AN ANALYSIS OF THE *INFOPAD* DOWNLINK

by

John Camagna

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ELECTRONICS RESEARCH LABORATORY

College of Engineering
University of California, Berkeley
94720

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CHAPTER 1

Introduction

The advancement of communication and computer technology has fundamentally and irrevocably altered the way people exchange information, do business, learn and make decisions. This will undoubtedly continue. Two general trends have been developing in the latter half of the twentieth century. First, the need for mobile communication so that people are no longer bound to specific location. Second, the need for high processing power to sift through the huge amount of information currently available.

The INFOPAD project at the University of California at Berkeley hopes to usher in a new era of mobile, high speed information processing by developing an indoor mobile multimedia terminal [1,2]. The INFOPAD will serve as a multipurpose telephone, computer and pager in one platform

While both high speed computers and mobile communication systems have been developed, combining the functionality of the two has yet to be done. The requirement of mobility greatly limits the size and weight of the mobile terminal. As a result, battery size and hence power consumption must be kept to a minimum.

Our solution is to provide low power mobile terminals linked to stationary computers that will perform the bulk of the computation. It is apparent that the design of the data link between the fixed computer and the mobile terminal is critical to the success of the project. The data needed by the mobile terminal or personal communication system (PCS) is estimated to be around 2 Mbits per second. This includes both video and system protocol information. The data transmitted by the terminal will consist of mainly pen and keystroke information and is estimated to be around 64 Kbits/sec.

It is expected that in a typical indoor environment there will be a user every 4-10 square meters[3]. With QAM symbol encoding the spectral density needed for the downlink will be 0.5bit/Hzm. To accommodate this high spectral density, large rooms will be divided into cells with a single base station within each cell. Fiber LANS are expected to have the flexibility to move data between the fixed computers and the base stations within each cell

Goals and Objectives

Our goal will be to analyze the performance of the wireless portion of the link from the base station to the multiple mobile terminals. Once this accomplished, a system level design for the base station transmitter and the mobile terminal will be developed. Analysis of an asynchronous Spread Spectrum communication in an indoor environment has been done by Misser [4,5] and Kavehrad [6,7] for Ricean and Rayleigh environments respectively. With a more realistic model for Ricean fading we will show that better performance is expected for broadcast transmission than that predicted for asynchronous transmission. In addition our analysis will include the effects of inter-cell interference.

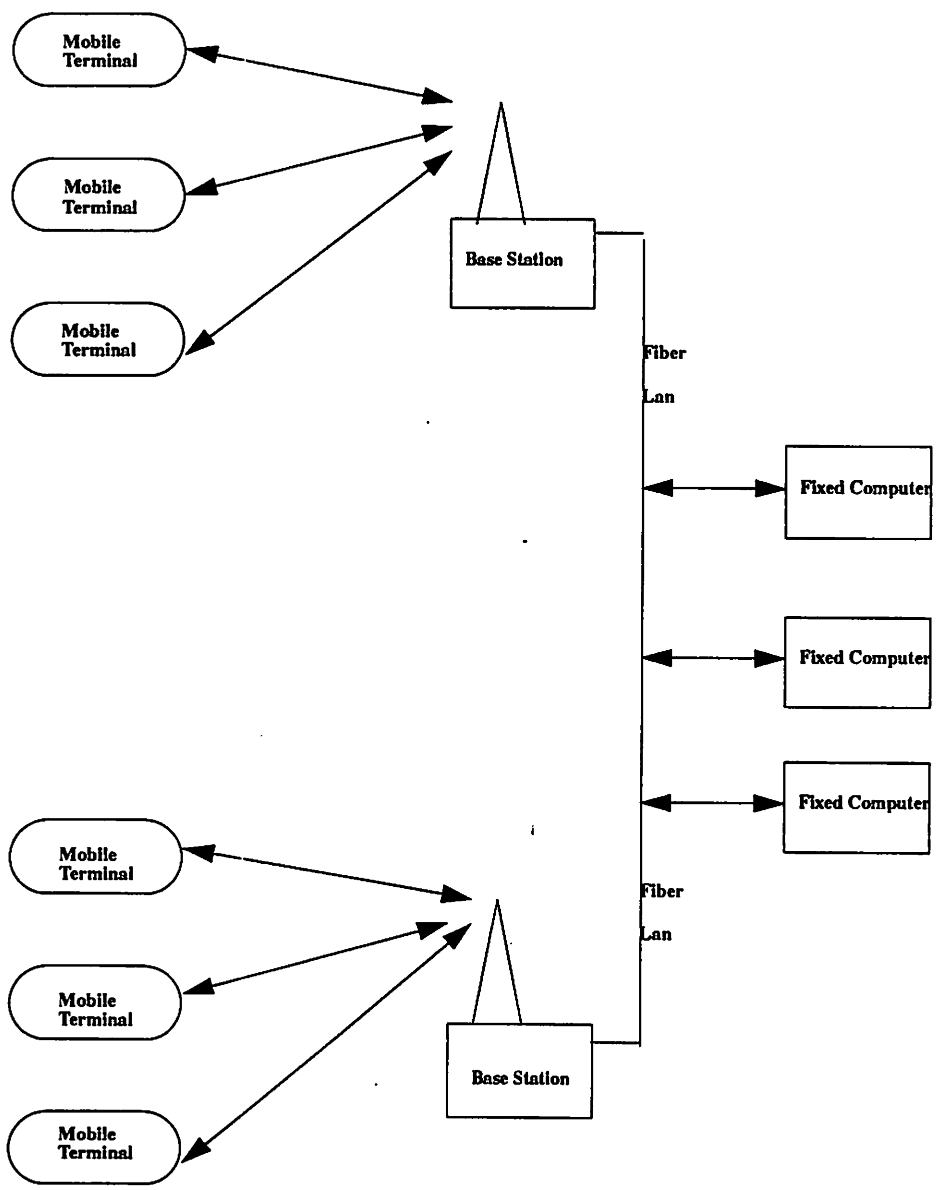
CHAPTER 2

System Overview

A block diagram of the overall system is shown in Figure 1. The bulk of the computation is done by the stationary computer which is connected to the mobile terminal via a fiber and wireless link. The mobile terminal is primarily a user interface which displays video signals and transmits the key and pen strokes of the user.

The data between the base station and the fixed computer can be carried by a fiber LAN system. This thesis will focus on the wireless link from the base station to the mobile terminals.

FIGURE 1. Overview of the INFOPAD Wireless Link



Indoor environment

In an indoor environment there are multiple signal paths between the base station and the mobile terminal. Reflections from various objects within the room constructively and destructively interfere at the antenna making the received signal amplitude vary dramatically or "fade" as the position of the mobile terminal changes. The base station is fixed. Furthermore, the fading occurs approximately independently for position shifts greater than half a wavelength of the carrier. The INFOPAD carrier frequency will be on the order of 1 GHz, so in our case this is a shift of about 0.15 meters.

If no line of sight (LOS) exists between the base station and the mobile terminals then the fading can be modeled as a Rayleigh fading. A more typical case occurs when a direct line of sight exists. In this case, the fading can be characterized as Ricean [8,9]. The Rayleigh case will provide a conservative lower bound on performance while the Ricean case will provide a more realistic scenario.

The environment will change slowly relative to the symbol rate. If we assume that people move on the order of 1m/s and that the carrier is 1GHz then the coherence time of the channel $\Delta (t_c)$ is approximately

$$\Delta(t_c) \cong \frac{0.5\lambda}{1\frac{m}{s}} = 0.15s \qquad \text{(EQ 1)}$$

With a symbol rate of 1MBaud we have a symbol duration of $T_s = 1\mu s$ and

$$\Delta(t_c) \cong 0.15s \gg T_s = 1\mu s \qquad \text{(EQ 2)}$$

As a result of [11] we can assume that the channel is fixed during one symbol time. An important measure of the multipath environment is the multipath delay spread. If $\Phi(t)$ is the delay power spectrum of the channel then the rms delay spread is given by

$$T_{ds} = \sqrt{\frac{\int_0^\infty t^2\Phi(t)\,dt}{\int_0^\infty \Phi(t)\,dt} - \left(\frac{\int_0^\infty t\Phi(t)\,dt}{\int_0^\infty \Phi(t)\,dt}\right)^2} \qquad \text{(EQ 3)}$$

The inverse of the delay spread is known as the coherence bandwidth F_c and is given by [12].

$$F_c = \frac{1}{2\pi T_{ds}} \qquad \text{(EQ 4)}$$

Signals separated by more than the coherence bandwidth will fade approximately independently of each other. One of the reasons spread spectrum is used in a multipath environment is that it spreads the signal over a bandwidth much greater than the coherence bandwidth. The probability that the larger channel will fade completely is much smaller than the probability that the narrowband channel will fade.

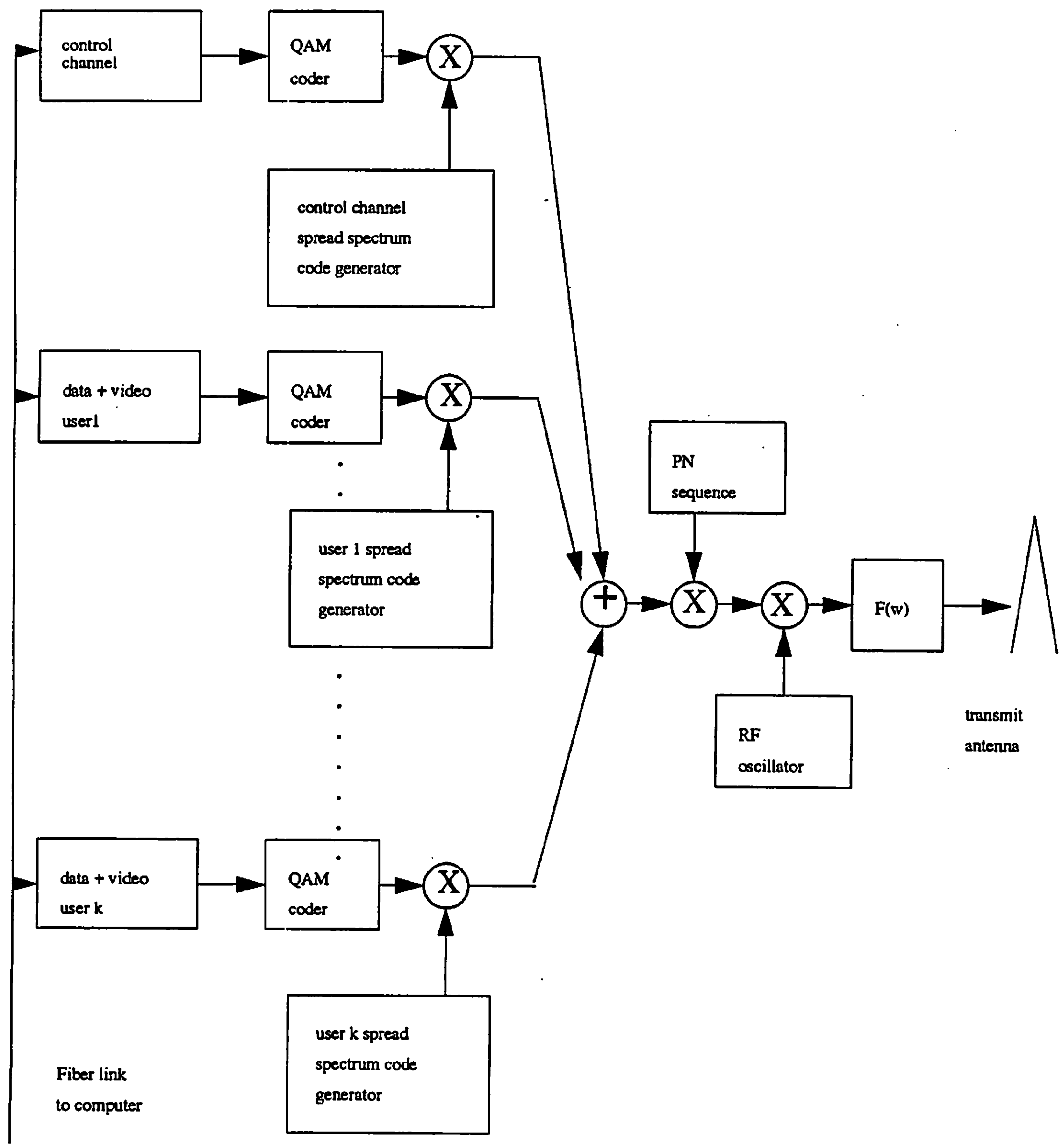
Base Station

A single base station in each cell will transmit to multiple users within the cell. In addition to the user data a common control channel or pilot tone is provided which fulfils two useful functions. About 50% of the time a known or reference signal is transmitted which will allow the mobile terminals to adapt to the channel and track changes in the environment. The other portion of the time the channel will carry control data. This will include a system time, code assignments for each user within the cell and the address for the current and adjacent cells. This information is necessary to handle cell hand-off. The control data will also include protocol information for the uplink, for instance acknowledging mobile terminal transmissions.

A block diagram is shown in Figure 2. Data from each user and the pilot tone first passes through a 4-QAM coder. At this point the symbol rate is 1MBaud. The data is then multiplied by a specific spread spectrum code. This code runs at the chip rate T_c and spreads the data over a much larger bandwidth than theoretically needed.

INFOPAD will use Walsh sequences for spreading codes which have the desirable property of complete orthogonality at a zero phase offset. It will be shown that this is a great advantage for broadcast transmission in a Ricean fading environment. Unfortunately, the auto correlation properties are very poor for other phase offsets. This means that the performance will be poor in a multipath environment. In addition to this, the Walsh codes do not spread the signal uniformly over the spectrum. To counteract this, the data streams are added together and multiplied by a pseudo random sequence which runs at the chip rate. This improves both the spectral properties of the transmitted signal as well as the autocorrelation properties for non zero phase shifts. Different base stations use the same PN sequence but at different phase offsets. This helps isolate the cells from one another.

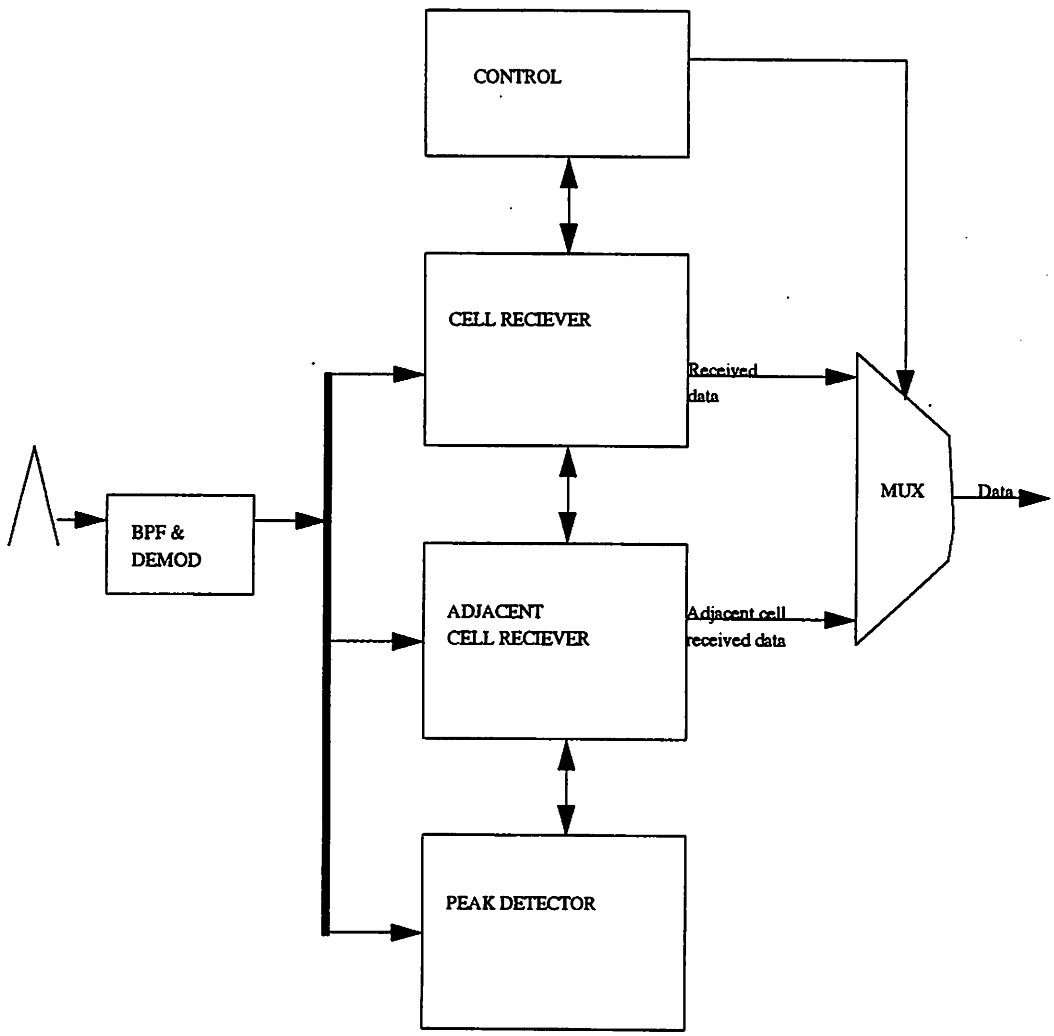
FIGURE 2. **Base Station Transmitter**



Downlink Receiver

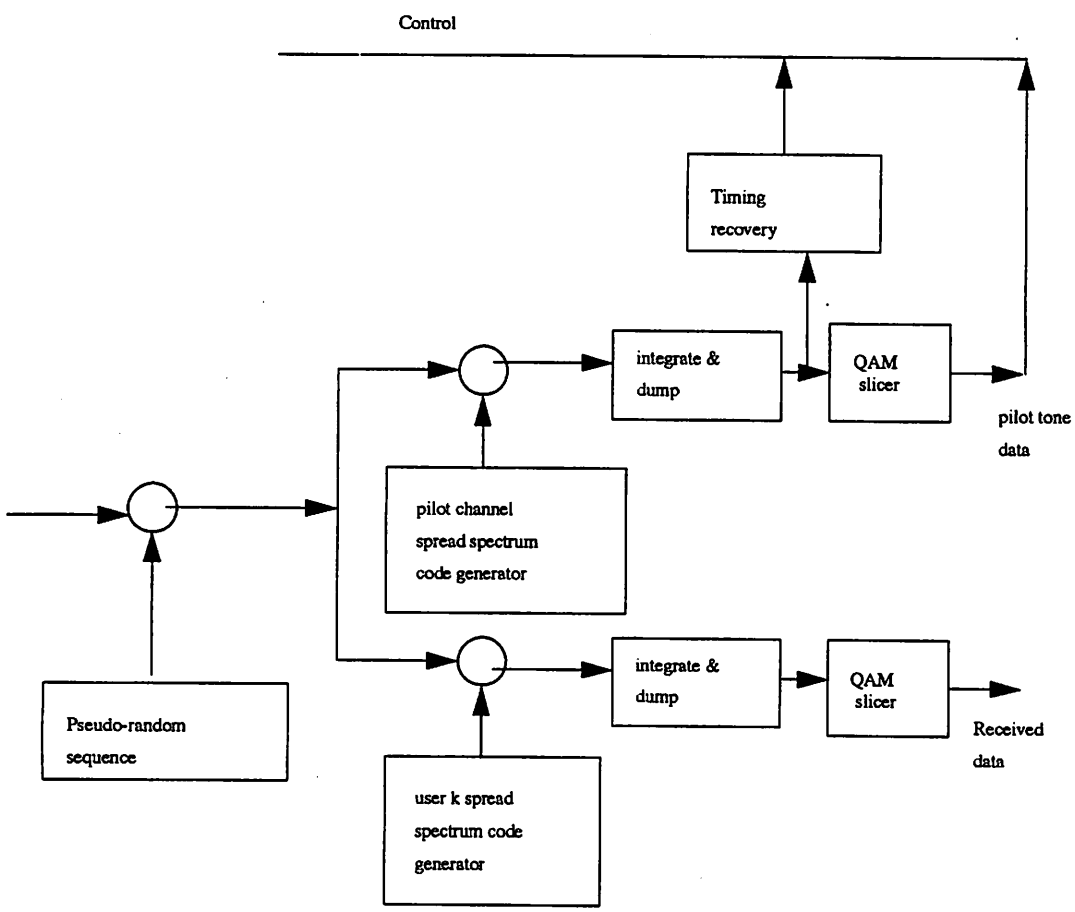
A block diagram of the downlink receiver in the mobile terminal is shown in Figure 3. The received signal passes through a band pass filter and is demodulated down to base-band. The cell receiver de-spreads both the user and control channel signals and outputs the data. As this is happening, the peak detector is continually scanning the power level of adjacent cells to determine if the mobile terminal is near a cell boundary. When it is near a cell boundary, it notifies the system via the uplink. The base station in the adjacent cell then transmits identical data. The adjacent cell receiver begins to track the data from the adjacent cell. While the PCS remains at the cell boundary the receiver dynamically switches between the strongest signal. This diversity reception decreases the chances of missing the cell hand-off. Once safely in the adjacent cell, the cell receiver takes over the reception.

FIGURE 3. **Downlink receiver in Mobile Terminal**



One of our primary concerns is with the trade-offs involved in building the cell receiver. A block diagram is shown below in Figure 4. The data is multiplied by the pseudo random sequence to reverse the original scrambling. The phase is determined using the fixed data within the control channel. This can be quite difficult in practice [13]. The control channel and user data are recovered by correlating with their respective codes and then running the output through a QAM slicer. Timing recovery is accomplished by adjusting the clock phase to maximize the absolute value of the output of the control channel correlator. This is a relatively simple receiver structure that does not take advantage of the diversity available, later it will be shown that more complex receiver structures will improve performance.

FIGURE 4. **Basic Cell Receiver For User *K***



CHAPTER 3

Modeling and Analysis

Transmission Model

The bit stream for the *k*th (*k*=1,...*K*) user is

$$b_k(t) = \sum_{i = \infty}^{\infty} b_{ki} P_{T_s}(t - iT_c),$$

(EQ 5)

where the rectangular pulse P_{T_s} is defined as

$$P_{T_s}(t) = \begin{pmatrix} 1, 0 \le t < T_s \\ 0, \text{elsewhere} \end{pmatrix}$$

(EQ 6)

and the user data b_{ki} has values

$$b_{ki} \in \{\pm 1 \pm j\} \quad \text{(EQ 7)}$$

The data bits are multiplied first by a Walsh spreading code and then by a pseudo random sequence which both run at the chip rate. Multiplication by both the Walsh code and the pseudo random sequence is equivalent to multiplication by a single sequence for each user. This sequence is given by

$$a_k(t) = \sum_{i=-\infty}^{\infty} a_{ik} P_{T_c}(t - iT_c) \quad \text{(EQ 8)}$$

Where the chips a_{ki} have values

$$a_{ki} \in \{\pm 1\} \quad \text{(EQ 9)}$$

Ignoring the filtering details, the transmitted signal for all users is then

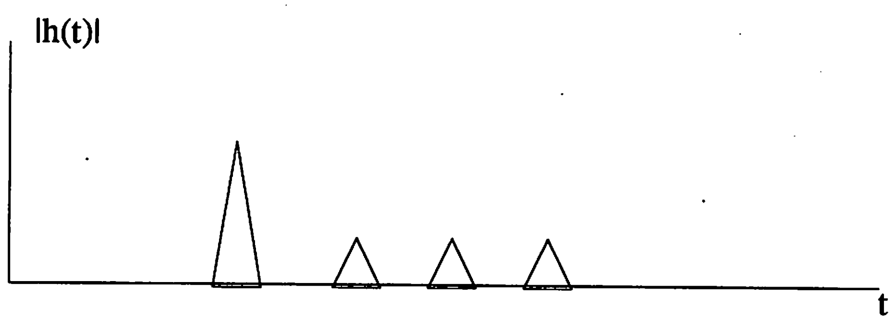
$$S(t) = \sum_{k=1}^{K} a_k(t) b_k(t) A e^{j\omega t} \quad \text{(EQ 10)}$$

where A is the amplitude of each signal and ω is the radial carrier frequency.

Channel Model

In general the impulse response spreads the energy of the signal over time. This is illustrated in Figure 5.

FIGURE 5. Typical Impulse Response of an Indoor Multipath Channel



In the INFOPAD system, the received signal will be band-limited to eliminate out of band noise, demodulated and sampled at the chip rate T_c. As a result, we can model the baseband channel as the sum of discrete impulses each separated by a chip time [11]

$$h(t) = \sum_{l=1}^{L} \beta_i \varsigma(t - lT) e^{j\theta_i} \qquad \text{(EQ 11)}$$

β_i, θ_i are random variables independent of one another. The amplitudes β_i 's are assumed to be either Rayleigh or Ricean distributed depending on the model that is used. The phases θ_i in both cases are uniform over the interval $[0, 2\pi]$. The multipath delay spread for the 1GHz range has been measured at somewhere between 20-50ns depending on the room size[10]. If the delay spread is known then the number of resolvable multipath components is given by

$$L = \mathrm{INT}\left(\frac{T_{ds}}{T_c}\right) + 1 \tag{EQ 12}$$

Where INT(x) takes the integer part of x.

Reception Model

After passing through the channel and being demodulated the received signal is

$$R(t) = \frac{A}{2}\sum_{l=1}^{L}\sum_{k=1}^{K}\beta_l e^{j\theta_l} a_k(t - lT_c) b_k(t - lT_c) + N(t) \tag{EQ 13}$$

where $N(t)$ is the filtered gaussian noise. This differs from the asynchronous case considered by Misser and Kavehrad[4,5,6,7]. In the asynchronous case, the signal for each user passes through an independent channel. In this case the received signal is

$$R'(t) = \frac{A}{2}\sum_{l=1}^{L}\sum_{k=1}^{K}\beta_{lk}e^{j\theta_{lk}}a_k(t-T_{lk})b_k(t-T_{lk})+N(t) \quad \text{(EQ 14)}$$

Where the amplitudes $\beta_{11}, \ldots, \beta_{LK}$ are i.i.d and Rayleigh distributed in Kavehrad's model and Ricean distributed in Misser's model. The phases θ_i in both cases are uniform over the interval $[0, 2\pi]$ and i.i.d. The delays are also i.i.d and uniform over the interval $[0, T_s]$.

With the timing recovery circuitry, the receiver will synchronize to one of the multipath components. The user then correlates with the corresponding multiple access (spread spectrum) code. Initially we will assume that the receiver randomly selects a path to synchronize with, so without loss of generality we will consider the case where the first user synchronizes to the first path component. After correlation with the first user's (k=1) spread spectrum sequence $a_1(t)$ the decision variable is

$$\varepsilon = \int_{t=T_1}^{t=T_1+T_b}\frac{A}{2}\sum_{l=1}^{L}\sum_{k=1}^{K}\beta_l e^{j(\theta_l-\theta_1)}a_k(t-lT_c)b_k(t-lT_c)a_1(t)dt+W \quad \text{(EQ 15)}$$

$$\varepsilon = \frac{A}{2}\sum_{l=1}^{L}\sum_{k=1}^{K}\beta_l e^{j(\Theta_l - \Theta_1)}\int_{t=T_1}^{t=T_1+T_b} a_k(t - lT_c)b_k(t - lT_c)a_1(t)dt + W \quad \text{(EQ 16)}$$

W is the filtered gaussian noise with

$$E(W) = 0, \mathrm{VAR}(W) = N_0 T_b \quad \text{(EQ 17)}$$

The variance of the channel noise at the input to the slicer is identical to the case when narrowband transmission is used. We can simplify (16) by noting that Walsh codes are orthogonal to each other at a zero phase offset. As a result,

$$\int_0^{T_b} a_k(t)b_k(t)a_j(t)dt = \begin{pmatrix} T_b b_0^j, i = j \\ 0, i \neq j \end{pmatrix} \quad \text{(EQ 18)}$$

where b_0^j is the current bit of user j. In addition let us define two useful correlation functions

$$R_{ki}(\tau) \equiv \int_0^{\tau} (a_k(t-\tau) \cdot a_i(t))\, dt \qquad \text{(EQ 19)}$$

$$\hat{R}_{ki}(\tau) \equiv \int_{\tau}^{T_b} (a_k(t-\tau) \cdot a_i(t))\, dt \qquad \text{(EQ 20)}$$

and define

$$\Theta_l = \theta_l - \theta_1 \qquad \text{(EQ 21)}$$

The decision variable can now be written as

$$\varepsilon = \beta_1 \frac{AT_b}{2} b_0^1 + \frac{A}{2} \sum_{l=2}^{L} \sum_{k=1}^{K} \beta_l e^{j\Theta_l} \left(b_{-1}^k R_{kl}(lT_c) + b_0^k \hat{R}_{kl}(lT_c)\right) + W \qquad \text{(EQ 22)}$$

Where b_0^j is the current bit of user j and b_{-1}^j is the previous bit of user j.

CHAPTER 4

Rayleigh Fading with Uniform Delay Profile

Gaussian Approximation

In order to explicitly calculate the error performance of the system we must take into account the effect of the channel noise and the interfering multipath components. In the case of Rayleigh fading with uniform delay profile each of the amplitudes of the impulse response are assumed Rayleigh distributed and i.i.d. The in-phase component is then gaussian. M. Kavehrad [6,7] used this to write the decision variable as

$$\varepsilon = \beta_1 \frac{AT_b}{2} \cdot b_0^1 + \zeta \qquad \text{(EQ 23)}$$

where ζ is zero mean and gaussian. Assuming Rayleigh fading and that each of the four possible QAM symbols is equally likely we have

$$E(\zeta) = \frac{A}{2}\sum_{l=2}^{L}\sum_{k=1}^{K} E(\beta_l)\, E(e^{j\Theta_l})\, E(b^k_{-1}R_{k1}(lT_c) + b^k_0\hat{R}_{k1}(lT_c)) + W \quad \text{(EQ 24)}$$

$$E(b^k_{-1}) = E(b^k_0) = 0 \rightarrow E(\xi) = 0 \quad \text{(EQ 25)}$$

For Rayleigh fading the real and imaginary components of ζ are i.i.d. and equal to one half of $VAR(\zeta)$ where

$$VAR(\zeta) = E(|\zeta|^2) = \frac{A^2}{4}\sum_{l=2}^{L}\sum_{k=1}^{K} E(\beta_l^2)\, E|\kappa_{lk}|^2 + E|W|^2 \quad \text{(EQ 26)}$$

or, using the i.i.d properties of the path amplitudes β_i

$$VAR(\zeta) = \frac{A^2}{4}(L-1)KE(\beta^2)\, E|\kappa_{11}|^2 + E|W|^2 \quad \text{(EQ 27)}$$

where we defined the code correlation factor

$$\kappa_{kl} = b^k_{-1}R_{k1}(lT_c) + b^k_0\hat{R}_{k1}(lT_c) \quad \text{(EQ 28)}$$

We interpret this result as follows: the multi-user interference and multi-path self-interference can be added to the channel noise. The variance of the interference term is proportional to the number of users K and the number of resolvable paths minus 1. This occurs because the first paths of all users are orthogonal.

Although Walsh codes are the underlying CDMA code and ensure orthogonality, the addition of the PN sequence makes the sequence pseudo-random. By approximating the cross correlation sequence as random for phases other than zero and assuming equal probabilities of 1,-1 we can express the real part of the cross correlation factor κ_{kl} as

$$\mathrm{Re}(\kappa_{kl}) = \int_0^{T_b} \sum_{i=1}^{N} X_i P_{T_c}(t - iT_c)\, dt,\ l \neq 1 \qquad \text{(EQ 29)}$$

simplifying we have

$$\mathrm{Re}(\kappa_{kl}) = \sum_{i=1}^{N} X_i T_c \qquad \text{(EQ 30)}$$

where the X_i 's are i.i.d. and

$$P((X_i = 1)) = 0.5,\ P((X_i = -1)) = 0.5 \qquad \text{(EQ 31)}$$

With this approximation the second moment of the real part is given by

$$E(\mathrm{Re}(\kappa_{kl})^2) = \sum_{i=1}^{N}\sum_{j=1}^{N} E(X_i X_j T_c^2) = \sum_{i=1}^{N} E(X_i^2 T_c^2) = N T_c^2 = \frac{T_b^2}{N} \quad \text{(EQ 32)}$$

The second moment of the imaginary part is identical to the second moment of the real part. This result is different than the case where the cross correlation is asynchronous. This result is given by reference [14] and is

$$E(\mathrm{Re}(\kappa'_{kl})^2) = \frac{2}{3}\frac{T_b^2}{N} \quad \text{(EQ 33)}$$

With the second moment of the cross correlation factor the variance of the interfering terms is

$$\mathrm{VAR}(\zeta) = 2\left(\frac{AT_b}{2}\right)^2 (L-1)\, K\mathrm{E}(\beta^2)\frac{1}{N} + N_0 T_b \quad \text{(EQ 34)}$$

The conditional signal-to-noise ratio in both the in-phase and quadrature components is

$$\Upsilon|\beta_1 = \frac{\left(\frac{\beta_1 A T_b}{2}\right)^2}{\left(\frac{AT_b}{2}\right)^2 (L-1)\, K\mathrm{E}(\beta^2)\frac{1}{N} + \frac{N_0 T_b}{2}} \quad \text{(EQ 35)}$$

$$\Upsilon | \beta_1 = \frac{\beta_1^2 E_s}{(L-1) K \left(\frac{1}{N}\right) \overline{E_s} + N_0} \qquad \text{(EQ 36)}$$

where the transmitted energy per symbol E_s and average energy per symbol $\overline{E_s}$ are defined as

$$E_s \equiv A^2 T_b, \overline{E_s} \equiv \mathrm{E}(\beta^2) E_b \qquad \text{(EQ 37)}$$

In a cellular system a significant portion of the multi user interference power comes from adjacent cells. For the Qualcomm mobile phone system approximately 35% according to [15]. To our knowledge, comparable indoor measurements have not been taken. We will assume that all cells have exactly the same number of users and that they all transmit at the same power. We will define α as the ratio of the total outer cell interference power to the power transmitted by one cell. The total power transmitted by one cell is $\mathrm{LK}\overline{\mathrm{E}_s}$. As a result, the conditional signal-to-noise ratio is given by

$$\Upsilon | \beta_1 = \frac{\beta_1^2 \mathrm{E}_s}{((1+\alpha) \mathrm{L} - 1) \mathrm{K} \left(\frac{1}{\mathrm{N}}\right) \overline{\mathrm{E}_s} + \mathrm{N}_0} \qquad \text{(EQ 38)}$$

This is different from the expression derived by Kavehrad for the asynchronous non-cellular case. In his case we have

$$\Upsilon'|\beta_1 = \frac{\beta_1^2 E_s}{(LK-1)\left(\frac{1}{N}\frac{2}{3}\right)\overline{E_s} + N_0} \qquad \text{(EQ 39)}$$

The real change in performance will occur in Ricean environments. If we assume that the symbols are gray encoded then the conditional probability of a bit error is

$$P(e|\beta_1) = \frac{1}{2}\operatorname{erfc}\left(\sqrt{\Upsilon|\beta_1}\right) \qquad \text{(EQ 40)}$$

No Diversity

Let us now consider 3 different ways of selecting our decision path. If we choose a path at random then the amplitude of the path is Rayleigh distributed. We can then average the probability of error by the distribution

$$P(e) = \int_0^{\infty} P(e|x) f_x(x)\, dx \qquad \text{(EQ 41)}$$

In the case of random path selection the integral becomes

$$P(e) = \int_0^{\infty} erfc\left(\sqrt{\frac{x^2 E_s}{((1+\alpha)L-1)K(\frac{1}{N})\overline{E_s}+N_0}}\right)\frac{1}{x_0}e^{-\frac{x}{x_0}}dx \qquad \text{(EQ 42)}$$

This integral is evaluated in [6] and is

$$P(e) = f(\Upsilon_0) = \left(1-\sqrt{\frac{\Upsilon_0}{\Upsilon_0+1}}\right)\frac{1}{2} \qquad \text{(EQ 43)}$$

Where the average signal-to-noise ratio is

$$\Upsilon_0 \equiv \frac{\overline{E_s}}{((1+\alpha)L-1)K(\frac{1}{N})\overline{E_s}+N_0} = \mathrm{E}(\Upsilon) \qquad \text{(EQ 44)}$$

We can simplify our expression for the signal to noise ratio if we note that the interference noise is usually much larger than the background noise in our system. In this case we have

$$\Upsilon_0 \cong \frac{N}{((1+\alpha)L-1)K} \qquad \text{(EQ 45)}$$

It is worth noting that increasing the transmitter power does not change the error performance of the system at high signal to noise ratios. Although this may seem counter-intuitive at first, it simply reflects the fact that the interference power increases along with the signal power.

Selection Diversity

A better alternative to the previous method is to select the largest signal path amplitude. This can be done by adjusting the system timing to track the largest correlation peak. It can also be done spatially by adding multiple antennas to the system. The antennas must be separated by at least half a wavelength in order for the signals to be approximately un-correlated. If NA is the number of antennas in the INFOPAD system and L is the number of discrete multipath components then the total number of signals to choose from is given by M = NA*L. M is referred to as the order of diversity. The p.d.f of the largest signal component is given by[4]

$$f_x(x) = M\left(\sum_{k=0}^{M-1}\binom{M-1}{k}\frac{(-1)^k}{x_0}exp\left(-\frac{x}{x_0}(k+1)\right)\right) \quad \text{(EQ 46)}$$

The probability of error is given by.

$$P(e) = \int_0^{\infty} P(e|x)f_x(x)dx \quad \text{(EQ 47)}$$

This integral is evaluated in ref[6]

$$P(e) = M\left(\sum_{k=0}^{M-1}\binom{M-1}{k}\frac{(-1)^k}{k+1}f\left(\frac{\Upsilon_0}{k+1}\right)\right) \quad \text{(EQ 48)}$$

Maximum Ratio Combining

The final case we wish to consider occurs when we take a weighted sum of the M signals available. The optimal reception will occur when the signal to noise ratio of the sum is maximized.

$$\Upsilon = \frac{\left|\mathrm{E}\left(\sum_{i=1}^{M} \alpha_i X_i\right)\right|^2}{\mathrm{VAR}\left(\sum_{i=1}^{M} \alpha_i X_i\right)} \qquad \text{(EQ 49)}$$

If the antennas receive the same average power and the power is distributed equally over the L path components then the variances are equal

$$\mathrm{VAR}\,(X_i) = \mathrm{VAR}\,(X_j),\ \forall i, j \qquad \text{(EQ 50)}$$

and the sum is maximized when

$$\alpha_i = E(X_i)^* = \beta_i e^{-j\theta_i} \qquad \text{(EQ 51)}$$

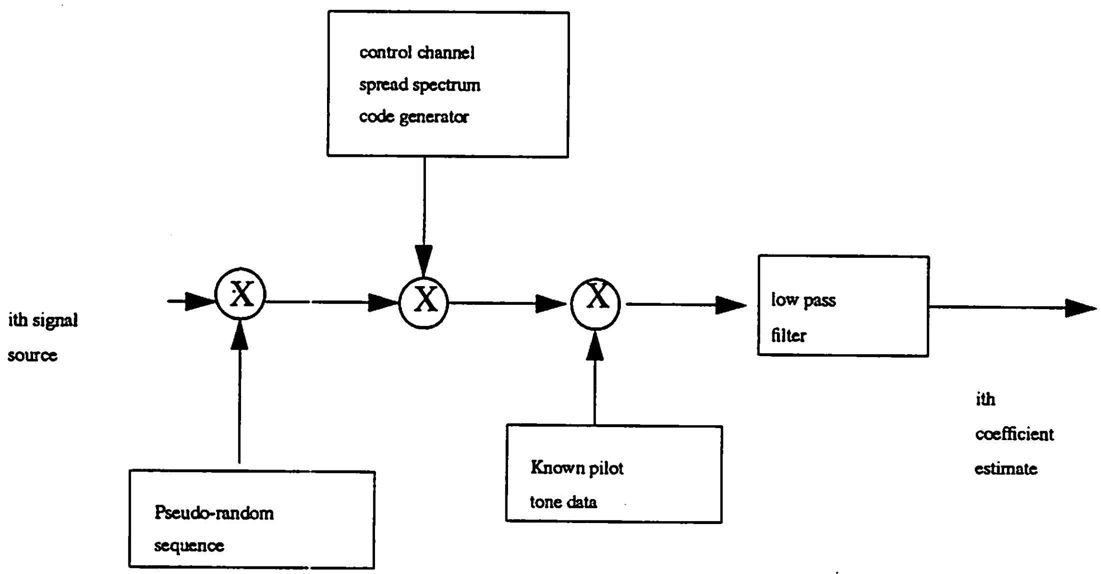
In this case the signal-to-noise ratio becomes

$$\Upsilon | \beta_1, \ldots, \beta_M = \frac{E_s \sum_{i=1}^{M} \beta_i^2}{((1+\alpha)L - 1)\,K\,(\frac{1}{N})\,\overline{E_s} + N_0} \qquad \text{(EQ 52)}$$

This is the optimal SNR that can be obtained. The coefficients can be determined using the known data in the control channel as shown in Figure 6. If each path is independent the probability of error is given by

$$P(e) = (f(\Upsilon_0))^M \left(\sum_{k=0}^{M-1} \binom{M+k-1}{k} \frac{(-1)^k}{k+1} (1 - f(\Upsilon_0))^k \right) \qquad \text{(EQ 53)}$$

FIGURE 6. Coefficient estimator for MRC Reception



CHAPTER 5 Ricean Fading

Channel Model

In his analysis of asynchronous indoor cellular systems Misser [4,5] assumed a channel impulse response of

$$h(t) = \sum_{i=1}^{L} \beta_i \delta(t - T_i) e^{j\theta_i} \qquad \text{(EQ 54)}$$

where the β_i's are assumed to be Ricean and i.i.d. We believe that a more realistic model for the INFOPAD system is

$$h(t) = \beta_d \delta(t) e^{j\theta_d} + \sum_{l=2}^{L} \beta_l \delta(t - lT_c) e^{j\theta_l} \qquad \text{(EQ 55)}$$

Where the channel response consists of a dominant resolvable path and several scattered paths. We believe this to be realistic because typically a LOS will exist between the transmitter and receiver in an indoor environment. Measurements [16] have indicated that the total power in the scattering paths is approximately 6-7 dB less than the power in the dominant path. We will define the total scattering path power to be a fraction γ of the dominant path power.

$$\sum_{i=2}^{L} \mathrm{E}(\beta_i^2) = \gamma \mathrm{E}(\beta_d{}^2) \qquad \text{(EQ 56)}$$

The distribution of the dominant resolvable path has been characterized as Ricean with I on the order of 7-11 dB [8].

$$I = \frac{(\mathrm{E}(\beta_d))^2}{\mathrm{VAR}(\beta_d) + \sum_{i=2}^{L} \mathrm{E}(\beta_i^2)} = \frac{0.5S^2}{\sigma^2} \qquad \text{(EQ 57)}$$

As a small approximation, we will assume that the resolvable path variance is much smaller than the total energy in the scattered paths.

$$I \cong \frac{E(\beta_d)^2}{\sum_{i=2}^{L} E(\beta_i^2)} = \frac{1}{\gamma} \qquad \text{(EQ 58)}$$

With our new model for Ricean fading we cannot assume symmetry of the amplitudes on the multipath components. Rather reference [16] suggests that the dominant resolvable path is significantly larger than the scattering components. As a result we will assume that we synchronize to the dominant path. As in (23) our received signal is

$$\varepsilon = \beta_d \frac{AT_b}{2} b_0^1 + \frac{A}{2} \sum_{l=2}^{L} \sum_{k=1}^{K} \beta_l e^{j\Theta_l} (b_{-1}^k R_{kl}(lT_c) + b_0^k R_{kl}(lT_c)) + w \qquad \text{(EQ 59)}$$

Note that due to the orthogonality of the spreading codes at zero phase offset a large portion of the multi-user power does not interfere with the desired signal. As before we can invoke the central limit theorem and model the interfering terms as gaussian noise.

$$\varepsilon = \beta_1 \frac{AT_b}{2} \cdot b_0^1 + \zeta \qquad \text{(EQ 60)}$$

Assuming independence of the scattering components and that each of the four possible QAM symbols is equally likely we have

$$E(\zeta) = \frac{A}{2}\sum_{l=2}^{L}\sum_{k=1}^{K} E(\beta_l)\, E(e^{j\Theta_l})\, E(b^k_{-1}R_{k1}(lT_c) + b^k_0\hat{R}_{k1}(lT_c)) + W \quad \text{(EQ 61)}$$

$$E(b^k_{-1}) = E(b^k_0) = 0 \rightarrow E(\zeta) = 0 \quad \text{(EQ 62)}$$

The variance of the noise is

$$\mathrm{VAR}(\zeta) = E(|\zeta|^2) = \frac{A^2}{4}\sum_{l=2}^{L}\sum_{k=1}^{K} E(\beta_l^2)\, E|\kappa_{lk}|^2 + E|W|^2 \quad \text{(EQ 63)}$$

As before

$$\mathrm{VAR}(\zeta) = E(|\zeta|^2) = \frac{A^2}{4}\sum_{k=1}^{K} (T_s^2 \frac{2}{N}) \sum_{l=2}^{L} E(\beta_l^2) + E|W|^2 \quad \text{(EQ 64)}$$

simplifying we have

$$\mathrm{VAR}(\zeta) = \frac{A^2}{4}\sum_{k=1}^{K} (T_s^2 \frac{2}{N})\, \gamma E(\beta_d{}^2) + E|W|^2 \quad \text{(EQ 65)}$$

with one final substitution using the fact that

$$\gamma E(\beta_d^2) = \sigma^2 \tag{EQ 66}$$

we have

$$VAR(\zeta) = \frac{A^2}{4} K \left(T_s^2 \frac{2}{N}\right) \sigma^2 + \frac{N_0 T_b}{2} \tag{EQ 67}$$

The conditional signal to noise ratio in both the in phase and quadrature components is

$$\Upsilon | \beta_d = \frac{\left(\frac{\beta_d A T_s}{2}\right)^2}{\left(\frac{AT_s}{2}\right)^2 \sigma^2 K \frac{1}{N} + \frac{N_0 T_b}{2}} \tag{EQ 68}$$

Normalizing the equations with (34) and

$$v = \frac{\beta}{\sigma}, s = \frac{S}{\sigma} = \sqrt{2I} \tag{EQ 69}$$

We have

$$\Upsilon | v_d = \frac{v_d^2}{\frac{K}{N} + \frac{N_0}{2\sigma^2 E_s}} \tag{EQ 70}$$

The average signal to noise ratio is given by

$$\Upsilon_0 \equiv E(\Upsilon | v_d) = \frac{E(v_d^2)}{K\frac{1}{N} + \frac{N_0}{2\sigma^2 E_s}} = \frac{1 + 0.5s^2}{K\frac{1}{N} + \frac{N_0}{2\sigma^2 E_s}} \quad \text{(EQ 71)}$$

This is significantly different from the signal to noise ratio in Misser's model. In his model he assumes L i.i.d. Ricean multipath components. In that case the conditional signal-to-noise ratio is given by.

$$\Upsilon' | v'_1 = \frac{v'^2_1}{(LK - 1)\frac{2}{3N}E(v'^2_1) + \frac{N_0}{2\sigma^2 E_s}} \quad \text{(EQ 72)}$$

The optimal average signal to noise ratio in Misser's case is given by using MRC reception. In that case it is given by.

$$\Upsilon'_0 \equiv E(\Upsilon' | v'_d) = \frac{E(v'^2_1)L}{(LK - 1)\frac{2}{3N}E(v'^2_1) + \frac{N_0}{2\sigma'^2 E_s}} \quad \text{(EQ 73)}$$

There is a significant difference in the performance predicted by the two models. If we compare the average signal to noise ratios we see that

$$\frac{\Upsilon_0}{\Upsilon'_0} = \frac{E(v_d^2)}{E(v'_1{}^2)L} \frac{(LK-1)\frac{2}{3N}E(v'_1{}^2) + \frac{N_0}{2\sigma^2 E_s}}{K\frac{1}{N} + \frac{N_0}{2\sigma^2 E_s}} \qquad \text{(EQ 74)}$$

We can simplify this ratio if we assume that the channel noise is small relative to the interference noise and that $KL \gg 1$ then we have

$$\frac{\Upsilon_0}{\Upsilon'_0} \cong (1 + 0.5s^2)\left(\frac{2}{3}\right) = (1 + \mathrm{I})\left(\frac{2}{3}\right) \qquad \text{(EQ 75)}$$

We see that there is a 6-7dB difference in the average signal to noise ratio between our model and Misser's for broadcast transmission. A numerical comparison of the bit error rate will be presented later.

Cellular System

A significant portion of interference comes from the adjacent cells. However, due to the highly directional antennas that will be used in the INFOPAD system, we expect that the scattering power and LOS power from adjacent cells will interfere differently with our transmission. Near the interior of the cell, LOS interference from other cells will not be present, only scattering interference. However, near cell boundaries this

LOS component will be significant. As a small approximation we will consider the dominant resolvable path power to be equal to the LOS power. We will define α to be the ratio of the total outer cell scattering interference power to the scattering power of 1 cell. We expect that this parameter will be solely a function of cell distance. The total scattering power in one cell is given by $K\gamma\mathrm{E}(v_d^2)E_s = K\sigma^2 E_s$.

We will define μ to be the ratio of the total outer cell LOS interference power to the LOS power of 1 cell. We expect this to be a function of distance and directional antenna gain. The total LOS power in one cell is given by $K\mathrm{E}(v_d^2)E_s = K(0.5S^2)E_s$. Furthermore we will make the simplifying assumption that all cells transmit at the same signal power and have the same number of users. In this case the signal to noise ratio becomes.

$$\Upsilon | v_d = \frac{v_d^2}{((1+\alpha) + \mu 0.5 s^2)\,\mathrm{K}\frac{1}{\mathrm{N}} + \frac{\mathrm{N}_0}{2\sigma^2 E_s}} \quad \text{(EQ 76)}$$

No Diversity

Diversity takes on a slightly different meaning with our model. By synchronizing to the dominant resolvable path, we are effectively using selection diversity among the multipath components. Minuscule gains can be obtained using true selection diversity or MRC reception in our model. Significant gains can still be obtained, however, by using multiple antennas. As a result we will define a different order of diversity with this model, where $M' = NA$. We will consider single antenna reception as non-diversity reception. We cannot derive a closed form expression for the error rate and, as a result, we must numerically integrate against the Ricean pdf.

$$P(e) = \int_0^\infty P(e|v)f_v(v)dv \qquad \text{(EQ 77)}$$

Where the pdf is given by

$$f_v(v) = v\exp\left(-\frac{v^2+s^2}{2}\right)I_0(sv) \qquad \text{(EQ 78)}$$

and $I_0(\ )$ is a zero order modified bessel function.

Selection Diversity

For selection diversity with m antennas, we must integrate against the pdf for the maximum of m i.i.d. ricean variables. This is given by

$$f_{vmax}(v) = m\,(F_v(v))^{m-1} f_v(v) \qquad \text{(EQ 79)}$$

where $F_v(\ )$ is the cumulative distribution function of $f_v(\ )$.

Maximum Ratio Combining

Finally, the pdf for the square root of the sum of the squares of m ricean variables is.

$$f_{mrc}(v) = v\left(\frac{v^2}{Ms^2}\right)^{\frac{M-1}{2}} exp\left(-\frac{v^2+Ms^2}{2}\right) I_{M-1}(sv) \qquad \text{(EQ 80)}$$

and $I_M(\ )$ is a Mth order modified bessel function.

CHAPTER 6 Results

We are now in a position to present some numerical results. All figures are generated assuming a 1 Mbaud symbol rate which will correspond to a total data rate of 2Mbits/s in our 4-QAM system. In addition, we will assume the following default parameters for all graphs unless otherwise stated.

$$\frac{E_s}{N_0} = 10^3, N = 64, NA = 2, \alpha = 0.35, \mu = 0.2$$

where $\frac{E_s}{N_0}$ is the symbol power to channel noise ratio, N is the spread factor, NA is the number of antennas, α is the ratio of outer-cell scattering interference power to the total scattering power within the cell and μ is the ratio of outer-cell LOS interference power to the total LOS power within the cell. We will first compare system performance under the two different model assumptions and show that there is a significant difference in performance. This is shown in Figure 7. Note that if our model is correct then we will get significantly better performance at a much lower hardware cost than expected for the asynchronous case. A summary of the differences in the various models is included below

TABLE 1.

MODEL	UPLINK/ DOWNLINK	DELAY PROFILE	INTERCELL INTERFERENCE	INTRACELL INTERFERENCE
INFOPAD	D	LOS + SCATTER	ORTHOGONAL LOS + RAYLEIGH SCATTER	RANDOM LOS + RAYLEIGH SCATTER
MISSER	U	I.I.D RICEAN		
KAVEHRAD	U	I.I.D RAYLEIGH		

The performance of the downlink depends heavily on environmental parameters which we cannot control, such as the LOS component in the signal (s) and the cellular interference parameters α, μ. The performance depends as well on several hardware parameters that we can control such as the type of receiver we use: No Diversity, Selection Diversity or Maximum Ratio Combining and factors such as the number of antennas and the spread factor. We will first look at the affect of the environmental parameters.

Environmental Parameters

The first case to be considered is the performance of a simple receiver with no diversity for different ricean fading environments. The spread factor is 128, the SNR is 30dB and there are two antennas. This is shown in Figure 8. Even in a ricean environment we can see that the performance is not that desirable. Figure 9 shows the error performance for selection diversity. We see that a moderate increase in hardware complexity can improve performance dramatically.

Finally, Figure10 shows the result of using Maximal ratio combining. This is the optimal receiver structure. The difficulty is that Maximal ratio combining requires perfect or close to perfect knowledge of the amplitudes and phases of each path while Selection diversity requires only knowledge of the relative signal power.

FIG8 - EFFECT OF RICEAN PARAMETER I ON NON DIVERSITY RECEPTION

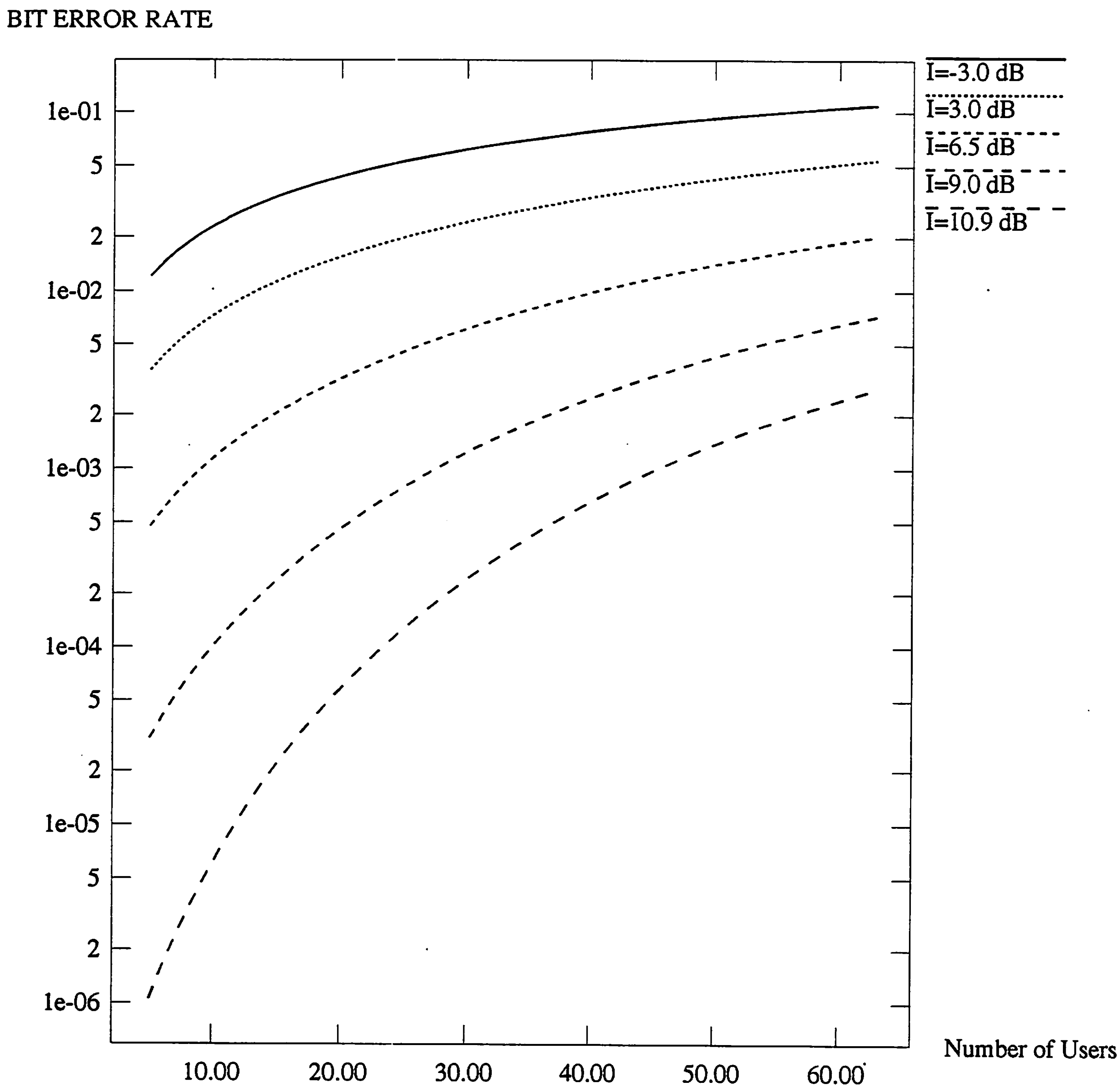


FIG9 - EFFECT OF RICEAN PARAMETER I ON SELECTION DIVERSITY RECEPTION

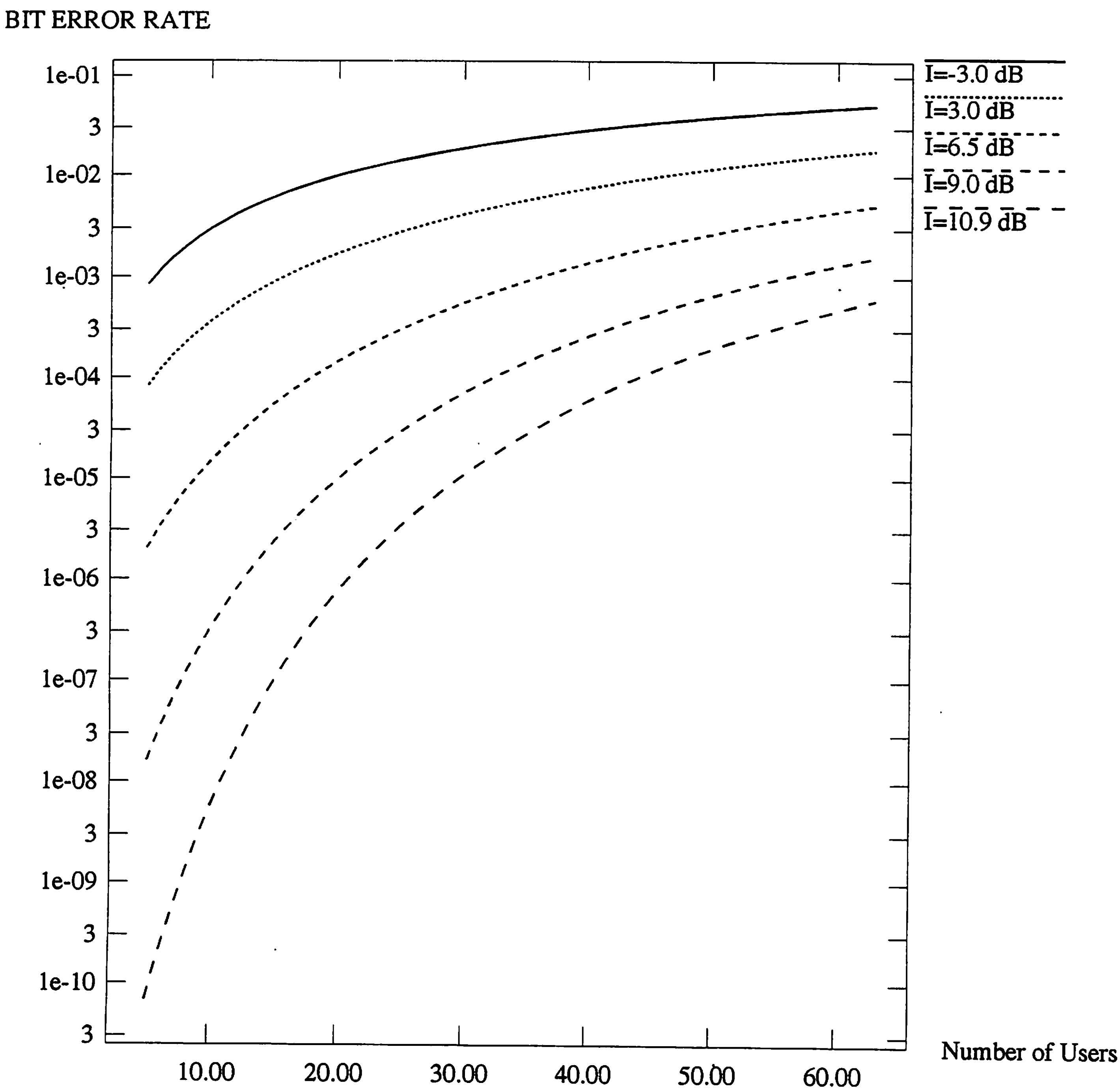


FIG10 - EFFECT OF RICEAN PARAMETER I ON MRC RECEPTION

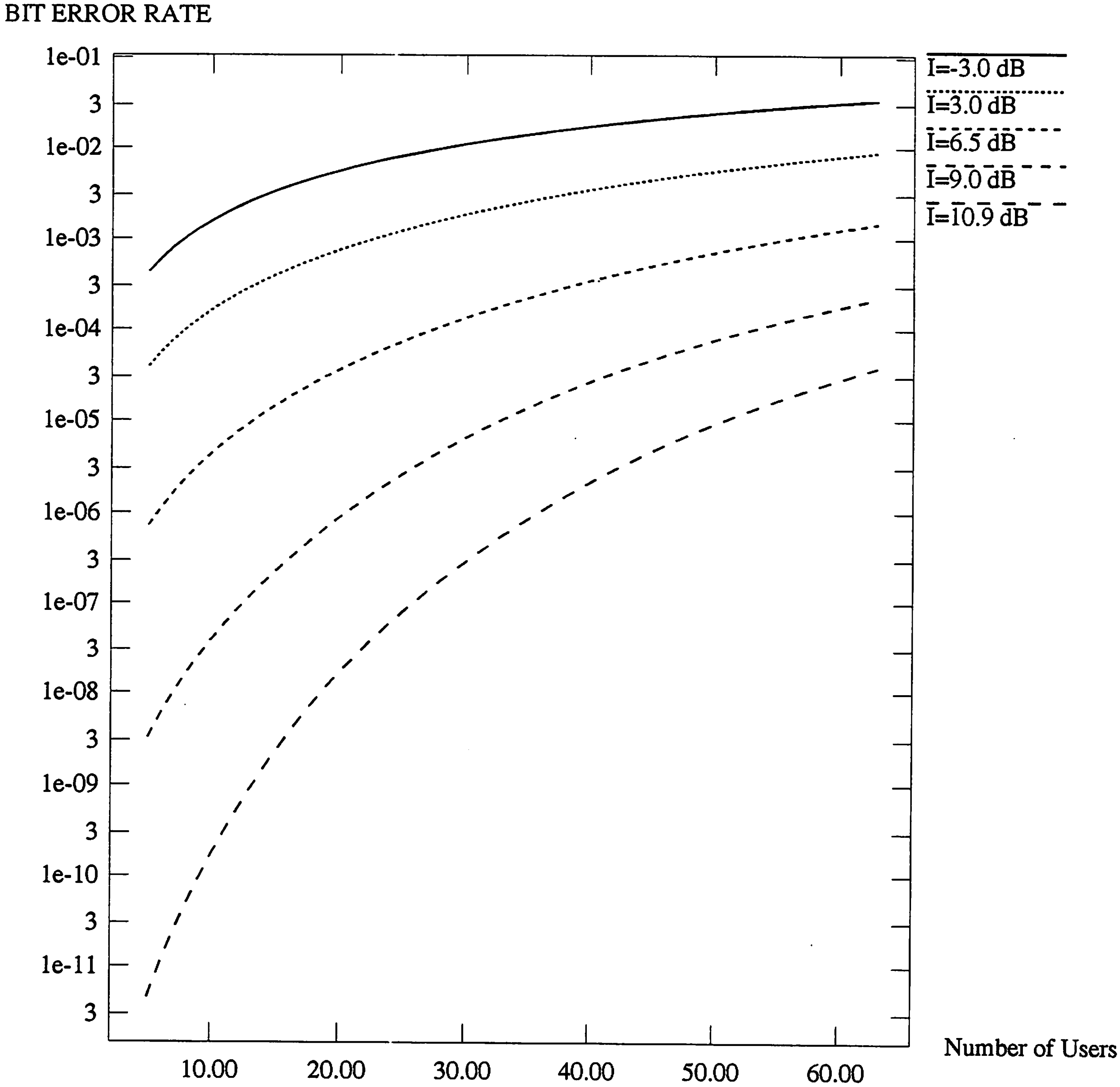
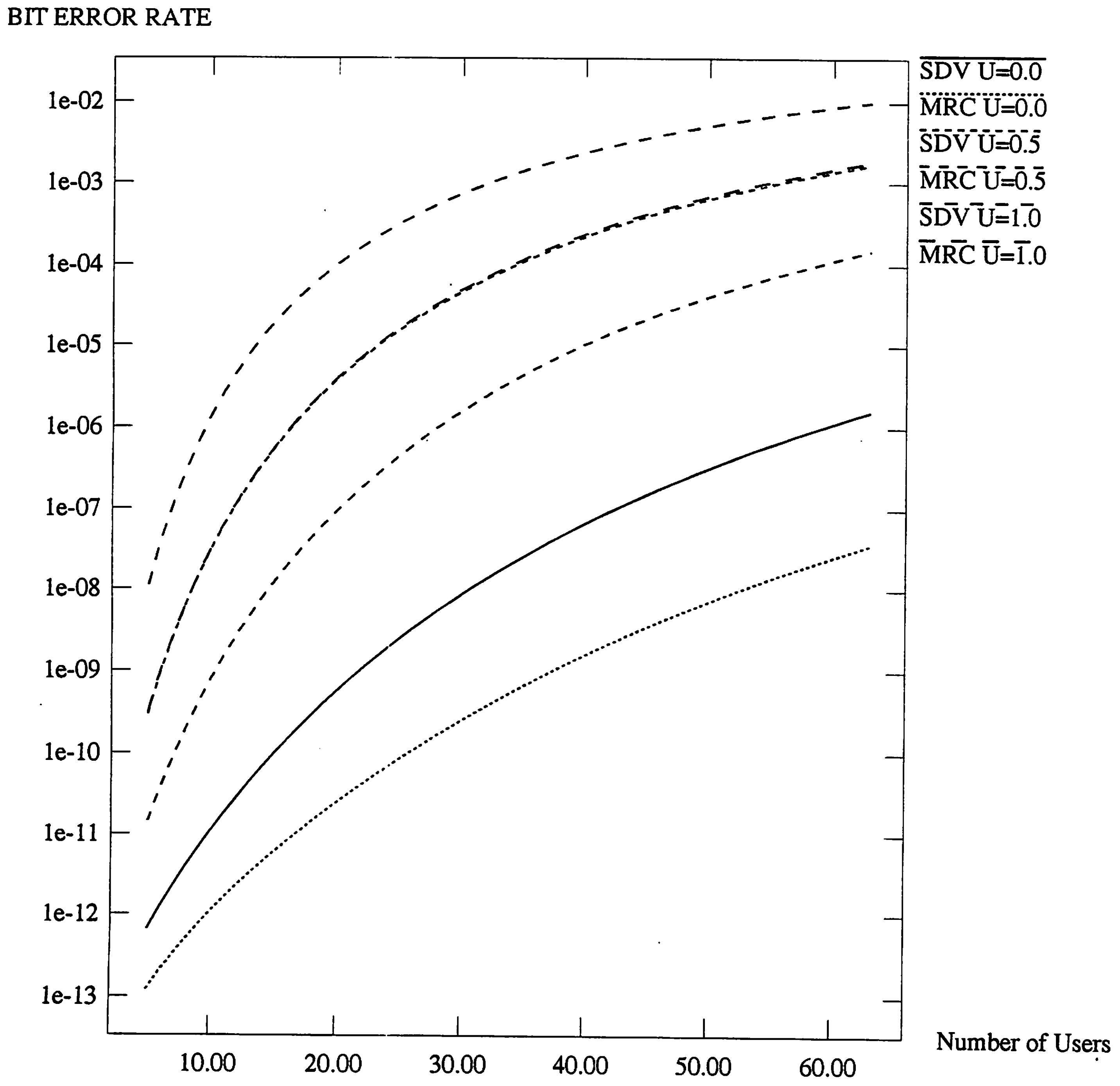


FIG12 - EFFECT OF CELLULAR PARAMETER U ON SYSTEM PERFORMANCE



The effect of the two cellular parameters α, μ is shown in Figures J,K. No estimates for these values exist at the moment.

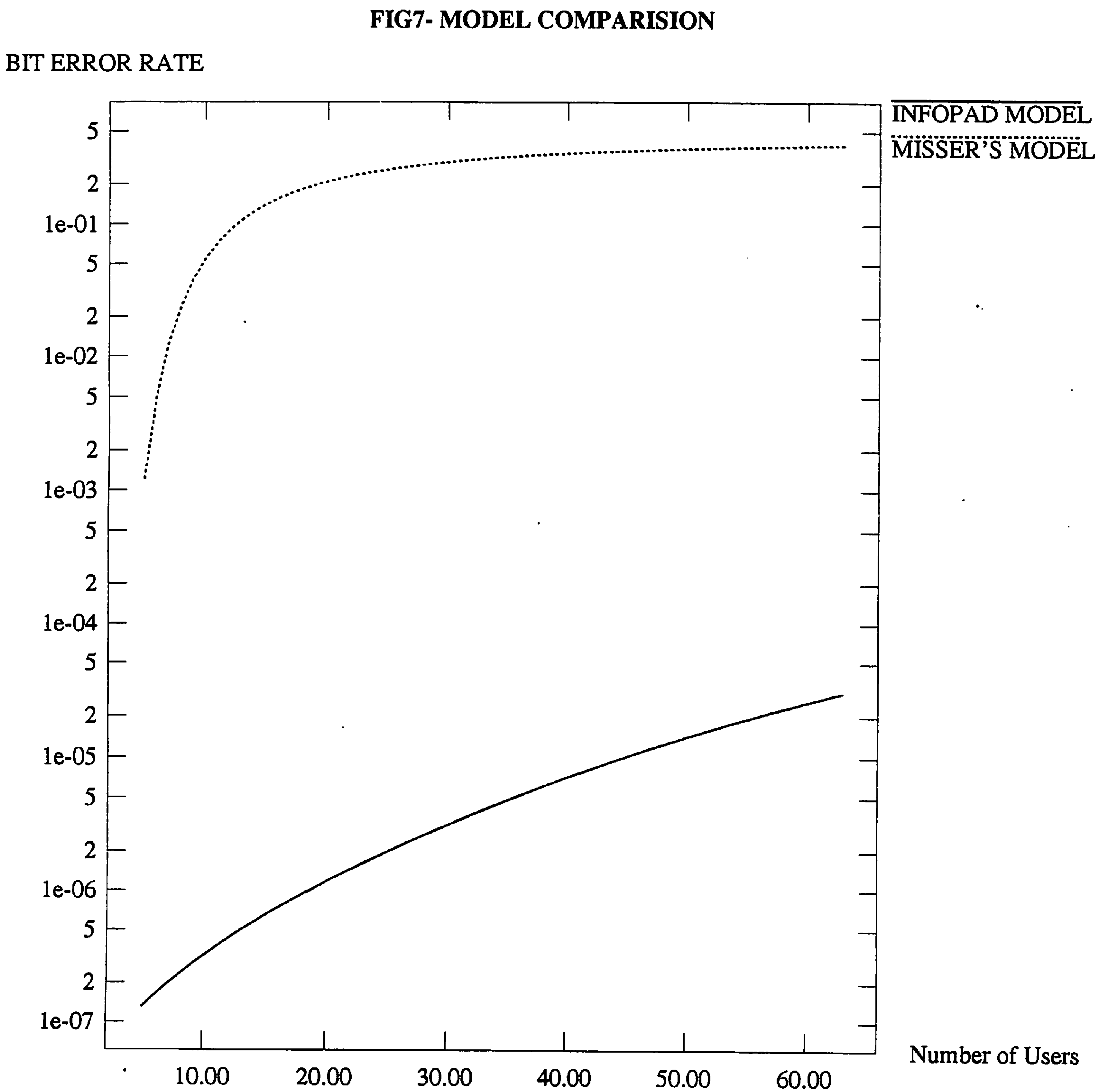
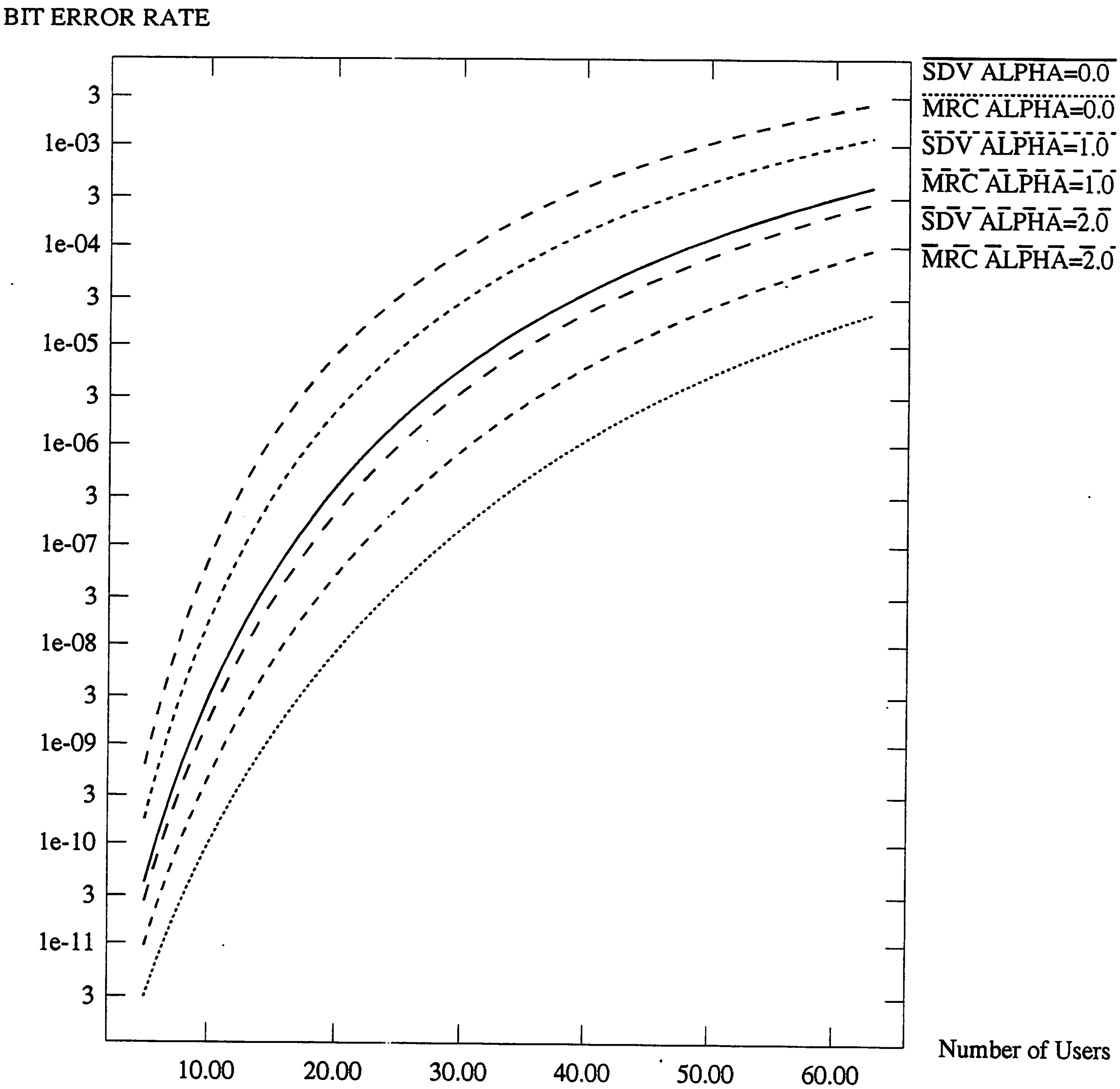


FIG7- MODEL COMPARISION
BIT ERROR RATE
INFOPAD MODEL
MISSER'S MODEL
5
2
1e-01
5
2
1e-02
5
2
1e-03
5
2
1e-04
5
2
1e-05
5
2
1e-06
5
2
1e-07
10.00
20.00
30.00
40.00
50.00
60.00
Number of Users

FIG11 - EFFECT OF CELLULAR PARAMETER ALPHA ON SYSTEM PERFORMANCE



Hardware Parameters

There are a few system parameters that we can control; such as the spread factor, number of antennas and receiver structure. If we increase the spread factor we will be able to improve the auto-correlation properties of the spreading codes and hence reduce the multi-user interference of the system for a fixed number of users. Increasing the spread factor however, also increases the bandwidth needed for transmission. In addition the correlator has to run at a higher rate so a substantial hardware penalty must be paid for improved performance. The effect of different spreading rates is shown in Figure 11. A doubling of the spread factor results in a lowering of the BER of about 70% in the case of selection diversity. The improvement is more pronounced in the case of MRC reception. In this case, a doubling of the spread factor results in about an order of magnitude decrease in the error rate.

Another way to improve INFOPAD's performance is to increase the number of antennas used to receive the signal. This improves the chances of receiving a high amplitude signal. More paths will be available at the receiver which will be able to take advantage of the multiple sources. The antennas must be separated by approximately 0.15 meters (half a wavelength of the carrier) for this method to be acceptable. This limits the number of antennas that we can actually have. The effect of multiple antennas is shown in Figure 12. A close look reveals that adding an additional antenna has about the same effect as doubling the spread factor for these parameters.

The last thing to be investigated is the change in performance for changes in the transmitter power. This is shown in figure 13. SNR in our case is defined as the ratio of the symbol power to the background noise power.

$$SNR \equiv \frac{E_s}{N_0} \qquad \text{(EQ 81)}$$

The behavior is atypical of a normal communications system in that an increase in transmitter power will not necessarily increase the error performance of the system. This is because the limiting noise component is not background noise but multi-user interference. This is important with a microwave carrier because we want to keep the signal power as low as possible.

FIG13 - EFFECT OF SPREAD FACTOR ON SYSTEM PERFORMANCE

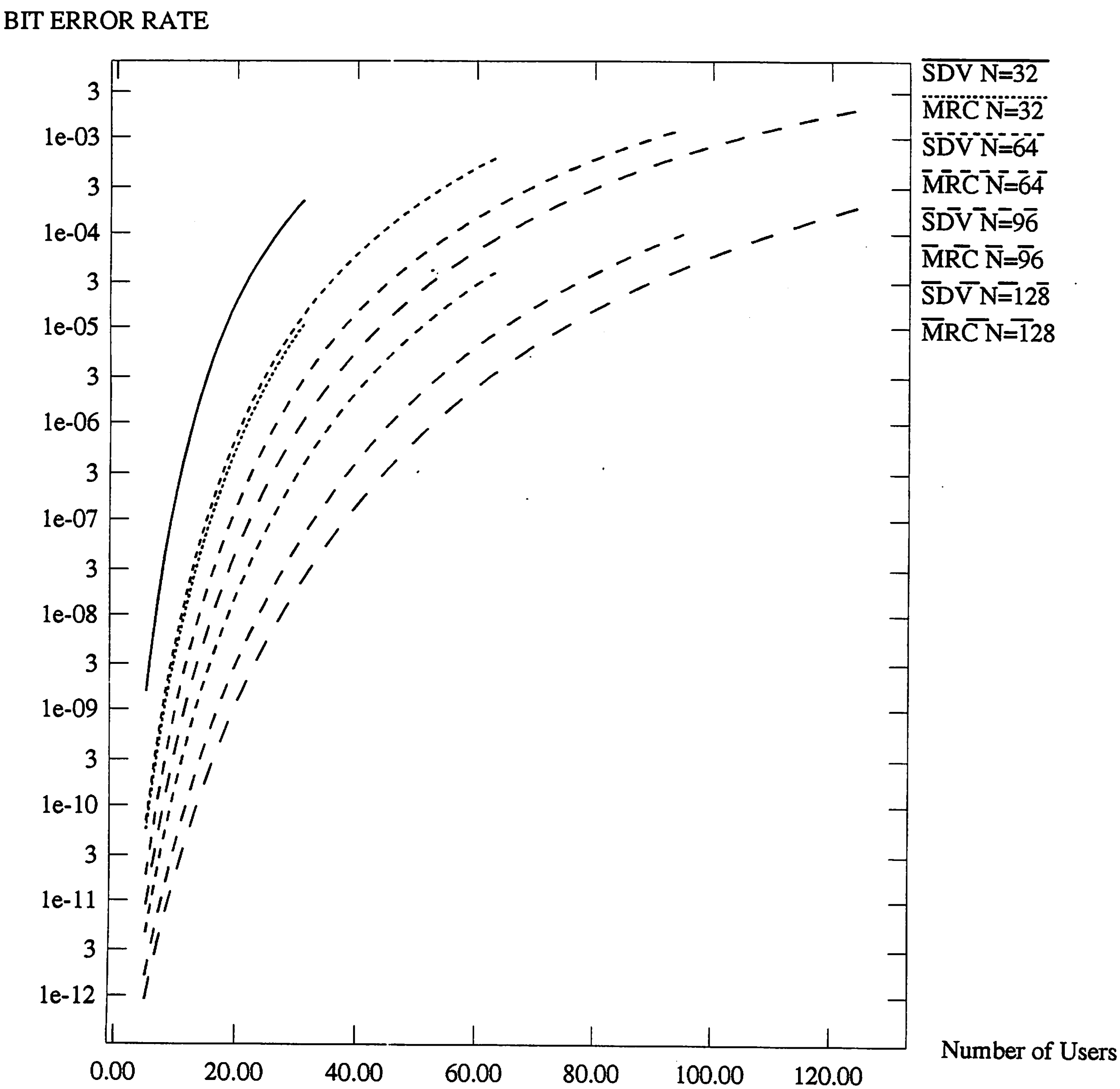


FIG15 - EFFECT OF SNR ON SYSTEM PERFORMANCE

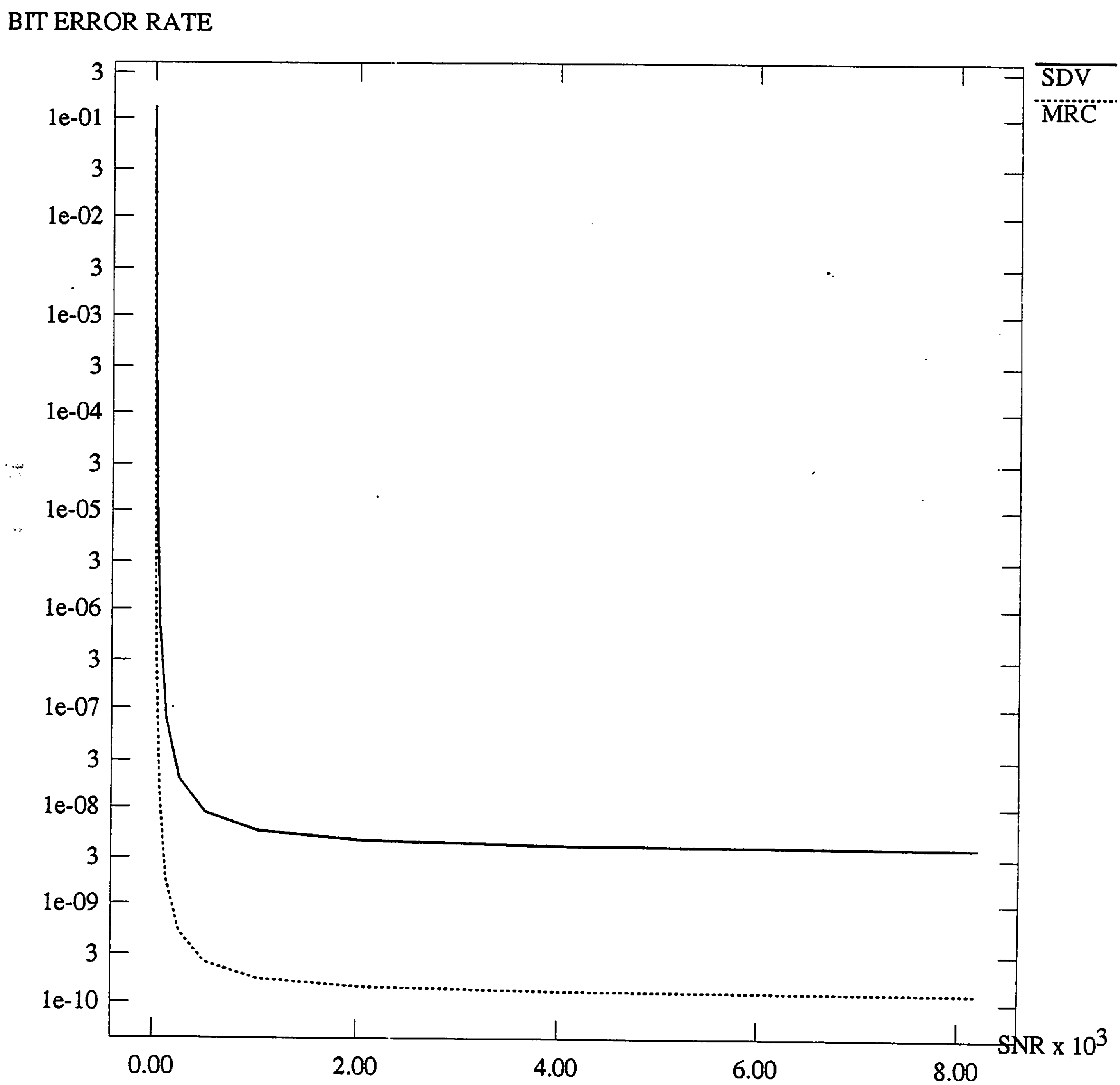
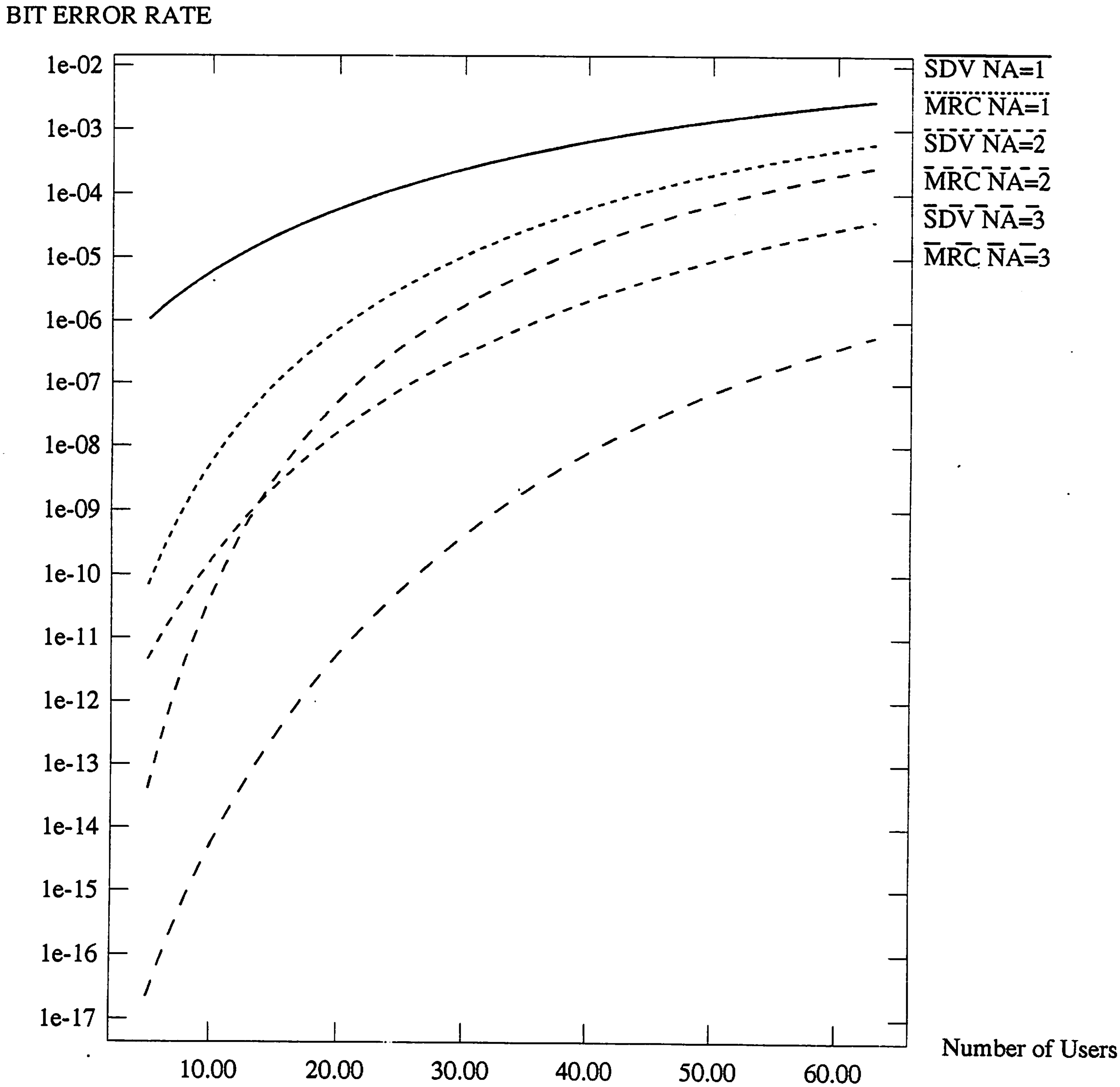


FIG 14 - EFFECT OF MULTIPLE ANTENNAS ON SYSTEM PERFORMANCE



Conclusion

These results provided some very useful design information. According to our propagation model, the INFOPAD downlink performance will be significantly better than expected with a much simpler hardware design. In addition it is critical that any spreading codes chosen should have zero cross correlation at a zero phase offset. Better performance will be obtained with the addition of additional antennas and/or increasing the spread factor of the system. The benefits from adding additional antennas and increasing the spread factor were explicitly calculated for specific environments. From these results, it is hoped, the hardware trade-offs can be decided. Finally, the addition of the adjacent cell receiver, while adding hardware complexity, will improve receiver performance near a cell boundary. The exact effect has not been calculated.

At this point it is expected that we could have two antennas on the mobile terminal and that the spreading factor will be either a factor of 32 or 64 depending on the difficulty of hardware implementation. We also expect to use a selection diversity receiver structure.

Future Work

Much work remains to be done. The most important is to determine if our model is realistic for indoor environments and if so, the critical multi-cell interference parameters α, μ must be determined or estimated. In addition, the same numbers of users have been assumed for all cells. This is unrealistic. The number of users in the adjacent cells should be assumed random and independent of the number of users within the current cell.

Another aspect of the system which was not investigated is the performance of the receiver when it is near a cell boundary. In this case it is expected that LOS components from the adjacent cells will interfere with the desired signal making the downlink performance worse. In this case the addition of the adjacent cell receiver will mitigate this effect. The overall performance of INFOPAD when it is near a cell boundary needs to be determined.

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CAPTIONS

Figure 7: The performance predicted by our model for synchronous transmission vs and the performance predicted by Misser's model for asynchronous transmission vs. the number of users with no intra-cellular interference, a ricean factor of 10.9dB, two antennas, a spread factor of 64 and a signal to noise ratio of 30dB.

$$\frac{E_s}{N_0} = 10^3, N = 64, NA = 2, \alpha = 0, \mu = 0, I = 10.9dB$$

Figure 8: The performance of non-diversity reception vs the number of users for various ricean environments with scattering interference power factor of .35, a LOS interference power factor of .2, two antennas, a spread factor of 64 and a signal to noise ratio of 30dB.

$$\frac{E_s}{N_0} = 10^3, N = 64, NA = 2, \alpha = 0.35, \mu = 0.2$$

Figure 9: The performance of selection diversity reception vs the number of users for various ricean environments with scattering interference power factor of .35, a LOS interference power factor of .2, two antennas, a spread factor of 64 and a signal to noise ratio of 30dB.

$$\frac{E_s}{N_0} = 10^3, N = 64, NA = 2, \alpha = 0.35, \mu = 0.2$$

Figure 10: The performance of MRC reception vs. the number of users for various ricean environments with a scattering interference power factor of .35, a LOS interference power factor of .2, two antennas, a spread factor of 64 and a signal to noise ratio of 30dB.

$$\frac{E_s}{N_0} = 10^3, N = 64, NA = 2, \alpha = 0.35, \mu = 0.2$$

Figure 11: The performance of MRC and selection diversity reception vs. the number of users for various scattering interference power factors, a LOS interference power factor of .2, two antennas, a spread factor of 64 and a signal to noise ratio of 30dB.

$$\frac{E_s}{N_0} = 10^3, N = 64, NA = 2, \mu = 0.2, I = 10.9dB$$

Figure 12: The performance of MRC and selection diversity reception vs. the number of users for various LOS interference power factors, a scattering interference power factor of .35, two antennas, a spread factor of 64 and a signal to noise ratio of 30dB.

$$\frac{E_s}{N_0} = 10^3, N = 64, NA = 2, \alpha = 0.35, I = 10.9dB$$

Figure 13: The performance of MRC and selection diversity reception vs. the number of users for various spread factors, an inter-cellular scattering interference power factor of .35, an inter-cellular LOS interference power factor of .2, two antennas, a spread factor of 64 and a signal to noise ratio of 30dB.

$\frac{E_s}{N_0} = 10^3, N = 64, NA = 2, \alpha = 0.35, I = 10.9dB$

Figure 14: The performance of MRC and selection diversity reception vs. the number of users for various numbers of antennas, an inter-cellular scattering interference power factor of .35, an inter-cellular LOS interference power factor of .2, two antennas, a spread factor of 64 and a signal to noise ratio of 30dB.

$\frac{E_s}{N_0} = 10^3, N = 64, NA = 2, \alpha = 0.35, I = 10.9dB$

Figure 15: The performance of MRC and selection diversity reception vs. the signal to noise ratio with an inter-cellular scattering interference power factor of .35, an inter-cellular LOS interference power factor of .2, two antennas, a spread factor of 64 and a signal to noise ratio of 30dB.

$\frac{E_s}{N_0} = 10^3, N = 64, NA = 2, \alpha = 0.35, I = 10.9dB$