(\tau_{lk}) P_{lk}, \\
E[U_k \hat{U}_k] &= E[V_k \hat{V}_k] = \sum_{l=1}^L R_{1k}(\tau_{lk}) \hat{R}_{1k}(\tau_{lk}) A_k^2 \rho_{lk} = \sum_{l=1}^L R_{1k}(\tau_{lk}) \hat{R}_{1k}(\tau_{lk}) P_{lk}.
\end{aligned} \tag{4.32}$$

Invoking the above properties and results we can easily show that the means of the complex random variables z_1 and z_2 are

$$E[z_1] = m_x = G_{j1} T b_1^0$$

and

$$E[z_2] = m_y = G_{j1} T b_1^{-1}. \tag{4.33}$$

Since $b_1^0 = b_1^{-1}$ (as the error probability, based on the decision variable, is $P_e = P[\Gamma_{maz} < 0 | b_1^0 b_1^{-1} = 1]$) then

$$m_x = m_y = m. \tag{4.34}$$

Similarly, we can show that second central moments of the complex Gaussian random variables z_1 and z_2 are

$$\begin{aligned} E[|z_1|^2] &= \mu_{xx} = 2E\left[\sum_{k=1}^K U_k^2 + \hat{U}_k^2 | \{\tau_{lk}\}, L\right] + 4E[U_1 \hat{U}_1 | \{\tau_{lk}\}, L] + 2\sigma_n^2, \\ E[|z_2|^2] &= \mu_{yy} = 2E\left[\sum_{k=1}^K U_k^2 + \hat{U}_k^2 | \{\tau_{lk}\}, L\right] + 2\sigma_n^2, \end{aligned}$$

and

$$E[z_1 z_2^*] = \mu_{xy} = 2E\left[\sum_{k=1}^K U_k \hat{U}_k + \hat{U}_1^2 | \{\tau_{lk}\}, L\right]. \quad (4.35)$$

where the above conditional expectations are defined from (4.29) to (4.32).

4.2.2.1 DS/SS Diversity and Space Diversity Combination

In the case of DS/SS diversity and space diversity combination, the average probability error expression can be given by (2.61) which is reproduced below as

$$\begin{aligned} P_b &= Q(a, b) - I_0(ab) \exp\left(-\frac{a^2 + b^2}{2}\right) \\ &+ \frac{I_0(ab) \exp\left[-(a^2 + b^2)/2\right]}{[2/(1-r)]^{2L_a-1}} \sum_{h=0}^{L_a-1} \binom{2L_a-1}{h} \left(\frac{1+r}{1-r}\right)^h \\ &+ \frac{\exp\left[-(a^2 + b^2)/2\right]}{[2/(1-r)]^{2L_a-1}} \sum_{n=1}^{L_a-1} I_n(ab) \left\{ \sum_{h=0}^{L_a-1-n} \binom{2L_a-1}{h} \right. \\ &\quad \times \left[\left(\frac{b}{a}\right)^n \left(\frac{1+r}{1-r}\right)^h - \left(\frac{a}{b}\right)^n \left(\frac{1+r}{1-r}\right)^{2L_a-1-h} \right] \Big\}, \quad L_a > 1; \quad (4.36) \end{aligned}$$

where

$$\begin{aligned} a &= \left(\frac{1}{2} \sum_{h=1}^{L_a} \left| \frac{m_x}{\sqrt{\mu_{xx}}} - \frac{m_y}{\sqrt{\mu_{yy}}} \right|^2 \right)^{1/2}, \\ b &= \left(\frac{1}{2} \sum_{h=1}^{L_a} \left| \frac{m_x}{\sqrt{\mu_{xx}}} + \frac{m_y}{\sqrt{\mu_{yy}}} \right|^2 \right)^{1/2}, \end{aligned} \quad (4.37)$$

and the other parameters are as defined in (4.23-26) with $Q(a, b)$ denoting the Marcum Q function and $I_n(\cdot)$ the n th order modified Bessel function of the first kind. We note that the overall selection diversity here is of the order of $M = L_a L$ where L_a is the number of antenna and L is the number of multipaths resolved by the DS/SS diversity receiver.

Having developed the system models and derived the system performance equations, we will proceed in the next chapter with a description of the simulation of the DS/SS-MA system used to determine numerically the system performance and discuss this performance under various scenarios.

V. DS/SS-MA SYSTEM SIMULATION

To simulate the IWC communication system we need to consider all the parameters that will ultimately influence the performance of the system. These parameters are given by the system models which are described in the preceding chapter. Thus to simulate and evaluate the BER performance (given by the system performance equations) of the IWC system, the following main steps (which correspond to the routines or subprograms of the simulator) are required.

We first generate information bits randomly to simulate the data source and differentially encode the data bits samples to simulate or represent the DPSK modulated signal transmitted by each k th user. For spreading these transmitted signals individual CDMA spreading codes are generated and assigned to the K users with the CDMA codes' parameters and input values as described in Chapter III. For the unique Gold and Kasami codes we will use initial loadings given in [21] and shown in Tables II.2 and II.4 respectively.

To simulate the multipath fading channel, the discrete channel model approach requires us to have a knowledge of the number of multipaths. In our study, the maximum number of multipaths L , given by (4.3), is based on the maximum multipath delay spread T_m where T_m is an input value based on expected or measured results of the channel. Thus we modelled the multipath signals as being comprised of " L " resolvable multipaths associated with β_{lk} , τ_{lk} and θ_{lk} (the gain, delay and phase respectively of the l th 'path' identified between the k th user and the base station) where:

- (a) τ_{lk} is simulated as a random variable that is uniformly distributed over $[0, T]$,
  - (b) the path phase θ_{lk} is simulated as a random variable uniformly distributed over $[0, 2\pi]$
- and
- (c) β_{lk} is simulated as a Rician random variable. We assume that the mean square value $2\rho_{lk}$ for the Rician path parameter β_{lk} is the same for all users and paths alike, i.e., $\rho_{lk} = \rho = \sigma^2 + \frac{1}{2}s^2$. The ratios $\kappa = \sqrt{2}\sigma/s$ (where the square of κ is the random to specular power ratio), are input values based on path measurements from [7].

We note that all the above random variables are assumed independent for all l and k . Subsequently from the system performance viewpoint, the predicted BER is sensitive to the following channel parameters:

- the multipath delay spread, and thus the number of received multipaths components, and
- the amount of fading.

The mean square value of the Rician random variable G_{lk} (formed from the products $A_k\beta_{lk}$), i.e., $E[G_{lk}^2] = A_k^2\rho = P_k$, is the parameter that we employ in mitigating the near/far problem. To exercise this average power control, we adjust the A_k values as, obviously ρ depends on the channel characteristics which is beyond the users control. We note that for the average received power (i.e., $P_k = P_{avg}$ for all k), the average received signal-to-noise ratio, γ_0 , is equal to $2\rho E_b/N_0 = \frac{P_{avg}T}{N_0}$.

The simulation of the received signal samples (including multipath fading signals) for the referenced user, will include addition or corruption by interfering multipath fading signals from the other users. As a result we need to compute the aperiodic crosscorrelations, described by (4.11), between the referenced user and other interfering users, i.e., between the regenerated referenced user code sequence and delayed version of the interfering codes. As this signal is further corrupted by AWGN noise we simulate random noise samples where the complex noise samples are independent zero mean Gaussian random variables with variances equal to N_0T . At the receiver we simulate the selection diversity rule as described in Chapter 4.

The evaluation of the BER performance of the system can be realized by the following two techniques:

- (A) Direct Monte Carlo Simulation and
- (B) Semi Analytical with Monte Carlo Simulation.

which are described next.

5.1 DIRECT MONTE CARLO SIMULATION TECHNIQUE

This approach is based on the probability of error given by the demodulation expression described in (4.20) and requires all the necessary routines or steps outlined above to simulate the IWC system. The main disadvantages of this technique are:

- a. the random noise samples are directly simulated,
- b. the simulation of a set of multipath channel characteristics, for each k th user and l th path, are required for each BER estimated and
- c. to attain reasonable accuracy in estimating the BER, a general rule of thumb is approximately 100 errors should be counted, which entails a computing run of $(100/\text{BER})$ bits. As an example, if a BER of 10^{-5} is computed a run length of 10^7 bits is required. Furthermore, if at least 100 different sets of multipath channel characteristics are required to provide a close approximation of the multipath effects then a run length of 10^9 bits is required.

Thus its not surprising that the major disadvantage of this approach is the horrendous amount of computing time required to estimate the BER, i.e., the bit errors are counted directly. Furthermore most computer systems are limited or finite in their dynamic storage user allocation. In our case, we are limited to estimating bit error rates not lower than 10^{-4} . Hence, mainly because of the inefficient computing time requirement and BER limit constraint we selected the alternate technique outlined in the next section for most of our simulation.

5.2 SEMI ANALYTICAL WITH DIRECT MONTE CARLO SIMULATION TECHNIQUE

This approach is based on the probability of error expression developed in Section 4.2.2. This technique reduces significantly the computation time as the direct simulation of the signals involved or decision variables is not required; rather we simply simulate randomly the selection of the parameters involved in the analysis of the probability of error done earlier. In effect we have determined the probability of error conditions on the variables $\{\tau_{lk}\}$, G_{j1} , K and L and simulating these random variables will allow us to numerically remove them as condition variables. To remove these conditions we proposed the following computational strategy:

- For any number of K users and L number of multipaths resolved the remaining random variables $\{\tau_{lk}\}$ and G_{j1} are removed by employing Monte-Carlo simulation.

The variance of the BER's (computed for each set of multipath channel values) will allow us to determine the confidence levels of the results. The average BER curves are then

available for further plotting with the 97.5 % confidence levels shown if necessary. To evaluate the Marcum Q-function, the technique described in [31] is employed. Although the number of random variables is sufficiently large its found that the number of iterations required is small in order to obtain a reliable value for the BER.

To implement the simulator, block programs of the abovementioned routines or steps were written in the FORTRAN 77 language, on a personal computer (16 bit machine) first. This modular programming approach enables the transportation of the overall simulation package to the mainframe environment. As a result the final standalone package is readily transportable to any size computer with an ANSI FORTRAN 77 compiler installed. Furthermore this modular approach facilitate or allow us to check and test the reliability and accuracy of the results the simulator produces.

A significant amount of resources were spent on modelling and verification. Specifically, two programs based on the two techniques were developed. The two program listings are attached as appendicies. Appendix A contains the program listing identified as "HD" based on the first technique while Appendix B contains the program listing identified as "BER" based on the second technique. Initially, the results from using the first program were used to compare with standard analytical BER expression for DPSK modulation, i.e., based on the trivial case of no multipaths and one user. A gradual increase in the complexity of the problem to check the deterioration in the BER results, i.e., by introducing multipath fading and multi-user interference, permit us to be confident of the results. Subsequently these results are compared with those obtained by using the second program. As there are significant shortcomings in the first technique, such as previously described, the results from using second program form the core of our simulation studies.

Having briefly described the mechanics of simulating the IWC DS/SS-MA system let us describe the measured channel or propagation results data that we use in our simulation next.

5.3 UTILIZATION OF MEASURED CHANNEL DATA

The BER performance expression, derived in Chapter 4 and implemented as described above, were used to determine the performance of the DS/SS-MA scheme for several different scenarios derived from some reported channel characteristics. Specifically, the performance of these schemes at bit rates of 32 kbits/s, 64 kbits/s, and 144 kbits/s were found.

These rates are those which are proposed for some of these systems [1]. As well, two code sequence sets were considered: $N = 127$ Gold codes and $N = 255$ Kassami codes. These systems were considered for situations where the rms delay spread was 100 ns and 250 ns (values reported to be worst case situations [3]) affecting the number of distinct paths that could be resolved by the receiver as shown in Table V.1 below.

Table V.1 Transmission parameters-maximum number of multipaths L

Data Rate (kB/s)	Period Length (N)	Chip Rate (MChips/s)	No. of paths (L) $T_m = 100\text{ns}$	No. of paths (L) $T_m = 250\text{ns}$
32	127	4.06	1	2
	255	8.16	1	3
64	127	8.13	1	3
	255	16.32	2	5
144	127	18.29	2	5
	255	36.72	4	10

The various combinations of these parameters were considered for two inbuilding Rician fading environments measured in [7]. The first set of measurements was recorded in a 30 year old brick building. The floor plan [7, Fig. 2] shows the building is partitioned into many small offices and laboratories. The second set of measurements was performed in another building of the same basic construction but has an open office concept. Distribution functions computed from the data recorded at cited buildings indicate Rician pattern with different κ values. These were found to correspond to a Rician fading channel with ratio $\kappa = 0.2089$ (for the classical building, identified as Building 1) and a channel with ratio $\kappa = 0.0794$ (for the "open office" environment identified as Building 2).

5.4 RESULTS AND DISCUSSIONS

To compare the results from the above-cited measurements values, the performance curves in [4] are reproduced as shown in Figure 5.1. The figure illustrates the performances of a DS/SS-MA scheme with selection diversity for two types of modulation scheme, namely DPSK and BPSK, at data rates of 32 kb/s and accomodating a maximum of 15 users (with

individual Kasami codes of periodic length equal to 255 assigned). The channel is assumed to have $L = 4$ resolvable multipaths (which corresponds to a maximum rms delay spread of about 490 ns) and $\rho = 0.019$ where ρ corresponds to the mean square value of the random path variable, β , characterized by a Rayleigh distribution.

Fig. 5.1. Performance of DS/SS-MA system shown for DPSK & CPSK modulation and selection diversity (from [4]): Kasami Code, $N = 255$, $T = 32$ kb/s, $K = 15$ users, $L = 4$ paths. $T_m = 490$ ns.

Figures 5.2–18 show the results for these channels for different levels of selection diversity M ($M = L_a L$, where L_a is the number of antennas being used to provide physical diversity), i.e., contrasting the performance of the scheme in the two building environments based on the following possible system configuration:

- a. No diversity ($M = 1$),
- b. DS/SS diversity only ($M = L$),
- c. Selection diversity, DS/SS diversity and space diversity ($L_a = 2$) combination, to give an overall diversity of $M = 2L$ and
- d. Selection diversity, DS/SS diversity and space diversity ($L_a = 3$) combination, to give an overall diversity of $M = 3L$

As previously described in Chapter 4, the order of diversity also depends on the number of L paths resolved by the DS/SS technique employed by the scheme. Thus, for example, in our Figure 5.2 if an order of $M = 2L$ is required to provide the appropriate BER threshold margin of $\approx 10^{-5}$ (given that the average received SNR is not less than 8 dB), we require 2 antennae. Since the discrete order of DS/SS diversity is $L_{SS} = L = 2$ the overall order of diversity provided is $M = 4$. As might be expected, the results demonstrate that, in most instances, the performance of the systems in the open office environment is better than in the more classical interior environment. This is particularly true in systems without diversity ($M = 1$) where we find the irreducible BER performance of the system, shown in all the figures, is significantly better in the open office environment. Since the BER performance of the IWC system depends on the number of users in operation, the information data rate, the type and length of CDMA codes employed, the selection diversity configuration, and most of all the channel characteristics in which it operates, let us focus and discuss on the results observed on the various possible scenarios.

5.4.1 Gold codes results

Figures 5.2 to 5.6 illustrates the performance of the IWC scheme where the codes employed are Gold codes of periodic length 127 and $K = 8$ users. Figures 5.2–5 illustrates the BER performance of the DS/SS-MA system as a function of the average received SNR ($5 \leq \text{Ave SNR (dB)} \leq 45$) for different data rates and time delay spreads. To summarize this section, the irreducible BER performance curves, as a function of the different data rates, for a maximum multipath delay spread of 250 ns, is shown in Figure 5.6.

Fig. 5.2 Performance of DS/SS-MA system with DPSK modulation and selection diversity (Gold Code, $N = 127$, $T = 32$ kb/s, $K = 8$ users, $L = 2$ paths, $T_m = 250$ ns).

The BER performance curves for data rate of $T = 32$ kb/s and $K = 8$ users for a maximum multipath delay spread of 250 ns are shown in Figure 5.2 above. We note that the maximum number of paths that can be resolved is 2 in this situation. From the figure, we observed that the irreducible BER (i.e., the BER saturates and any increase in the average SNR will not improve the BER) is approached at a SNR of about 30 dB. As expected, the irreducible BER of the IWC scheme in Building 1 is reduced, by about an order of magnitude, when DS/SS diversity ($M = L$) is employed in contrast to not employing any diversity at all. The performance is further increased significantly, by three orders of magnitude, for selection diversity of order $M = 2L$, i.e., $M = 4$. The performance increases sharply by another five orders of magnitude as M increases to $3L$, i.e., $M = 6$. On the other hand, the irreducible BER of the IWC scheme in Building 2 is reduced, by only a factor of two, as M increases from 1 to L . The performance is further increased, by four orders of magnitude, for selection diversity of order $M = 2L$. As M increases to $3L$ the performance increases sharply by another six orders of magnitude. To contrast

the irreducible BER in both buildings, we observed that for all cases the scheme performs better in the open office environment than in the classical building. The difference in performances of the scheme in both buildings increases as higher order of diversities are employed. The difference is about one order of magnitude when no diversity is employed and increases to about four order of magnitude for selection diversity of $M = 6$.

Fig. 5.3 Performance of DS/SS-MA system with DPSK modulation and selection diversity (Gold Code, $N = 127$, $T = 64$ kb/s, $K = 8$ users, $L = 3$ paths, $T_m = 250$ ns).

Figure 5.3 above illustrates the BER performance of the DS/SS-MA system for data rate of $T = 64$ kb/s and $K = 8$ users for a maximum multipath delay spread of 250 ns. The maximum number of paths that can be resolved is 3 in this situation. We see that the irreducible BER is approached at a SNR of 20 dB. When DS/SS diversity ($M = L$) is employed, the irreducible BER performance of the IWC scheme in Building 1 is observed to improve by a factor of four. This performance is further increased significantly, by two orders of magnitude, for selection diversity of order $M = 2L$, i.e., $M = 6$. The performance improves by another three orders of magnitude as M increases to $3L$, i.e., $M = 9$. On the other hand, the irreducible BER of the IWC scheme in Building 2 is reduced, by only a

small amount as M increases from 1 to L . However, as M increases to $2L$ the performance increased significantly by two orders of magnitude. As M increases to $3L$ the performance increases by another three orders of magnitude. To contrast the irreducible BER of the scheme in both buildings, we observed that for all cases the scheme performs better in the open office environment than in the classical building as in the previous case study. The difference here in this case study is about a factor of four when no diversity is employed and decrease to less than a factor when DS/SS diversity is used. The difference increases to about a factor of five when $M = 6$ and increases to about a factor of six for selection diversity of $M = 9$.

Figure 5.4 illustrates the BER performance of the DS/SS-MA system for data rate of $T = 144$ kb/s and $K = 8$ users for a maximum multipath delay spread of 100 ns. From the figure, we observed that the plot is identical to Figure 5.2. We expect this as the maximum number of paths that can be resolved is 2 which is identical to that resolved for the case shown in Figure 5.2. Obviously, the results are sensitive to L , the number of multipaths, which is a function of the chip rate and the delay spread.

Fig. 5.4 Performance of DS/SS-MA system with DPSK modulation and selection diversity (Gold Code, $N = 127$, $T = 144$ kb/s, $K = 8$ users, $L = 2$ paths, $T_m = 100$ ns).

Figure 5.5 shows the BER performance of the DS/SS-MA system for data rate of $T = 144$ kb/s and $K = 8$ users for a maximum multipath delay spread of 250 ns.

Fig. 5.5 Performance of DS/SS-MA System with DPSK modulation and selection diversity (Gold Code, $N = 127$, $T = 144$ kb/s, $K = 8$ users, $L = 5$ paths, $T_m = 250$ ns).

The maximum number of paths that can be resolved is 5 in this situation. We observed that the irreducible BER is approached at a SNR of 25 dB. The irreducible BER of the IWC scheme in Building 1 is better, by a factor of seven, when DS/SS diversity ($M = L$) is employed in contrast to not employing any diversity at all. The performance is further increased by an order of magnitude for selection diversity of order $M = 2L$, i.e., $M = 10$. The performance increases by another two orders of magnitude as M increases to $3L$, i.e., $M = 15$. The irreducible BER of the IWC scheme in Building 2 is reduced, by only a factor of four as M increases from 1 to L . The performance is increased significantly, by an order of magnitude, for selection diversity of order $M = 2L$. As M increases to $3L$ the performance increases by another two orders of magnitude. To compare the irreducible BER of the scheme in both buildings, we observed that the scheme performs better, as expected, in the open office environment than in the classical building in the case of no

diversity at all. However, with selection diversity of $M \geq 5$ the scheme performs slightly better in Building 1 than in Building 2. The difference here in this case study is about a factor of two or less when no diversity is employed and decreases to less than a factor when selection diversity is used.

Based on the results shown in Figures 5.2, 5.3 and 5.4, we see significant improvement in the BER performance when the system employs selection diversity of order $M \geq 2L$. If the system's specification calls for a typical BER performance of not worse than 10^{-4} , for the cases shown, then a diversity order greater or equal to $2L$ is required. We note that the minimum average received SNR should be greater or equal to 5 dB here.

A summary of the irreducible BER of the DS/SS-MA system as a function of the different data rates for $K = 8$ users and a maximum multipath delay spread of 250 ns is shown in Figure 5.6. We note that the maximum number of paths that can be resolved depends on the data rates. In general, for a fixed multipath delay spread of 250 ns we observed that the performance of the scheme, in either buildings, improve with higher order of diversities. As the data rates increase, this advantage is reduced as shown in Figure 5.6. The main point to be made here is that with selection diversity the difference in performance in both buildings decrease as higher data rates are sought till we obtained almost similar performance at about 120 kb/s. At data rate of 144 kb/s the scheme performs slightly better in Building 1 than in Building 2 with higher order of selection diversity ($M \geq 2L$). We attribute this to the higher random to specular power ratio in Building 1 where we surmised that at higher data rates (the number of multipaths increase for a fixed multipath delay spread) the scheme exploits the advantage of using selection diversity. We conclude that for the scheme to be highly effective, especially in Building 1, space diversity is to be augmented to the DS/SS-MA scheme.

Fig. 5.6. Irreducible BER performance of DS/SS-MA system with DPSK modulation and selection diversity at different data rates (Gold Code, $N = 127$, $K = 8$ users, $T_m = 250$ ns).

5.4.2 Kasami codes results

In this section we discussed the performance of the IWC system where the CDMA codes employed are Kasami codes of length 255. Figures 5.7–12 illustrates the BER performance of the DS/SS-MA system as a function of the average received SNR ($5 \leq \text{Ave SNR (dB)} \leq 45$) for different data rates and time delay spreads. Figures 5.13–14 illustrates the irreducible BER performance of the DS/SS-MA system as a function of the number of multipaths for data rate of 144 kb/s and $K = 15$ users. Figures 5.15–17 illustrates the irreducible BER of the DS/SS-MA system as a function of the number of users for different data rates and $T_m = 250$ ns. Finally, to complete this section, the irreducible BER performance curves, as a function of the different data rates, for a maximum multipath delay spread of 250 ns and $K = 15$ users, is shown in Figure 5.18.

Figure 5.7 shows the BER performance of the DS/SS-MA system for data rate of $T = 32$ kb/s and $K = 15$ users for a maximum multipath delay spread of 250 ns.

Fig. 5.7 Performance of DS/SS-MA system with DPSK modulation and selection diversity (Kasami Code, $N = 255$, $T = 32$ kb/s, $K = 15$ users, $L = 3$ paths, $T_m = 250$ ns).

We note that the maximum number of paths that can be resolved is 3 in this situation. From the figure, we observed that the irreducible BER is approached at a SNR of 30 dB. The irreducible BER of the IVC scheme in Building 1 is reduced, by a factor of seven, when DS/SS diversity ($M = L$) is employed. This performance is further increased significantly, by two orders of magnitude, for selection diversity of order $M = 2L$, i.e., $M = 6$. The performance increases similarly by another two orders of magnitude as M increases to $3L$, i.e., $M = 9$. On the other hand, the irreducible BER performance of the IVC scheme in Building 2 is reduced, by only a factor of four, as M increases from 1 to L . The performance is further increased, by two orders of magnitude, for selection diversity of order $M = 2L$. As M increases to $3L$ the performance increases another two orders of magnitude. To contrast the performance in both buildings, we observed that for all cases the scheme performs better in the open office environment than in the classical building. The difference in performances of the scheme in both buildings increase as higher order of diversities are employed. The difference is about a factor of eight when no diversity is

employed and decrease to about a factor of four to five factors for selection diversity of $M = 9$.

Figure 5.8 illustrates the BER performance of the DS/SS-MA system for data rate of $T = 32$ kb/s and $K = 15$ users for a maximum multipath delay spread of 490 ns. We note that the maximum number of paths that can be resolved is 4 in this situation. From the figure, we observed that the irreducible BER is approached at a SNR of 25 dB. As expected, the irreducible BER of the IWC scheme in Building 1 is reduced, by a factor of seven, when DS/SS diversity ($M = L$) is employed in contrast to not employing any diversity at all. The performance is further increased significantly, by two orders of magnitude, for selection diversity of order $M = 2L$, i.e., $M = 8$. The performance increases increases similarly by another two orders of magnitude as M increases to $3L$, i.e., $M = 12$. On the other hand, the irreducible BER of the IWC scheme in Building 2 is reduced, by only a factor of two, as M increases from 1 to L . The performance is further increased, by two orders of magnitude, for selection diversity of order $M = 2L$. As M increases to $3L$ the performance increases another two orders of magnitude. We observed that for all cases the scheme performs better in the open office environment than in the classical building. We observed that the difference in performances of the scheme in both buildings decrease as higher order of diversities are employed. The difference is about a factor of eight when no diversity is employed and decreases to about a factor of four for selection diversity of $M = 12$.

The primary reason for including this figure is to compare the results with that of the Rayleigh performance curves [4] reproduced in Figure 5.1 (with other parameters remaining the same). In Figure 5.1, we observed that, the irreducible BER is better by a factor of two if the scheme employs CPSK rather than DPSK modulation in the case of no diversity. This factor of improvement remains more or less constant even with selection diversity of order $M = 8$. If we compare Figures 5.1 and 5.8 we observed that the DS/SS-MA scheme performs better in the the buildings characterized by Rician channels than in the building (for sake of argument identified as Building 0) characterized by Rayleigh fading channel. We observed that the difference in the irreducible BERs of the scheme (with DPSK modulation) between Building 1 and Building 0 is about a factor of four in the case of no diversity ($M = 1$) and increases significantly to about an order of magnitude with selection diversity of $M = 8$.

Fig. 5.8 Performance of DS/SS-MA system with DPSK modulation and selection diversity (Kasami Code, $N = 255$, $T = 32$ kb/s, $K = 15$ users, $L = 4$ paths, $T_m = 490$ ns).

The BER performance of the DS/SS-MA system for data rate of $T = 64$ kb/s and $K = 15$ users for a maximum multipath delay spread of 100 ns is shown in Figure 5.9. The maximum number of paths that can be resolved is 2 in this case. From the figure, we observed that the irreducible BER begins at about a SNR of 35 dB. In Building 1, the irreducible BER of the IWC scheme improves by an order of magnitude, when DS/SS diversity ($M = L$) is employed. This performance is further increased significantly, by seven orders of magnitude, for selection diversity of $M = 2L$, i.e., $M = 4$. On the other hand, the irreducible BER of the IWC scheme in Building 2 is reduced, by about a factor of two, as M increases from 1 to L . The performance is further increased tremendously, by ten orders of magnitude, as M increases to $2L$. To contrast the performance in both buildings, we observed that for all cases the scheme performs better in the open office environment than in the classical building. The difference in performances of the scheme in both buildings increase as higher order of diversities are employed. The difference is about a couple of orders of magnitude when no diversity is employed and increases to about four orders of magnitude for selection diversity of $M = 4$.

Fig. 5.9 Performance of DS/SS-MA system with DPSK modulation and selection diversity (Kasami Code, $N = 255$, $T = 64$ kb/s, $K = 15$ users, $L = 2$ paths, $T_m = 100$ ns).

Figure 5.10 illustrates the BER performance of the DS/SS-MA system for data rate of $T = 64$ kb/s and $K = 15$ users for a maximum multipath delay spread of 250 ns. We note that the maximum number of paths that can be resolved is 5 in this situation. From the figure, we observed that the irreducible BER is approached at a SNR of 25 dB. As expected, the irreducible BER of the IWC scheme in Building 1 is reduced, by a factor of six, when DS/SS diversity ($M = L$) is employed in contrast to not employing any diversity at all. The performance is further increased significantly, by two orders of magnitude, for selection diversity of order $M = 2L$, i.e., $M = 10$. The performance increases another two orders of magnitude as M increases to $3L$, i.e., $M = 15$. On the other hand, the irreducible BER of the IWC scheme in Building 2 is reduced, by about a factor of four, as M increases from 1 to L . The performance is further increased tremendously, by two orders of magnitude, for selection diversity of order $M = 2L$. As M increases to $3L$, the performance increases another two orders of magnitude. To contrast the performance in both buildings, we observed that for all cases the scheme performs better in the open office

environment than in the classical building. The difference in performances of the scheme in both buildings increase as higher order of diversities are employed. The difference is about a factor of two when no diversity is employed and increases to about a factor of six for selection diversity of $M = 15$.

Fig. 5.10 Performance of DS/SS-MA System with DPSK modulation and selection diversity (Kasami Code, $N = 255$, $T = 64$ kb/s, $K = 15$ users, $L = 5$ paths, $T_m = 250$ ns).

Figure 5.11 illustrates the BER performance of the DS/SS-MA system for data rate of $T = 144$ kb/s and $K = 15$ users for a maximum multipath delay spread of 100 ns. From the figure, we observed that the plot is identical to Figure 5.8. The maximum number of paths that can be resolved is 4 which is identical to that resolved for the case shown in Figure 5.8.

Fig. 5.11. Performance of DS/SS-MA system with DPSK modulation and selection diversity (Kasami Code, $N = 255$, $T = 144$ kb/s, $K = 15$ users, $L = 4$ paths, $T_m = 100$ ns).

Figure 5.12 illustrates the BER performance of the DS/SS-MA system for data rate of $T = 144$ kb/s and $K = 15$ users for a maximum multipath delay spread of 250 ns. The maximum number of paths that can be resolved is 10 in this case. From the figure, we see that the irreducible BER begins at SNR of about 25 dB. The irreducible BER of the IWC scheme in Building 1 is better, by an order of magnitude, when DS/SS diversity ($M = L$) is employed in contrast to not employing any diversity at all. This performance is increased by another three orders of magnitude, for selection diversity of order $M = 2L$, i.e., $M = 20$. The performance increases further by another three orders of magnitude as M increases to $3L$, i.e., $M = 30$. On the other hand, the irreducible BER of the IWC scheme in Building 2 improves by a factor of three as M increases from 1 to L . This performance is increased significantly, by two orders of magnitude, for selection diversity of order $M = 2L$. As M increases to $3L$ the performance increases by another two orders of magnitude. The performance of the scheme in both buildings, performs better, by a factor of five, in the open office environment than in the classical building in the case

of no diversity at all. However, with selection diversity of $M \geq 10$ the scheme performs better in Building 1 than in Building 2. The difference in these cases is about a factor of one (for $M = 10$) and increases to a factor of six for selection diversity of $M = 30$.

Fig. 5.12. Performance of DS/SS-MA system with DPSK modulation and selection diversity (Kasami Code, $N = 255$, $T = 144$ kb/s, $K = 15$ users, $L = 10$ paths, $T_m = 250$ ns).

Figures 5.13 and 5.14 illustrates the irreducible BER performance of the DS/SS-MA system as a function of the maximum number of multipaths (or function of the amount of time delay spread as shown) for data rate of 144 kb/s and $K = 15$ users. In Figure 5.13, the deterioration in the irreducible BER of the system is gradual as the number of multipaths increase, about 1 to 2 orders of magnitude, as the number of paths increase from 3 to 10 (corresponding to increase from 60 ns to 250 ns multipath delay spread).

Fig. 5.13. Irreducible BER performance of DS/SS-MA system with DPSK modulation and selection diversity at different multipath delay spreads (Kasami Code, $N = 255$, $T = 144$ kb/s, $K = 15$ users).

However, in the “lower limb” region, as shown in Figure 5.14, we observe that the deterioration increases sharply, especially with higher order of diversities, as the number of multipaths increase from 2 to 3, i.e., up to twenty three orders of magnitude. Thus, we can conclude that multipath fading has a deleterious effect on system performance but with further increase in the multipath delay spread (values shown in the inset of plots) the deterioration rate tapers off. Alternately, we can argue that with 3 or less multipaths (corresponding to delay spread of less than 60 ns), the impact of deploying higher order of diversities is very great.

Fig. 5.14. Irreducible BER performance of DS/SS-MA system with DPSK modulation and selection diversity at different multipath delay spreads or small number of paths. (Kasami Code, $N = 255$, $T = 144$ kb/s, $K = 15$ users)

Figure 5.15 illustrates the irreducible BER performance of the DS/SS-MA system as a function of the maximum number of users that it can accommodate at information data rate of $T = 32$ kb/s based on a maximum multipath delay spread of 250 ns. The maximum number of paths that can be resolved is 3. In reality, we note that the system degrades softly as the number of users increase which is typical of DS/SS-MA system. From the figure, we observed that with DS/SS diversity and at least two antennae diversity the system can accommodate up to a maximum of 15 users based on a specification that the irreducible BER will not be worse than 10^{-5} .

Fig. 5.15. Irreducible BER performance of DS/SS-MA system with DPSK modulation and selection diversity as a function of the number of users (Kasami Code, $N = 255$, $T = 32$ kb/s, $L = 3$ paths, $T_m = 250$ ns).

Figure 5.16 illustrates the irreducible BER performance of the DS/SS-MA system as a function of the maximum number of users that it can accommodate at information data rate of $T = 64$ kb/s based on a maximum multipath delay spread of 250 ns. The maximum number of paths that can be resolved is 5 in this situation. Similarly as in the previous case, we observed that, with selection diversity of order $M = 2L$, the system can accommodate up to a maximum of 15 users based on a specification that the irreducible BER will not be worse than 10^{-5} .

Fig. 5.16. Irreducible BER performance of DS/SS-MA system with DPSK modulation and selection diversity as a function of the number of users (Kasami Code, $N = 255$, $T = 64$ kb/s, $L = 5$ paths, $T_m = 250$ ns).

Figure 5.17 illustrates the irreducible BER performance of the DS/SS-MA system as a function of the maximum number of users that it can accommodate at information data rate of $T = 144$ kb/s based on a maximum multipath delay spread of 250 ns. The maximum number of paths that can be resolved is 10. Similarly, as in the previous two cases, the system can accommodate up to a maximum of 10 users, with order $M = 2L$, based on a specification that the irreducible BER will not be worse than 10^{-5} .

Fig. 5.17. Irreducible BER performance of DS/SS-MA system with DPSK modulation and selection diversity as a function of the number of users (Kasami Code, $N = 255$, $T = 144$ kb/s, $L = 10$ paths, $T_m = 250$ ns).

Figure 5.18 summarizes the irreducible BER performances, for various data rates and $K = 15$ users at a maximum multipath delay spread of 250 ns. We note that the maximum number of paths that can be resolved depends on the data rates. In general, for a fixed multipath delay spread of 250 ns we observed that the performance of the scheme improve with higher order of diversities, however this advantage is reduced as the data rates increase as shown. The main point to be made here is that with selection diversity the difference in performance in both buildings decrease as higher data rates are sought till we obtained almost similar performance at about 120 kb/s. At data rate of 144 kb/s the scheme performs better in Building 1 than in Building 2 with selection diversity $M \geq 2L$. We attribute this to the higher random to specular power ratio in Building 1 where we surmised that at higher data rates (the number of multipaths increase for a given multipath delay spread) the scheme exploits the advantage of using selection diversity more effectively.

Fig. 5.18. Irreducible BER performance of DS/SS-MA system with DPSK modulation and selection diversity at different data rates. (Kasami Code, $N = 255$, $K = 15$ users, $T_m = 250$ ns.)

5.4.3 Comparison between employing Gold and Kasami codes sequence sets

Table V.2 compares the performance of the DS/SS-MA system using two code sequence sets: $N = 127$ Gold codes and $N = 255$ Kasami sets. It shows the performance results (for $K = 8$ users and multipath delay spread, $T_m = 250$ ns, for two different data rates) of the DS/SS-MA system employing these codes. At 64 kb/s we observe that the DS/SS-MA system performs slightly better with Kasami codes of periodic length $N = 255$ instead of Gold codes of periodic length $N = 127$. At 144 kb/s we observe that the scheme performs significantly better (viz., by two orders of magnitude in the case of $M = 2L$) in either building environment when it employs the Kasami codes instead of the Gold codes with a shorter periodic length. The penalty is the larger RF bandwidth requirement. However this is compensated by the larger set of individual codes which can accommodate more potential users.

Table V.2 BER performance of DS/SS-MA system for two sets of CDMA codes

$K = 8$ users, $T = 64$ kb/s, $T_m = 250$ ns.			
Building No. (CDMA Code)	BER ($M = 1$)	BER ($M = L$)	BER ($M = 2L$)
<u>Building 1</u>			
Gold $N = 127$	5.479×10^{-3}	1.810×10^{-3}	1.017×10^{-5}
Kasami $N = 255$	5.211×10^{-3}	4.693×10^{-4}	2.208×10^{-6}
<u>Building 2</u>			
Gold $N = 127$	1.658×10^{-3}	1.241×10^{-3}	5.625×10^{-6}
Kasami $N = 255$	5.324×10^{-4}	3.034×10^{-4}	1.374×10^{-6}

$K = 8$ users, $T = 144$ kb/s, $T_m = 250$ ns.			
Building No. (CDMA Code)	BER ($M = 1$)	BER ($M = L$)	BER ($M = 2L$)
<u>Building 1</u>			
Gold $N = 127$	2.814×10^{-2}	5.636×10^{-3}	1.543×10^{-4}
Kasami $N = 255$	3.453×10^{-2}	1.469×10^{-3}	1.892×10^{-6}
<u>Building 2</u>			
Gold $N = 127$	1.219×10^{-2}	6.629×10^{-3}	1.799×10^{-4}
Kasami $N = 255$	9.604×10^{-3}	2.697×10^{-3}	4.077×10^{-6}

VI. SUMMARY AND SUGGESTIONS FOR FURTHER RESEARCH

6.1 SUMMARY

In this thesis we have investigated the performance of a DS/SS-MA scheme in an indoor environment characterized by Rician multipath fading channels. The results of these investigations are summarized below.

By comparing the performance of the scheme under various scenarios we conclude that DS/SS diversity (i.e., $M = L$) improves the performance of the scheme. However, we show that the combination of DS/SS diversity and space selection diversity ($M \geq 2L$) provides greater and significant advantages. In most cases, this selection diversity offers BER performance better than 10^{-4} . These performances are attained even at high data rates of up to 144 kb/s with multipath delay spread of 250 ns (worst case situation) with a CDMA scheme capacity of 15 users. These performance results indicate that the realization of the DS/SS-MA scheme proposed in this study is feasible in the newly proposed CADN architecture.

The results demonstrate that, in most instances, the IWC system has a better performance in the open office environment than in the more classical interior environment. Their performances dovetailed at higher data rates, for $M \geq L$, with the scheme having similar or slightly better performance in the classical building at data rates greater than 120 kb/s. However, it should be noted that these comparisons are made on the basis of a common multipath delay profile. In reality, the multipath delay spread in the classical interior building will be wider than in the open office environment.

We compared the performance of the scheme for two different code sequence sets ($N = 127$ Gold and $N = 255$ Kasami codes) and determine that there is significant advantage in deploying the $N = 255$ Kasami sets. The advantage is particularly prominent at high data rates with significant improvement in performance (viz. two orders of magnitude at 144 kb/s for $M = 2L$ as shown in Table V.2). The penalty for uprating the code rate is a larger RF bandwidth requirement. However, as discussed, this is compensated by the

larger set of individual codes which can accomodate more potential users within the indoor cell site.

The performances of the scheme in the two inbuilding Rician channels were compared with the Rayleigh channel described in [4]. Its shown that the system has a better performance in the Rician channels. The difference is about an order of magnitude for diversity order $M \geq 2L$ as shown in Figures 5.1 and 5.8. We conclude that the Rayleigh model can, in this instance, give a conservative estimate of the IWC system under study as we would expect.

We compiled the performance results for the various case studies based on values of the parameters shown in Table V.1 and runs based on other relevant parameter values as an extension of our study. As a result we were able to provide irreducible BER's performance of the scheme as a function of

- (a.) the number of users (K) as shown in Figures 5.15–17,
- (b.) the number of multipaths L (or the amount of the time delay spread) as shown in Figures 5.13 and 5.14, and
- (c.) the information data rate T (Figures 5.6 and 5.18).

The irreducible BER's values are derived based on the irreducible BER threshold (which approached average SNR values ranging from 20 to 30 dBs). For our purpose, we consider average SNR values of about 20-25 dBs (corresponding to those irreducible BER's) as there is not much difference in performance between average SNR's of 20 and 30 dB. These average SNR values of 20-25 dB are attainable based on system processing gain G_p of 23.5 dB (for $N = 255$ codes). We allowed for system losses which can be recovered, viz. antenna gains, and assumed that average power control will preclude jamming margin (M_j) requirements. As a result of this study, we show that the BER performances of the scheme are sensitive to the following channel parameters

- the multipath delay spread T_m , and thus the number of received multipaths components L , and
- the amount of fading κ .

The parameters (L and T) in items (b) and (c) above, are related to the resolution of the number of multipaths expression given in (4.3). We note that the chip rate is proportional to the information data rate for a given code length. Hence, we show that this is another

one of the advantage of the discrete channel model approach which allow us to extrapolate our results.

As an example, let us consider the DS/SS-MA system (with $N = 255$ Kasami codes, 15 users simultaneously online) operating in the open interior building with a worst case situation corresponding to a delay spread of 250 ns. To achieve a BER performance of 10^{-5} with 64 kb/s data rate access, from Figure 5.10, we observed that this performance is attainable with selection diversity $M \geq 2L$. If the maximum delay spread is actually less than 120 ns, we can show that the data rate access can be uprated to 144 kb/s as shown in Figure 5.13. This extrapolation can be shown to be equally applicable for the scheme operating in the classical interior building.

Similarly, we can extrapolate the results shown in Figure 5.8 which illustrates the performance of the scheme operating in both buildings with a multipath delay spread of 490 ns with users having 32 kb/s data rate access. We can obtain the same performances by increasing the data rates to 144 kb/s, as shown in Figure 5.13, if the maximum multipath delay spread is actually less than 100 ns. Since we compared the results between the Rayleigh and Rician channels at 32 kb/s, the comparison results given in the previous chapter are equally applicable in this example, i.e., between Rayleigh ($T = 32$ kb/s, $T_m = 490$ ns) and Rician ($T = 144$ kb/s, $T_m = 100$ ns) channels.

In summarizing our conclusions, we note that we limit our diversity studies to the combination of DS/SS and space diversity. We did not consider other diversity schemes for the following reason. We have shown that space diversity reception when operating in tandem with DS/SS modulation produce significant improvement in the performance of the DS/SS-MA scheme with BER better than 10^{-4} at data rates of 144 kb/s with two antenna diversity. Furthermore, in practice, space diversity is more efficient and is always less complex than maximal ratio combining especially when two antennae is configured for diversity reception.

6.2 SUGGESTIONS FOR FURTHER RESEARCH

Our investigations centred on two code lengths ($N = 127$ and 255) based on practical code rate of 50—100 MHz (hardware limitations) and assuming that 80 MHz of RF bandwidth is allocated at 900 MHz range spectrum. This limits our processing gain to

less than 23.5 dB. Assuming that there are no hardware limitations and RF bandwidth allocations are relaxed, CDMA codes with length greater than 255 are required for further study. This will maximize the processing gain and hence increase system performance and capacity.

In situations where the *Doppler* effects are predominant, we should take into account the manner in which it impacts on the phase characteristics of the Rician channel model. In our study, the phase of the specular component of the Rician fading channel is assumed to be random. This is a reasonable assumption as suggested in [26]

Further work on other binary signalling scheme, such as BPSK, can be contemplated since the infrastructure for deriving the performance and models of a DS/SS-MA system, for any binary signalling scheme, are in place as a result of this thesis. Bandwidth-efficient schemes such as MSK (minimum shift keying) should be considered for increasing code rate (with increase system capacity in mind) and hence maximizing the processing gain to enhance system performance.

APPENDIX A

This Appendix contains the listing of the program "HD" based on the Direct Monte Carlo Simulation Technique.

```

C-----
C  PROGRAM: DEVELOPED BY E.C.LIM ,VERSION 2.0C
C           UNIVERSITY OF OTTAWA,ELEC. ENGINEERING, AUG'87
C
C  PURPOSE: SIMULATE DIRECT SEQUENCE SPREAD SPECTRUM WITH DPSK
C           MODULATION AND CODE DIVISION MULTIPLE ACCESS FOR
C           INDOOR WIRELESS COMMUNICATIONS.
C           DIRECT MONTE CARLO SIMULATION TECHNIQUE
C-----
C
C-----VARIABLES DEFINITIONS-----
C
C  NK   :SYSTEM CAPACITY (K USERS)
C  NL   :NO. OF PATHS (L)
C  NCG  :CODE LENGTH (N)
C  NB   :NO OF BITS REQD FOR ANALYSIS
C  NOTME:NO. OF TIMES OR ITERATIONS FOR ANALYSIS
C  NSNR :NO. OF SNR VALUES REQUIRED FOR ANALYSIS
C  NSRA :NO. OF SHIFT REGISTERS (CODES' GENERATION)
C  NTPA :TAP COEFFICIENTS OF SHIFT REGISTERS (CODES' GENERATION)
C  ALDGS: INITIAL LOADINGS (CODES' GENERATION)
C  NSRB :NO. OF SHIFT REGISTERS (USERS' BITS GENERATION)
C  NTPA :TAP COEFFICIENTS OF SHIFT REGISTERS (USERS'BITS GENERATION)
C  BLDGS: INITIAL LOADINGS (USERS'BITS GENERATION)
C  DB   :DIFFERENTIALLY ENCODED DATA
C  TB   :INFORMATION DATA RATE(1/BRT)
C  TC   :CHIP RATE (1/CPRT)
C  TM   :MAX. MULTIPATH DELAY SPREAD IN NSECS
C  VAR  :RANDOM PATH POWER COMPONENT
C  RHO  :AVERAGE PATH POWER COMPONENT
C  AMPL :TRANSMITTED SIGNAL AMPLITUDE TO BE REPLACED
C        FOR DIFFERENT AVERAGE SNR REQUIREMENTS.
C  NMR  :NO. OF ANTENNAE/SPACE DIVERSITY=(NMR-1)
C  NML  :NO. OF RESOLVED MULTIPATHS/S.S. DIVERSITY
C  MD   :NO. OF SELECTION DIVERSITY ( MD=NML*(NMR-1) )
C
C  EB   :ENERGY PER BIT
C  AWGN :NOTE QUAD. COMP ARE ZERO MEAN G.R.V. WITH VARIANCES NOT
C        No NORMALIZED TO UNITY.
C-----
C
C  INTEGER NK,NL,NCG,NB,NOTME,MD,NMR,NSNR,NSRA,NSRB,NTPA,NTPB
C
C  PARAMETERS FOR GOLD CODES
C
C  PARAMETER(NK=8,NL=4,NCG=127,NB=22000,MD=20,
C  1NSRA=14,NSRB=15,NTPA=14,NTPB=15)
C  INTEGER A(NCG,NK),B(NB,NK),DB(NB,NK)
C  INTEGER ALDGS(NSRA,NK),BLDGS(NSRB,NK)
C  INTEGER IATAPS(NTPA,NK),IBTAPS(NTPB,NK)
C

```

```

C      PARAMETERS FOR KASAMI CODES
C
C      PARAMETER(NK=14,NL=5,NCG=255,NB=22000,MD=20,
C 1NSRA=12,NSRB=15,NTPA=12,NTPB=15)
C      INTEGER DB(NB,NK),A(NCG,NK)
C      REAL BETMAX(MD),BET(NL,NK),TAU(NL,NK),PHI(NL,NK)
C      REAL R1K(NL,NK),RH1K(NL,NK)
C      INTEGER IDX(NL,NK)
C      COMPLEX GN(NB)
C      REAL N0
C
C      DOUBLE PRECISION DSEED
C      DSEED=123457.D0
C
C-----INITIALIZE-----
C
C
C
C      WRITE(8,9005)
C 9005 FORMAT(' CASE STUDY # ')
C      READ(8,*) ICASE
C
C      WRITE(8,9105)
C 9105 FORMAT(' # OF SNR STUDIES ')
C      READ(8,*) NSNR
C      NSNR=5
C
C      FOR DIVERSITY PURPOSE
C      WRITE(8,9115)
C 9115 FORMAT(' NO. OF ORDER OF DIVERSITIES FOR ANALYSIS ')
C      READ(8,*) NMR
C
C      NMR=4
C
C      FOR TOTAL NUMBER OF TIMES NB BITS REQUIRED FOR ANALYSIS
C      WRITE(8,9118)
C 9118 FORMAT(' NO. OF TIMES NB BITS REQUIRED FOR ANALYSIS ')
C      READ(8,*) NBT
C
C      WRITE(8,9125)
C 9125 FORMAT(' INPUT DATA RATE IN kB/S ,VAR,ANX,ANY ')
C      READ(8,*) BRT,VAR,ANX,ANY
C      READ(8,*) BRT,VAR
C      ANX=SQRT(2.0)/2.0
C      ANY=ANX
C      WRITE(8,9825)
C 9825 FORMAT(' MAX M/PATH TIME DELAY SPREAD IN NSECS')
C      READ(8,*) TM
C
C      WRITE(8,9135)
C 9135 FORMAT(' INPUT # OF USERS AND # OF PATHS ')
C      READ(8,*) NMK,NML
C

```

```

      TB=1.0/BRT/1000.0
      CPRT=FLOAT(NCG)*BRT
      TC=1.0/CPRT/1000.0
      NO=1.0
C
C      DETERMINE AVERAGE PATH POWER?
C
      RHO=VAR+0.5*(ANX**2+ANY**2)
C
C-----INITIAL OUTPUT SECTION-----
C
      WRITE(6,8105) ICASE
8105  FORMAT(' CASE STUDY NUMBER      = ',I8)
C
      WRITE(6,8125) BRT
8125  FORMAT(' INFO DATA RATE        = ',F8.1,' Kbits/s ')
C
      CPMRT=CPRT/1000.0
      WRITE(6,8135) CPMRT
8135  FORMAT(' CHIP RATE              = ',F10.6,' MHz ')
C
      WRITE(6,8138) RHO,ANX,ANY
8138  FORMAT(' AVERAGE PATH POWER = ',F8.6,' X-MEAN= ',F8.6,
&' Y-MEAN = ',F8.6)
C
C-----
C
C      READ INPUT DATA TO GENERATE INFORMATION BITS RANDOMLY FOR
C      K USERS
C
C      READ(5,*) ( (BLDGS(I,J),I=1,NSRB),J=1,NK )
C      READ(5,*) ( (IBTAPS(I,J),I=1,NTPB),J=1,NK )
C
C      READ INPUT DATA TO GENERATE DIRECT SEQUENCE I.E. CHIP SEQUENCE
C      PER GOLD CODE PROPERTIES
C
C      READ(5,*) ( (ALDGS(I,J),I=1,NSRA),J=1,NK )
C      READ(5,*) ( (IATAPS(I,J),I=1,NTPA),J=1,NK )
C
C-----CALL SUBROUTINES-----
C
C      CALL PNSBG (NK,NSRB,NTPB,NB,IBTAPS,BLDGS,B)
C      CALL DIFENC (NK,NB,B,DB)
C      CALL PNSBG (NK,NSRA,NTPA,NCG,IATAPS,ALDGS,A)
C      CALL PATHS (TC,TM,NML)
C
C      READ INPUT CHIP SEQUENCE FOR NK USERS
C
      DO 83 J=1,NCG
83      READ(4,*) (A(J,K),K=1,NK)
C      DO 84 J=1,NCG
C84      WRITE(7,1108) (A(J,K),K=1,NK)

```

```

C      READ INPUT BINARY INFORMATION DATA FOR NK USERS
887    FORMAT(' INPUT BINARY DATA FOR NK USERS ')
C      WRITE(7,887)
C
C      DO 87 J=1,NB
C87    READ(3,*) (DB(J,K),K=1,NK)
C      DO 88 J=1,NB
C88    WRITE(7,1108) (DB(J,K),K=1,NK)
1108   FORMAT(2X,14(I2,', '))
C
C-----
C
C      CALL CHANN(DSEED,TB,TC,BET,BMAX,TAU,IDX,PHI,NL,NK,MD,VAR,ANX,ANY)
C
C      CALL APERC(TC,TAU,IDX,A,R1K,RH1K,NL,NK,NCG)
C
C
C      CALL AWGN(DSEED,NB,GN,N0,TB)
C
C
C      WRITE(6,114)
114    FORMAT('0',3H.K.,3H.L.,5H.BETA,5H..PHI,10H...TAU....,4H.IDX,
&12H....R1K.....,12H.....RH1K...)
C
C      WRITE(6,113)
113    FORMAT(52(1H-))
C
C      DO 1110 K=1,NK
C      DO 2220 L=1,NL
C      WRITE(6,115) K,L,BET(L,K),PHI(L,K),TAU(L,K),IDX(L,K),
&R1K(L,K),RH1K(L,K)
2220   CONTINUE
1110   CONTINUE
115    FORMAT(2I3,2F5.2,E10.3,I4,2(2X,E10.3))
C
C-----
C
C      NM=1
C      DO 3000 IM=1,NMR
C
C
C      CALL SELDIV(NM,MD,NL,NK,BET,BMAX)
C
C      INITIAL SNR VALUE (IN DB) CONSIDERED
C
C      SNR=5.0
C
C      DO 2000 JJ=1,NSNR
C
C      CALL POWER (SNR,N0,TB,RHO,AMPL)
C
C

```

```

      CALL HARDEC (NB,NBT,NK,NL,AMPL,BET,PHI,DB,R1K,RH1K,GN,
1N0,TB,DSEED)
C
C      REWIND DATA
      REWIND (3)
C
      DSEED=123457.D0
C
      SNR=SNR+10.0
C
2000  CONTINUE
C
      WRITE(8,*) NM
      NM=NML*IM
C
3000  CONTINUE
C
      STOP
      END
C
      END
C
C-----
C

```

```

C-----SUBROUTINE HARDEC-----
C
C   PURPOSE: MAIN SUBROUTINE TO EMULATE DEMODULATION AND DECISION
C             SEE SECTION 4.2 OF THESIS
C-----
C
C   SUBROUTINE HARDEC(NB,NBT,NK,NL,AMPL,BET,PHI,DB,R1K,RH1K,GN,
C   IN0,TB,DSEED)
C
C   INTEGER DB(NB,NK),NDB
C   REAL BET(NL,NK),PHI(NL,NK)
C   REAL R1K(NL,NK),RH1K(NL,NK)
C   DOUBLE PRECISION GX1,GX2,GY1,GY2,DSEED
C   COMPLEX Z1,Z2,GN(NB)
C
C   TPER=0.0
C   TSPER=0.0
C
C   DO 2000 I=1,NBT
C   ERROR=0.0
C   COUNT=0.0
C
C   READ FURTHER INPUT BINARY INFORMATION DATA FOR NK USERS
C   WRITE(7,887)
C887  FORMAT(' INPUT BINARY DATA FOR NK USERS ')
C
C   DO 87 J=1,NB
C87   READ(3,*) (DB(J,K),K=1,NK)
C
C   DO 88 J=1,NB
C88   WRITE(7,1108) (DB(J,K),K=1,NK)
C1108  FORMAT(2X,14(I2,' '))
C
C   CALL AWGN(DSEED,NB,GN,N0,TB)
C
C   DO 1000 J=3,NB
C
C   GX1=0.0
C   GY1=0.0
C   GX2=0.0
C   GY2=0.0
C
C   DO 200 K=1,NK
C   B2=FLOAT(DB(J-2,K))
C   B1=FLOAT(DB(J-1,K))
C   B0=FLOAT(DB(J,K))
C
C   DO 210 L=1,NL
C
C   AC=AMPL*BET(L,K)*COS(PHI(L,K))
C   AS=AMPL*BET(L,K)*SIN(PHI(L,K))
C   WRITE(6,*)AC,AS

```

```

C
GX1=GX1 + AC*( B1*R1K(L,K) + B0*RH1K(L,K) )
GY1=GY1 + AS*( B1*R1K(L,K) + B0*RH1K(L,K) )
C
GX2=GX2 + AC*( B2*R1K(L,K) + B1*RH1K(L,K) )
GY2=GY2 + AS*( B2*R1K(L,K) + B1*RH1K(L,K) )
C
C   WRITE(6,1205) GX1,GY1,GX2,GY2
C
210  CONTINUE
C
200  CONTINUE
C   WRITE(6,1205) GX1,GY1,GX2,GY2
C
GX1=GX1+REAL(GN(J))
GY1=GY1+AIMAG(GN(J))
GX2=GX2+REAL(GN(J-1))
GY2=GY2+AIMAG(GN(J-1))
C
Z1=CMPLX(GX1,GY1)
Z2=CMPLX(GX2,GY2)
1205  FORMAT('0',8(1X,E9.2))
C
C
DECS=REAL( Z1*CONJG(Z2) )
C   WRITE(6,1215) J,DB(J,1),DECS
1215  FORMAT('0',2I8,2X,E10.3 )
C
NDB=DB(J,1)*DB(J-1,1)
IF(NDB.EQ.1) THEN
IF(DECS.LT.0.0) ERROR=ERROR+1.0
ELSE
IF(DECS.GT.0.0) ERROR=ERROR+1.0
ENDIF
COUNT=COUNT+1.0
C
1000  CONTINUE
C
PERR=ERROR/COUNT
SPER=(PERR)**2
TPER=TPER+PERR
TSPER=TSPER+SPER
C   WRITE(6,1285) PERR
1285  FORMAT('0',' P-ERROR= ',E10.5)
C   WRITE(*,1285) PERR
C   WRITE(6,1295) SPER
1295  FORMAT('0',' SQ P-ERROR= ',E10.5)
C   WRITE(*,1295) SPER
C
C   READ FURTHER INPUT BINARY INFORMATION DATA FOR NK USERS
C
2000  CONTINUE
C

```



```

      DIVNB=FLOAT(NBT)
      BER=TPER/DIVNB
      VBER=TSPEP/DIVNB-(BER)**2
      STD=SQRT(VBER)/(DIVNB-1)
C
      WRITE(6,1385) BER
1385  FORMAT('0',' BER = ',E10.5)
      WRITE(8,1385) BER
      WRITE(6,1485) VBER
1485  FORMAT('0',' VBER = ',E10.5)
      WRITE(8,1485) VBER
      WRITE(6,1585) STD
1585  FORMAT('0',' STD = ',E10.5)
      WRITE(8,1585) STD
      RETURN
      END

```

C

C-----

C

```

C
C-----SUBROUTINE PNSBG-----
C
C   PURPOSE: GENERATES A PN SEQUENCE ( OF DEGREE UP TO NSR)
C             BASED ON TAP POSITIONS AND THE NUMBER OF SHIFT REGISTERS
C             FOR NK USERS.AS A BONUS THIS ROUTINE GENERATES THE CHIP
C             SEQUENCE.
C             SEE SECTION 3.2 FOR GOLD AND KASAMI CODES GENREATION
C
C-----
C
C   SUBROUTINE PNSBG (NK,NSR,NTP,NCL,ITAPS,LDGS,C)
C   INTEGER C(NCL,NK),LDGS(NSR,NK)
C   INTEGER ITAPS(NTP,NK)
C   INTEGER SUM,OUTPUT
C
C   NM1 = NSR - 1
C
C   DO 10 K = 1,NK
C
C   DO 20 IJ = 1,NCL
C
C   SUM = 0
C
C   IT = 0
C
C   30   IT= IT + 1
C       IF(ITAPS(IT,K).EQ.0) GOTO 40
C       SUM = SUM + LDGS(ITAPS(IT,K),K)
C       GOTO 30
C
C   40   OUTPUT=MOD(SUM,2)
C40   OUTPUT=SUM-INT(SUM/2)*2
C
C   REMAP ( 0 -> 1,1 -> -1 )
C
C   IF (OUTPUT.EQ.0) THEN
C   C(IJ,K) = 1
C   ELSE
C   C(IJ,K) = -1
C   END IF
C
C   SHIFT REGISTERS TO RIGHT
C   DO 50 J=1,NM1
C   50   LDGS(NSR+1-J,K)=LDGS(NSR-J,K)
C
C   20   LDGS(1,K) = OUTPUT
C
C   10   CONTINUE
C
C   RETURN
C   END
C
C-----

```

```

C
C-----SUBROUTINE APERC-----
C
C      PURPOSE: COMPUTE APERIODIC CROSSCORRELATIONS  R1K AND
C                RH1K i.e BETWEEN REFERENCE USER AND INTERFERERS
C                SEE SECTION 4.2 OF THESIS
C
C-----
C
C      SUBROUTINE APERC(TC,TAU,IDX,A,R1K,RH1K,NL,NK,NCG)
C
C      REAL TAU(NL,NK),R1K(NL,NK),RH1K(NL,NK)
C      INTEGER A(NCG,NK)
C      INTEGER IDX(NL,NK)
C
C      DO 100 K=1,NK
C
C      DO 200 L=1,NL
C
C      XA=0.0
C
C      NX=IDX(L,K)-NCG
C
C      IF(IDX(L,K).EQ.0) GOTO 20
C
C      DO 10 I=1,NCG+NX
C
10    XA=XA + FLOAT( A(I-NX,K)*A(I,1) )
C
20    NX=NX+1
C
C      XB=0.0
C
C      DO 30 I=1,NCG+NX
30    XB=XB + FLOAT( A(I-NX,K)*A(I,1) )
C
C      R1K(L,K)=XA*TC + (XB-XA)*( TAU(L,K)-FLOAT( IDX(L,K) )*TC )
C
C      NX=IDX(L,K)
C      XHA=0.0
C
C
C
C      DO 40 I=1,NCG-NX
C
40    XHA=XHA + FLOAT( A(I,K)*A(I+NX,1) )
C
C      NX=NX+1
C
C      XHB=0.0
C
C      DO 50 I=1,NCG-NX
50    XHB=XHB + FLOAT( A(I,K)*A(I+NX,1) )
C

```

```

C      WRITE(6,1315) K,L,XA,XB,XHA,XHB
1315  FORMAT(2I3,4(2X,F10.1))
C
C      RH1K(L,K)=XHA*TC+(XHB-XHA)*( TAU(L,K)-FLOAT( IDX(L,K) ) *TC )
C
200   CONTINUE
C
100   CONTINUE
C
      RETURN
      END
C
C-----
C

```

```

C
C-----SUBROUTINE DIFENC-----
C
C   PURPOSE: DIFFERENTIAL ENCODING OF DATA BITS
C
C-----
C
C   SUBROUTINE DIFENC (NK,NB,B,DB)
C
C   INTEGER B(NB,NK),DB(NB,NK)
C
C   DO 10 K=1,NK
C
C       DB(1,K)=B(1,K)
C       DO 20 J=2,NB
20      DB(J,K)=DB(J-1,K)*B(J,K)
C
C   10  CONTINUE
C
C       RETURN
C       END
C
C
C-----
C
C

```

```

C-----SUBROUTINE POWER-----
C
C   PURPOSE: EXERCISE AVERAGE POWER CONTROL
C             AS DESCRIBED IN SECTIONS 3.3 AND 4.1 OF THESIS
C
C-----
C
C   SUBROUTINE POWER (SNR,N0,TB,RHO,AMPL)
C   REAL N0
C   EB=N0*(10.0**(0.1*SNR))/(2.0*RHO)
C   AMPL=SQRT(2.0*EB/TB)
C   EBN0DB=10.0*ALOG10(EB/N0)
C   WRITE (6,105) SNR
C   WRITE(8,105) SNR
C   WRITE (6,119) EBN0DB
C   WRITE(8,119) EBN0DB
C   WRITE(8,108) AMPL
C
105  FORMAT( ' SNR   = ',F8.1,' DB ')
119  FORMAT(' EB/NO = ',F8.1,' DB ')
108  FORMAT(' AMPL  = ',F8.1)
C
C
C   PEDPSK=EXP(-SNR)/2.0
C   WRITE(6,1995)PEDPSK,AMPL
1995 FORMAT(' PEDPSK= ',E10.5,' AMPL = ',E10.5)
C   RETURN
C   END
C
C-----
C

```

```

C-----SUBROUTINE PATHS RESOLUTION-----
C
C      PURPOSE: DETERMINE MAX. NO. ON MULTIPATHS
C              SEE CHANNEL MODEL SECTION 4.1.2 OF THESIS
C
C-----
C
C      SUBROUTINE PATHS (TC,TM,NML)
C
C      TM  ( MULTIPATH TIME DELAY SPREAD ) IS IN NANoseconds
C      TCN ( NORMALIZED CHIP RATE )      IS IN NANoseconds
C
C      TCN=TC*(10.0**9.0)
C      NML=INT(TM/TCN)+1
C
C
C      WRITE(6,8139) NML,TM
8139  FORMAT(' MAX. NUMBER OF RESOLVED PATHS IS = ',I4,
&'FOR A M/PATH DELAY SPREAD OF',F6.1,' nsecs ')
C      WRITE(8,8139) NML,TM
C
C      RETURN
C      END
C
C-----
C

```

```

C-----SUBROUTINE CHANN-----
C
C   PURPOSE: PROVIDES CHANNEL CHARACTERISTICS AS DESCRIBED IN
C             REPORT ON CHANNEL MODEL IN SECTION 4.1 OF THESIS
C
C-----
C
C   SUBROUTINE CHANN(DSEED,TB,TC,BET,BMAX,TAU,IDX,PHI,NL,NK,MD,VAR,
C   LANX,ANY)
C
C   REAL BET(NL,NK),TAU(NL,NK),PHI(NL,NK),BMAX(MD)
C   INTEGER IDX(NL,NK)
C   RHO: AVERAGE PATH POWER
C   DOUBLE PRECISION DSEED
C
C   DO 10 K=1,NK
C
C   DO 20 L=1,NL
C
C   BET(L,K)=GRV(DSEED,VAR,ANX,ANY)
C   TAU(L,K)=TB*GGUBFS(DSEED)
C   IDX(L,K)=INT(TAU(L,K)/TC)
C   PHI(L,K)=8.0*ATAN(1.0)*GGUBFS(DSEED)
C
C 20  CONTINUE
C
C 10  CONTINUE
C
C   ASSUMED DESIRED PATH VALUES BELOW
C   TAU(1,1)=0.0
C   IDX(1,1)=0
C   PHI(1,1)=0.0
C   DO 30 M=1,MD
C 30  BMAX(M)=GRV(DSEED,VAR,ANX,ANY)
C
C   SORT BETA VALUES IN ASCENDING ORDER:REF. USER
C
C   DO 301 J=1,MD-1
C
C   DO 401 L=1,MD-J
C   IF(BMAX(L).GT.BMAX(L+1)) THEN
C   TEMP=BMAX(L)
C   BMAX(L)=BMAX(L+1)
C   BMAX(L+1)=TEMP
C   END IF
C 401 CONTINUE
C
C 301 CONTINUE
C
C   RETURN
C   END
C
C-----

```



```

C-----SUBROUTINE SELDIV-----
C
C   PURPOSE: PROVIDES SELECTION DIVERSITY IN CHANNEL MODEL
C
C-----
C
C   SUBROUTINE SELDIV(NM,MD,NL,NK,BET,BMAX)
C   REAL BET(NL,NK),BMAX(MD)
C
C   DO 10 L=1,NL
10  BET(L,1)=BMAX(L)
C
C   TEMP=BET(1,1)
C   BET(1,1)=BMAX(NM)
C   BET(NM,1)=TEMP
C
C   RETURN
C   END
C-----
C

```

```

C-----FUNCTION:CHI-SQUARE DISTRIBUTION-----
C
C      PURPOSE: GENERATES Rician DISTRIBUTION RANDOM VARIABLES
C               AS SUM OF TWO INDEPENDENT GAUSSIAN RANDOM
C               VARIABLES WITH COMMON VARIANCE, VAR, AND MEANS
C               ANX & ANY.
C      NOTE:    THIS FUNCTION IS PROVIDED TO EASE COMPATIBILITY FOR ALL
C               MACHINES. A RANDOM SEED IS REQUIRED ONLY.
C
C-----
C
C      FUNCTION GRV(DSEED,VAR,ANX,ANY)
C
C      DOUBLE PRECISION DSEED
C
C      NX=12
C      DNX=FLOAT(NX)/2.0
C      STD=SQRT( FLOAT(NX)/12.0*VAR )
C
C      X=0.0
C      Y=0.0
C
C      DO 20 INX=1,NX
C      X=X+GGUBFS(DSEED)
C      Y=Y+GGUBFS(DSEED)
20  CONTINUE
C      X=(X-DNX)*STD + ANX
C      Y=(Y-DNX)*STD + ANY
C      GRV=CABS( CMPLX(X,Y) )
C      RETURN
C      END
C-----
C

```

```

C
C
C-----SUBROUTINE AWGN-----
C
C  PURPOSE:  GENERATES COMPLEX AWGN ,QUADRATURE COMPONENTS
C            WITH ZERO MEANS  AND VARIANCES NoTb
C  NOTE:    THIS FUNCTION IS PROVIDED TO EASE COMPATIBILTY FOR ALL
C            MACHINES. A RANDOM SEED IS REQUIRED ONLY.
C-----
C
C      SUBROUTINE AWGN(DSEED,NB,GN,N0,TB)
C      REAL N0
C      COMPLEX GN(NB)
C      DOUBLE PRECISION DSEED
C
C      NX=12
C      ANX=FLOAT(NX)/2.0
C      STD=SQRT(N0*TB)
C
C      DO 10 J=1,NB
C
C      EX=0.0
C      VX=0.0
C
C      DO 20 INX=1,NX
C      EX=EX+GGUBFS(DSEED)
C      VX=VX+GGUBFS(DSEED)
20  CONTINUE
C      GN(J)=CMPLX( (EX-ANX), (VX-ANX) ) *STD
10  CONTINUE
C      RETURN
C      END
C-----
C

```

APPENDIX B

This Appendix contains the listing of the program " BER" based on the Semi-Analytical
with Monte Carlo Simulation Technique.

```

C-----
C  PROGRAM: DEVELOPED BY E.C.LIM ,VERSION 3.0C      ^
C           UNIVERSITY OF OTTAWA,ELEC. ENGINEERING AUG'87      ^
C  PURPOSE: SIMULATE DIRECT SEQUENCE SPREAD SPECTRUM WITH DPSK  ^
C           MODULATION AND CODE DIVISION MULTIPLE ACCESS FOR   ^
C           INDOOR WIRELESS COMMUNICATIONS. RICIAN              ^
C           MULTIPATH FADING SIMULATION.                      ^
C           BER TECHNIQUE.                                     ^
C  INPUT   :                                                  ^
C-----
C
C
C
C-----VARIABLES DEFINITIONS-----
C
C  NK      :SYSTEM CAPACITY (K USERS)
C  NL      :NO. OF PATHS (L)
C  NCG     :CODE LENGTH (N)
C
C  NB      :NO OF BITS REQD FOR ANALYSIS
C  NOTME   :NO. OF TIMES OR ITERATIONS FOR ANALYSIS
C  NSNR    :NO. OF SNR VALUES REQUIRED FOR ANALYSIS
C
C  NSRA    :NO. OF SHIFT REGISTERS (CODES' GENERATION)
C  NTPA    :TAP COEFFICIENTS OF SHIFT REGISTERS (CODES' GENERATION)
C  ALDGS   :INITIAL LOADINGS (CODES' GENERATION)
C
C  NSRB    :NO. OF SHIFT REGISTERS (USERS' BITS GENERATION)
C  NTPA    :TAP COEFFICIENTS OF SHIFT REGISTERS (USERS'BITS GENERATION)
C  BLDGS   :INITIAL LOADINGS (USERS'BITS GENERATION)
C
C  TB      :INFORMATION DATA RATE(1/BRT)
C  TC      :CHIP RATE (1/CPRT)
C  TM      :MAX. MULTIPATH DELAY SPREAD IN NSECS
C
C  VAR     :RANDOM PATH POWER COMPONENT
C  RHO     :AVERAGE PATH POWER COMPONENT
C  AMPL    :TRANSMITTED SIGNAL AMPLITUDE TO BE REPLACED
C           FOR DIFFERENT AVERAGE SNR REQUIREMENTS.
C
C  NMA     :NO. OF ANTENNAE/SPACE DIVERSITY
C  NMR     : (NMA+1) VARIABLE
C  NML     :NO. OF RESOLVED MULTIPATHS/S.S. DIVERSITY
C  MD      :NO. OF SELECTION DIVERSITY (MD=NML*NMA)
C
C  EB      :ENERGY PER BIT
C  AWGN    :NOTE QUAD. COMP ARE ZERO MEAN G.R.V. WITH VARIANCES Not
C           No NORMALIZED TO UNITY.
C-----
C

```

```

      INTEGER NK,NL, NCG, NOTME, MD, NMR, NSNR, NSRA, NTPA
C     PARAMETERS FOR KASAMI CODE
      PARAMETER (NK=15, NL=10, NCG=255,
1MD=40, NMR=4, NMA=3, NSNR=5,
2NSRA=12, NTPA=12)

C
C     PARAMETERS FOR GOLD CODE
C
C     PARAMETER (NK=8, NL=5, NCG=127,
C     1MD=20, NMR=4, NMA=3, NSNR=5,
C     2NSRA=14, NTPA=14)
C
      INTEGER A (NCG, NK)
      INTEGER ALDGS (NSRA, NK)
      INTEGER IATAPS (NTPA, NK)
      INTEGER IDX (NL, NK)

C
      REAL BMAX (NL, NMA), BET (NL, NK), TAU (NL, NK), PHI (NL, NK), TEM (NL)
      REAL R1K (NL, NK), RH1K (NL, NK)
      REAL N0
      REAL PBER (NSNR, NMR), SPER (NSNR, NMR), TPER (NSNR, NMR)
      REAL TSPER (NSNR, NMR), VBER (NSNR, NMR), STD (NSNR, NMR)

C
      DOUBLE PRECISION DSEED
      DSEED=123457.D0

C
C     READ INPUT DATA TO GENERATE DIRECT SEQUENCE I.E. CHIP SEQUENCE
C     PER GOLD/KASAMI CODE PROPERTIES
C
      READ (5, *) ( (ALDGS (I, J), I=1, NSRA), J=1, NK )
      READ (5, *) ( (IATAPS (I, J), I=1, NTPA), J=1, NK )

C
      WRITE (8, 9005)
9005  FORMAT (' CASE STUDY # ')
      READ (8, *) ICASE
      WRITE (8, 9103)
9103  FORMAT (' # OF ITERATIONS ')
      READ (8, *) NOTME

C
C
      WRITE (8, 9125)
9125  FORMAT (' INPUT DATA RATE IN kB/S , VAR, ANX, ANY ')
      READ (8, *) BRT, VAR, ANX, ANY

C
      WRITE (8, 9825)
9825  FORMAT (' MAX M/PATH TIME DELAY SPREAD IN NSECS ')
      READ (8, *) TM

C
      WRITE (8, 9135)
9135  FORMAT (' INPUT # OF USERS ')
      READ (8, *) NMK

```

```

      TB=1.0/BRT/1000.0
      CPRT=FLOAT(NCG)*BRT
      TC=1.0/CPRT/1000.0
C
      NO=1.0
C
C      DETERMINE AVERAGE PATH POWER
C
      RHO=VAR+0.5*(ANX**2+ANY**2)
C
C
C-----INITIAL OUTPUT SECTION-----
C
      WRITE(6,8105) ICASE
8105  FORMAT(' CASE STUDY NUMBER      = ',I8)
C
C
      WRITE(6,8125) BRT
8125  FORMAT(' INFO DATA RATE        = ',F8.1,' Kbits/s ')
C
C
      CPMRT=CPRT/1000.0
      WRITE(6,8135) CPMRT
8135  FORMAT(' CHIP RATE              = ',F10.6,' Mcps ')
C
C
      WRITE(6,8138) RHO,ANX,ANY
8138  FORMAT(' AVERAGE PATH POWER = ',F8.6,' X-MEAN= ',F8.6,
&' Y-MEAN = ',F8.6)
C
C
C-----CALL SUBROUTINES-----
C
      CALL PNSBG (NK,NSRA,NTPA,NCG,IATAPS,ALDGS,A)
C
      CALL PATHS (TC,TM,NML)
C
C
C
      DO 77 I=1,NSNR
C
      DO 777 J=1,NMR
C
      TPER(I,J)=0.0
      TSPER(I,J)=0.0
C
777  CONTINUE
C
77  CONTINUE
C

```

```

C-----
C
C      DO 4000 NT=1,NOTME
C
C
C      CALL CHANN(DSEED,TB,TC,BET,BMAX,TAU,IDX,PHI,NL,NK,MD,VAR,ANX,ANY,
INML,NMA,TEM)
C
C      CALL APERC(TC,TAU,IDX,A,R1K,RH1K,NL,NK,NCG)
C
C      DO 3000 IM=1,2
C
C      INITIAL SNR VALUE CONSIDERED
C
C      SNR=-5.0
C
C      DO 2000 JJ=1,NSNR
C
C      EXERCISE AVERAGE POWER CONTROL
C
C      CALL POWER (SNR,N0,TB,RHO,AMPL)
C
C      WRITE(6,114)
C114  FORMAT(3H.K.,3H.L.,5H.BETA,5H..PHI,10H...TAU....,4H.IDX,
C      &12H....R1K.....,12H.....RH1K...)
C      WRITE(6,113)
C113  FORMAT(54(1H-))
C
C      DO 1110 K=1,NK
C      DO 2220 L=1,NL
C      WRITE(6,115) K,L,BET(L,K),PHI(L,K),TAU(L,K),IDX(L,K),
C      &R1K(L,K),RH1K(L,K)
C2220  CONTINUE
C1110  CONTINUE
C115  FORMAT(2I3,2F5.2,E10.3,I4,2(2X,E10.3))
C
C      CALL BER(NL,NK,NML,NMK,R1K,RH1K,RHO,BET(1,1),AMPL,N0,TB,PER)
C
C      SPER(JJ,IM)=(PER)**2
C      TPER(JJ,IM)=TPER(JJ,IM)+PER
C      TSPER(JJ,IM)=TSPER(JJ,IM)+SPER(JJ,IM)
C      WRITE(6,1285) PER
C1285  FORMAT(' P-ERROR= ',E10.5)
C      WRITE(8,1285) PER
C      WRITE(6,1295) SPER
C1295  FORMAT(' SQ P-ERROR= ',E10.5)
C      WRITE(8,1295) SPER
C
C      SNR=SNR+5.0
C
C      2000  CONTINUE

```



```

C-----
C   WRITE(8,*) NM
C   NM=NML*IM
C
C   IF (IM.GT.1) GOTO 3000
C
C   CALL SELDIV(NML,IM,NL,NK,NMA,BET,BMAX)
C
3000 CONTINUE
C
C   INITIAL SNR VALUE CONSIDERED
C   IF(NOTME.EQ.10)GOTO 4000
C
C   SNR=-5.0
C
C   DO 2200 JJ=1,NSNR
C
C   EXERCISE AVERAGE POWER CONTROL
C
C   CALL POWER (SNR,N0,TB,RHO,AMPL)
C
C   DO 3200 IM=3,NMR
C
C   CALL MCBER(NL,NK,NML,NMK,R1K,RH1K,RHO,BET(1,1),AMPL,N0,TB,PER,
1IM-1)
C
C   SPER(JJ,IM)=(PER)**2
C   TPER(JJ,IM)=TPER(JJ,IM)+PER
C   TSPER(JJ,IM)=TSPER(JJ,IM)+SPER(JJ,IM)
C
3200 CONTINUE
C
C   SNR=SNR+5.0
C
2200 CONTINUE
C
4000 CONTINUE
C
C   DO 9909 IM=1,NMR
C
C   DO 9919 JJ=1,NSNR
C
C   PBER(JJ,IM)=TPER(JJ,IM)/FLOAT(NOTME)
C   VBER(JJ,IM)=TSPER(JJ,IM)/FLOAT(NOTME)-PBER(JJ,IM)**2
C   STD(JJ,IM)=SQRT(VBER(JJ,IM))/(FLOAT(NOTME)-1)
C
C   WRITE(6,3385) PBER(JJ,IM)
3385 FORMAT(' PER = ',E10.5)
C
C   WRITE(8,3385) PBER(JJ,IM)
C

```

```

C      WRITE(6,4485) VBER(JJ,IM)
4485  FORMAT(' VBER = ',E10.5)
C      WRITE(8,4485) VBER(JJ,IM)
C
C      WRITE(6,5585) STD(JJ,IM)
C
C
C
5585  FORMAT(' STD = ',E10.5)
C      WRITE(8,5585) STD(JJ,IM)
C
9919  CONTINUE
C
9909  CONTINUE
C
      DO 6999 I=1,NMR
C
6999  WRITE(6,5995) ( STD(J,I),J=1,NSNR)
C
      DO 5999 I=1,NMR
C
      WRITE(8,5995) ( PBER(J,I),J=1,NSNR)
C
5999  WRITE(6,5995) ( PBER(J,I),J=1,NSNR)
C
5995  FORMAT(5(E10.5,3X))
C
      DO 7999 I=1,NMR
C
      WRITE(6,*) I
C
      DO 7888 J=1,NSNR
C
7888  WRITE(6,6995) PBER(J,I)
C
7999  CONTINUE
6995  FORMAT(E10.5)
C
      DO 7909 I=1,NMR
C
      WRITE(6,*) I
C
      DO 7808 J=1,NSNR
C
7808  WRITE(6,6995) STD(J,I)
C
7909  CONTINUE
C
      STOP
      END
C

```

```

C-----SUBROUTINE BER-----
C
C      PURPOSE: EVALUATE BER EXPRESSION FOR SELECTION DIVERSITY (L.E.'L')
C              SEE SECTION 4.2 OF THESIS FOR BER EXPRESSION
C              AND SECTION 5.2 FOR TECHNIQUE
C-----
C
C      SUBROUTINE BER(NL,NK,NML,NMK,R1K,RH1K,RHO,BETMAX,A,N0,T,PER)
C
C      REAL R1K(NL,NK),RH1K(NL,NK),N0,PER
C      DOUBLE PRECISION U1,U2,U12,BI0,Q
C
C      AX1SQ=0.0
C      AXH1SQ=0.0
C      AX1PR=0.0
C      U2=0.0
C      U12=0.0
C
C      IF (NML.LT.2) GOTO 999
C
C      DO 10 L=2,NML
C
C      AX1SQ=AX1SQ + R1K(L,1)**2 * RHO
C      AXH1SQ=AXH1SQ + RH1K(L,1)**2 * RHO
C      AX1PR=AX1PR + R1K(L,1)*RH1K(L,1)*RHO
C      WRITE(6,111) R1K(L,1),RH1K(L,1),RHO,AX1SQ,AXH1SQ,AX1PR
111  FORMAT(6(2X,E10.4))
C
10  CONTINUE
C
999  CONTINUE
C
C      DO 20 K=2,NMK
C
C      DO 30 L=1,NML
C
C      U2=U2 + (R1K(L,K)**2 + RH1K(L,K)**2) * RHO
C      U12=U12 + R1K(L,K)*RH1K(L,K)*RHO
C
C      WRITE(6,222) R1K(L,K),RH1K(L,K),RHO,U2,U12
222  FORMAT(5(2X,E10.4))
C
30  CONTINUE
C
20  CONTINUE
C
C      U2=(U2+AX1SQ+AXH1SQ)*2.0*A*A + 2.0*(N0*T)
C      U1= U2 + 4.0*A*A*AX1PR
C      U12=(U12+AX1PR + AXH1SQ)*2.0*A*A + 2.0*(N0*T)
C      WRITE(6,111) A,N0,T,U1,U2,U12
C      ZM=A*BETMAX*T
C

```

```

      SA=ZM*DABS(1.0/DSQRT(U1) - 1.0/DSQRT(U2))/SQRT(2.0)
      SB=ZM*DABS(1.0/DSQRT(U1) + 1.0/DSQRT(U2))/SQRT(2.0)
C
33   FORMAT(4(2X,E10.4))
C
      CALL QFUNCT(SA,SB,Q)
C
      CALL BESI(SA*SB,0,BI0,IER)
C
      WRITE(6,33) SA,SB,Q,BI0
C
      PER=Q - 0.5 *( 1.0 + U12/DSQRT(U1*U2) ) *
1 BI0 * EXP( -(SA*SA+SB*SB) /2.0 )
C
      WRITE(6,1385) PER
1385 FORMAT(' PER = ',E10.5)
C
      WRITE(8,1385) PER
C
      RETURN
      END
C
C-----
C

```

```

C-----SUBROUTINE MCBER-----
C
C   PURPOSE: EVALUATE BER EXPRESSION FOR SELECTION DIVERSITY FOR
C             MULTICHANNEL SIGNALLING, I.E., G.E. '2L'
C             SEE SECTION 4.2 OF THESIS FOR BER EXPRESSION
C             AND SECTION 5.2 FOR TECHNIQUE
C-----
C
C
C
C
C   SUBROUTINE MCBER(NL,NK,NML,NMK,R1K,RH1K,RHO,BETMAX,A,N0,T,PER,
1MJ)
C
C   REAL R1K(NL,NK),RH1K(NL,NK),N0,PER
C   REAL SMA,SMB,ABM,EAB,BIO,UMU,PMU,TMU
C   DOUBLE PRECISION U1,U2,U12,BI0,Q
C   AX1SQ=0.0
C   AXH1SQ=0.0
C   AX1PR=0.0
C   U2=0.0
C   U12=0.0
C
C   IF (NML.LT.2) GOTO 999
C
C   DO 10 L=2,NML
C
C     AX1SQ=AX1SQ + R1K(L,1)**2 * RHO
C     AXH1SQ=AXH1SQ + RH1K(L,1)**2 * RHO
C     AX1PR=AX1PR + R1K(L,1)*RH1K(L,1)*RHO
C
C     WRITE(6,111) R1K(L,1),RH1K(L,1),RHO,AX1SQ,AXH1SQ,AX1PR
111  FORMAT(6(2X,E10.4))
C
C   10  CONTINUE
C
C   999  CONTINUE
C
C     DO 20 K=2,NMK
C
C       DO 30 L=1,NML
C
C         U2=U2 + (R1K(L,K)**2 + RH1K(L,K)**2) * RHO
C         U12=U12 + R1K(L,K)*RH1K(L,K)*RHO
C
C         WRITE(6,222) R1K(L,K),RH1K(L,K),RHO,U2,U12
222  FORMAT(5(2X,E10.4))
C
C   30  CONTINUE
C
C   20  CONTINUE
C

```

```

      U2=(U2+AX1SQ+AXH1SQ)*2.0*A*A + 2.0*(N0*T)
      U1= U2 + 4.0*A*A*AX1PR
      U12=(U12+AX1PR + AXH1SQ)*2.0*A*A + 2.0*(N0*T)
C
C      WRITE(6,111) A,N0,T,U1,U2,U12
C
      ZM=A*BETMAX*T
C
      SA=ZM*DABS(1.0/DSQRT(U1) - 1.0/DSQRT(U2))/SQRT(2.0)
      SB=ZM*DABS(1.0/DSQRT(U1) + 1.0/DSQRT(U2))/SQRT(2.0)
C
      UMU=U12/DSQRT(U1*U2)
      PMU=1.+UMU
      TMU=1.-UMU
C
C      SMA=SA*FLOAT(MJ)
      SMB=SB*FLOAT(MJ)
      ABM=SMA*SMB
      EAB=EXP(-(SMA*SMA+SMB*SMB)/2.)
C
      EVALUATE FIRST TERM
C
      CALL QFUNCT(SMA,SMB,Q)
C
      PE1=Q
C      WRITE(6,2114)
C114  FORMAT(' PE1/Q= ' )
C      WRITE(6,213)PE1
C
      COMPUTE ZEROth ORDER/ MODIFIED BESSEL FUNCTION OF FIRST KIND
C
      CALL BESI(ABM,0,BI,IER)
C
      BI0=BI
C      WRITE(6,2214)
C214  FORMAT(' BI0= ' )
C      WRITE(6,213)BI0
C213  FORMAT(6(2X,E10.4))
C
      EVALUATE SECOND TERM
C
      SUM2=BI0*EAB
      PE2=SUM2
C
C      WRITE(6,2215)
C215  FORMAT(' PE2= ')
C      WRITE(6,213)PE2
C      EVALUATE THIRD TERM
C
      SUM3=0.0
C

```

```

DO 330 K=1,MJ
C
SUM3=FLOAT( MCOM(2*MJ-1,K-1) ) *( (PMU/TMU)**(K-1) ) + SUM3
C
C
C WRITE(6,2216)
C216 FORMAT(' SUM3,PMU,TMU= ')
C WRITE(6,213)SUM3,PMU,TMU
C
C
C
C330 CONTINUE
PE3=SUM2/( (2.0/TMU)**(2*MJ-1) ) * SUM3
C
C WRITE(6,2217)
C217 FORMAT(' PE3= ')
C WRITE(6,213)PE3
C
C
C EVALUATE FOURTH TERM
C
SUM5=0.0
C
DO 440 N=1,MJ-1
C
BAN=(SMB/SMA)**N
ABN=(SMA/SMB)**N
C
C WRITE(6,2218)
C218 FORMAT(' BAN,ABN,N= ')
C WRITE(6,213)BAN,ABN
C
SUM4=0.0
C
DO 450 K=1,MJ-N
C
PTUK=(PMU/TMU)**(K-1)
PTUM=(PMU/TMU)**(2*MJ-K)
COM=FLOAT( MCOM(2*MJ-1,K-1) )
SUM4=COM*(BAN*PTUK-ABN*PTUM) + SUM4
C
C WRITE(6,2219)
C219 FORMAT(' SUM4,PTUK,PTUM= ')
C WRITE(6,213)SUM4,PTUK,PTUM
C
C450 CONTINUE
C
CALL BESI(ABN,N,BI,IER)
C
SUM5=BI*SUM4+SUM5
C
C440 CONTINUE
C

```

```

      PE4=EAB/( (2.0/TMU)**(2*MJ-1) ) *SUM5
      PER=PE1-PE2+PE3+PE4
C
C      WRITE(6,2239)
C239  FORMAT(' SUM5,BI,SUM4,PER= ')
C      WRITE(6,213)SUM5,BI,SUM4,PER
C
106   FORMAT(F8.5)
      RETURN
      END
C
C-----
C

```



```

C
C-----SUBROUTINE PNSBG-----
C
C   PURPOSE: GENERATES A PN SEQUENCE ( OF DEGREE UP TO NSR)
C             BASED ON TAP POSITIONS AND THE NUMBER OF SHIFT REGISTERS
C             FOR NK USERS.AS A BONUS THIS ROUTINE GENERATES THE CHIP
C             SEQUENCE.
C             SEE SECTION 3.2 FOR GOLD AND KASAMI CODES GENREATION
C
C-----
C
C   SUBROUTINE PNSBG (NK,NSR,NTP,NCL,ITAPS,LDGS,C)
C   INTEGER C (NCL,NK) , LDGS (NSR,NK)
C   INTEGER ITAPS (NTP,NK)
C   INTEGER SUM,OUTPUT
C
C   NM1 = NSR - 1
C
C   DO 10 K = 1,NK
C
C   DO 20 IJ = 1,NCL
C
C   SUM = 0
C
C   IT = 0
C
C30  IT= IT + 1
C   IF (ITAPS (IT,K) .EQ.0) GOTO 40
C   SUM = SUM + LDGS (ITAPS (IT,K) ,K)
C   GOTO 30
C
C40  OUTPUT=MOD (SUM,2)
C40  OUTPUT=SUM-INT (SUM/2) *2
C
C   REMAP ( 0 -> 1,1 -> -1 )
C
C   IF (OUTPUT.EQ.0) THEN
C   C (IJ,K) = 1
C   ELSE
C   C (IJ,K) = -1
C   END IF
C
C   SHIFT REGISTERS TO RIGHT
C   DO 50 J=1,NM1
C50  LDGS (NSR+1-J,K) =LDGS (NSR-J,K)
C
C20  LDGS (1,K) = OUTPUT
C
C10  CONTINUE
C
C

```

```

      RETURN
      END
C
C-----
C
C-----SUBROUTINE APERC-----
C
C      PURPOSE: COMPUTE APERIODIC CROSSCORRELATIONS R1K AND
C                RH1K i.e BETWEEN REFERENCE USER AND INTERFERERS
C                SEE SECTION 4.2 OF THESIS
C
C-----
C
C      SUBROUTINE APERC(TC,TAU,IDX,A,R1K,RH1K,NL,NK,NCG)
C
C      REAL TAU(NL,NK),R1K(NL,NK),RH1K(NL,NK)
C      INTEGER A(NCG,NK)
C      INTEGER IDX(NL,NK)
C
C      DO 100 K=1,NK
C
C      DO 200 L=1,NL
C
C      XA=0.0
C
C      NX=IDX(L,K)-NCG
C
C      IF (IDX(L,K).EQ.0) GOTO 20
C
C      DO 10 I=1,NCG+NX
C
C      10  XA=XA + FLOAT( A(I-NX,K)*A(I,1) )
C
C      20  NX=NX+1
C
C      XB=0.0
C
C      DO 30 I=1,NCG+NX
C
C      30  XB=XB + FLOAT( A(I-NX,K)*A(I,1) )
C
C      R1K(L,K)=XA*TC + (XB-XA)*(TAU(L,K)-FLOAT( IDX(L,K) )*TC )
C
C      NX=IDX(L,K)
C      XHA=0.0
C
C
C      DO 40 I=1,NCG-NX
C
C      40  XHA=XHA + FLOAT( A(I,K)*A(I+NX,1) )
C
C
C

```

```

C
    NX=NX+1
C
    XHB=0.0
C
C
    DO 50 I=1,NCG-NX
50    XHB=XHB + FLOAT( A(I,K)*A(I+NX,1) )
C
    WRITE(6,1315) K,L,XA,XB,XHA,XHB
1315  FORMAT(2I3,4(2X,F10.1))
C
C
    RH1K(L,K)=XHA*TC+(XHB-XHA)*( TAU(L,K)-FLOAT( IDX(L,K) ) *TC )
C
200  CONTINUE
C
100  CONTINUE
C
    RETURN
    END
C
C-----
C
C-----SUBROUTINE POWER-----
C
C    PURPOSE: EXERCISE AVERAGE POWER CONTROL
C             AS DESCRIBED IN SECTIONS 3.3 AND 4.1 OF THESIS
C
C-----
C
    SUBROUTINE POWER (SNR,N0,TB,RHO,AMPL)
    REAL N0
    EB=N0*(10.0**(0.1*SNR))/(2.0*RHO)
    AMPL=SQRT(2.0*EB/TB)
    EBN0DB=10.0*ALOG10(EB/N0)
C
    WRITE (6,105) SNR
C
    WRITE(8,105) SNR
C
    WRITE (6,119) EBN0DB
C
    WRITE(8,119) EBN0DB
C
    WRITE(8,108) AMPL
C
105  FORMAT( ' SNR   = ',F8.1,' DB ')
119  FORMAT(' EB/NO = ',F8.1,' DB ')
108  FORMAT(' AMPL  = ',F8.1)
C
C
    PEDPSK=EXP(-SNR)/2.0
C
    WRITE(6,1995) PEDPSK,AMPL
1995  FORMAT(' PEDPSK= ',E10.5,' AMPL = ',E10.5)
    RETURN
    END
C

```

```

C
C-----SUBROUTINE PATHS RESOLUTION-----
C
C      PURPOSE: DETERMINE MAX. NO. ON MULTIPATHS
C              SEE CHANNEL MODEL SECTION 4.1.2 OF THESIS
C
C-----
C
C      SUBROUTINE PATHS (TC,TM,NML)
C
C      TM  ( MULTIPATH TIME DELAY SPREAD ) IS IN NANOSECONDS
C      TCN ( NORMALIZED CHIP RATE )      IS IN NANOSECONDS
C
C      TCN=TC*(10.0**9.0)
C      NML=INT(TM/TCN)+1
C
C
C      WRITE(6,8139) NML,TM
8139  FORMAT(' MAX. NUMBER OF RESOLVED PATHS IS = ',I4,
&'FOR A M/PATH DELAY SPREAD OF',F6.1,' nsecs ')
C      WRITE(8,8139) NML,TM
C
C      RETURN
C      END
C
C-----
C
C-----SUBROUTINE CHANN-----
C
C      PURPOSE: PROVIDES CHANNEL CHARACTERISTICS AS DESCRIBED IN
C              REPORT ON CHANNEL MODEL IN SECTION 4.1 OF THESIS
C
C-----
C
C      SUBROUTINE CHANN(DSEED,TB,TC,BET,BMAX,TAU,IDX,PHI,NL,NK,MD,VAR,
1ANX,ANY,NML,NMA,TEM)
C
C      REAL BET(NL,NK),TAU(NL,NK),PHI(NL,NK),BMAX(NL,NMA),TEM(NL)
C      INTEGER IDX(NL,NK)
C      RHO: AVERAGE PATH POWER
C      DOUBLE PRECISION DSEED
C
C      DO 10 K=1,NK
C
C      DO 20 L=1,NL
C
C      BET(L,K)=GRV(DSEED,VAR,ANX,ANY)
C      TAU(L,K)=TB*GGUBFS(DSEED)
C      IDX(L,K)=INT(TAU(L,K)/TC)
C      PHI(L,K)=8.0*ATAN(1.0)*GGUBFS(DSEED)
C
C      20 CONTINUE
C

```

```

C
10  CONTINUE
C
C    ASSUMED DESIRED PATH VALUES BELOW
    TAU(1,1)=0.0
    IDX(1,1)=0
    PHI(1,1)=0.0
C
    DO 40 L=1,NL
C
    DO 30 M=1,NMA
C
30    BMAX(L,M)=GRV(DSEED,VAR,ANX,ANY)
C
40    CONTINUE
C
    DO 501 M=1,NMA
C
C    SORT BETA VALUES IN ASCENDING ORDER:REF. USER
C    TO SIMULATE PATH SELECTION
C
    DO 301 J=1,NL-1
C
    DO 401 L=1,NL-J
C
    IF (BMAX(L,M).GT.BMAX(L+1,M)) THEN
    TEMP=BMAX(L,M)
    BMAX(L,M)=BMAX(L+1,M)
    BMAX(L+1,M)=TEMP
    END IF
C
C
401  CONTINUE
C
301  CONTINUE
C
501  CONTINUE
C
C    SORT BETA VALUES IN ASCENDING ORDER:REF. USER
C    TO SIMULATE ANTENNAE SELECTION
C
    DO 801 J=1,NMA-1
C
    DO 701 M=1,NMA-J
C
    IF (BMAX(NML,M).GT.BMAX(NML,M+1)) THEN
C
    DO 601 L=1,NL
    TEM(L)=BMAX(L,M)
    BMAX(L,M)=BMAX(L,M+1)
C
C
C

```

```

601   BMAX(L,M+1)=TEM(L)
C
      END IF
C
701   CONTINUE
C
801   CONTINUE
C
      BET(1,1)=BMAX(1,1)
C
      DO 14 L=1,NL
C
        WRITE(8,*) (BMAX(L,M),M=1,NMA)
14    CONTINUE
      RETURN
      END
C
C-----
C
C
C
C-----SUBROUTINE SELDIV-----
C
C   PURPOSE: PROVIDES SELECTION DIVERSITY IN CHANNEL MODEL
C
C-----
C
      SUBROUTINE SELDIV(NML,IM,NL,NK,NMA,BET,BMAX)
      REAL BET(NL,NK),BMAX(NL,NMA)
C
      BET(1,1)=BMAX(NML,IM)
C
      DO 10 L=1,NL
C10   BET(L,1)=BMAX(L)
C
      TEMP=BET(1,1)
C
      BET(1,1)=BMAX(NM)
C
      BET(NM,1)=TEMP
C
      RETURN
      END
C

```

```

C-----
C
C
C
C-----FUNCTION:CHI-SQUARE DISTRIBUTION-----
C
C  PURPOSE: GENERATES RICIAN DISTRIBUTION RANDOM VARIABLES
C            AS SUM OF TWO INDEPENDENT GAUSSIAN RANDOM
C            VARIABLES WITH COMMON VARIANCE, VAR, AND MEANS
C            ANX & ANY.
C  NOTE:     THIS FUNCTION IS PROVIDED TO EASE COMPATIBLTY FOR ALL
C            MACHINES. A RANDOM SEED IS REQUIRED ONLY.
C
C-----
C
C  FUNCTION GRV(DSEED,VAR,ANX,ANY)
C
C  DOUBLE PRECISION DSEED
C
C  NX=12
C  DNX=FLOAT(NX)/2.0
C  STD=SQRT( FLOAT(NX)/12.0*VAR )
C
C  X=0.0
C  Y=0.0
C
C  DO 20 INX=1,NX
C    X=X+GGUBFS(DSEED)
C    Y=Y+GGUBFS(DSEED)
20  CONTINUE
C    X=(X-DNX)*STD + ANX
C    Y=(Y-DNX)*STD + ANY
C    GRV=CABS( CMPLX(X,Y) )
C    RETURN
C  END
C
C-----
C
C
C
C-----SUBROUTINE QFUNCT-----
C
C  PURPOSE: EVALUATE MARCUM Q FUNCTION
C            SEE [31] OF BIBLIGRAPHY
C
C-----
C
C  SUBROUTINE QFUNCT(SA,SB,QFUN)
C
C  DOUBLE PRECISION QFUN,AL(100),BE(100),D(100)
C

```

```

C      EVALUATE Q FUNCTION
C
      P=1000000.0
      AL(1) = 0.0
      BE(1) = 0.0
      BE(2) = 0.5
C
      IF (SA.GE.SB) GOTO 210
C
      AL(2) = 1.0
      D(3) = SA/SB
C
      GOTO 220
C
210    AL(2) = 0.0
      D(3) = SB/SA
C
220    DO 225 J = 3,100
      AL(J) = D(J) + 2.0*FLOAT(J-2)*AL(J-1)/SA/SB + AL(J-2)
      BE(J) = 1.0 + 2.0*FLOAT(J-2)*BE(J-1)/SA/SB + BE(J-2)
C
      IF (BE(J).GE.P) GOTO 230
C
      D(J+1) = D(J)*D(3)
C
      WRITE(7,39)
39     FORMAT(' ALPHA,BETA,D ')
C      WRITE(7,111) AL(J),BE(J),D(J+1)
111    FORMAT(3(E13.7,3X))
      GOTO 225
C
230    QFUN= AL(J)/BE(J)/2.0*EXP(-(SA-SB)**2/2.0)
      GOTO 250
C
225    CONTINUE
C
250    IF (SA.GE.SB) QFUN=1.0-QFUN
C
      WRITE(7,75) QFUN,SA,SB
C
75     FORMAT(' Q VALUE IS ',3(2X,E13.7))
C
      RETURN
      END
C

```



```

C-----
C
C
C-----SUBROUTINE BESI -----
C
C      PURPOSE: COMPUTE BESSEL FUNCTIONS OF NTH ORDER
C
C-----
C
C      SUBROUTINE BESI(X,N,BI,IER)
C
C      CHECKS FOR ERRORS IN N AND X AND EXIT IF ANY ARE PRESENT
C
C      IER=0
C      BI=1.0
C      IF (N)150,15,10
C
C10      IF(X)160,20,20
C15      IF(X)160,17,20
C17      RETURN
C
C      DEFINE TOLERANCE
C20      TOL=1.E-6
C      IF ARGUMENT GT 12 AND GT N, USE ASYMPTOTIC FORM
C      IF (X-12.)40,40,30
C30      IF(X-FLOAT(N))40,40,110
C
C      COMPUTE FIRST TERM SERIES AND SET INITIAL VALUE OF THE SUM
C40      XX=X/2.
C50      TERM=1.0
C      IF(N)70,70,55
C55      DO 60 I=1,N
C      FI=I
C      IF (ABS(TERM)-1.E-68)56,60,60
C56      IER=3
C      BI=0.0
C      RETURN
C60      TERM=TERM*XX/FI
C70      BI=TERM
C      XX=XX*XX
C      COMPUTE TERMS, STOPPING WHEN ABS(TERM) LE ABS(SUM OF TERMS)
C      TIMES TOLERANCE
C      DO 90 K=1,1000
C      IF (ABS(TERM)-ABS(BI*TOL))100,100,80
C80      FK=K*(N+K)
C      TERM=TERM*(XX/FK)
C90      BI=BI+TERM
C      RETURN BI AS ANSWER
C100      RETURN
C      X GT 12 AND X GT N, SO USE ASYMPTOTIC APPROXIMATION
C110      FN=4*N*N

```

```

      IF (X-170.0)115,111,111
111   IER=4
      RETURN
115   XX=1./(8.*X)
      TERM=1.
      BI=1.

C
C
C
C

      DO 130 K=1,30
      IF (ABS (TERM) -ABS (TOL*BI))140,140,120
120   FK=(2*K-1)**2
      TERM=TERM*XX*(FK-FN)/FLOAT(K)
130   BI=BI+TERM
C     SIGNIFICANCE LOST AFTER 30 TERMS, TRY SERIES
      GO TO 40
140   PI=3.141592653
      BI=BI*EXP(X)/SQRT(2.*PI*X)
      GO TO 100
150   IER=1
      GO TO 100
160   IER=2
      GO TO 100
      END

C
C-----
C
C
C
C
C-----FUNCTION:COMBINATION-----
C
C   PURPOSE: COMPUTE COMBINATIONS
C
C   NOTE:   THIS FUNCTION IS PROVIDED TO EASE COMPATIBILTY FOR ALL
C           MACHINES. NO LIBRARY ROUTINE IS REQUIRED
C
C-----
C
C
C
C   FUNCTION MCOM(N1,N2)
C
C   IF (N2.GT.0) THEN
C     GOTO 110
C   ELSE
C     MCOM=1
C   ENDIF
C   RETURN
C
C
110   N1N2=N1-N2

```

```

      N1N2F=1
C
      DO 10 J=1,N1N2
10    N1N2F=N1N2F*J
C
      NF=1
      DO 20 K=1,N2
20    NF=NF*K
      N2F=NF
C
      DO 30 L=N2+1,N1
30    NF=NF*L
C
      MCOM=NF/N2F/N1N2F
      RETURN
      END
C
C-----
C

```

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