PILOT SYMBOL ASSISTED SUCCESSIVE INTERFERENCE CANCELLING RECEIVER FOR DIRECT SEQUENCE/CODE DIVISION MULTIPLE ACCESS SYSTEMS

by

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Abstract

Direct Sequence/Code Division Multiple Access systems employing conventional receiver are severely limited in their capacity, because of non-zero cross correlations of the spreading sequences and the high variation of the received powers at the base station.

Interference cancelling receiver have offered a solution for these problems at the expense of a higher system complexity. Among the various schemes that have been proposed so far, the successive interference cancellation (SIC) receiver offers a good compromise scheme in terms of capacity and complexity. Proper operation of these receivers in a mobile environment requires them to perform the estimation of the fading channel gain which is needed for the data detection, regeneration and cancellation of the received signals. This channel estimation can be performed by using Pilot Symbol Assisted Modulation (PSAM).

This thesis investigates the performance of successive interference cancelling receiver with PSAM operating in a Rayleigh fading environment. Different receiver concepts will be presented and their performance will be discussed. In particular, we will present a hybrid receiver that combines a low complexity decorrelator to reduce the interference in the channel gain estimates with a successive interference cancellation scheme for the data detection. The simulation results indicate an increase in the system capacity at a moderate level of implementation complexity.

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List of Abbreviations

AWGN	Additive White Gaussian Noise
BER	Bit Error Rate
BPSK	Binary Phase Shift Keying
CDMA	Code Division Multiple Access
DPSK	Differentially coherent Phase Shift Keying
DS	Direct Sequence
FDMA	Frequency Division Multiple Access
FH	Frequency Hopping
MAI	Multiuser Access Interference
MSE	Mean Square Error
NRZ	Non-Return to Zero
PN	Pseudorandom Noise
PSA	Pilot Symbol Assisted
PSAM	Pilot Symbol Assisted Modulation
PSD	Power Spectral Density
PSK	Phase Shift Keying
TDMA	Time Division Multiple Access

Chapter 1

Introduction

1.1 Introduction

Over the past twenty years, the demand for mobile communications has significantly increased and the bandwidth requirements impose a problem to be solved by the telecommunications industry. Moreover, users are demanding a greater variety of services to be implemented in future mobile telecommunication systems, like facsimile or transmission of images. To fulfill these demands, it is essential to have a flexible multiple access method that maintains both the ability to handle different data rates and a high capacity. Furthermore, the fading mobile radio environment and its impact on the system performance imposes another difficult task that has to be taken into consideration.

One solution for these various tasks is believed to be code division multiple access (CDMA). This technique is already used in a commercial system by Qualcomm, Inc., USA [29] and has become the North American communications standard IS-95.

Two of the most important factors that limit the capacity of a multiuser

DS/CDMA system are the power variations of the received signals at the base station and the non-zero cross-correlations of the users' spreading sequences.

The conventional multiuser detector consists of a bank of matched filters, with each filter matched to a specific signature sequence. Following the filter, the output would be sampled at the symbol rate and decisions would be made on the sampled filter outputs. In a system where the transmissions over the channel are only corrupted by additive white Gaussian noise (AWGN), this receiver is the optimal detector [30]. However, spreading sequences in the system are non-orthogonal and transmissions of the user will introduce multiuser access interference (MAI) whichs yields to an irreducible error floor [31]. Lack of orthogonality will occur if the transmissions of the users are asynchronous or when the channel is time dispersive. The system performance will severely degrade unless the cross-correlations between users are low enough and the received powers are nearly equal. However, user in a mobile communications system will change their positions to the receiver and the received signal powers are not equal at all. The phenomena is usually known as the *near-far* problem [5]. Another reason for this problem is fading and shadowing. A solution to that problem can be found by the use of stringent power control [29], [39] or the use of near-far resistent receiver. Recently, a lot of attention has been given to these receiver concepts, e.g. [1], [2], [3], [14].

Sergio Verdú [1] related the multiple access channel to a periodically time varying, single user, intersymbol interference channel and derived the optimal multiuser detector. The complexity of this detector increases exponentially with the number of users and subsequent research was done in the area of suboptimal solutions with reduced complexity e.g. [3], [16], [13], [14]. One part of this thesis will discuss some suboptimal solutions in more detail.

For the proper operation of these receivers, it is necessary to estimate the fading complex gains of the channel. Research in the area of suboptimal multiuser receiver has often assumed these channel parameters are known at the receiver site. In reality, these parameters have to be estimated. One method often applied is the use of pilot symbols as it was done in [17] for a single user BPSK and QPSK system.

The use of pilot symbols in DS/CDMA systems was considered in [18] which focused on the perfomance of a pilot symbol DS/CDMA receiver without any interference cancellation.

Work in the area of pilot symbol assisted MAI cancelling receiver was reported in [19], where the performance of a successive interference cancellation scheme with pilot symbols was investigated. In [6] the scheme was modified for a quasi-synchronous CDMA system and in [7] the scheme was investigated in a multi-cell environment.

1.2 Contributions of the Thesis

This thesis focuses on the performance of a multiuser CDMA receiver which estimates the complex fading gains by means of pilot symbols and cancels out multiuser access interference successively. Three different methods of inserting the pilot symbols into the data stream will be discussed and a hybrid receiver concept will be presented. We will compare our proposed concept to an already existing scheme and will show that our scheme can improve the capacity in the considered system.

1.3 Thesis Outline

The thesis is organized in the following way:

Chapter 2 will give a review and background information, which includes a brief introduction into spread spectrum systems, code division multiple access, the radio channel model description and channel estimation strategies. It also includes a partial literature review of multiuser detection schemes and will introduce some concepts by using a simple model.

In chapter 3 the model, introduced in chapter 2, will be generalized and the investigated receiver concepts will be discussed in detail. Three different schemes of inserting pilot symbols into the data stream will be presented and some baseline systems will be introduced.

Chapter 4 gives a description of the computer simulation that was used in these studies, and we will present results for the three different schemes and the different receiver concepts under various conditions. This chapter will be concluded by a discussion of these ideas.

Chapter 5 will conclude the thesis with some suggestions for further work.

Chapter 2

Review and Background

2.1 Spread Spectrum Communications

One can distiguish between two different spread-spectrum (SS) techniques: *direct sequence* (DS) and *frequency hopping* (FH) spread spectrum. In a FH-SS system a pseudorandom-code is used to select different carrier frequencies. The carrier frequency is switched over a wide frequency range. These carriers are then modulated by the data. FH-SS systems can be further classified into slow hopping, where the carrier is switched only once every couple of symbols, or fast hopping, where several carrier switches in one symbol time occur. Today, only technology for slow hopping is available.

In DS-spread spectrum, the data sequence is multiplied with a pseudonoise random sequence prior to modulation. The rate of the pseudonoise sequence is much higher than that of the data sequence which spreads out the power spectrum of the signal and makes it indistinguishable from the background noise. This attribute makes it, for example, possible to overlay narrowband applications on the code division multiple access channel, as will be discussed later.

In this thesis we will only focus on DS-spread spectrum and particular DS/CDMA applications. The interested reader can find further information about FH-SS for example in [34].

2.1.1 Direct Sequence Spread Spectrum Systems

To illustrate some basic concepts consider the example of an uncoded coherent direct sequence, binary phase shift keying (DS/BPSK) system.

In principle, the transmitter would multiply the data sequence with a high rate pseudonoise sequence and would apply the multiplied signal to the modulator to modulate it onto the carrier. Both operations, the multiplication with the pseudonoise sequence and the modulation, are linear operations and thus interchangeable. For analysis reasons it is more convenient to consider a system as shown in Fig. 2.1.

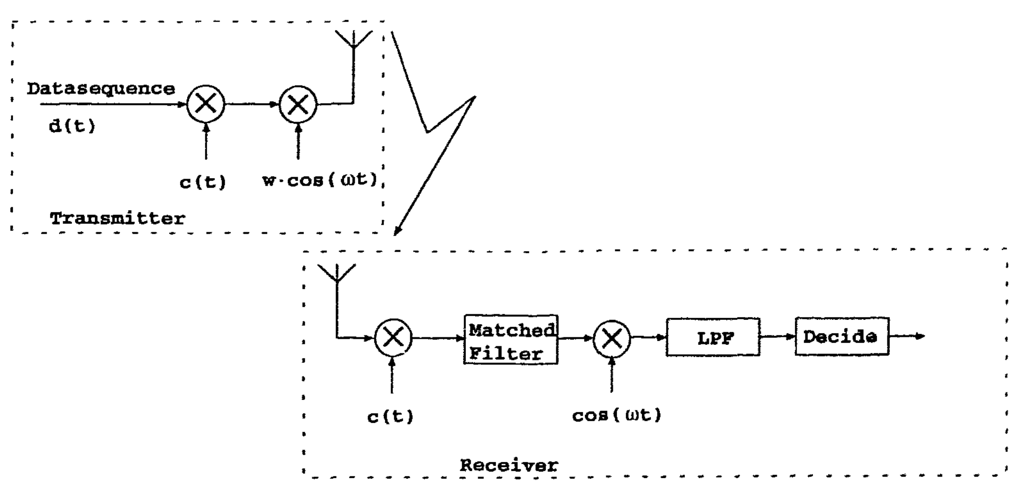

Figure 2.1: Overview of a spread spectrum system

The transmitted signal can be expressed as

$$s(t) = w\, d(t)\, c(t)\, \cos(\omega t) \tag{2.1}$$

where w is the transmitted amplitude, $d(t)$ is the data sequence with a rate $R = \frac{1}{T}$, $c(t)$ is the PN sequence with rate $R_c = \frac{1}{T_c}$, and ω is the carrier frequency of the system. Bits in the PN sequence are called *chips* and the rate is referred to as the *chiprate* R_c. The ratio

$$N = \frac{T}{T_c} \tag{2.2}$$

is called the *spreading factor* or *processing gain* of the system.

The signal in (2.1) is sent over a radio channel which we assume at this moment as ideal[1]. If we now multiply the received signal with the PN-sequence again, we would obtain

$$\begin{aligned} r(t) &= c(t) \cdot s(t) \\ &= c(t)^2 \cdot w\, d(t) \cos(\omega t) \\ &= w\, d(t) \cos(\omega t) \end{aligned} \tag{2.3}$$

since $c(t)^2 = 1$. The signal $r(t)$ would then be demodulated in a manner similiar to a narrowband system. It is clear that the PN sequence on the receiver site has to be properly aligned to the transmitted PN sequence in order to despread the received signal properly.

Fig. 2.2 shows the data signal before and after the multiplication with the PN sequence and the PN sequence itself. The bandwidth required to transmit the data signal with a rate R would have been approximately $\frac{2}{T}$, where it is now, after the spreading, $\frac{2}{T_c}$. Since T_c is usually orders of a magnitude smaller than T, the spread signal occupies a much wider bandwidth. The power spectral density (PSD) is spread out to an extent, that the signal is hidden under the additive white noise of the channel. Unintended user will therefore not notice the existence of a signal. Even knowing the presence of the signal, decoding is impossible without knowledge about the particular PN sequence used for spreading. The despreading process would on the other hand spread an interfering narrowband signal[2] at the receiver site to a spectral density below the noise level. After lowpass filtering of the despread data signal, only a small portion of the narrowband interfering signal would pass this filter and would

[1] the channel adds no noise or delay and the signal is not subject to fading
[2] jamming signal or overlaying narrowband application

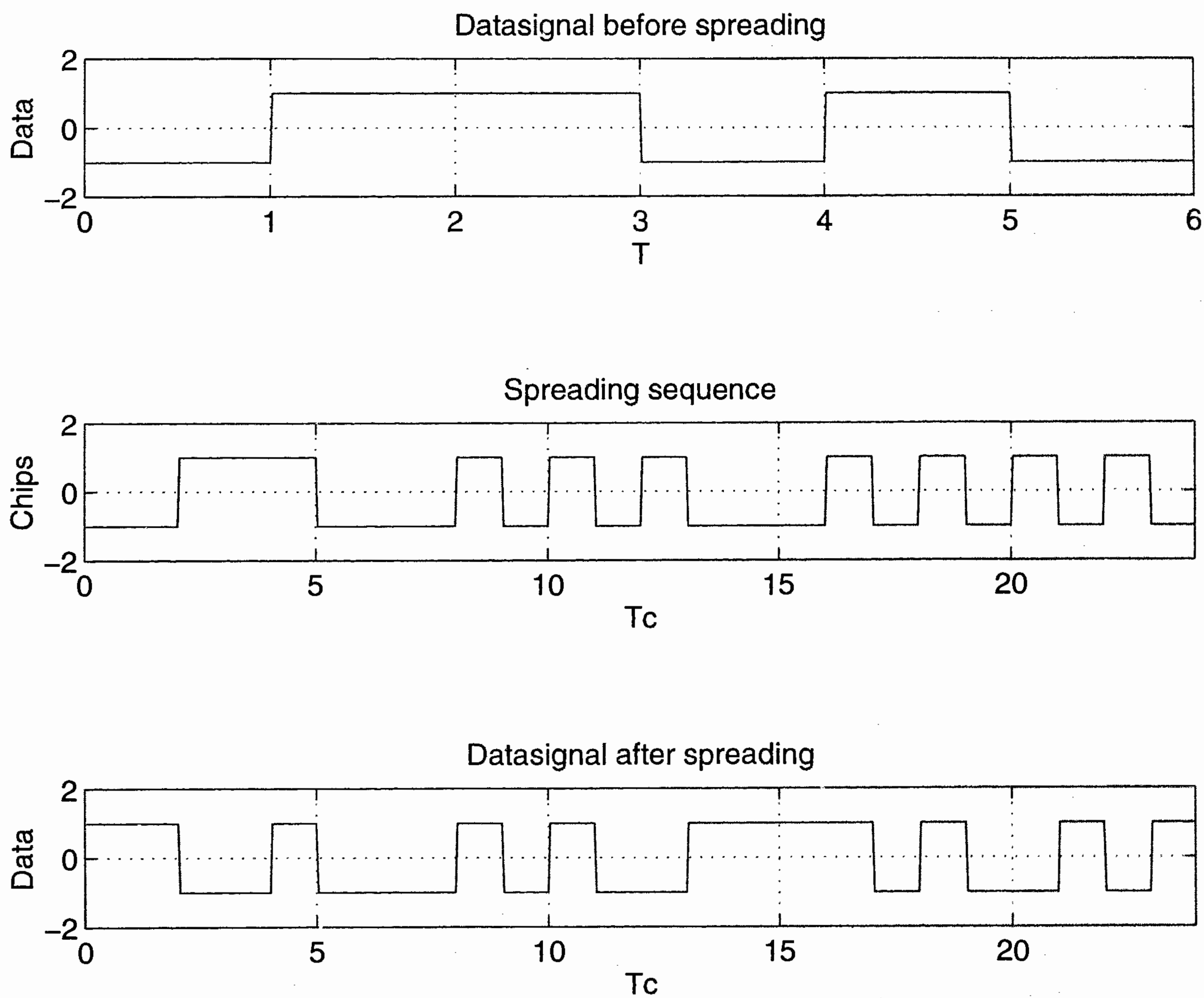

Figure 2.2: The effect of spreading on the data sequence

cause just a minor noise enhancement. These two properties are the reason why the first use of spread spectrum was made in military applications.

Pseudonoise sequences

A spread spectrum system requires a pseudonoise code sequence for the spreading and despreading of the data signal. Some properties of the sequences are important:

I) The sequences should be easy to generate.

II) They appear to be random to an unintended user.

III) They are either aperiodic or have very long periods.

IV) They should have good autocorrelation properties.

V) They should have low cross correlation properties.

Properties *II)* and *III)* are important to protect a spread spectrum system from jamming and unintended listeners. Property *IV)* is important for synchronizing the receiver's local generated PN-sequence with the time shifted received version. This procedure is called acquisition (coarse alignment) and tracking (fine alignment) and is further discussed in [34]. *V)* is needed for a wideband multiple access scheme where it is desired to have low crosscorrelation properties to reduce the interference of users in the system. Moreover, in an asynchronous CDMA scheme, it would be desirable to have these good crosscorrelation properties for all possible time shifts in the system. However, combining all these properties in one code set is impossible and one has to compromise.

One common way to obtain PN sequences is to use a linear feedback shift register as shown in figure 2.3. The coefficients $a_0, a_1, \ldots, a_p$ are either 0 or 1. The contents in the register are modulo-2 added and feedback to the input of the register. Particular combinations of the coefficients $a_0, a_1, \ldots, a_p$ generate maximum length sequences of length $2^p - 1$ [10]; all states will be included in the sequence, except for the all zero

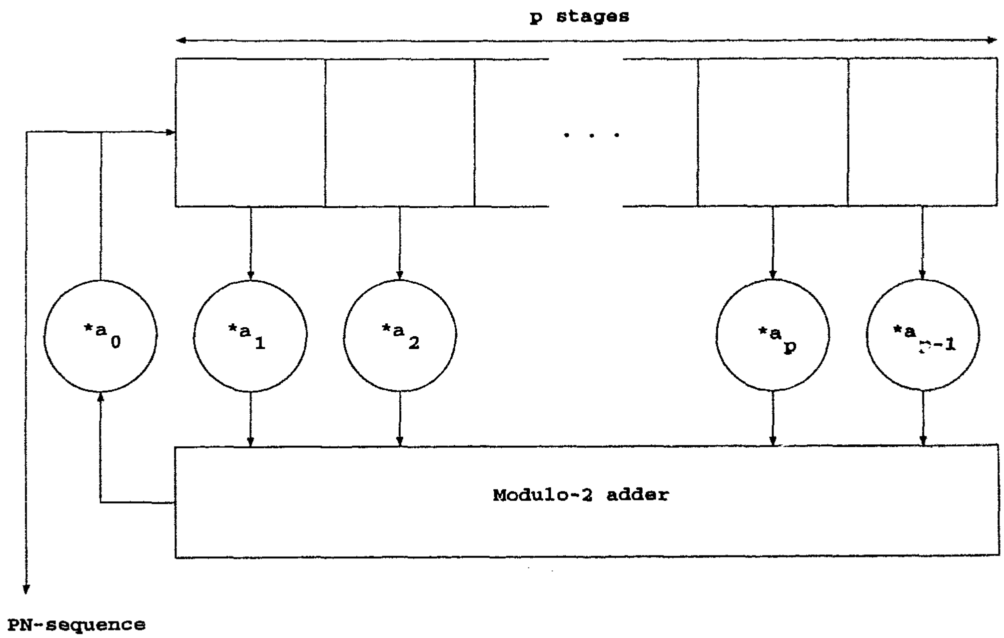

Figure 2.3: Linear Feedback Shift Register for M-sequence generation

state. Although m-sequences fulfill property *III)*, from the security point of view they do not offer much protection. It is possible to determine the feedback coefficients by analyzing a sequence of $2p$ consecutive known bits in the sequence as for instance is shown in [10].

Another popular set of PN-sequences are the Gold sequences, which offer better cross-correlation properties than the M-sequences. These and other methods for obtaining PN-sequences are discussed in the literature, e.g. [34].

2.2 DS/Code Division Multiple Access

One can use the spread spectrum principle for dividing the resource spectrum between different users. This is known as *code division multiple access* or CDMA and offers a lot of advantages for a mobile communication system. We will discuss these interesting properties in this section.

There are different ways of sharing bandwidth between different users. One, commonly used, is *frequency division multiple access* or FDMA. The radio spectrum is devided into narrow frequency channels, each of them to be used by one user. In *time division multiple access* or TDMA, the bandwidth is divided into time periodic slots in which users are allowed to transmit in a consecutive order. TDMA is for example used in the European GSM digital cellular standard. Combinations of these schemes are possible and are used in practice.

In *code division multiple access* or CDMA all users share the same frequency band at all times. The separation between users is done by spreading each user's data sequence with its distinct pseudonoise spreading sequence. After spreading, all users are transmitting on the same radio channel simultaneously. The basestation has knowledge of all the PN-sequences and time delays and can therefore despread and demodulate all users' transmissions.

Although all three schemes are theoretically able to accomodate the same number of user per bandwidth, CDMA has some interesting features that increase the capacity of the system. We will discuss these features in the following paragraphs.

Soft Capacity Limit

The number of channels in an FDMA or the number of timeslots in a TDMA system is fixed. If all channels or slots in a cell are occupied, a new caller will get a busy signal and his request has to be rejected. Moreover, since users are mobile, they will cross the cell borders and calls have to be passed on to the new cell. If the new cell has no space to accomodate the call to be handed over, it will be dropped. FDMA and TDMA have a hard capacity limit; either they can accept a user or they have to reject him. In CDMA such a limit does not exist. Starting from the second user on, every active call introduces some interference to all other calls and the overall interference increases gradually. It is therefore always possible to allow another user to enter the cell, by increasing the bit error rate equally for all user in the system. Definitely, CDMA has also a capacity limit, however the system degrades gracefully.

Soft Handoff

The cellular idea was one solution to solve the problem of the limited bandwidth resource. The coverage area is divided into different regions called *cells*. Each cell uses one set of frequencies. All its immediate neighbouring cells use different frequencies band to avoid interference between cells. A frequency band can be reused in an other cell far enough from a cell using the same band. Depending on the area, frequencies and antennas used, the number of cells per area can be increased, offering an almost unlimited capacity for the system. However, each cell needs a base station and the number of cells determine greatly the costs of the system. Moreover, user are crossing the cell boundaries during conversations and calls have to be passed on to the new cell. This process is referred to as *hand off*. The mobile station will monitor the signal of the original and the new cell and will switch, when the signal from the new cell is stronger and exceeds a certain threshold. If the signal of the original cell was just temporarily weaker because of fading or shadowing, the mobile will switch back and probably forth again, causing what is known as the *ping-ponging* effect. If a new cell cannot offer a free time slot or frequency channel, the call from the original cell has to be dropped.

One feature of CDMA cellular systems is, that the frequency reuse factor is one[3] as compared to 7 in a TDMA or FDMA system, or in other words, the same frequency band is used in all cells and separation between cells is done by different PN-sequence shifts. Therefore, no switching of frequencies, but also no frequency planning for the design of the system is necessary. A mobile travelling in the boundary of two cells will be served by both cells until it has established a firm link to the new cell. This procedure, make-before-break, in combination with the soft capacity limit previously discussed, reduces significantly the probability of a dropped call.

[3]at least in theory

Voice Activity

In a typical two-way conversation, the duty cycle of each voice is at most 50 percent. To exploit this voice activity factor in a FDMA or TDMA system is difficult, because of the problem of reassigning the unused capacity of the channel or time slot to a different user, and the control scheme that would be necessary for this. In a CDMA system, the transmission rate, when there is no speech, can be reduced resulting in an overall decreased interference for all other user in the system, without the need of having any control scheme. The CDMA system capacity is directly related to the interference of other user in the system and can therefore be increased. As stated in [29], the system capacity can be increased by approximately a factor of two, through the exploitation of the voice activity factor.

Capacity for the CDMA system

We denote the received signal and interference power by S and I, respectively. The ratio S/I can then be described as

$$\frac{S}{I} = \frac{R_d \cdot E_b}{W \cdot N_0} \tag{2.4}$$

where R_d is the data transmission rate, E_b is the signal energy per bit, W is the system transmission bandwidth and N_0 is the power spectral density of the interference. If we assume that all users' received signals have equal power and other sources of interference are neglected, the interference power can be expressed as

$$I = S \cdot (K - 1) \tag{2.5}$$

where K is the number of users in the system. This results in the following equation (2.6) for the capacity of a single cell CDMA system.

$$K = \frac{W \cdot N_0}{R_d \cdot E_b} + 1 \tag{2.6}$$

An important parameter for reliable transmission is the ratio $\frac{E_b}{N_0}$. It has been shown, that with interleaving and error correcting codes, an SNR of 7dB is sufficient [29]. $\frac{W}{R_b}$ is generally referred to as the *processing gain* and we can see that the capacity increases with this parameter.

Privacy

The multiplication of the signal with the PN sequence adds an additional feature to the system. Without knowledge of the sequence a decoding of the data transmissions is impossible. The encryption of the messages is inherent in the system design.

New Services

Future mobile communication systems will have to be able to support various services, like image transmission, file transfers and facsimile. These services will require the implementation of different data rates in the system. The CDMA scheme allows user to send data asynchronously, and there are no restrictions on the used data rates, because each user's transmission appears like noise to other users. The performance of different CDMA schemes to support multiple data rates has, for example, been investigated in [11], [13].

Frequency Diversity

The mobile radio channel is usually modelled as a frequency selective Rayleigh fading channel. Different parts of the spectrum are subject to independent fading as long as these parts are separated by at least the coherent bandwidth of the fading process [30]. A wideband system, like CDMA, is capable of separating different path components and can recombine them to improve the performance as it is done in the *RAKE* receiver [30]. If the different path components are not resolved by the system, they will still be rejected and add up as additional interference as long as their time delays are longer than one chip time T_c.

Drawbacks of CDMA

Besides many desired features, there are some problems involved with the use of CDMA. One is the problem of PN-sequence acquisition and tracking. For the receiver to despread the signal it has to align its PN-sequence to the transmitted PN-sequence. This process is called acquisition (coarse alignment) and tracking (fine adjustment). The tracking has to be done with high accuracy to achieve a good system performance. This is particularly difficult because of the short chip time T_c. Another issue is to obtain a set of spreading sequences with good properties for all possible time shifts. Generally, there is some cross correlation between all users in the system. This spurious cross-talk can cause problems if the received users' powers are very different. Transmitters closely located to the basestation make reception of weak users, usually further apart from the receiver, impossible. This scenario is known as the *near-far* problem. One solution is the use of stringent power control [29]; another approach, which will be presented in the following section, is the use of a more sophisticated receiver.

2.3 Multiuser Detection

The issue of multiuser detection in a DS/CDMA system has been a research topic over the past 10 years and numerous publications in that area have been made, e.g. [1], [3], [14], [16]. This section gives a brief insight into some of the ideas suggested. The interested reader can find a good starting point in [26] and [15] and its references.

Many important issues have to be considered in a mobile communication system and the question of how to achieve a low bit error rate is only one of it. Aspects like the size of the mobile units, battery restrictions and a low price are other important aspects. These issues yield to the topic of complexity and the realizeability of many promising ideas. The next section will introduce some ideas of multiuser detection schemes for DS/CDMA systems.

To show the principle of some multiuser receivers, a simple DS/CDMA system model is introduced at this point. A generalized asynchronous system model will be introduced in the next chapter "System Description".

2.3.1 Two-User Synchronous DS/CDMA System Model

Consider a baseband system as shown in figure 2.4. This system represents a two-user DS/CDMA system using binary phase shift keying. It is assumed that the amplitudes of both transmissions are known, that the phase shifts are zero and both user operate fully synchronized on the common channel.

Although this system is simple, it allows us to explain the main principles of different receiver concepts and can later be generalized to the K user case.

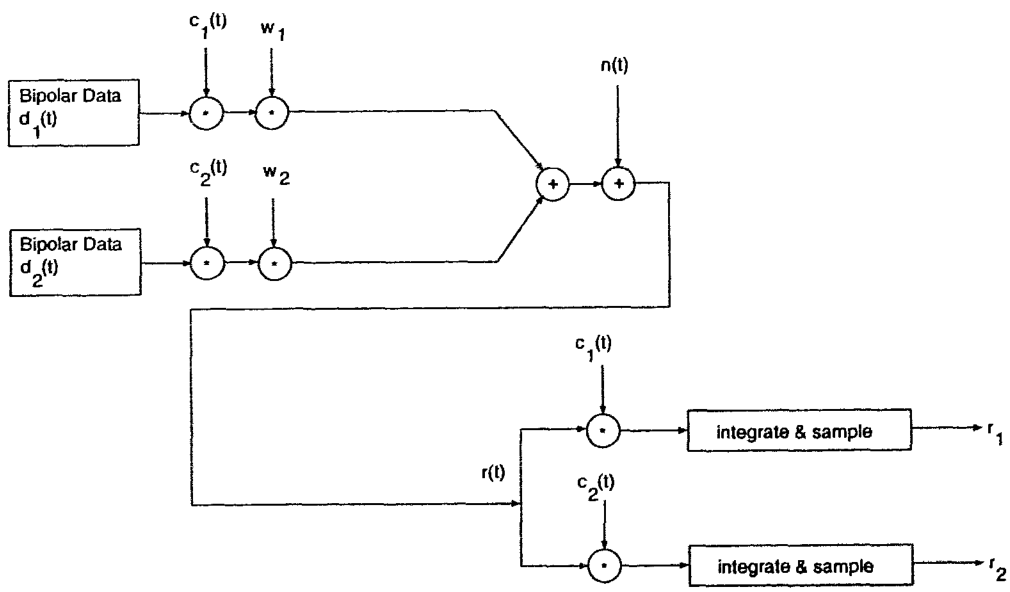

Figure 2.4: Two user DS/CDMA system

The transmitted signals of user one and two can be described by the following equations

$$s_1(t) = d_1(t)\, c_1(t)\, w_1 \tag{2.7}$$

and

$$s_2(t) = d_2(t)\, c_2(t)\, w_2 \tag{2.8}$$

where the w_k are the amplitudes of the transmissions, the $d_k(t)$ are the data sequences and the $c_k(t)$ are the spreading sequences and are defined as follows

$$d_k(t) = \sum_{n=-\infty}^{\infty} d_k[n] \cdot p(t - n \cdot T) \tag{2.9}$$

and

$$c_k(t) = \sum_{n=-\infty}^{\infty} c_k[n] \cdot p_c(t - n \cdot T_c) \tag{2.10}$$

where $d_k[n]$ and $c_k[n]$ are the k-th user data and spreading bit for the n-th time interval. They are random variables chosen from the set $[-1, +1]$ with equal probability. The duration of one bit or chip is denoted by T and T_c, respectively. The functions $p(t)$ and $p_c(t)$ are rectangular pulses of unit height and duration T and T_c, respectively. The spreading sequences in (2.10) are assumed to be periodic in T.

The signals share a common channel, which adds white Gaussian noise. The received baseband signal is therefore

$$r(t) = \sum_{k=1}^{2} d_k(t)\, c_k(t)\, w_k + n(t) \tag{2.11}$$

The term $n(t)$ represents an additive Gaussian noise term with a power spectral density (PSD) of N_0.

The receiver would despread the received signal with the replica of the spreading sequences and would apply them to an integrator and sample the output at T. We are considering the n-th time interval. The output of the integrator for the n-th data bit for both users can then be described by

$$\begin{aligned} r_1[n] &= \frac{1}{T}\int_{n\cdot T}^{(n+1)\cdot T} r(t)\cdot c_1(t)\,dt \\ &= w_1\,d_1[n] + w_2\,d_2[n]\,b(1,2) + e_1[n] \end{aligned} \tag{2.12}$$

and

$$\begin{aligned} r_2[n] &= \frac{1}{T}\int_{n\cdot T}^{(n+1)\cdot T} r(t)\cdot c_2(t)\,dt \\ &= w_2\,d_2[n] + w_1\,d_1[n]\,b(1,2) + e_2[n] \end{aligned} \tag{2.13}$$

where $b(i,j)$ is the correlation between the i-th and j-th user spreading sequences and is defined as

$$b(i,j) = \frac{1}{T}\int_0^T c_i(t-\tau_i)\,c_j(t-\tau_j)\,dt \tag{2.14}$$

The term $e_k[n]$ is the k-th matched filter noise output at time instant n and can be written as

$$e_k[n] = \frac{1}{T}\int_{n\cdot T}^{(n+1)\cdot T} n(t)\cdot c_k(t-\tau_k)\,dt \tag{2.15}$$

We can see that the correlator outputs for both users consist of three different terms. Consider for example (2.12). The first term is the desired signal component for user number one. Term two represents the interference of user number two's transmitted bit for the time interval n. Finally, the third term is a noise term. Term two shows that the interference is caused by the non-zero spreading sequences. In a scenario with different received power levels, the second term can block out the signal completely, making it therefore impossible to detect the desired signal reliably.

The receiver would now base its decisions on (2.12) and (2.13) and this can be done either directly, which is performed in the conventional receiver, or after processing these outputs further.

In the following sections we will consider four different receiver concepts and will briefly discuss their advantages and disadvantages. However, as already pointed out, all receivers present a tradeoff and there is no optimal solution for all aspects a system designer has to take into account in a practical implementation. Furthermore, this collection can only represent a small number of ideas presented in the literature.

2.3.2 The Conventional Receiver

The conventional detector treats other users' transmissions simply as noise and detects every user as if it were the only one in the system. In a coherent BPSK system it would take the filter outputs in (2.12) and (2.13) and would decide simply on the sign of these, i.e.

$$\hat{d}_k[n] = sign(r_k[n]) \tag{2.16}$$

where $sign(\bullet)$ represents the signum function.

This receiver is simple and would be the optimal one, if interference of other users could be perfectly treated as white Gaussian noise. However, in most practical situations, the Gaussian assumption is far too optimistic [15] and does not model the reality with the needed accuracy. Spurious crosscorrelations between spreading sequences and an imbalance in received powers severely degrades the performance for low power users. A solution for this can only be found by using low cross correlation spreading sequences in combination with stringent power control schemes as what is done in IS-95 [29]. However, as stated in [1] even under this conditions, the performance of the conventional receiver is only marginally satisfactory.

In summary, although the conventional receiver is easy to implement, its performance is poor and can only be improved by using a sophisticated power control in the system in combination with a proper spreading sequence set.

2.3.3 Optimal Multiuser Detector

The optimal detector for the signal described in (2.11) was presented in [15]. In this receiver the following operation would be performed

$$\min_{\substack{d_1[n]\epsilon\{-1,+1\} \\ d_2[n]\epsilon\{-1,+1\}}} \int_{nT}^{(n+1)T} |r(t) - \sum_{k=1}^{2} w_k\, d_k(t)\, c_k(t)|^2 dt \tag{2.17}$$

i.e. the receiver has to find the minimum noise energy for all four possible combinations of $d_1[n]$ and $d_2[n]$. This result will be the solution to the optimum demodulation problem. The optimal receiver is required to have knowledge about the received signal energies w_1 and w_2. For the more general case of K user the receiver would have to perform a search over 2^K possible combinations, which is clearly a computational burden that is difficult to handle for a large user population.

2.3.4 Decorrelating Receiver

Consider again (2.11) and represent the signal in matrix form. We could then describe the integrator outputs for both user by

$$\mathbf{r}(n) = \mathbf{\Phi}\, \mathbf{D}(n)\, \mathbf{w} + \mathbf{e}(n) \tag{2.18}$$

where the vector of matched filter outputs is

$$\mathbf{r}(n) = (r_1[n], r_2[n])^t \tag{2.19}$$

and the cross correlation matrix $\mathbf{\Phi}$ is defined as

$$\mathbf{\Phi} = \begin{pmatrix} b(1,1) & b(1,2) \\ b(2,1) & b(2,2) \end{pmatrix} \tag{2.20}$$

and $b(i,j)$ is defined in (2.14). The data matrix $\mathbf{D}(n)$ can be written as

$$\mathbf{D}(n) = \begin{pmatrix} d_1[n] & 0 \\ 0 & d_2[n] \end{pmatrix} \tag{2.21}$$

Furthermore, the amplitude vector can be written as

$$\mathbf{w} = (w_1, w_2)^t \tag{2.22}$$

and

$$\mathbf{e}(n) = (e_1[n], e_2[n])^t \tag{2.23}$$

represents the noise vector. The notation $(\bullet)^t$ stands for the transpose of a vector.

The system of linear equations in (2.18) can be solved by inverting the crosscorrelation matrix $\mathbf{\Phi}$. The decision variables can then be denoted as

$$
\begin{aligned}
\mathbf{z}(n) &= \mathbf{\Phi}^{-1}\,\mathbf{r}\,(n) \\
&= \mathbf{D}\,(n)\,\mathbf{w} + \mathbf{\Phi}^{-1}\,\mathbf{e}\,(n) && (2.24) \\
&= (z_1[n], z_2[n])^t && (2.25)
\end{aligned}
$$

The decisions are then made in a similiar way as for the conventional receiver, i.e.

$$\hat{d}_k[n] = sign(z_k[n]) \tag{2.26}$$

We can see from equation (2.24), that the decision variables in $\mathbf{z}(n)$ are interference free. The only requirement for this receiver to work is that the spreading sequences of the different users are non-identical. Compare this to the conventional receiver, which can only deliver an interference free signal for orthogonal spreading sequences, a much stricter requirement [3]. However, the channel noise at the output of the decorrelator will be amplified depending on the crosscorrelation properties of the spreading sequences and the new noise variances are now

$$\mathbf{e}(n) = (N_0 \cdot \phi_{1,1}^{-1}, N_0 \cdot \phi_{2,2}^{-1})^t \tag{2.27}$$

where $\phi_{k,k}^{-1}$ represents the k-th element on the diagonal of $\mathbf{\Phi}^{-1}$. This receiver obtains the maximum likelihood decisions, if the energies of the different transmissions are unknown [15].

The decorrelating receiver for the synchronous case can be extended to the asynchronous case as shown in [15].

The computational burden of this algorithm lies in the computation of the inverse of the crosscorrelation matrix $\mathbf{\Phi}$, which is of order $O(K^3)$. In a dynamic system the crosscorrelation matrix changes because users are moving[4] or the number of users in

[4]this results in different propagation delays τ_i and τ_j in (2.14)

the cell is changing. For the synchronous case, the computational complexity can be kept on a moderate level, as will be discussed later, by using an efficient algorithm [4]. Section 3.5.1 will focus more on the complexity of these algorithms.

2.3.5 Successive Interference Canceller

One Stage Canceller

The successive interference canceller (SIC) uses a sequential approach instead of the parallel cancellation of the interference that is done in the decorrelator. The main idea is that users are detected according to their received power. The argument is that a decision on a strong user, even under interference introduced by other users transmissions, can be reliably done by just deciding on the matched filter output in (2.12) or (2.13).

After the decision for the strongest user is done, its signal will be regenerated. Assume for now that the first user is the strongest one. Its regenerated signal is then

$$\hat{s}_1(t) = \hat{w}_1\,\hat{d}_1(t)\,c_1(t) \tag{2.28}$$

where $(\hat{\bullet})$ stands for the estimate of a value. In the ideal case, the regenerated signal in (2.28) would be identical to the original signal in (2.7).

The signal would then be subtracted from the composite signal in (2.11) and the new received signal after the first iteration would be

$$\begin{aligned} r_{-1}(t) &= \sum_{k=1}^{2} d_k(t)\,c_k(t)\,w_k + n(t) - \hat{s}_1(t) \\ &= d_2(t)\,w_2\,c_2(t) + d_1(t)\,w_1\,c_1(t) - \hat{w}_1\hat{d}_1(t)\,c_1(t) + n(t) \end{aligned} \tag{2.29}$$

where $(\bullet)_{-k}$ indicates that all signals from the first user up to the k-th user's signal are cancelled from the original received signal. After processing the signal and passing it through the second filter the output would be

$$r_2[n] = d_2[n]\,w_2 + (d_1[n]\,w_1 - \hat{d}_1[n]\,\hat{w}_1)\cdot b(1,2) + e_2[n] \tag{2.30}$$

where $b(i,j)$ is defined in (2.14). Now, if we assume that the amplitudes of the transmissions are known at the receiver site, i.e. $\hat{w}_k = w_k$, then

$$r_2[n] = d_2[n]\,w_2 + w_1\,(d_1[n] - \hat{d}_1[n]\,)\cdot b(1,2) + e_2[n] \tag{2.31}$$

which means that, if the right decision for user one is made, the interferer can be cancelled out completely.

In a practical system $\hat{w}_k \neq w_k$ and the cancellation will be imperfect, even if decisions on the transmitted data were correct. Wrong decisions in preceeding stages of the algorithm are also a cause for errors, since these decisions are used to regenerate the signal. A multistage approach can improve the performance, as will be shown in the next section.

Two-stage Canceller

In the first stage, the strongest user had to be detected in the presence of the second user's interfering signal. If this interference is strong, a wrong decision might have been made and a wrong signal was regenerated and subtracted.

A second stage interference canceller can be implemented that redoes possibly wrong cancellations of the first stage.

Consider the residual signal after the cancellation of all users' signals in the system, i.e. in this case after user two's signal has been subtracted.

$$r_{-2}(t) = \sum_{k=1}^{2} d_k(t)\,c_k(t)\,w_k - \sum_{k=1}^{2} \hat{d}_k(t)\,c_k(t)\,\hat{w}_k + n(t) \tag{2.32}$$

The residual signal in (2.32) contains possibly wrong cancellations. One can now redo the previously done cancellations one by one, starting again from the strongest user. The previously subtracted signal (2.28) for user one will now be added and the new composite signal will be processed again.

The output of the integrator for the first user is then

$$\begin{aligned} r_1^{II}[n] &= \frac{1}{T}\int_{nT}^{(n+1)T} \left(r_{-2}(t) + c_1(t)\,\hat{w}_1\,\hat{d}_1(t)\right) dt \\ &= d_1[n]\,w_1 + d_2[n]\,w_2\,b(1,2) - \hat{d}_2[n]\,\hat{w}_2\,b(1,2) + e_1[n] \end{aligned} \tag{2.33}$$

where $(\bullet)^{II}$ stands for the second stage. A decision is made for the second time by

$$\hat{d}_1^{II}[n] = sign(r_1^{II}[n]) \tag{2.34}$$

The signal will be regenerated again

$$\hat{s}_1^{II}(t) = \hat{d}_1^{II}(t)\,c_1(t)\,\hat{w}_1 \tag{2.35}$$

Finally this new signal estimate can be subtracted again and the new residual signal would then be

$$\begin{aligned} r_{-1}^{II}(t) &= \sum_{k=1}^{2} d_k(t)\,c_k(t)\,w_k - \hat{d}_2(t)\,c_2(t)\,\hat{w}_2 + n(t) - \hat{s}_1^{II}(t) \\ &= \sum_{k=1}^{2} d_k(t)\,c_k(t)\,w_k - \hat{d}_2(t)\,c_2(t)\,\hat{w}_2 + n(t) - \hat{d}_1^{II}(t)\,c_1(t)\,\hat{w}_1 \end{aligned} \tag{2.36}$$

and the same procedure can be repeated for the second user or, in a system with more users, until all user are detected for a second time.

Multi-stage Canceller

If desired, the steps presented in the foregoing section can be implemented for more stages to improve the receiver performance.

To summarize this section, the *Conventional Receiver* is unbeaten in terms of complexity but its performance is quite poor. On the other hand the *Optimal Multiuser Detector* [1] offers a very good performance with the drawback of a complexity almost impossible to implement. Two schemes have been presented, the *Successive Interference Canceller* e.g. [16] and the *Decorrelating Detector* e.g. [3], that offer a compromise that is realizable in a practical application.

A question remains, as how to obtain the channel gain estimates for the different algorithms presented here since the knowledge of them is necessary for these concepts to work. Section 2.5 will offer some solutions to that problem.

2.4 Transmission Channel Model

The channel characteristics for different communication systems are different. For example, a telephone line can be considered as a static channel with additive noise, whereas a mobile radio channel is subject to fading and shadowing and can be modelled by adding noise and multiplying the signal with a gain factor. A satellite channel for a mobile communication link has an additional direct path, which the model has to take into account.

In this section we will introduce a model for a typical mobile radio channel, which consists of short-term Rayleigh fading.

2.4.1 Fading Mobile Radio Channel

A signal transmitted from a transmitter and received by either a mobile or portable unit would propagate over a particular terrain configuration between the two ends. The effect of the terrain configuration generates a long-term fading characteristic which follows a log-normal variation appearing on the envelope of the signal, as can be seen in figure 2.5. Since the antenna of a mobile or portable unit is close to the ground, three effects can be observed [36]. First, the signal received is not only from the direct path but also from the strong reflected path due to the fact that the antennas of the mobile units are close to ground. These two paths create an

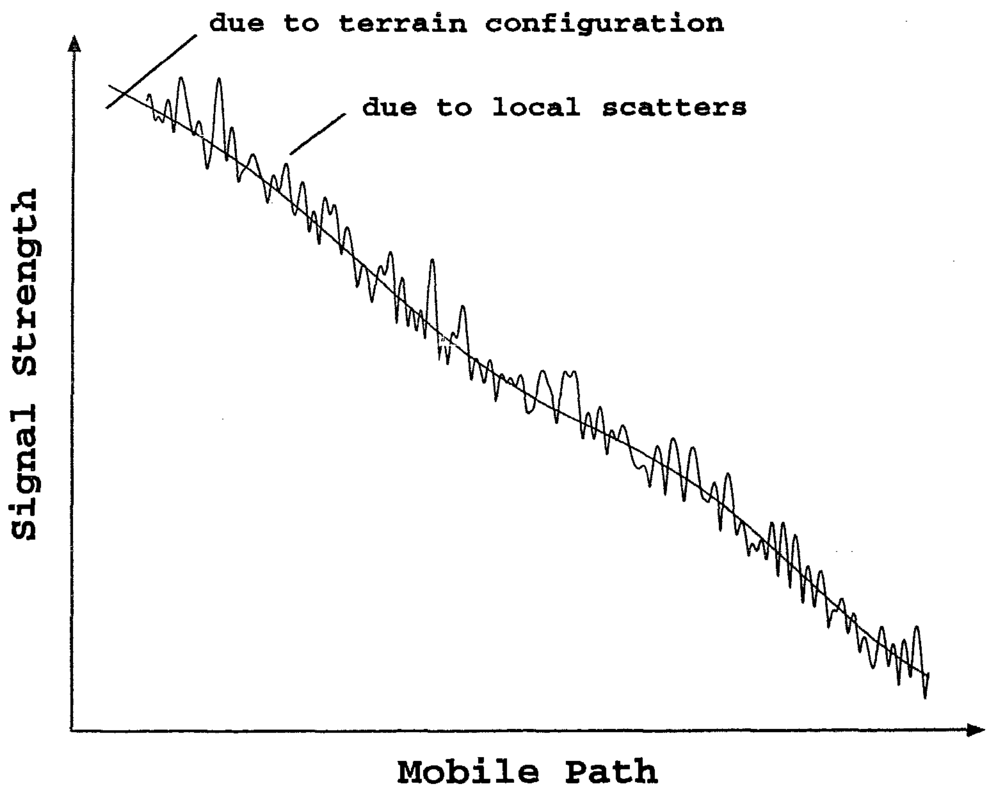

Figure 2.5: Fading signal in the mobile environment

excess path loss which is 40 dB/dec[5], i.e. doubling the path loss in decibels of the free-space path loss. Second, under the low antenna height condition at the mobile units, the man-made structures surrounding them would generate multipath fading on the received signal called Rayleigh fading as shown in figure 2.5. The third factor to mention, is a time delay spread phenomenon which is due to a time dispersive medium. In a mobile radio environment, a single symbol transmitted from one end and received at the other end, receives not only its own symbol but also many echoes of its symbol. The time delay spread intervals are measured from the first symbol to the last detectable echo, which is different for different human-made environments [33]. Figure 2.6 illustrates the two parts of the mobile radio environment. If the time-delay spread is small compared to one symbol time, the fading is called frequency non-selective (or *flat fading*). On the other hand, a large delay spread causes a frequency

[5]4-th power law applied

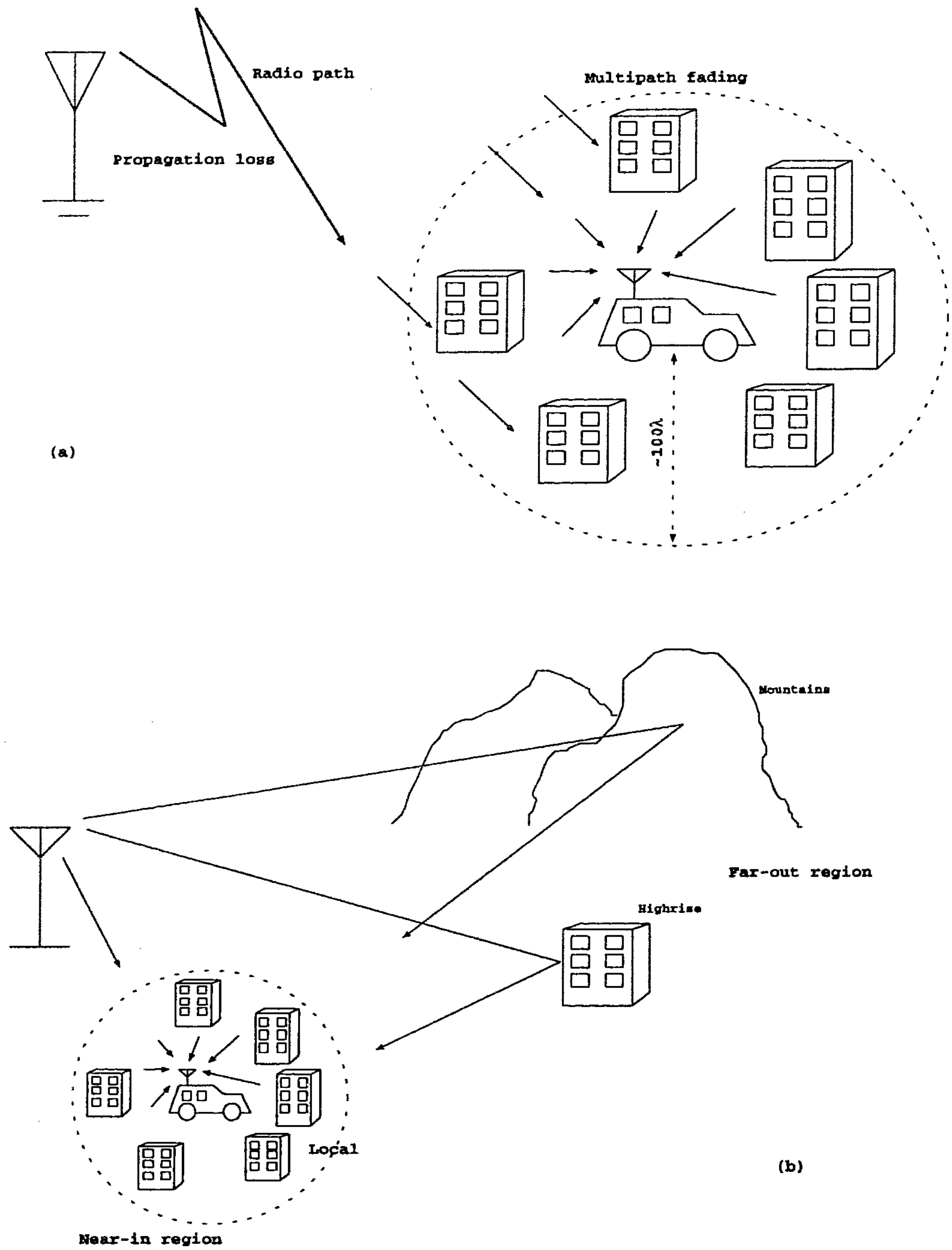

Figure 2.6: Two parts of the mobile fading environment (a) Propagation loss and multipath fading. (b) Time-delay spread scenario.

selective environment, which means that different parts of the frequency band are affected unequally by the attenuation of the channel.

The channel model considered in this thesis, will only take the second factor mentioned, the Rayleigh fading, into account. We assume, that the communication system uses a slow power control to overcome the affects of the path loss and furthermore only a flat fading channel will be considered. The baseband received signal of a signal transmitted over this channel can be described by the following equation

$$r(t) = g(t) \cdot s(t) + n(t) \tag{2.37}$$

where $s(t)$ is the the transmitted signal, $g(t)$ is the multiplicative distortion and $n(t)$ is the additive complex white Gaussian noise with zero mean and a power sprectral density (PSD) of N_0.

The fading is caused by the multipath phenomenon as previously described. The fading gain $g(t)$ can be modelled as a complex Gaussian random process [35]. Its magnitude $|g(t)|$ is Rayleigh distributed and the phase $\arg(g(t))$ is uniformly distributed over $[-\pi, \pi]$. The normalized autocorrelation function of the gain is

$$\frac{1}{2} \cdot E[g(t) \cdot g^*(t-\tau)] = J_0(2 \cdot \pi f_d \tau) \tag{2.38}$$

where $J_0(\bullet)$ is the modified Bessel function of the first kind and zero-th order and f_d is the maximum Doppler frequency or fade rate. The fade rate f_d can be expressed in terms of the vehicle speed v and the carrier frequency f_c as

$$f_d = \frac{v}{c} f_c \tag{2.39}$$

where $c = 3 \cdot 10^8 \frac{m}{s}$ is the speed of light. The normalized fade rate is defined as $f_d \cdot T$ where T is the symbol time.

2.5 Channel estimation

Consider a communication system where the received signal can be described by (2.37), i.e. assuming in the mean time the absence of multiple access interference. Now, $s(t)$ is the baseband equivalent of the transmitted signal and can be written as

$$s(t) = \sum_{n=-\infty}^{\infty} d[n] \cdot p(t - T \cdot n) \tag{2.40}$$

where $d[n]$ is the n-th data symbol and $p(t)$ is a rectangular pulse shape of unit height and duration T. For binary phase shift keying (BPSK) the data symbols $d[n]$ are chosen from the set $[-1, +1]$.

The output of the integrator at the receiver at time instant k can by written as

$$r[k] = d[k] \cdot g[k] + e[k] \tag{2.41}$$

where $g[k]$ is the complex fading gain in the k-th interval[6] and $e[k]$ is the complex Gaussian noise at the output of the filter at time k.

To make a decision on the transmitted symbol, the receiver has to derotate the output $r[k]$ in (2.41), take the real part and decide on the sign of that. The derotation needs the receiver to know the phase shift $\arg(g[k])$ introduced by the channel or $g[k]$ for amplitude modulation schemes or interference cancelling receiver.

The receiver has to estimate this fading gain by means of either one of the following methods, referred to as *Differential Phase Shift Keying, Pilot Tone Assisted Channel Estimation* or *Pilot Symbol Assisted Channel Estimation.*

The following sections will give a brief introduction into these methods.

[6]assumed to be constant over the duration of one symbol

2.5.1 Differential Phase Shift Keying

One method to obtain the phase shift introduced by the channel, which is often used in a practical system, is differential phase shift keying (DPSK). Instead of directly transmitting the absolute phase of the signal, the symbols will be encoded into phase differences between consecutive symbols. For example, in binary PSK, a data bit 1 may be transmitted by shifting the phase of the carrier by 180° relative to the previous carrier phase, while a zero data bit is transmitted by a zero phase shift relative to the phase in the previous signaling interval.

The received signal is demodulated to one of the two possible transmitted phases. Following the demodulator is a phase comparator that compares the demodulated signal phases over two consecutive intervals to extract the original information [30].

Although DPSK offers the advantage of simplicity and introduces no overhead in the transmissions, its performance is about 3 dB worse then that of coherent detection [30]. It also has a significant error floor for moderate to fast fading [17].

A solution to reduce the error floor is to transmit a known signal along with the data. The receiver can extract this signal, generate a complex fading gain estimate and use it for demodulation. Two general methods are used in practice and further improvements on these methods have been found.

We will discuss the two main ideas in the subsequent sections.

2.5.2 Pilot Tone Assisted Channel Estimation

A tone signal is transmitted along with the data signal, that can be extracted by the receiver to generate the phase reference. The spectrum has to have a notch to accommodate the tone signal. The tone signal can then be separated from the received signal by low pass filtering. The problem is to accommodate the tone, since data signals usually do not have a notch in the center of the spectrum. Secondly, creating a notch and including the tone increases the dynamic range of the signal and might yield to a need for amplifier linearisation.

2.5.3 Pilot Symbol Assisted Channel Estimation

In this technique, known pilot symbols are added periodically into the data stream. The frame structure is shown in figure 2.7, where M is the frame size, and P and D denote pilot and data symbols.

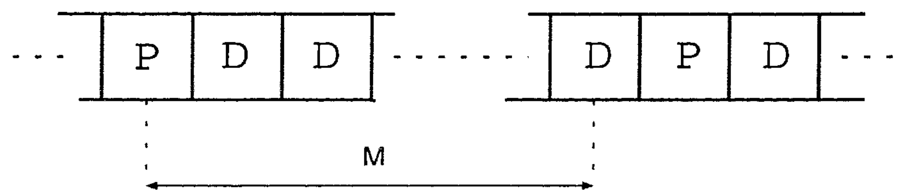

Figure 2.7: Transmitted frame stucture in PSAM

The receiver extracts these pilot symbols and obtains channel estimates at pilot symbol time intervals. An FIR interpolation filter is then used to obtain the fading gain estimates for all other time instants. In [17] the optimal interpolation filter using the mean square error criterion has been presented. The pilot symbols have to be added according to the worst case fade rate, i.e.

$$M < \frac{1}{2 f_d T} \tag{2.42}$$

where f_d is the fade rate and T is the symbol time. Pilot symbol assisted modulation reduces significantly the error floor, whereas the data spectrum remains unchanged and there is no increase in the dynamic range.

The drawbacks of using embedded references is the increase in complexity on the receiver site, the delay due to estimation and some bandwidth or transmission time is spent on the reference signals.

Despite true disadvantages, we will use in this study the pilot symbol technique for channel estimation.

2.6 Summary

This chapter has given some review and background information that is needed in the subsequent chapters. A simple two-user DS/CDMA system model was introduced to explain the main principles of multiuser detection schemes. This model will be generalized and extended to the flat fading environment in the next chapter. The discussion about interference cancelling receiver will be done in more depth, when a pilot symbol assisted cancellation scheme will be introduced.

Chapter 3

System Description

This chapter will give a detailed description of the system that was considered in our studies. For this purpose the simple synchronous two-user model from chapter 2 will by generalized to the asynchronous K–user case. Other issues that will be discussed are a successive interference cancellation receiver, the decorrelator and the pilot symbol assisted channel estimation scheme.

3.1 General Model of a DS/CDMA System

We will extend now the synchronous two-user system in figure 2.4 to the K user asynchronous case and will also introduce the fading channel into the model. We are considering the reverse link (mobile to base station) of a single cell DS/CDMA system. The modulation scheme is assumed to be binary phase shift keying (BPSK) and the number of user in the system is K.

The transmitted signal for the k-th user can be described by

$$s_k(t) = A_k \cdot d_k(t) \cdot c_k(t) \cdot \cos(\omega\, t) \tag{3.1}$$

where A_k is the k-th user transmitted amplitude, $d_k(t)$ is the k-th user data sequence, $c_k(t)$ is the k-th user spreading sequence and ω is the common carrier frequency. The spreading and data waveforms have been defined in (2.9) and (2.10). The spreading sequences are assumed to be periodic in T, i.e.

$$c_k(t) = c_k(t + T \cdot i) \tag{3.2}$$

where i is an arbitrary integer.

All users share a common channel which introduces flat fading and additive white Gaussian noise (AWGN). The base station would multiply the received superposition of all time shifted signals with $\cos(\omega t)$ and $\sin(\omega t)$ followed by a low pass filter to obtain the inphase and quadrature components of the baseband signal. The block diagram of the equivalent baseband system can be seen in figure 3.1.

The received baseband signal at the base station can be expressed as

$$r(t) = \sum_{k=1}^{K} g_k(t - \tau_k) \cdot d_k(t - \tau_k) \cdot c_k(t - \tau_k) + n(t) \tag{3.3}$$

where $g_k(t)$ is defined as

$$g_k(t) = A_k \cdot h_k(t) \tag{3.4}$$

The A_k is the amplitude of the transmitted k-th user's signal, the $h_k(t)$ is the complex fading gain experienced by the k-th user and τ_k is the k-th user's time shift, which is due to the different locations of the transmitters in the cell, and assumed to be a random variable in $[0, T)$. The term $n(t)$ is an additive white Gaussian noise (AWGN) term with a power spectral density of N_0. The number of users in the system is assumed to be K. The complex fading gains $h_k(t)$ are independent, each having an autocorrelation function of $J_0(2\pi f_k \tau)$, were J_0 is the zero-th order Bessel function and f_k is the maximum Doppler frequency of the k-th user. The $g_k(t)$ are therefore independent with an autocorrelation function of $\sigma_k^2\, J_0(2\pi f_k \tau)$ and σ_k^2 is the variance of $g_k(t)$ which is

$$\sigma_k^2 = A_k \cdot A_k^* \tag{3.5}$$

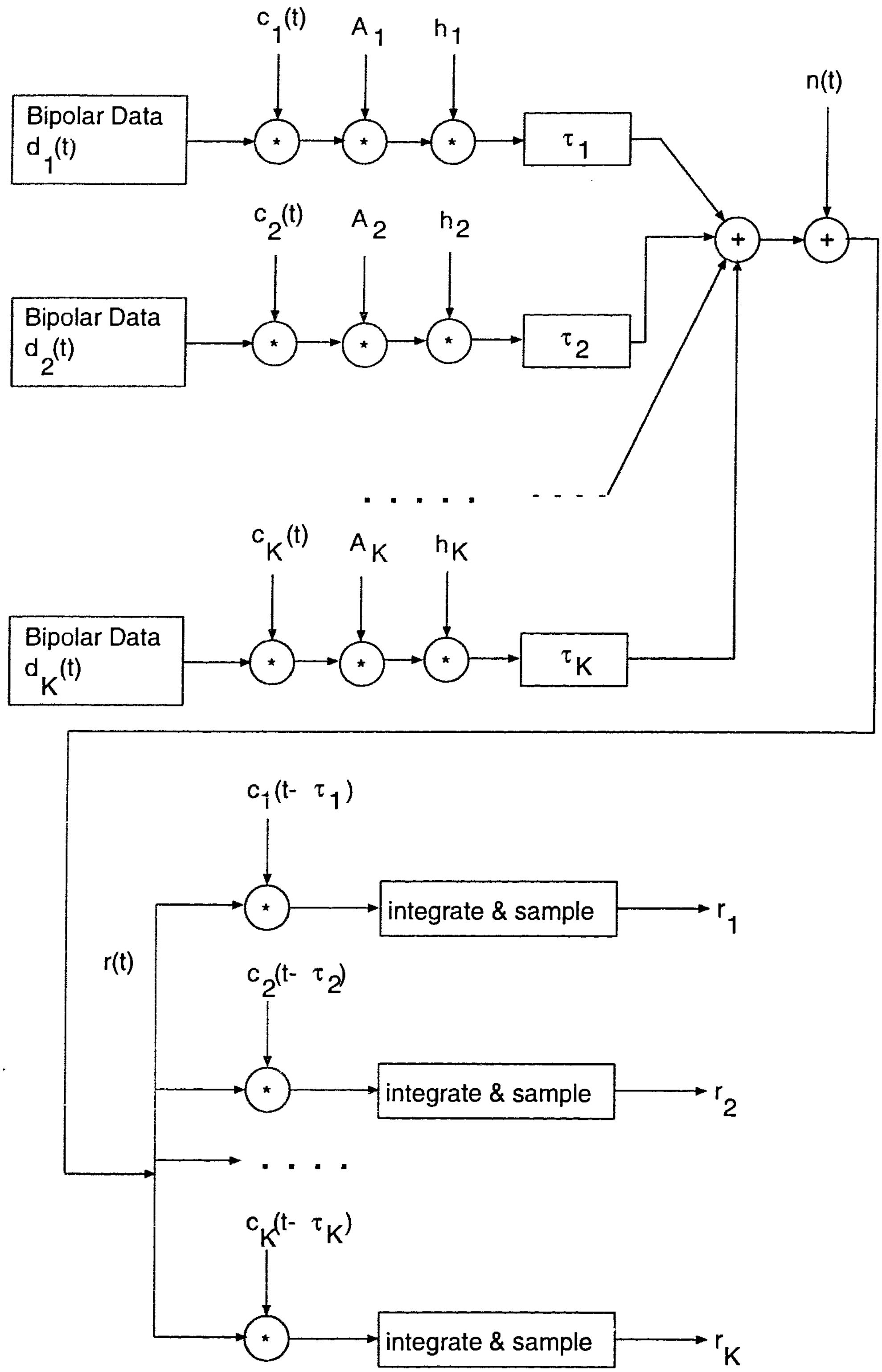

Figure 3.1: General DS/CDMA system block diagram

Consider, without loss of generality, that the users are ordered according to their time delays, i.e. $\tau_1 \leq \tau_2 \leq \ldots \leq \tau_k$.

The receiver's task is now to detect all data transmissions in the received signal in (3.3). It would perform the following operations (figure 3.1) for every user, here described for the k-th user. The signal would be multiplied with the properly time aligned[1] k-th user spreading sequence followed by an integrator. The integrator output for the n-th time interval is then

$$\begin{aligned}
r_k[n] &= \frac{1}{T}\int_{nT+\tau_k}^{(n+1)T+\tau_k} c_k(t-\tau_k)\cdot r(t)dt \\
&= g_k[n]\cdot d_k[n] \\
&\quad + \sum_{j=1}^{k-1} g_j[n]\cdot d_j[n]\, b(j,k) + g_j[n+1]\cdot d_j[n+1]\, a(j,k) \\
&\quad + \sum_{j=k+1}^{K} g_j[n-1]\cdot d_j[n-1]\, a(k,j) + g_j[n]\cdot d_j[n]\, b(k,j) \\
&\quad + e_k[n] \qquad (3.6)
\end{aligned}$$

where $a(k,j)$ and $b(k,j)$ are cross correlations of the spreading sequences of user k and j and can be defined as

$$a(i,j) = \frac{1}{T}\int_{\tau_i}^{\tau_j} c_i(t-\tau_i)\, c_j(t-\tau_j)\, dt \qquad (3.7)$$

and

$$b(i,j) = \frac{1}{T}\int_{\tau_j}^{T+\tau_i} c_i(t-\tau_i)\, c_j(t-\tau_j)\, dt \qquad (3.8)$$

The output of the filter consists of four different terms. The first term in (3.6) is due to the desired signal component and the second and the third term is interference introduced by users with shorter and longer time delay, respectively. Figure 3.2 shows the time delayed transmissions and the crosscorrelations between different users' transmissions. The noise term $e_k[n]$ is defined in (2.15) and reproduced here for convenience,

$$e_k[n] = \frac{1}{T}\int_{n\cdot T+\tau_k}^{(n+1)\cdot T+\tau_k} n(t)\cdot c_k(t-\tau_k)\, dt \qquad (3.9)$$

We will introduce the matrix notation at this point for the received signal structure

[1]which we assume, has been done by the acquisition and tracking process

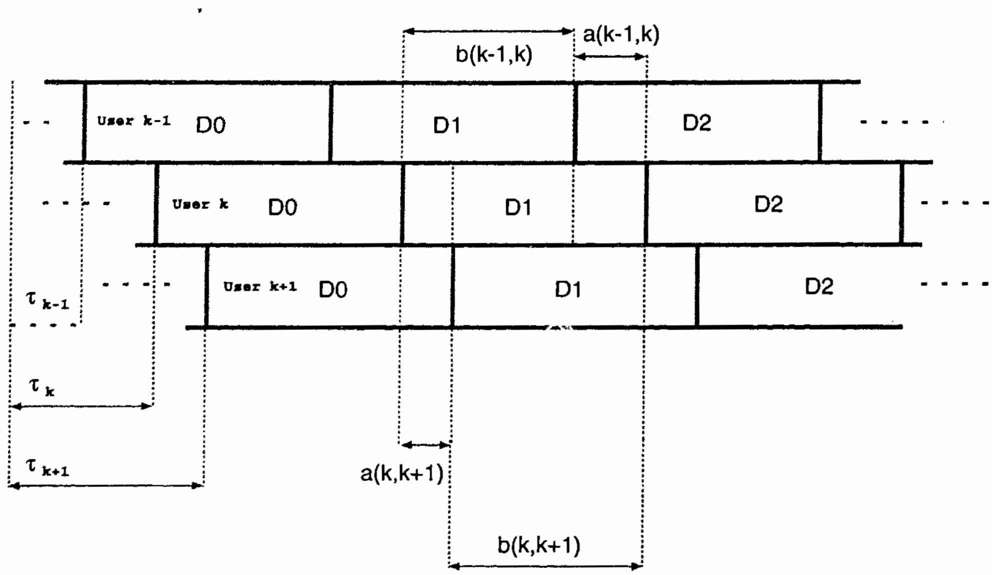

Figure 3.2: Delayed transmissions and crosscorrelations in asynchronous CDMA

of the asynchronous received transmissions and can express (3.6) for all users as

$$\mathbf{R}(n) = \sum_{m=-1}^{1} \mathbf{\Phi}_m \, \mathbf{D}(n-m) \, \mathbf{G}(n-m) + \mathbf{e}(n) \tag{3.10}$$

where

$$\mathbf{R}(n) = (r_1(n), r_2(n), \ldots, r_K(n))^t \tag{3.11}$$

is the received vector,

$$\mathbf{D}(n) = \begin{pmatrix} d_1[n] & 0 & 0 & \ldots & 0 \\ 0 & d_2[n] & 0 & \ldots & 0 \\ 0 & 0 & d_3[n] & \ldots & 0 \\ \vdots & \vdots & \vdots & \vdots & \vdots \\ 0 & 0 & 0 & \ldots & d_K[n] \end{pmatrix}, \tag{3.12}$$

is the n-th diagonal data matrix,

$$\mathbf{G}(n) = (g_1[n], g_2[n], \ldots, g_K[n])^t \tag{3.13}$$

is the n-th fading gain vector with $g_k[n]$ being the value of $g_k(t)$ in the interval $[nT, (n+1)T]$ [2], and

$$\mathbf{e}(n) = (e_1[n], \ldots, e_K[n])^t \tag{3.14}$$

[2]We assume a piecewise constant model for the fading process

is the n-th noise vector.

The matrices $\mathbf{\Phi}_{-1}$, $\mathbf{\Phi}_0$, and $\mathbf{\Phi}_1$ are correlation matrices of the spreading sequences with

$$\mathbf{\Phi}_{-1} = \begin{pmatrix} 0 & 0 & 0 & \dots & 0 \\ a(1,2) & 0 & 0 & \dots & 0 \\ a(1,3) & a(2,3) & 0 & \dots & 0 \\ \vdots & \vdots & \vdots & \vdots & \vdots \\ a(1,K) & a(2,K) & a(3,K) & \dots & 0 \end{pmatrix}, \tag{3.15}$$

being a lower triangular matrix,

$$\mathbf{\Phi}_0 = \begin{pmatrix} b(1,1) & b(1,2) & b(1,3) & \dots & b(1,K) \\ b(1,2) & b(2,2) & b(2,3) & \dots & b(2,K) \\ b(1,3) & b(2,3) & b(3,3) & \dots & b(3,K) \\ \vdots & \vdots & \vdots & \vdots & \vdots \\ b(1,K) & b(2,K) & b(3,K) & \dots & b(K,K) \end{pmatrix}, \tag{3.16}$$

a symmetric matrix, and

$$\mathbf{\Phi}_1 = \mathbf{\Phi}^t_{-1} \tag{3.17}$$

an upper triangular matrix. The terms $a(i,j)$ and $b(i,j)$ are partial correlations of the spreading waveforms and are defined as in (3.7) and (3.8). Finally, the covariance matrix of the noise vector $\mathbf{e}(n)$ is

$$\mathbf{\Phi}_{ee} = N_o\mathbf{\Phi}_0 \tag{3.18}$$

Equation (3.10) represents a vector channel with a time-varying (matrix) impulse response of duration equal to three samples. In the case of synchronous reception, the $a(i,j)$'s are all zero and $b(i,j)$ is the full correlation over one symbol interval and $\mathbf{R}(n)$ reduces to

$$\mathbf{R}(n) = \mathbf{\Phi}_0\mathbf{D}(n)\mathbf{G}(n) \tag{3.19}$$

When long PN sequences are used for spreading, then the matrices $\mathbf{\Phi}_{-1}$, $\mathbf{\Phi}_0$, and $\mathbf{\Phi}_1$ become time-varying.

3.1.1 Quasi-synchronous DS/CDMA System

In most parts of this thesis, a quasi-synchronous system has been considered. This section will describe what is meant by that.

In a quasi-synchronous system, it is assumed that all mobile users in the cell are synchronized by the base-station. However, users are operating at different locations in the cell and therefore, the composite received signal at the base station is asynchronous, i.e. the τ_k in equation (3.3) are non-zero. By quasi-synchronous we simply mean that the maximum time delay τ_{max} of all users in the system does not exceed a certain limit. This can be guaranteed by limiting the cell size to a certain radius. For example, if for any reason we do not want the time delays to exceed a value of T, then the cell radius should not exceed

$$R = \frac{T \cdot c}{2} \tag{3.20}$$

where $c = 3 \cdot 10^8 \frac{m}{s}$ is the speed of light. If we assume a data rate of $9600 \frac{bit}{s}$, we could operate in a cell cite of radius $15.6\,km$. The reason for considering a quasi-synchronous system is that the assumption assures the partial or complete overlap of pilot symbol transmissions of different users which can be useful for the potential receiver.

3.2 Pilot Symbol Assisted Channel Estimation

As mentioned in Section 2.5, a receiver has to perform channel estimation to detect the received signal properly. For a system using PSK the estimation of the phase shift introduced by the channel would be sufficient. However, if the modulation scheme uses some kind of amplitude modulation, like quadrature amplitude modulation (QAM), a simple derotation of the signal vector is insufficient and a scaling operation has to be done in addition. The same statement can be made for an interference cancelling receiver, since the detection of the signal is not the only task that has to be performed. The signal has also to be regenerated and removed, which would be impossible without knowledge of both the amplitudes and phases[3].

One method that can be used for the gain estimation is the *Pilot Symbol Assisted Estimator*. One well known publication that deals with this kind of estimation technique is by Cavers [17]. In this paper the optimal pilot symbol assisted estimator in terms of the mean square error (MSE) is derived and analyzed.

For the purpose of channel estimation pilot symbols are inserted periodically into the data stream. The users are transmitting their data in frames of length M, with M_d data symbols and $M_p = M - M_d$ pilot symbols. Usually, the number of inserted pilot symbols per frame is $M_p = 1$. However, the problem of multiuser access interference is also present in the pilot symbol interval and in a quasi-synchronous multiuser CDMA system the insertion of more pilot symbols might be beneficial to reduce this interference and supply the receiver with better gain estimates. Three different pilot symbol insertion techniques have been considered in the study and will be discussed in more detail.

Insertion of One Pilot Symbol

The transmitted frame structure can be seen in figure 3.3. The frame length will be denoted by M and the pilot symbols are inserted at time instants $n = \ell M$ with ℓ an integer. The letters P and D in figure 3.3 represent a pilot or data symbol, respectively. This method of inserting pilots has the advantage of low overhead, since only one symbol per frame is used for the channel estimation. However, the case of only one pilot symbol per frame in an asynchronous system also means that an SIC receiver is needed for the interference suppression in the pilot interval which limits the accuracy of the estimates. The performance of such a receiver will later be investigated in chapter 4.

[3]the decorrelator requires only the knowledge of the phases

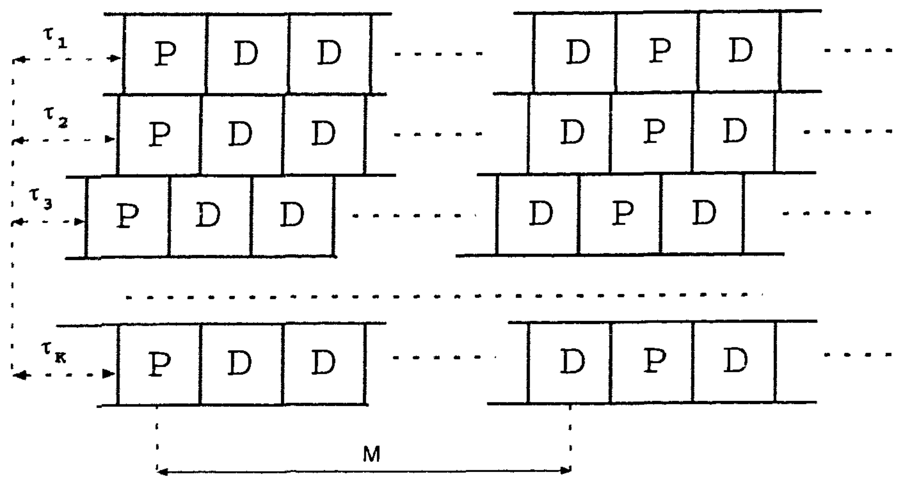

Figure 3.3: Transmitted Frame Structure, One Pilot Symbol

Three Pilot Symbols

The maximum likehood receiver for a synchronous channel, if the signal energies are unknown, is the synchronous decorrelator [3] as introduced in chapter 2. Its complexity is comparatively low since it only involves a matrix inversion of order $K \cdot K$. Furthermore, efficient updating algorithms for synchronous decorrelator are available [4], [26] and, since the decorrelator can be seen as a zero-forcing equalizer, the readily available technology for high-speed zero-forcing equalizers makes the decorrelator one of the simplest IC techniques to implement [26]. Keeping these points in mind, using this receiver concept seems therefore beneficial to decorrelate the pilot symbols and obtain an interference free signal.

Lets consider equation (3.10) again. If the fading is reasonably slow and the noise power is small, then $\mathbf{G}(n-1) \approx \mathbf{G}(n) \approx \mathbf{G}(n+1)$ and

$$\mathbf{R}(n) \approx \left(\sum_{m=-1}^{+1} \mathbf{\Phi}_m \mathbf{D}(n-m) \right) \mathbf{G}(n) \tag{3.21}$$

Furthermore, if all three data matrices are identical, then

$$\mathbf{R}(n) = \mathbf{\Phi}\, \mathbf{D}(n)\, \mathbf{G}(n) \tag{3.22}$$

where

$$\mathbf{\Phi} = \mathbf{\Phi}_{-1} + \mathbf{\Phi}_0 + \mathbf{\Phi}_{+1} \tag{3.23}$$

is the full correlation between spreading sequences. This suggests that for channel estimation purpose in a quasi-synchronous system with $\tau_{max} \leq T$ we can use $M_p = 3$ identical pilot symbols per frame and the estimated fading gain vector can then be obtained by first decorrelating the signals of the users, followed by by pilot-symbols removal

$$\hat{\mathbf{G}}(\ell M) = \mathbf{P}^{-1}\,\mathbf{\Phi}^{-1}\,\mathbf{R}(\ell M) \tag{3.24}$$

where it is assumed that the pilot symbols are located at $n = \ell M + m$, $m = -1, 0, +1$ and ℓ an integer, and $\mathbf{P} = \mathbf{D}(\ell M)$ is the pilot symbol matrix. Since, $\mathbf{P}$ is a diagonal matrix with each diagonal element either $+1$ or -1, this means $\mathbf{P} = \mathbf{P}^{-1}$. The transmitted frame structure is shown in figure 3.4, where it is already considered that the system is quasi-synchronized and $\tau_{max} \leq T$. The disadvantage of this pilot

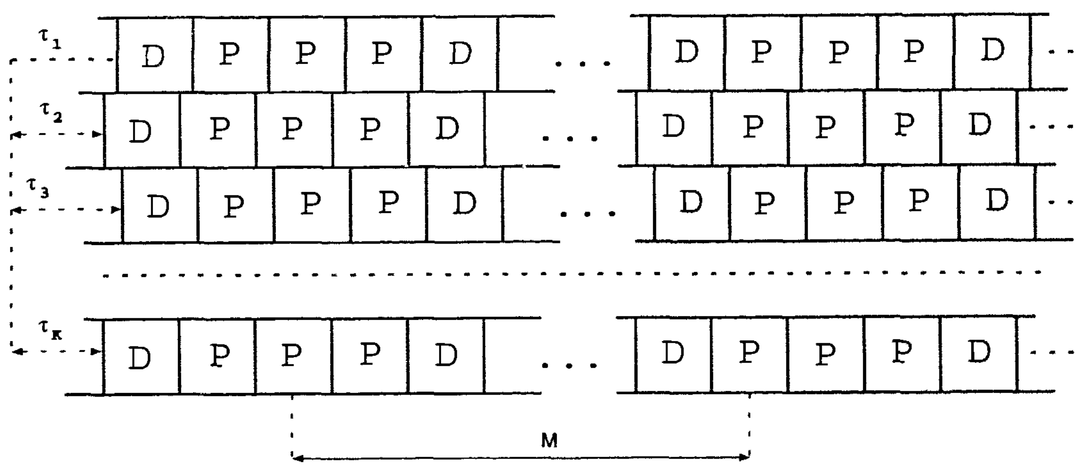

Figure 3.4: Transmitted Frame Structure, Three Pilot Symbols

insertion technique is obviously the large overhead compared to the one-pilot symbol transmission frames.

One Pilot Symbol with Guard Intervals

A third method of inserting pilot symbols, also considered in the quasi-synchronous case, can be described as ollows. Pilot symbols will be inserted at time intervals $n = \ell M$ and the transmitters will be turned off for time instants $n = \ell M - 1$ and

$n = \ell M + 1$, i.e. we will insert a pilot symbol preceded and followed by a guard interval of time T. Reconsidering (3.10) with $\mathbf{D}(n-1) = \mathbf{D}(n+1) = 0$, this means

$$\mathbf{R}(n) = \mathbf{\Phi}_0 \, \mathbf{D}(n) \, \mathbf{G}(n) \tag{3.25}$$

The estimated fading gain vector can now be obtained through

$$\hat{\mathbf{G}}(\ell M) = \mathbf{P}^{-1} \, \mathbf{\Phi}_0^{-1} \, \mathbf{R}(\ell M) \tag{3.26}$$

It can be seen from figure 3.5, that in a quasi-synchronous system the interference during the pilot symbol time can be reduced and a performance increase can be expected as compared to the insertion of one pilot symbol described before. Furthermore, the

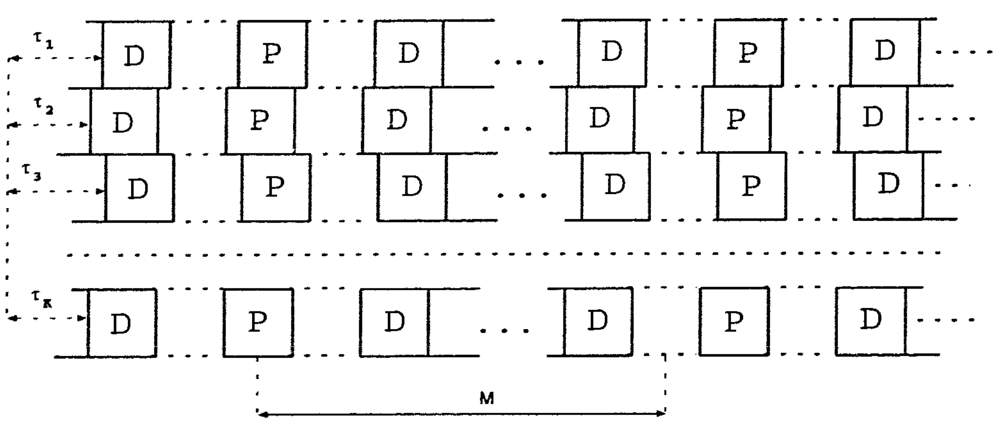

Figure 3.5: Transmitted Frame Structure, One Pilot Symbol, Guard Intervals

decorrelator performance is independent of the fading gain change over three symbol intervals and a better performance can be expected for fast fading channels, i.e. for $\mathbf{G}(n-1) \not\approx \mathbf{G}(n) \not\approx \mathbf{G}(n+1)$. The frame structure presented here was also introduced in [6] where it was used in a double SIC scheme to be described later on.

One Pilot Symbol and Shorter Integration Time

The synchronous decorrelator can also be used in a quasi-synchronous system using the one-pilot frames as described before, if we shorten the integration time in the pilot symbol time interval as shown in figure 3.6. Specifically, the receiver would process the signal in the k-th branch as

$$r_k[n] = \frac{1}{T} \int_{nT+\tau_{max}}^{(n+1)T+\tau_{min}} c_k(t-\tau_k) \cdot r(t)\, dt \tag{3.27}$$

where τ_{min} and τ_{max} represent the minimum and maximum time shift of users in the system.

In matrix notation the signal can now (in the absence of noise) be written as

$$\mathbf{R}(n) = \mathbf{\Phi}_s\, \mathbf{D}(n)\, \mathbf{G}(\mathbf{n}) \tag{3.28}$$

where the cross correlation matrix $\mathbf{\Phi}_s$ is defined as

$$\mathbf{\Phi}_s = \begin{pmatrix} a(1,1) & a(1,2) & a(1,3) & \dots & a(1,K) \\ a(1,2) & a(2,2) & a(2,3) & \dots & a(2,K) \\ a(1,3) & a(2,3) & a(3,3) & \dots & a(3,K) \\ \vdots & \vdots & \vdots & \vdots & \vdots \\ a(1,K) & a(2,K) & a(3,K) & \dots & a(K,K) \end{pmatrix}, \tag{3.29}$$

and the cross correlations are now

$$a(i,j) = \frac{1}{T} \int_{\tau_{max}}^{T+\tau_{min}} c_i(t-\tau_i)\, c_j(t-\tau_j)\, dt \tag{3.30}$$

The gain estimates are obtained accordingly

$$\hat{\mathbf{G}}(\ell M) = \mathbf{P}^{-1}\, \mathbf{\Phi}_s^{-1}\, \mathbf{R}(\ell M) \tag{3.31}$$

The shorter integration time leads to an SNR decrease [12] and this approach can only be applied to a medium cell size. Since the interference in the pilot symbols can be completely removed, the pilot symbols could be transmitted with increased power, without introducing additional interference to other users in the system and therefore bringing the SNR during the pilot interval reception to the original level.

The channel estimation process will then be completed by passing on the gain estimates at pilot symbol instants to the Wiener estimation filter [17]. The interpolation filter is a function of the channel SNR and other system parameters. Also, in the presence of multiuser access interference, as in the case when only one pilot symbol is

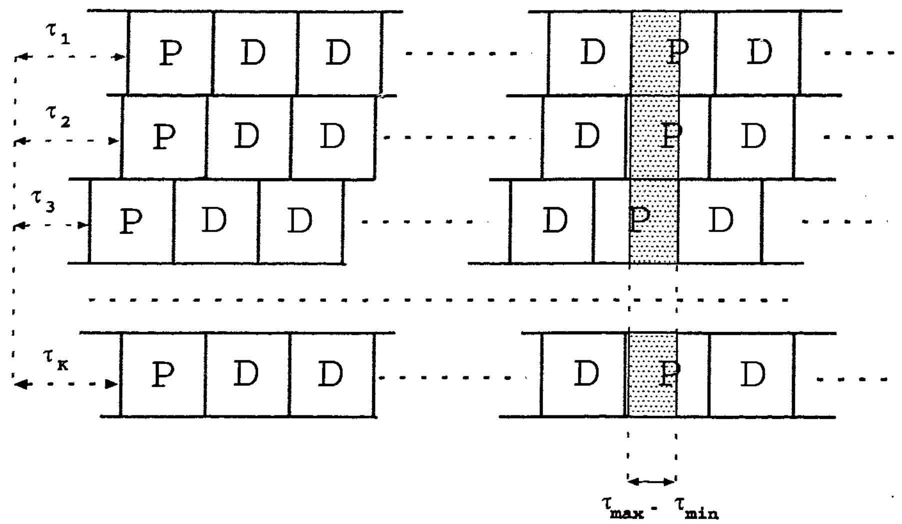

Figure 3.6: Shortened integration time in a quasi-synchronous system for the pilot symbol interval

used, this interference level should be taken into account when designing the interpolator. However, in a practical implementation, one would choose one operating point and optimize the filter for this condition. A mismatch of the filter would then lead to a degradation of the performance as was shown in [17] for the single-user case.

In the investigated system, the filter was optimized to the Doppler frequency and the additive white Gaussian noise. However, the multiuser access interference was not taken into account.

3.3 Successive Interference Cancelling Receiver

The processing in an interference cancelling pilot symbol assisted receiver for a multiuser DS/CDMA channel can be divided into three steps. Since the multiuser access interference problem is also present during the pilot symbol intervals, the receiver has, first of all, to make attempts to reduce the interference for these time instants. Afterwards the channel estimation will be performed and in the third step the actual data will be detected.

One solution for step one was already presented in the foregoing section by using a decorrelator for this purpose. Another approach could be the use of a successive cancellation scheme discussed in this section. For data detection the decorrelator cannot be used in the presented form and the modifications would result in a complexity increase [15]. Therefore, in this thesis a successive cancelling scheme will be used for data detection purpose as well.

This section will describe this scheme and some modifications will be made in the sequel to use this idea for processing step one and three in the receiver.

All SIC make the same assumption that users have to be detected and cancelled out according to their received powers. The argument is simply that a stronger user can be more reliably detected, even in the presence of interference, than a weaker user's signal.

Thus before detecting any data, the receiver has to perform a ranking according to the signal powers. If the SIC would perform in a synchronous system, there would be no ambiguity about how to compare the different users' powers and the comparison could simply be done on a symbol-by-symbol basis.

However, in an asynchronous channel the question of comparing which symbols of one user with which symbols of other users is ambiguous because of the time shifts and the proposed solution for that was the consideration of a block of received symbols [16]. The received powers in this block will be averaged over the frame and a decision about the rank of the users will be made. All symbols in this block will then be detected and the signal afterwards subtracted from the received signal in that block. The grouping of the bits is shown in figure 3.7. The group consists of q symbols for each user, where the maximum time between the first and the last bit is $(q+1)$ symbol times.

We will reconsider the received baseband signal defined in (3.3),

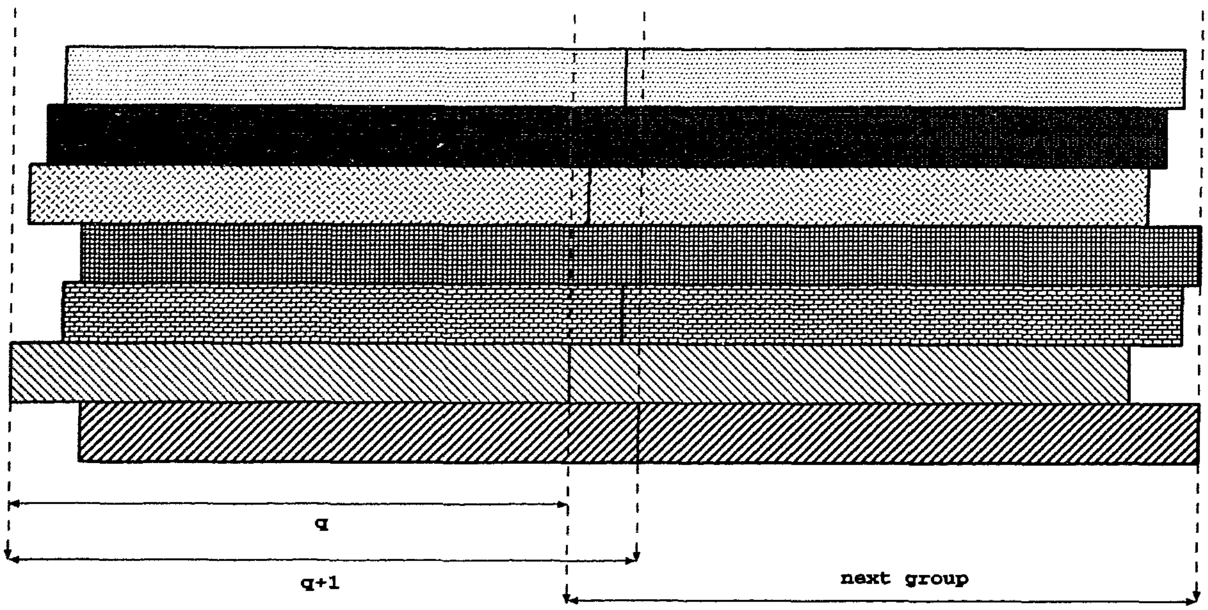

Figure 3.7: Grouping of bits for ranking in an asynchronous channel

$$r(t) = \sum_{k=1}^{K} g_k(t-\tau_k) \cdot d_k(t-\tau_k) \cdot c_k(t-\tau_k) + n(t) \tag{3.32}$$

One assumption made in this section is that the receiver has perfect knowledge about the time delays τ_k of the transmissions. Moreover, we assume the users are ordered according to their time delays, i.e. $\tau_1 \leq \tau_2, \ldots \leq \tau_K$. Finally, we assume the first user has the strongest signal, followed by the second one and so on, i.e. $g_1(t) \geq g_2(t) \geq \ldots \geq g_K(t)$.

3.3.1 One-Stage SIC

Reconsider now the matched filter output in (3.6) for the n-th time interval for the first user which can be represented as

$$\begin{aligned} r_1[n] &= \frac{1}{T}\int_{nT+\tau_1}^{(n+1)T+\tau_1} c_1(t-\tau_1) \cdot r(t)dt \\ &= g_1[n] \cdot d_1[n] \end{aligned}$$

$$
\begin{aligned}
&+\sum_{j=2}^{K} g_j[n-1]\cdot d_j[n-1]\,a(1,j)+g_j[n]\cdot d_j[n]\,b(1,j)\\
&+e_1[n] \qquad (3.33)
\end{aligned}
$$

The output of the filter in (3.33) has then to be derotated which, for a PSK constellation, can be done by a multiplication with $\hat{g}_1^*[n]$, the receiver's estimate of $g_1^*[n]$. The resulting decision variable $z_1[n]$ for the first user is

$$z_1[n] = r_1[n]\cdot \hat{g}_1^*[n] \qquad (3.34)$$

The decision is then made on the real part of $z_1[n]$ by applying the signum function

$$\hat{d}_1[n] = sign(\mathcal{RE}[z_1[n]]) \qquad (3.35)$$

where $\mathcal{RE}[\bullet]$ stands for the real part of a complex value and $sign(\bullet)$ is the signum function. The data for each user will be detected in frames of length q, for reasons that were discussed before. The signal in the block of length q symbols is then regenerated

$$\hat{s}_1(t-\tau_1) = c_1(t-\tau_1)\,\hat{d}_1(t-\tau_1)\,\hat{g}_1(t-\tau_1) \qquad (3.36)$$

where $\hat{d}_1(t-\tau_1)$ is defined over the block of q symbols as

$$\hat{d}_1(t-\tau_1) = \sum_{n=x}^{x+q}\hat{d}_1[n]\,p_d(t-n\cdot T-\tau_1) \qquad (3.37)$$

and x stands for the block number and $p_d(t)$ is the rectangular pulse as used in (2.9). The signal (3.36) is then subtracted from the composite received baseband signal in (3.32)

$$
\begin{aligned}
r_{-1}(t) &= r(t)-\hat{s}_1(t-\tau_1)\\
&= r(t)-c_1(t-\tau_1)\,\hat{d}_1(t-\tau_1)\,\hat{g}_1(t-\tau_1)\\
&= \sum_{l=2}^{K} g_l(t-\tau_l)\cdot d_l(t-\tau_l)\cdot c_l(t-\tau_l)\\
&\quad +g_1(t-\tau_1)\,d_1(t-\tau_1)\,c_1(t-\tau_1)\\
&\quad -\hat{g}_1(t-\tau_1)\,\hat{d}_1(t-\tau_1)\,c_1(t-\tau_1)\\
&\quad +n(t) \qquad (3.38)
\end{aligned}
$$

The notation $(\bullet)_{-k}$ denotes that all signals up to the k-th user's signal have been cancelled from the composite signal.

The signal in (3.38) is now processed by the correlator for the second strongest user, in this case user two, and the output can be written as

$$\begin{aligned} r_2[n] &= g_2[n] \cdot d_2[n] \\ &\quad +(g_1[n] \cdot d_1[n] - \hat{g}_1[n] \cdot \hat{d}_1[n])\, b(1,2) \\ &\quad +(g_1[n+1] \cdot d_1[n+1] - \hat{g}_1[n+1] \cdot \hat{d}_1[n+1])\, a(1,2) \\ &\quad +\sum_{j=3}^{K} g_j[n-1] \cdot d_j[n-1]\, a(2,j) + g_j[n] \cdot d_j[n]\, b(2,j) \\ &\quad +e_2[n] \end{aligned} \tag{3.39}$$

From (3.39) the second decision variable is formed through

$$z_2[n] = r_2[n] \cdot \hat{g}_2^*[n] \tag{3.40}$$

and the decision rule is

$$\hat{d}_2[n] = sign(\mathcal{RE}[z_2[n]]) \tag{3.41}$$

In a similiar manner, after k-1 user have been cancelled out, the received baseband signal $r_{-(k-1)}(t)$ can then be expressed as

$$\begin{aligned} r_{-(k-1)}(t) &= \sum_{l=k}^{K} g_l(t-\tau_l) \cdot d_l(t-\tau_l) \cdot c_l(t-\tau_l) \\ &\quad + \sum_{l=1}^{k-1} [g_l(t-\tau_l)\, d_l(t-\tau_l)\, c_l(t-\tau_l) - \hat{g}_l(t-\tau_l)\, \hat{d}_l(t-\tau_l)\, c_l(t-\tau_l)] \\ &\quad + n(t) \end{aligned} \tag{3.42}$$

The signal consists of three terms, where the first contains the users' transmissions which have not yet been detected, the second is due to imperfect cancellations and the third is the additive white Gaussian noise.

The correlator output for user k can then be represented through

$$\begin{aligned} r_k[n] &= g_k[n] \cdot d_k[n] \\ &\quad + \sum_{l=1}^{k-1} [(g_l[n] \cdot d_l[n] - \hat{g}_l[n] \cdot \hat{d}_l[n])\, b(l,k) \\ &\qquad +(g_l[n+1] \cdot d_l[n+1] - \hat{g}_l[n+1] \cdot \hat{d}_l[n+1])\, a(l,k)] \\ &\quad + \sum_{j=k+1}^{K} g_j[n-1] \cdot d_j[n-1]\, a(k,j) + g_j[n] \cdot d_j[n]\, b(k,j) \\ &\quad + e_k[n] \end{aligned} \tag{3.43}$$

and the k-th user decision variable is obtained by

$$z_k[n] = r_k[n] \cdot \hat{g}_k^*[n] \tag{3.44}$$

and the decision rule is

$$\hat{d}_k[n] = sign(\mathcal{RE}[z_k[n]]) \tag{3.45}$$

3.3.2 Multi-Stage SIC

The performance of the one-stage canceller can be improved by adding more stages to the scheme. This concept has been presented considering the perfect knowledge of the phases in e.g. [13] or in quite similiar form in [6].

Consider the residual received signal after all users' signals have been cancelled, i.e.

$$\begin{aligned} r_{-K}(t) &= \sum_{l=1}^{K} [g_l(t-\tau_l)\, d_l(t-\tau_l) - \hat{g}_l(t-\tau_l)\, \hat{d}_l(t-\tau_l)] \cdot c_l(t-\tau_l) \\ &\quad + n(t) \end{aligned} \tag{3.46}$$

This signal consists only of imperfect cancellations and the AWGN term. Ideally, if all the cancellations could be done perfectly, equation (3.46) would only contain the white Gaussian noise term $n(t)$.

In the first stage, all users, except the weakest one, had to be cancelled in the presence of interference, because all weaker signals had not been removed at the time of detection. The regeneration of signals is dependent on the decisions that the receiver have made before, and so (3.46) may contain wrongly cancelled signals.

In the second stage, a previously cancelled signal will be added again to redo the first stage operation. The order for this operation is again in decreasing signal powers, i.e. user one's cancellation will be reversed first, its signal detected, the signal regenerated and subtracted again, followed by the same operation for user two and so on.

Reconsider equation (3.36) which represents user one's signal that was cancelled in the first stage. This signal would now be added again to the residual in (3.46) to obtain

$$\begin{aligned} r^{II}(t) &= r_{-K}(t) + \hat{s}_1(t-\tau_1) \\ &= c_1(t-\tau_1)\, d_1(t-\tau_1)\, g_1(t-\tau_1) \\ &\quad + \sum_{l=2}^{K} [g_l(t-\tau_l)\, d_l(t-\tau_l) - \hat{g}_l(t-\tau_l)\, \hat{d}_l(t-\tau_l)] \cdot c_l(t-\tau_l) \\ &\quad + n(t) \end{aligned} \tag{3.47}$$

This signal would then be passed through the first user's processing branch again and the new filter output for user one and time interval n would be

$$\begin{aligned} r_1^{II}[n] &= g_1[n] \cdot d_1[n] \\ &\quad + \sum_{j=2}^{K} (g_j[n-1]\, d_j[n-1] - \hat{g}_j[n-1]\, \hat{d}_j[n-1]) \cdot a(1,j) \\ &\qquad + (g_j[n] \cdot d_j[n] - \hat{g}_j[n] \cdot \hat{d}_j[n]) \cdot b(1,j) \\ &\quad + e_1[n] \end{aligned} \tag{3.48}$$

A decision on $d_1[n]$ would be made for the second time according to

$$\hat{d}_1^{II}[n] = sign(\mathcal{RE}(r_1^{II}[n] \cdot \hat{g}_1[n])) \tag{3.49}$$

Based on the newly made decision, the first user's signal will be regenerated for a second time and subtracted from the signal in (3.47) to obtain

$$\begin{aligned} r_{-1}^{II}(t) &= r^{II}(t) - \hat{s}_1^{II}(t-\tau_1) \\ &= \sum_{l=2}^{K} (g_l(t-\tau_l)\, d_l(t-\tau_l) - \hat{g}_l(t-\tau_l)\, \hat{d}_l(t-\tau_l)) \cdot c_l(t-\tau_l) \\ &\quad + (g_1(t-\tau_1)\, d_1(t-\tau_1) - \hat{g}_1(t-\tau_1)\, \hat{d}_1^{II}(t-\tau_1)) \cdot c_1(t-\tau_1) \\ &\quad + n(t) \end{aligned} \tag{3.50}$$

In a similar fashion, the filter output for the k-th user can be obtained as

$$\begin{aligned} r_k^{II}[n] &= g_k[n] \cdot d_k[n] \\ &\quad + \sum_{j=k+1}^{K} (g_j[n-1]\, d_j[n-1] - \hat{g}_j[n-1]\, \hat{d}_j[n-1]) \cdot a(k,j) \end{aligned} \tag{3.51}$$

$$
\begin{aligned}
&\quad +(g_j[n]\, d_j[n] - \hat{g}_j[n]\, \hat{d}_j[n]) \cdot b(k,j) \\
&+ \sum_{j=1}^{k-1} (g_j[n]\, d_j[n] - \hat{g}_j[n]\, \hat{d}_j^{II}[n]) \cdot b(j,k) \qquad (3.52) \\
&\quad +(g_j[n+1]\, d_j[n+1] - \hat{g}_j[n+1]\, \hat{d}_j^{II}[n+1]) \cdot a(j,k) \\
&+ e_k[n] \qquad (3.53)
\end{aligned}
$$

and the residual signal after the second stage cancellation of the k-th user would be

$$
\begin{aligned}
r_{-k}^{II}(t) &= r_{-(k-1)}^{II}(t) - \hat{s}_k^{II}(t-\tau_1) \\
&= r_{-(k-1)}^{II}(t) - c_k(t-\tau_k)\, \hat{d}_k^{II}(t-\tau_1)\, \hat{g}_k(t-\tau_k) \\
&= \sum_{l=k+1}^{K} (g_l(t-\tau_l)\, d_l(t-\tau_l) - \hat{g}_l(t-\tau_l)\, \hat{d}_l(t-\tau_l))\, c_l(t-\tau_l) \\
&\quad + \sum_{l=1}^{k} (g_l(t-\tau_l)\, d_l(t-\tau_l) - \hat{g}_l(t-\tau_l)\, \hat{d}_l^{II}(t-\tau_l)) \cdot c_l(t-\tau_l) \\
&\quad + n(t) \qquad (3.54)
\end{aligned}
$$

The operations described above can be implemented in more stages, to improve the performance of the algorithm.

In the next section, we will look into the theoretical case of perfect channel estimation, followed by one idea using the correlator outputs as amplitude estimates.

3.3.3 SIC receiver with ideal channel estimates

This section considers an SIC which has perfect knowledge of the complex channel gains. Although this case is rather idealistic, it can be used as an upper performance limit.

Assuming a black box perfect channel estimator, the estimated fading process becomes

$$\hat{g}_k(t) = g_k(t) \qquad (3.55)$$

for $k = 1, \ldots, K$, and the k-th user integrator output would be

$$r_k[n] = g_k[n] \cdot d_k[n]$$

$$
\begin{aligned}
&+ \sum_{j=k+1}^{K} g_j[n-1] \cdot d_j[n-1]\, a(k,j) + g_j[n] \cdot d_j[n]\, b(k,j) \\
&+ \sum_{l=1}^{k-1} [g_l[n]\, (d_l[n] - \hat{d}_l[n])\, b(l,k) + g_l[n+1]\, (d_l[n+1] - \hat{d}_l[n+1])\, a(l,k)] \\
&+ e_k[n] \qquad (3.56)
\end{aligned}
$$

and the k-th user decision variable is

$$z_k[n] = r_k[n] \cdot g_k^*[n] \qquad (3.57)$$

and the decision rule is (3.45). It can be seen from (3.56) that the detected user's signals can be completely removed, if the decision about the data bits were correct.

A multi-stage receiver can be implemented, similarly to what was shown in equation (3.46)-(3.54) by replacing $\hat{g}_k(t)$ by $g_k(t)$.

3.3.4 SIC using correlator outputs as channel estimates

One idea described in [16] would use the integrator output of the preceding user as an estimate for the product $g_k[n] \cdot d_k[n]$. This product can be used in the cancelling receiver to regenerate and subtract the signal, even without knowledge about the actual data that was transmitted. However, if besides the cancelling operation a decision on the data has to be made, the knowledge of at least the received signal phase is necessary[4]. This phase has to be obtained by other means, such as PSAM or DPSK.

The output of user's first processing branch would be as described in (3.33). Using this combined estimate for the data and the channel gain, the receiver would regenerate and subtract the signal and the receiver part for user k would process the following signal

$$
\begin{aligned}
r_{-(k-1)}(t) &= r_{-(k-2)}(t) - \hat{s}_{k-1}(t - \tau_{k-1}) \\
&= r_{-(k-2)}(t) - c_{k-1}(t - \tau_{k-1}) \cdot \sum_{n=x}^{x+q} r_{k-1}[n] \cdot p_d(t - n \cdot T - \tau_{k-1})
\end{aligned}
$$

[4]considering a PSK system

$$\begin{aligned}
&= \sum_{l=k}^{K} g_l(t-\tau_l)\cdot d_l(t-\tau_l)\cdot c_l(t-\tau_l) \\
&\quad + \sum_{l=1}^{k-1} g_l(t-\tau_l)\cdot d_l(t-\tau_l)\cdot c_l(t-\tau_l) \\
&\qquad\quad - c_l(t-\tau_l)\cdot \sum_{n=x}^{x+q} r_l[n]\cdot p_d(t-n\cdot T-\tau_l) \\
&\quad + n(t) \qquad (3.58)
\end{aligned}$$

where x represents again the block number and q is the block length over which the signals are cancelled. Subsequently, the k-th user integrator output is then

$$\begin{aligned}
r_k[n] &= g_k[n]\cdot d_k[n] \\
&\quad + \sum_{l=1}^{k-1}[(g_l[n]\cdot d_l[n]-r_l[n])\, b(l,k) \\
&\qquad\quad +(g_l[n+1]\cdot d_l[n+1]-r_l[n+1])\, a(l,k)] \\
&\quad + \sum_{j=k+1}^{K} g_j[n-1]\cdot d_j[n-1]\, a(k,j)+g_j[n]\cdot d_j[n]\, b(k,j) \\
&\quad + e_k[n] \qquad (3.59)
\end{aligned}$$

An analysis of this idea was done for the case of perfect phase recovery in [16]. Multi-stage canceller can be similarly implemented as in the foregoing sections.

3.4 PSA Successive Interference Canceller

This section discusses all successive interference canceller that use pilot symbol assisted modulation and that were implemented in the simulations. The data detection part in all investigated receivers is done by an SIC scheme presented in section 3.3. Different interference cancellation strategies have been used for reducing interference in the gain estimates at pilot symbol time intervals and have been discussed before. The block diagram of both PSA-SIC receiver is shown in figure 3.8. Both receiver are identical except for the part denoted as *Pilot Symbol Interference Cancellation and Channel Estimation* that will be shown in the next two sections.

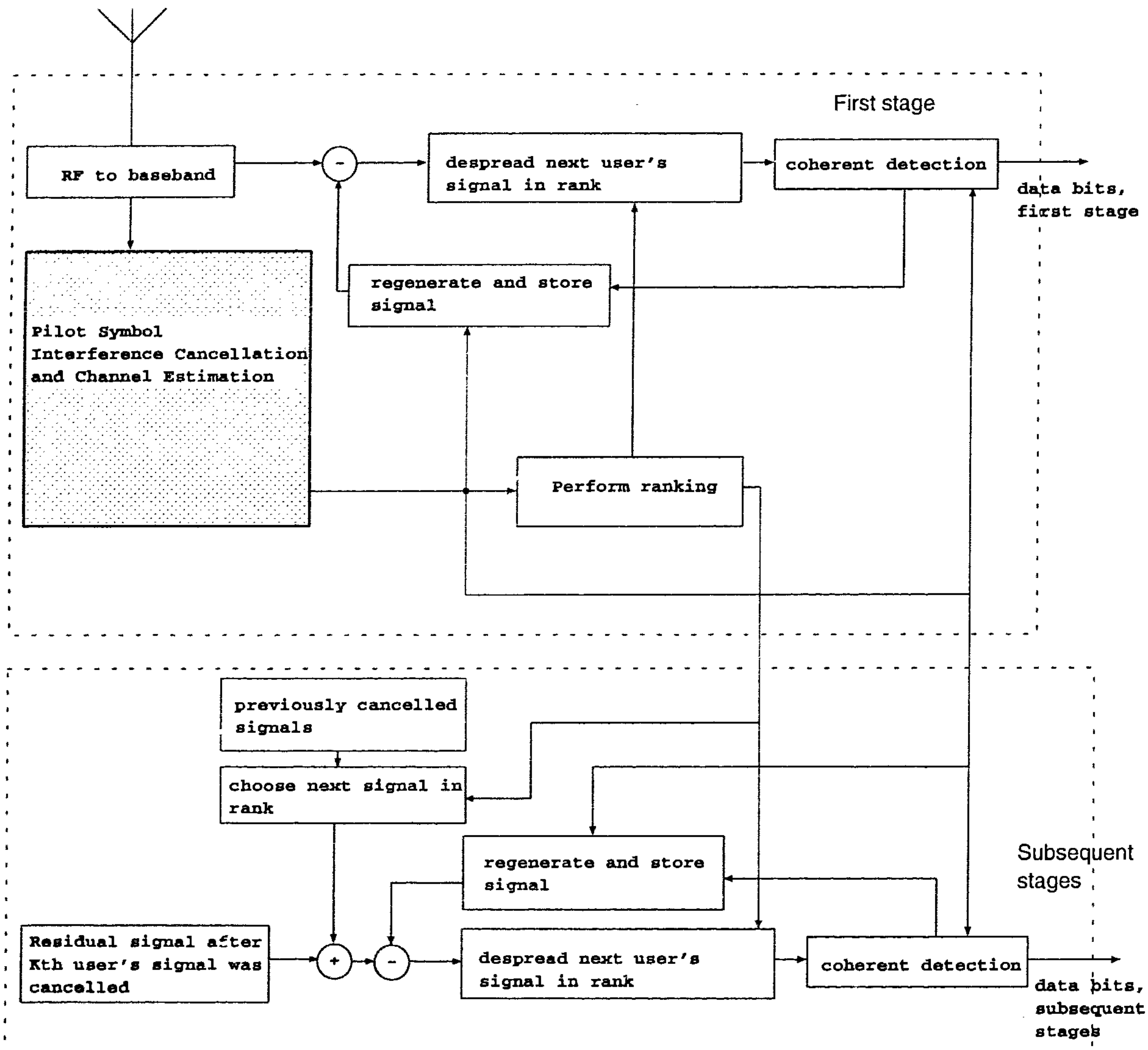

Figure 3.8: Block Diagram of the SIC Receiver with channel estimator

3.4.1 Double SIC Scheme

The double SIC scheme uses a MAI canceller employing correlator outputs for the signal regeneration as discussed in section 3.3.4 to decrease the interference in the channel estimates. The channel estimator consists of a Wiener interpolator optimized to the signal to additive white noise ratio of the signal and the fade rate in the system. For the purpose of data detection a successive scheme, as discussed in section 3.3.1 and 3.3.2, is used. Figure 3.9 shows a block diagram of the channel estimator part and the rest of the receiver was shown in figure 3.8. Out of the possible transmission frame structures introduced in section 3.2, the receiver would operate on the structure with one inserted pilot symbol and the frame with one pilot symbol with preceding and following guard intervals. Clearly, the use of three consecutive inserted pilot symbols offers no advantage for this receiver and it is better instead to reduce the frame size by a factor of three to reduce the estimation error for the gain estimates.

3.4.2 Hybrid SIC Scheme

The hybrid PSAM cancelling receiver consists of a decorrelator for a synchronous channel, as described in section 3.2, to reduce the MAI in the pilot intervals. This operation enhances the noise in the gain estimates which was taken into account in the Wiener interpolator. Again, for the data detection a successive scheme, as discussed in section 3.3.1 and 3.3.2, was employed. The block diagram of the hybrid receiver is shown in figure 3.8 combined with figure 3.10 for the channel estimation part.

The scheme presented here is able to operate on the transmission frame structure with three pilot symbols, the frame with one pilot symbol and guard intervals and on a transmitted frame with one pilot symbol, if the integration time in the pilot interval is shortened. These transmission frame structures have been discussed before in section 3.2.

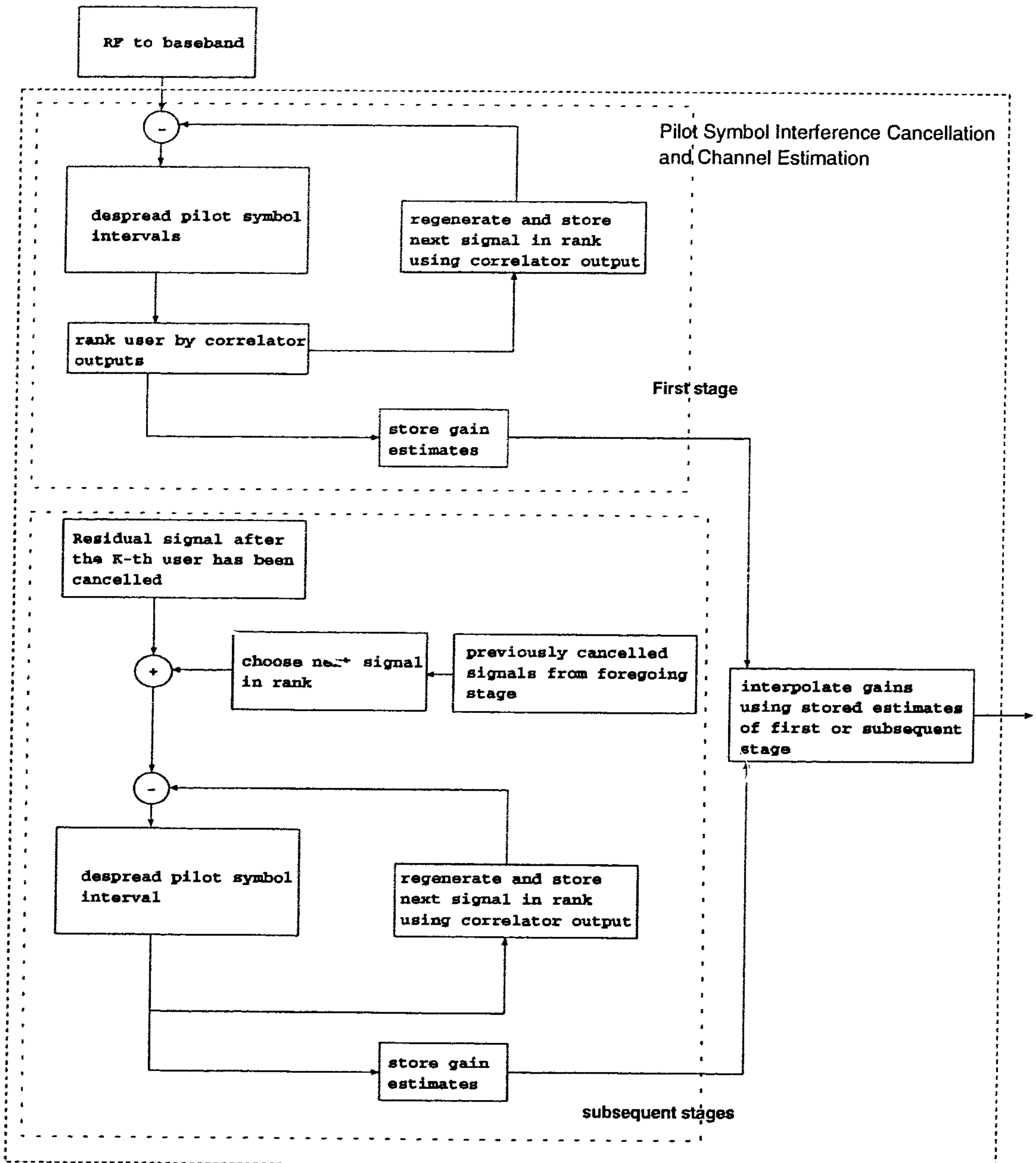

Figure 3.9: Block Diagram of the Channel Estimator using SIC

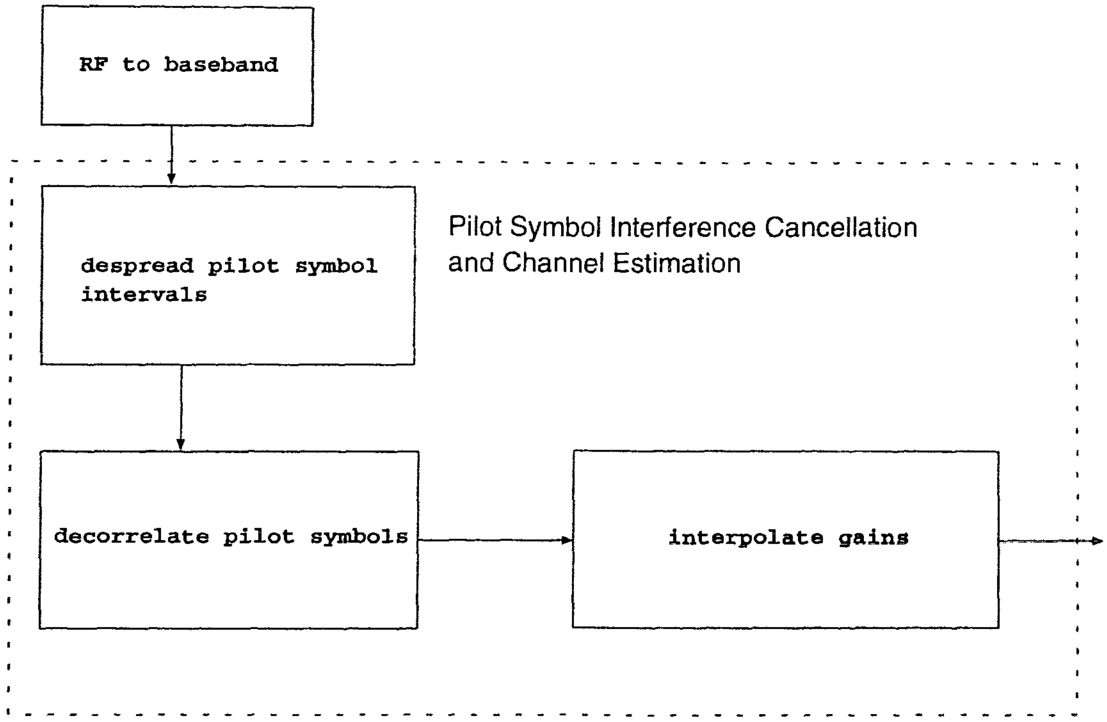

Figure 3.10: Block Diagram of the Channel Estimator using a Decorrelator

3.5 Implementation Issues

3.5.1 Complexity

The issue of complexity is an important point to consider, since a theoretical idea should perform later in a practical system. If the computational load cannot be handled by the hardware, the system can clearly not be realised. More complexity generally means a higher failure rate of the device, more maintainance is necessary and higher costs are involved. Also, especially in the area of mobile communications, the mobile devices should be light, cheap and reliable but a higher complexity usually means higher power consumption and larger batteries.

If we would like to compare the different PSAM receiver schemes introduced before, we have to make some considerations. Since we are using the same receiver concept for the data detection, the SIC in one- or multi-stage implementation, this part of the receiver has not to be taken into consideration. Our attention will therefore only focus on the pilot symbol time instants and the operations necessary to process this part of the signal.

The computational burden in using a decorrelator is always to obtain the matrix inverse of the crosscorrelation matrix. The crosscorrelations change in a dynamic system, since users are changing locations in the cell and the number of active users in the cell changes. Therefore, the inverse of the matrix would have to be recalculated. An initial inversion of the matrix of order $K \cdot K$ involves computations of the order K^3. However, as shown in [4] once the inverse is optained, the updates can be calculated by an efficient algorithm and involve only $2 \cdot K^2$ real multiplications and $2.5 \cdot K^2$ real additions per frame, if we assume that one user is leaving and one user is entering the cell during that frame[5]. The actual detection involves then $2 \cdot K^2$ real additions and multiplications. In total, the pilot decorrelation involves $4 \cdot K^2$ real multiplications and $4.5 \cdot K^2$ real additions. From the implementation point of view the decorrelator is one of the simplest interference cancellation techniques to implement, since technology for high-speed zero-forcing equalizer is readily available [26].

The SIC would have to take the correlator output of user one and multiply this with the crosscorrelation coefficient between user one and two. The results would be subtracted from user two's correlator output. These operations involve 2 real multiplications for the scaling by the crosscorrelation and the subtraction needs two real additions. These steps have then to be performed for $K-1$ times by user one, $K-2$ times by user two and so on and therefore the total number of operations is

$$\begin{aligned} o &= \sum_{k=1}^{K-1} k \cdot (2 \text{ real mults} + 2 \text{ real adds}) \\ &= \frac{K^2 - K}{2} \cdot (2 \text{ real mults} + 2 \text{ real adds}) \\ &\approx \frac{K^2}{2} \cdot (2 \text{ real mults} + 2 \text{ real adds}) \\ &\approx K^2 \cdot (1 \text{ real mults} + 1 \text{ real adds}) \end{aligned} \tag{3.60}$$

if we are considering the interference removal in one symbol which is only sufficient

[5] one frame is about $2.5ms$ for a frame length of 25 symbols and a symbol rate of $9600\frac{bit}{s}$

in the synchronous case. In a quasi-synchronous system the interference has to be removed in 2 symbols[6] and the complexity would be twice as high. Therefore, the complexity of the decorrelator, as compared to the one-stage SIC for the pilot interval, is higher and involves double the number of multiplications and 2.25 times the number of additions.

Every additional stage in the scheme would account for 2 real additions for adding the removed signal back to the residual and again 2 real multiplications and 2 real additions for scaling the integrator output by the cross correlation and subtracting it. The second and subsequent stages would also perform on 2 symbols, if we consider the quasi-synchronous case. Therefore, for K users the number of operations required in the second stage is

$$o \approx 2 \cdot K^2 \cdot (1 \text{ real mults} + 2 \text{ real adds}) \tag{3.61}$$

and similar for the following stages. If we now compare the complexity of the decorrelator with the two-stage SIC, the decorrelator would require the same amount of multiplications and 75 percent the number of additions. Furthermore, because of the shorter frame length, the SIC would have to perform the operations 3 times as often as the decorrelator.

Using a third stage SIC would make the hybrid approach even more favourable in terms of the complexity.

3.5.2 Cancelling Frame Position and Length

In an asynchronous system the transmitted signals are cancelled out in frames as was discussed in section 3.3. At the border of cancellation frames, a user with a lower received power and a longer time delay will experience a higher bit error rate as can be seen in figure 3.11. If a user with a stronger received signal has a shorter time delay, parts of its next transmitted frame (denoted as interval 'A' in the figure) will interfere with other user's transmissions who have a longer delay. Therefore, it can

[6]if one pilot symbol is inserted without guard intervals

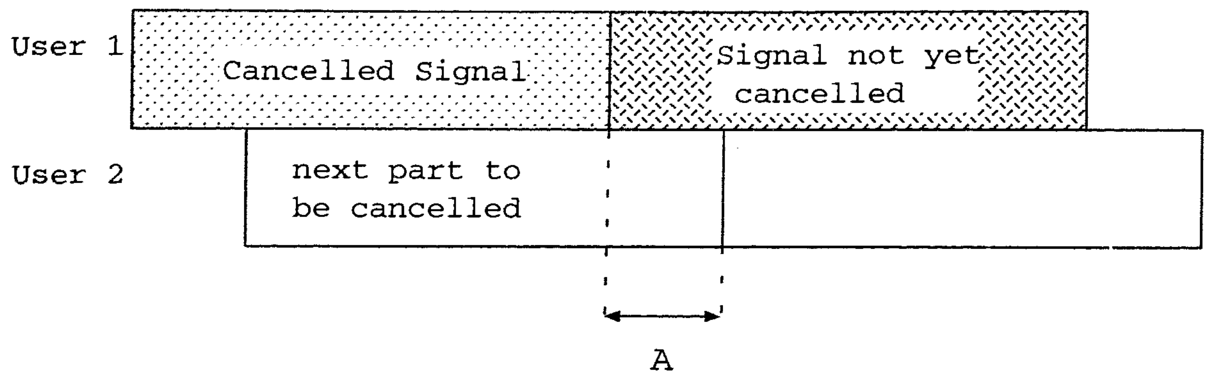

Figure 3.11: Frame Border Effects on the Receiver Performance

be expected that the number of borders affects the performance of the cancellation scheme. If longer frames are choosen, less frames have to be cancelled for the same number of detected symbols and less degradation of the receiver is expected. Also, the influence of the borders on the performance decreases with a decreasing maximum time delay in the system and for a synchronous channel no border effects have to be considered. Another issue that should be mentioned here, is the position of the frame borders. Choosing the borders such that the first symbol of the next frame to be cancelled is a pilot symbol[7] can reduce or eliminate the border effects, because the pilot symbols are known and can be removed before the data detection with high accuracy. Figure 3.12 shows this situation where the pilot symbols have been removed prior to the data detection reducing the frame border effects.

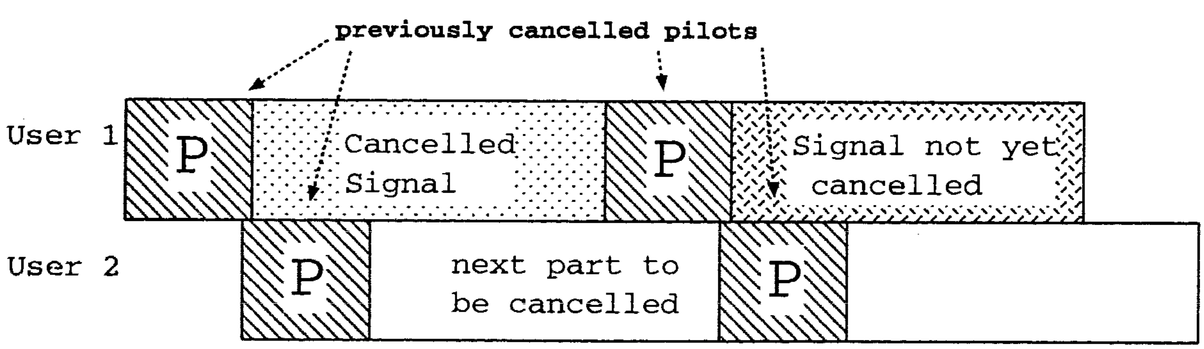

Figure 3.12: Reduced Frame Border Effects

[7]or guard interval

3.5.3 Pilot Insertion Factor

Inserting pilot symbols into the data stream reduces the energy that can be used per transmitted information bit since some energy is wasted on the pilots. On the other hand, increasing the pilot insertion factor increases also the estimation error. There exists an optimal value for the pilot insertion factor M as it was discussed in [17] for the single-user case. However, the multi-user case involves also the consideration of the MAI to evaluate the optimal insertion factor and furthermore, the optimal value might change with a change in the system parameters like user population etc. In most cases this parameter has to be choosen through the use of simulations.

3.6 Analysis versus Simulations

An analysis of pilot symbol assisted successive interference cancellation schemes has so far only partially been done. In [19] an analysis has been done for the one-stage idea, however, it was indicated that the analysis for the multi-stage approach would involve computations that are as intensive as the actual simulations. The way the system performance was evaluated in this thesis was therefore through simulations. The wide bandwidth of the CDMA signals imposed a problem due to the high number of samples that had to be processed and long running times of the program were usually the case.

3.7 Summary

This section introduced a general model of the system investigated. The implemented receiver have been discussed in more detail and a new idea for a hybrid pilot symbol assisted receiver operating in a quasi-synchronous system was presented. Some attention was also focused on the complexity of the receivers presented. The next section will discuss the simulation results and will make some comparisons of the concepts presented here.

Chapter 4

Simulation Results

Simulations of a CDMA system are very time intensive, because of the high bandwidth of the signal and the large number of samples that have to be processed. Running times of more than a day on a *SPARC 20* for one data point, depending on the system parameters, were not unusual. Another problem of investigating the receiver performance in a CDMA system are the large number of parameters to be considered, e.g. number of user, spreading factor, time delays of the transmissions etc., and their coupling effect on the receiver performance.

The results shown here are generated with a simulation written in *C*. A *Matlab* simulation, programmed before, has proven to be too slow, and the running time was far too long. Also, the implementation in *C* was easily portable to different machines and the computational load could therefore be spread onto a variety of work stations in different departments[1].

[1]including all the problems involved with the use of other people's computer...!

4.1 Assumptions

The simulations were considering an uncoded single-cell DS/CDMA system using BPSK as the modulation scheme. The spreading factor in the system was 31 (unless otherwise specified) and the average received power of the different users' transmissions was assumed to be the same for all user. This means that shadowing or path loss was already taken care of by means of a slow power control. Therefore, the signals were only subject to the fast Rayleigh fading. The normalized fade rate was $f_d \cdot T = 0.01$ for all of the simulations.

The interpolator length in the system was 11 and the frame length was 27 or 9 symbols, depending on the number of inserted pilot symbols, to make the throughput of the different schemes comparable. The cancellation frame length and the pilot insertion factor (9 or 27) were identical. Inserting pilot symbols yield to a loss in energy per transmitted information bit and this loss was taken into account in the results. Three different ways of inserting pilot symbols have been simulated, which have been described before in chapter 3.2. In addition, different receiver concepts have been implemented. Table 4.1 shows all receiver/frame combinations that have been simulated. The double SIC receiver is implemented in our system in one-,

Simulated Receiver/Frame Combinations			
	Receiver		
Frametype	Double SIC	Hybrid	Conventional
one pilot	9 sym	9 sym	7, 9, 27 sym
one pilot w/ guard intervals	27 sym	27 sym	27 sym
three pilots	n/a	27 sym	n/a

Table 4.1: Simulated Receiver/Frame Combinations

two- and three-stages and is denoted as **Receiver A**, **Receiver B** and **Receiver C**, respectively. **Receiver D** operates with two-stages on the pilot symbols and uses a one-stage approach for the data detection and approach **E** is using three (pilot) and two (data) stages. The ideal SIC receiver with perfect knowledge about the fading

Legend for the Graphs			
Receiver	**Pilot**	**Data**	**Linestyle**
A	SIC, one-stage	SIC, one-stage	dash-dot and +
B	SIC, two-stages	SIC, two-stages	solid and +
C	SIC, three-stages	SIC, three-stages	dash and +
D	SIC, two-stage	SIC, one-stage	dash-dot and *
E	SIC, three-stage	SIC, two-stage	dash and *
F	ideal	SIC, one-stage	dash-dot and x
G	ideal	SIC, two-stages	solid and x
H	ideal	SIC, three-stages	dash and x
I	Decorrelator	SIC, one-stage	dash-dot and o
J	Decorrelator	SIC, two-stages	solid and o
K	Decorrelator	SIC, three-stages	dash and x
L	ideal	conventional	solid
M	conventional	conventional	dash

Table 4.2: Legend for the Graphs

gains is also implemented and is denoted as **Receiver F** (one-stage), **Receiver G** (two-stages) and **Receiver H** (three-stages). The Hybrid Receiver introduced in section 3.2 is denoted by **Receiver I** in one-stage, **Receiver J** in two-stage and **Receiver K** for the three-stage implementation. The conventional receiver with ideal (**Receiver L**) and estimated gains (**Receiver M**) is also displayed in the graphs. Table 4.2 shows the legend for the graphs appearing in this chapter. If results for different frame lengths are shown in one graph, the notation is extended by the frame size, i.e. A9 stands for the double SIC receiver in one-stage implementation operating on a frame size of $M = 9$.

4.2 Conventional Receiver

One of the comparison systems in the simulations was the conventional receiver. We will present simulation results for the one pilot frame and for frame sizes of 7, 9 and

27 symbols.

4.2.1 Single-User Case

Figure 4.1 shows the performance of a single-user conventional receiver operating on a flat Rayleigh fading channel. The pilot insertion factor was $M = 7$. We can see that the performance gap between the ideal coherent receiver and the receiver with estimated channel gains is about 2 dB, which matches the results that have been reported in [17] for this kind of simulation scenario.

4.2.2 Multi-User Case

Figure 4.2 shows results for the conventional receiver with perfect knowledge of the channel gains and with implemented channel estimator operating in a multiuser environment. Both receiver have a poor performance as the bit error rate is greater than 10^{-2}. Different frame sizes of 9 and 27 symbols per frame had just a minor influence on the receiver performance. The multiuser access interference plays the major role in the limitations of this receiver concept.

The average bit error rate (BER) versus the SNR in a system with 20 active user for both conventional schemes is shown in figure 4.3. Both conventional receivers develop a high error floor due to the multiuser access interference.

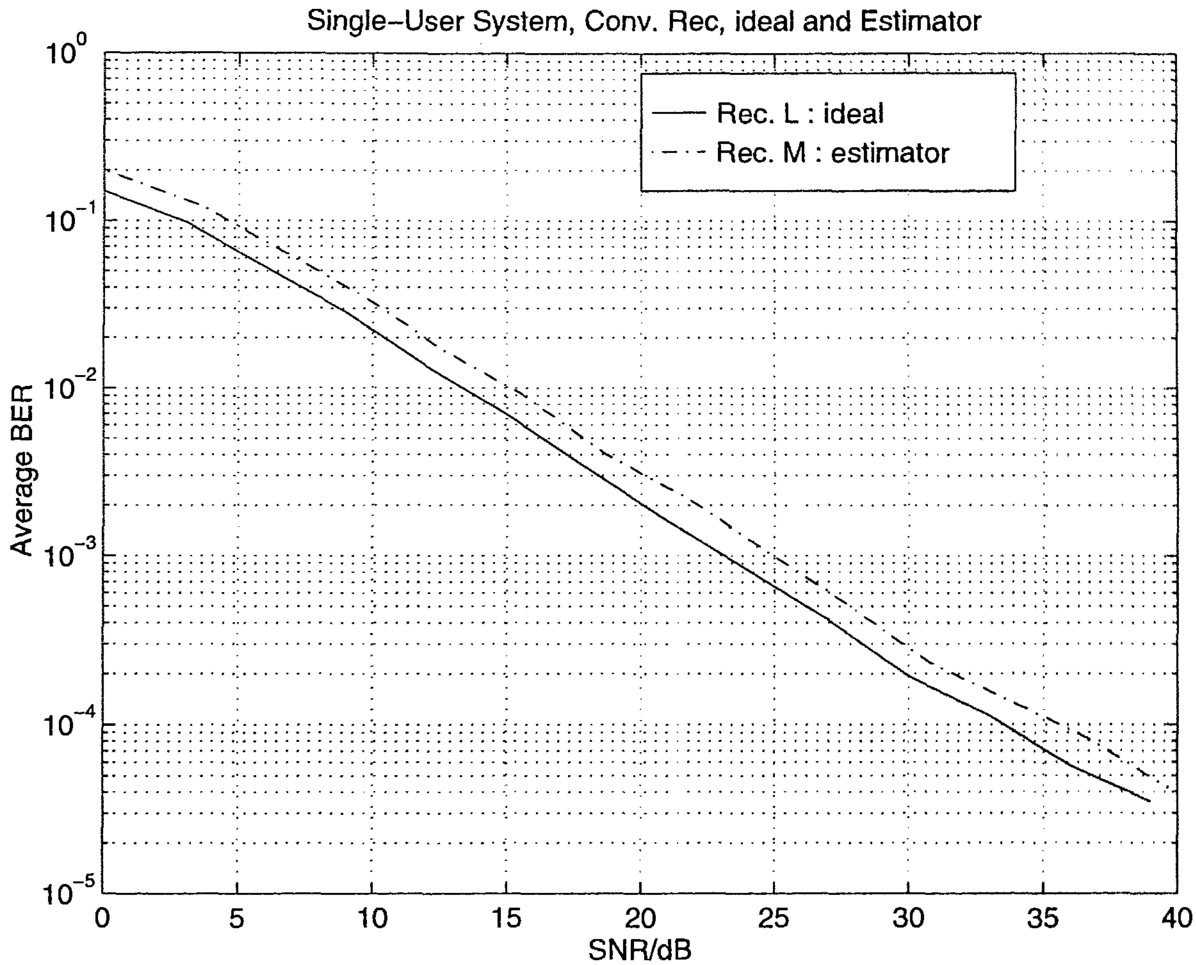

Figure 4.1: Conventional Receiver in a single user system

Parameter for Simulation	
Parameter	**Value**
Spreading Factor	1
Despread SNR/dB	variable
Pilot Block	One Pilot
Interpolator length	11
Data-Frame Length	7
Number of User	1
$f_d T$	0.01
τ_{max}/T	n/a

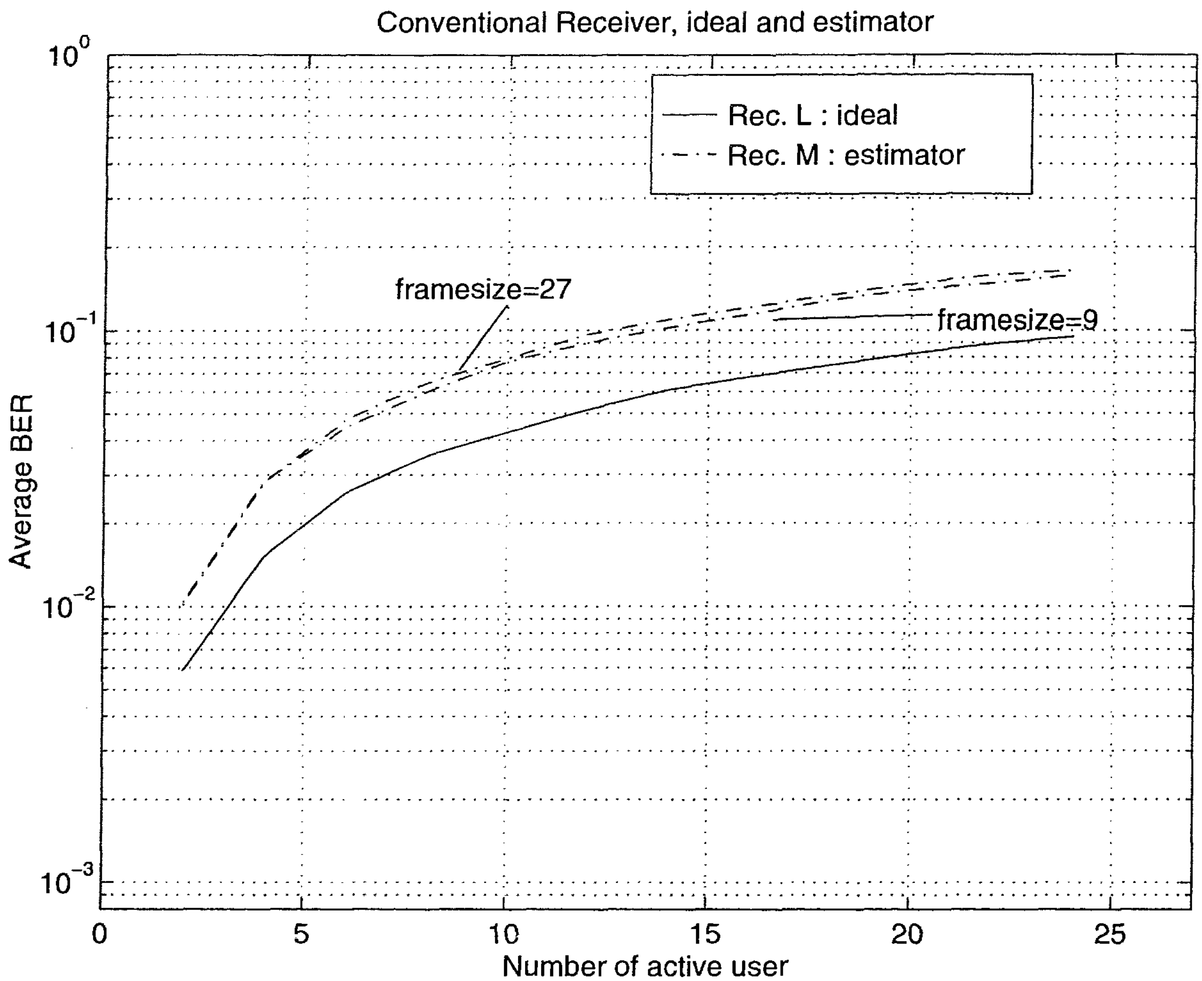

Figure 4.2: Conventional Receiver in a multi-user system

Parameter for Simulation	
Parameter	**Value**
Spreading Factor	31
Despread SNR/dB	25
Pilot Block	One Pilot
Interpolator length	11
Data-Frame Length	9 & 27
Number of User	variable
$f_d T$	0.01
τ_{max}/T	1.0

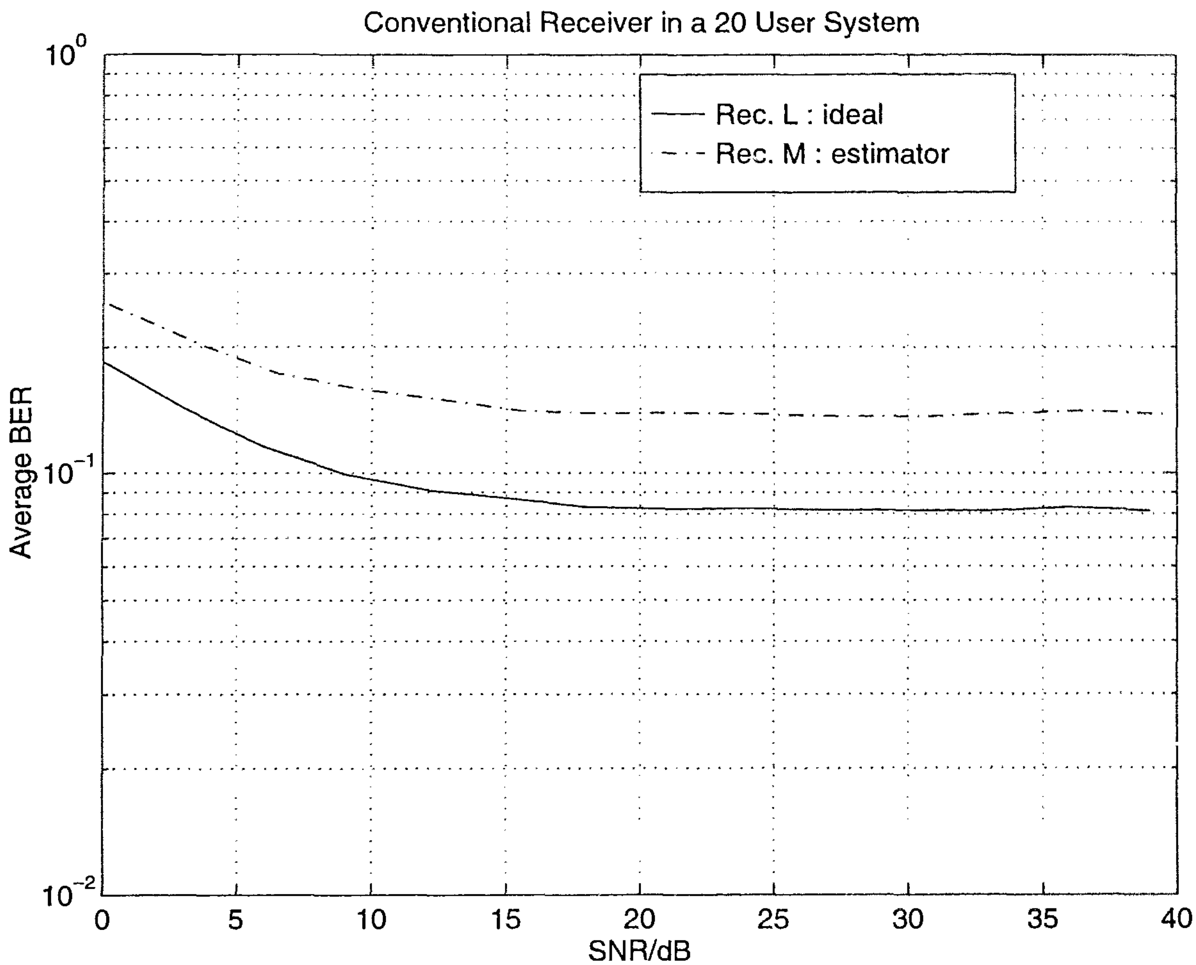

Figure 4.3: Error floor of the conventional receiver, 20-user system

Parameter for Simulation	
Parameter	**Value**
Spreading Factor	31
Despread SNR/dB	variable
Pilot Block	One Pilot
Interpolator length	11
Data-Frame Length	9
Number of User	20
$f_d T$	0.01
τ_{max}/T	1.0

4.3 Double Successive Interference Canceller

4.3.1 One Pilot Symbol

We will consider a system where one pilot symbol per frame is inserted into the transmitted data streams. The pilot insertion rate was $M = 9$. The graph 4.4 shows results for four different SIC receiver concepts. In addition, the results for the conventional receiver M and the ideal receiver in one (F), two (G) and three stages (H) can be seen. The ideal receiver sets the performance limit that can be achieved for the choosen parameters by applying perfect channel estimates to the data detection SIC. It can be seen that implementing the ideal receiver in a second stage does not offer much performance improvement for the set of parameters choosen. A three stage receiver (H) does not yield to any improvement over the two stage implementation (G). The double SIC receiver with one-stage (Receiver A) shows already a significant improvement over the conventional receiver (M). Additional performance can be achieved by implementing the double SIC scheme in two-stages (Receiver B). Adding a third stage to the double SIC scheme (Receiver C) yield again to a performance increase and the error floor was reduced.

Finally, a receiver using a two-stage algorithm only for the pilot symbols and performing the data detection with one-stage (Receiver D) gives already a similar performance as a two-stage scheme for data and pilot symbols This indicates the importance that the gain estimate has on the performance of the cancellation technique used and it shows furthermore that if more emphasis is placed on obtaining better channel estimates, less emphasis can be placed on the actual data detection to achieve a similar error rate. The same behaviour was observed for the three-stage SIC's C and E which had almost exactly the same BER. Therefore, in the sequel only the results for receiver C are presented. Similiar statements can be made for the ideal receiver G and H which had the same performance under all conditions considered in the simulations and therefore results will be shown only for either one of them.

The next two figures, 4.5 and 4.6, present the average BER for 25 dB and 15 dB SNR when the number of users in the system varies. We see the improvement over the conventional receiver (M) by using an SIC receiver (A, B, C, D). For both SNR considered, a second stage receiver (B, D) offers an additional advantage and the three-stage approach (C) can further improve the performance. Receiver D which operates only on the pilot symbols with a two-stage scheme, shows an almost similar performance as compared to receiver B and offers in addition a lower complexity. The receiver with perfect knowledge of the complex channel gains (F, G, H) show the limit that could be achieved and that there exists a sizeable gap between the ideal receiver and the schemes employing a channel estimator.

4.3.2 One Pilot Symbol with Guard Intervals

The next case that was considered were pilot symbol assisted receiver operating in a quasi-synchronous system with $\tau_{max} = T$. One pilot symbol was used preceded and followed by a guard interval with no transmission for one symbol time. The authors in [6], were proposing that idea and used a double SIC receiver (denoted as Receiver A27, B27, C27), as described in section 3.2. The reasoning for using this frame structure was that the interference in the pilot intervals can be reduced, therefore better channel state information can be obtained and this should yield to a decreased error rate.

To make a comparision with results in section 4.3.1, the frame length was now increased by a factor of three ($M = 27$ symbols) to compare systems with the same throughput. Figure 4.7 shows the BER for the SIC receiver A, B, C and D. The curves are labelled with 'x27', where 'x' stands for a letter corresponding to one of the receivers and the 27 indicates the longer frame length. To simplify the comparison between the SIC operating on the one pilot frame of size 9, which where shown in the preceding section 4.3.1, we have reproduced the result for receiver C in the graph and denoted it by C9.

For a signal to noise ratio of up to approx. 17 dB all SIC receiver operating on the long frame size are doing slightly worse than the one operating on the frame with one pilot symbol and a pilot insertion factor of $M = 9$. Up to this SNR the influence of the

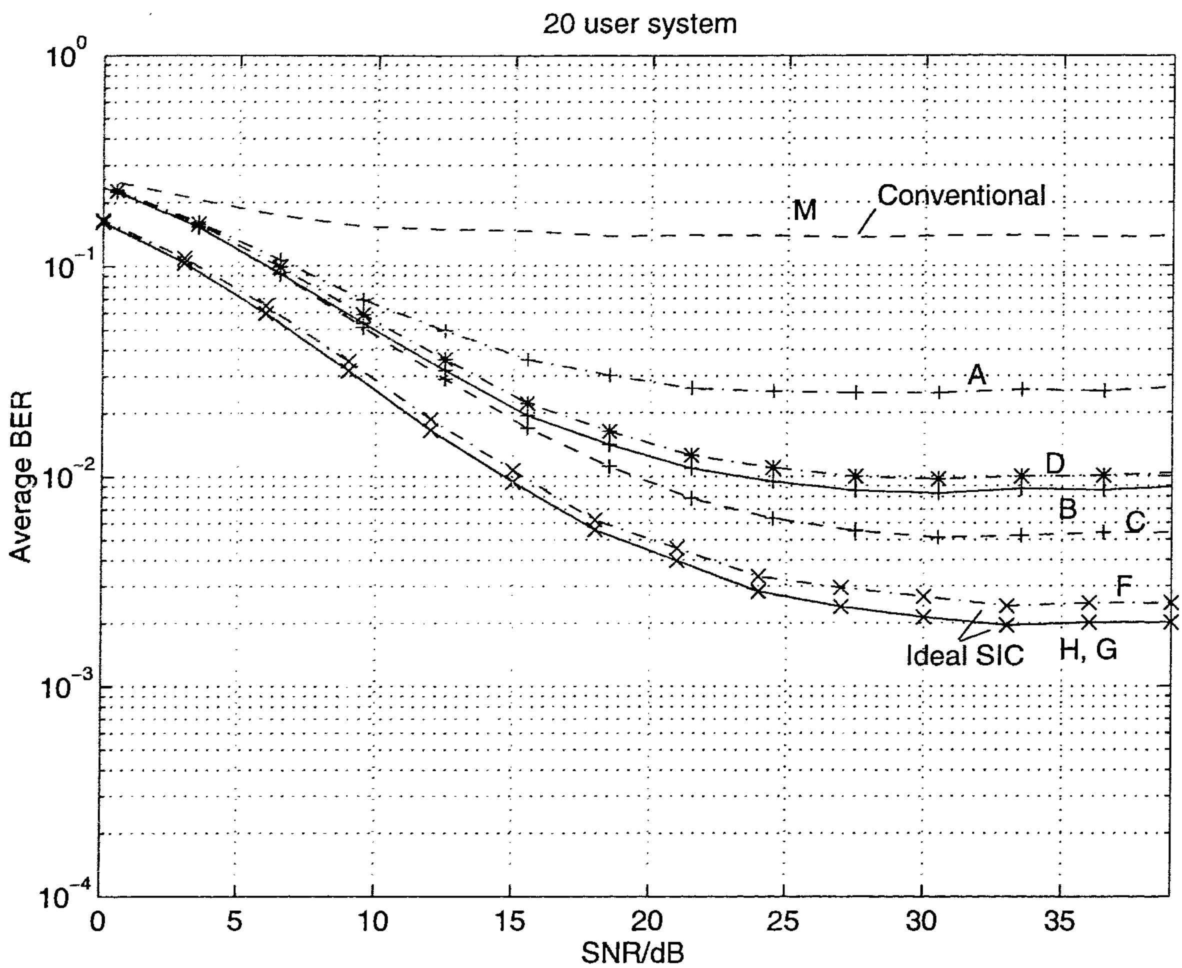

Figure 4.4: One pilot, SIC, BER vs. SNR, 20 User System

Parameter for Simulation	
Parameter	**Value**
Spreading Factor	31
Despread SNR/dB	variable
Pilot Block	One Pilot
Interpolator length	11
Data-Frame Length	9
Number of User	20
$f_d T$	0.01
τ_{max}/T	1.0

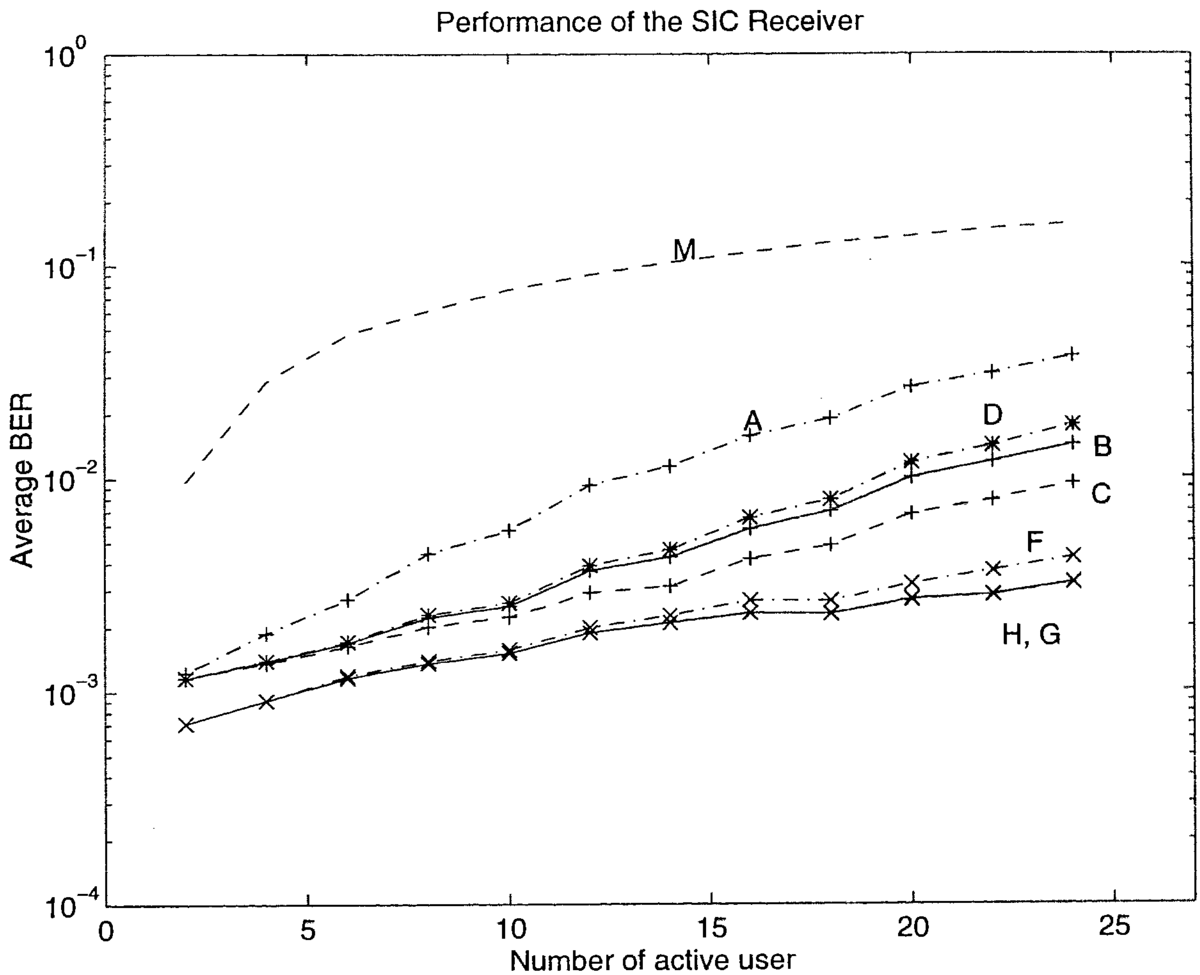

Figure 4.5: One pilot, SIC, BER vs. user, SNR 25dB

Parameter for Simulation	
Parameter	**Value**
Spreading Factor	31
Despread SNR/dB	25
Pilot Block	One Pilot
Interpolator length	11
Data-Frame Length	9
Number of User	variable
$f_d T$	0.01
τ_{max}/T	1.0

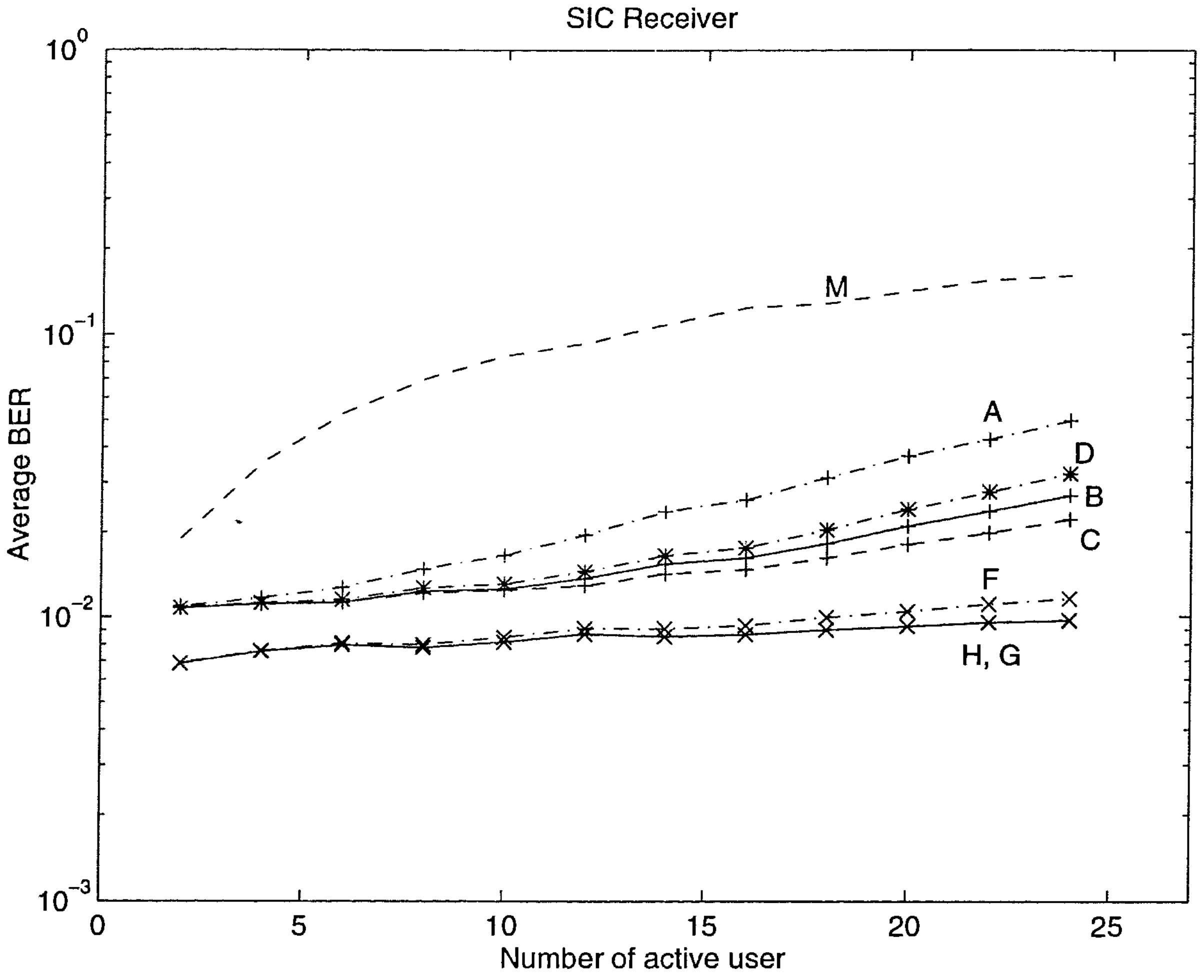

Figure 4.6: One pilot, SIC, BER vs. user, SNR 15dB

Parameter for Simulation	
Parameter	**Value**
Spreading Factor	31
Despread SNR/dB	15
Pilot Block	One Pilot
Interpolator length	11
Data-Frame Length	9
Number of User	variable
$f_d T$	0.01
τ_{max}/T	1.0

AWGN of the channel on the error rate is more significant than the MAI introduced by the other users. The one-stage SIC receiver (A27) cannot benefit at all from this frame structure and its overall performance is slightly worse. Receiver D27 shows a similar behaviour. Receiver B27 and C27, the two- and three-stage SIC Receiver, can benefit from this frame structure and decrease the error rate in the system. It can be seen that the two-stage SIC B27 is performing similar to the three-stage receiver C9 at a lower complexity. The three-stage receiver C27 improves again the bit error rate in the system. The results for receiver E27 were identical with receiver C27 and its results are not shown in the graph.

Figure 4.8 shows results for a channel with a fixed SNR of 25 dB and the number of active user as the parameter. Results from section 4.3.1 are again reproduced in the graph and are denoted by X9, where the X represents a letter corresponding to one of the receivers. The one-stage SIC A27 does not benefit from the frame stucture chosen and it shows a worse performance than receiver A9. However, if a second stage is added to the receiver the BER will be decreased and it can be seen that for receiver with two or three stages the pilot symbol transmissions with preceding and following guard intervals are beneficial. Using a three-stage approach (C27) reduces the BER again and more user could be allowed in the system for a given error rate.

For a channel with an SNR of 15 db it would be better to use the short frame of size 9 for all investigated receiver, since the influence of the AWGN on the channel estimates is more significant than the degradation by the multiuser access interference.

4.4 The Hybrid Receiver

This section presents results for the Hybrid Receiver I, J and K. They are designed to work either on the one pilot frame with preceding and following guard intervals, the frame structure with three consecutive pilot symbols or on a frame with one pilot symbol and shortened integration time as described in chapter 3.2. We will present simulation results for all three frames and will compare the receiver performance under

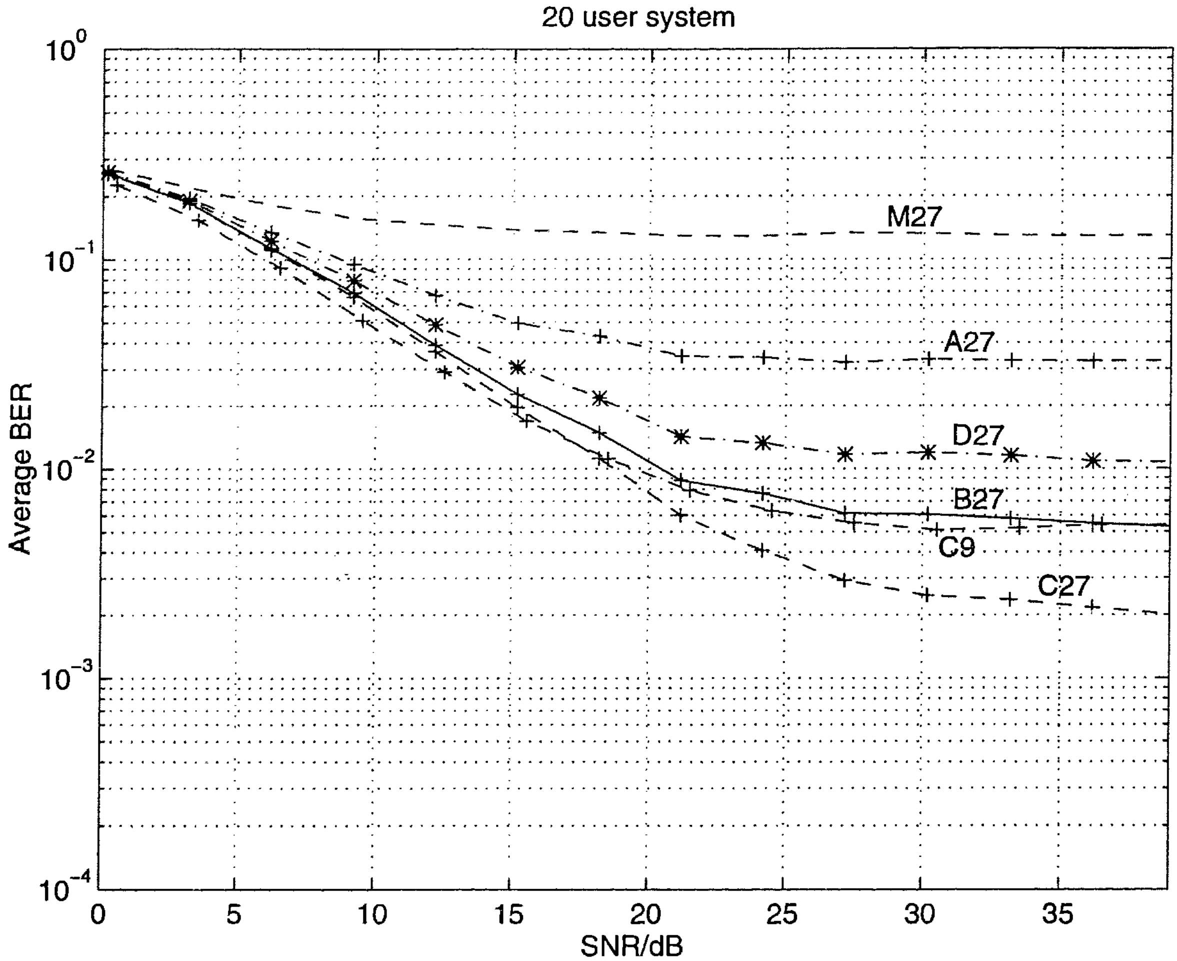

Figure 4.7: One pilot w/ guard intervals, SIC, BER vs. SNR, 20 user system

Parameter for Simulation	
Parameter	**Value**
Spreading Factor	31
Despread SNR/dB	variable
Pilot Block	One Pilot with Guard Intervals
Interpolator length	11
Data-Frame Length	27
Number of User	20
$f_d T$	0.01
τ_{max}/T	1.0

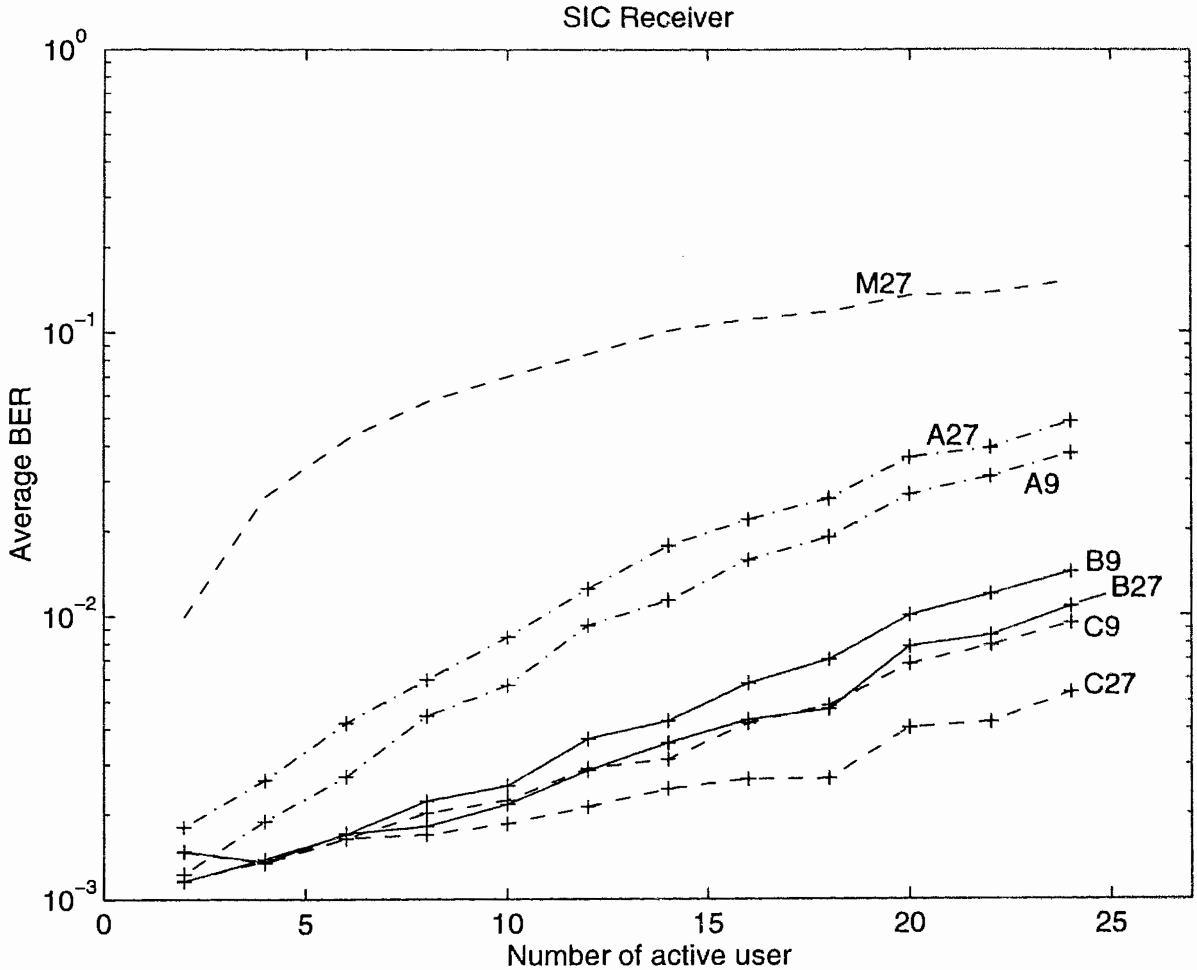

Figure 4.8: One pilot w/ guard intervals, SIC, BER vs. user, SNR 25 dB

Parameter for Simulation	
Parameter	**Value**
Spreading Factor	31
Despread SNR/dB	25
Pilot Block	One Pilot with Guard Intervals
Interpolator length	11
Data-Frame Length	27
Number of User	variable
$f_d T$	0.01
τ_{max}/T	1.0

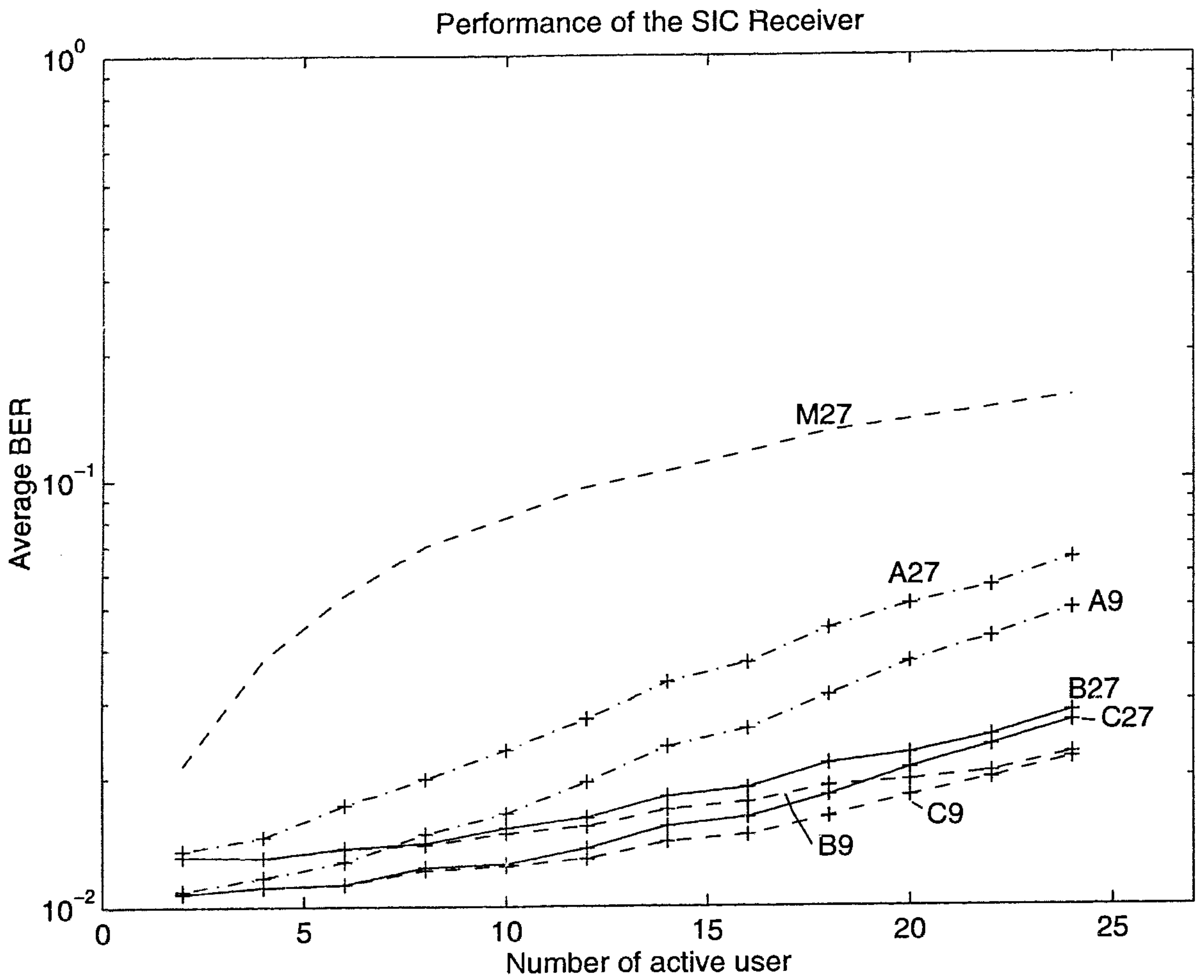

Figure 4.9: One pilot w/ guard intervals, SIC, BER vs. user, SNR 15 dB

Parameter for Simulation	
Parameter	**Value**
Spreading Factor	31
Despread SNR/dB	15
Pilot Block	One Pilot with Guard Intervals
Interpolator length	11
Data-Frame Length	27
Number of User	variable
$f_d T$	0.01
τ_{max}/T	1.0

different conditions.

4.4.1 One Pilot Symbol with Guard Interval

Figure 4.10 displays results for the hybrid receiver I, J and K. In addition, results from the sections 4.3.1 and 4.3.2 have been reproduced in the graphs and be denoted by C9 (one pilot frame and insertion factor 9) and C27 (one pilot frame with guard intervals and insertion factor 27).

It can be seen that receiver J and K show similiar performance and improve the receiver performance as compared to the double SIC receiver C27 for all SNRs. Comparing the hybrid schemes with receiver C9 from section 4.3.1 we can see that for SNR higher than 15 db the hybrid scheme performs better and decreases the BER. Receiver J with decorrelator and two-stage SIC for the data detection outperforms the three-stage SIC scheme C27 operating on the same structure. Receiver I with only one-stage SIC for the data detection performs as well as the double SIC scheme C9. However, as was shown in Section 3.5.1 the complexity of receiver I is lower than that of receiver C9, especially for longer frames.

Figure 4.11 presents results for a despread SNR of 25 dB. If we compare the results with the SIC C9 from figure 4.5 that operates on the one pilot frame with $M = 9$, the hybrid receiver J offers an advantage from a population of 6 user on. In comparison with receiver C27 the hybrid approach allows a larger capacity in the system for the same BER.

For a fixed error rate of $3 \cdot 10^{-3}$, receiver J can accomodate 24 user whereas receiver C27 is only capable of allowing 18 user. Receiver C9 reaches this BER for a user population of 13 user.

Comparing results for both receiver types, the double SIC (C27) and the hybrid receiver operating on the same frame size, the hybrid receiver (J, K) outperforms all PSAM schemes. For the case of 15 dB SNR the hybrid approach is still performing better than the SIC C27 operating on the same frame. However, the SIC C9 with the short frame shows now a slight advantage over the other schemes considered. The ideal SIC receiver H is also shown in the graphs to set the limit that can be achieved by using perfect channel estimates at the receiver.

4.4.2 Three Pilot Symbols

Hybrid receiver operating on the frame structure with three consecutive pilot symbols are expected to perform worse than if they would operate on the pilot frame with guard intervals as considered before in section 4.3.2. The reason is simply that the fading gain $\mathbf{G}(n)$ in equation 3.10 is not truly constant over three symbol intervals and therefore the decorrelator cannot completely decouple the pilot symbols. This will result in some residual interference in the fading gain estimates and therefore to a lower performance. Indeed, figure 4.13 shows a degraded performance as compared to the same receiver operating on the frames with inserted pilot symbol and guard intervals and the error floor develops at a higher bit error level. If the results are compared to the SIC operating on the 9 symbol frame (C9) the system performance is worse up to an SNR of approx. 22 dB. For higher SNR the hybrid receiver (J, K) can lower the error floor and offers an advantage.

For a system with an SNR of 25 dB eight users in the cell are neccessary for the hybrid approach to perform equally well as receiver C9. For a cell population of 20 user receiver C9 could serve 16 user for the same average BER. For an SNR of 15 db in the system the hybrid receiver performs worse than the SIC receiver C9 operating on the shorter frame length of 9.

4.4.3 One Pilot Symbol and Shortened Integration Time

This section presents results for receiver I, J and K operating with a shorter integration time in the pilot symbol interval as presented in section 3.2. The amplitude in the pilot symbol transmissions is increased to account for the SNR loss introduced by the

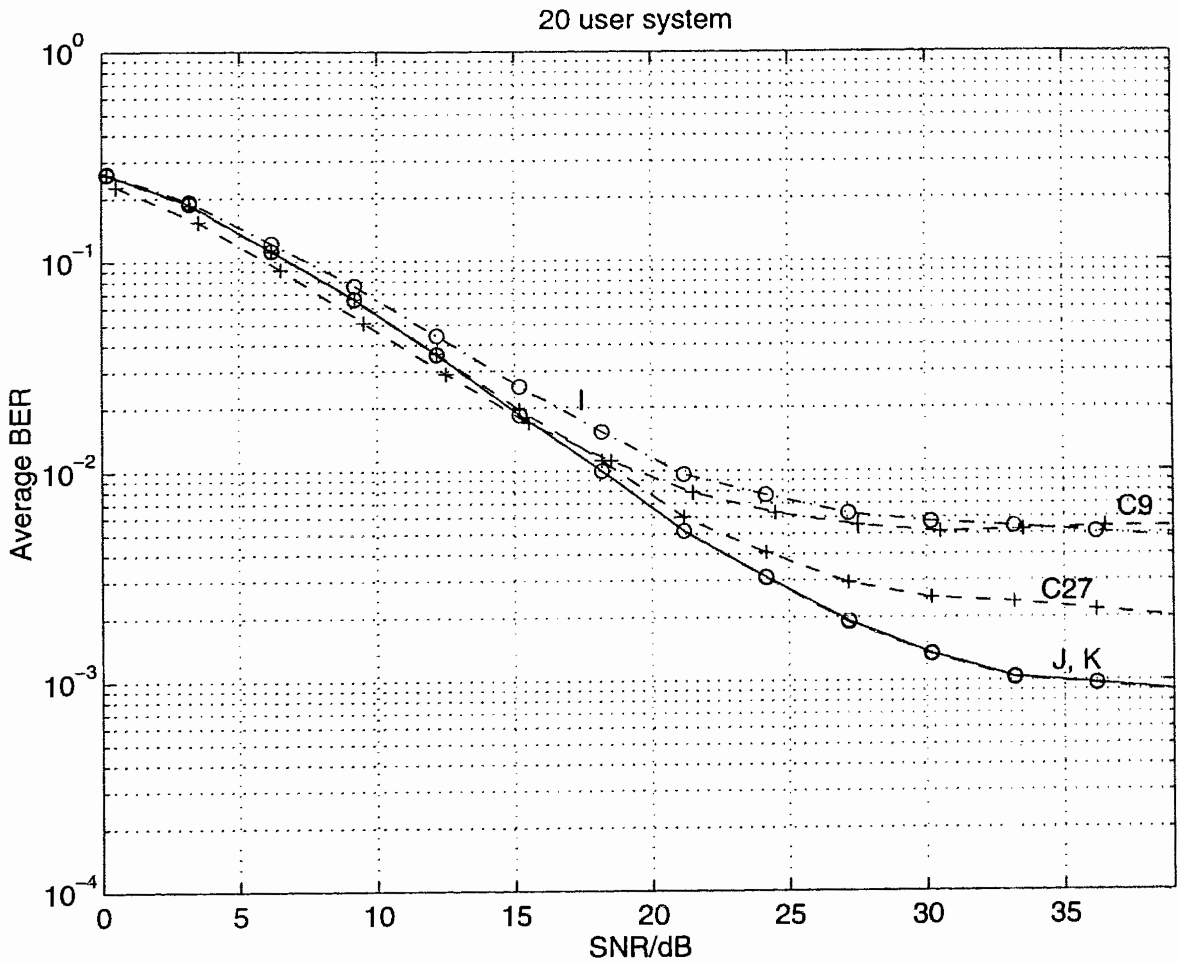

Figure 4.10: One pilot w/ guard intervals, Hybrid Rec., BER vs. SNR, 20 User System

Parameter for Simulation	
Parameter	**Value**
Spreading Factor	31
Despread SNR/dB	15
Pilot Block	One Pilot with Guard Intervals
Interpolator length	11
Data-Frame Length	27
Number of User	variable
$f_d T$	0.01
τ_{max}/T	1.0

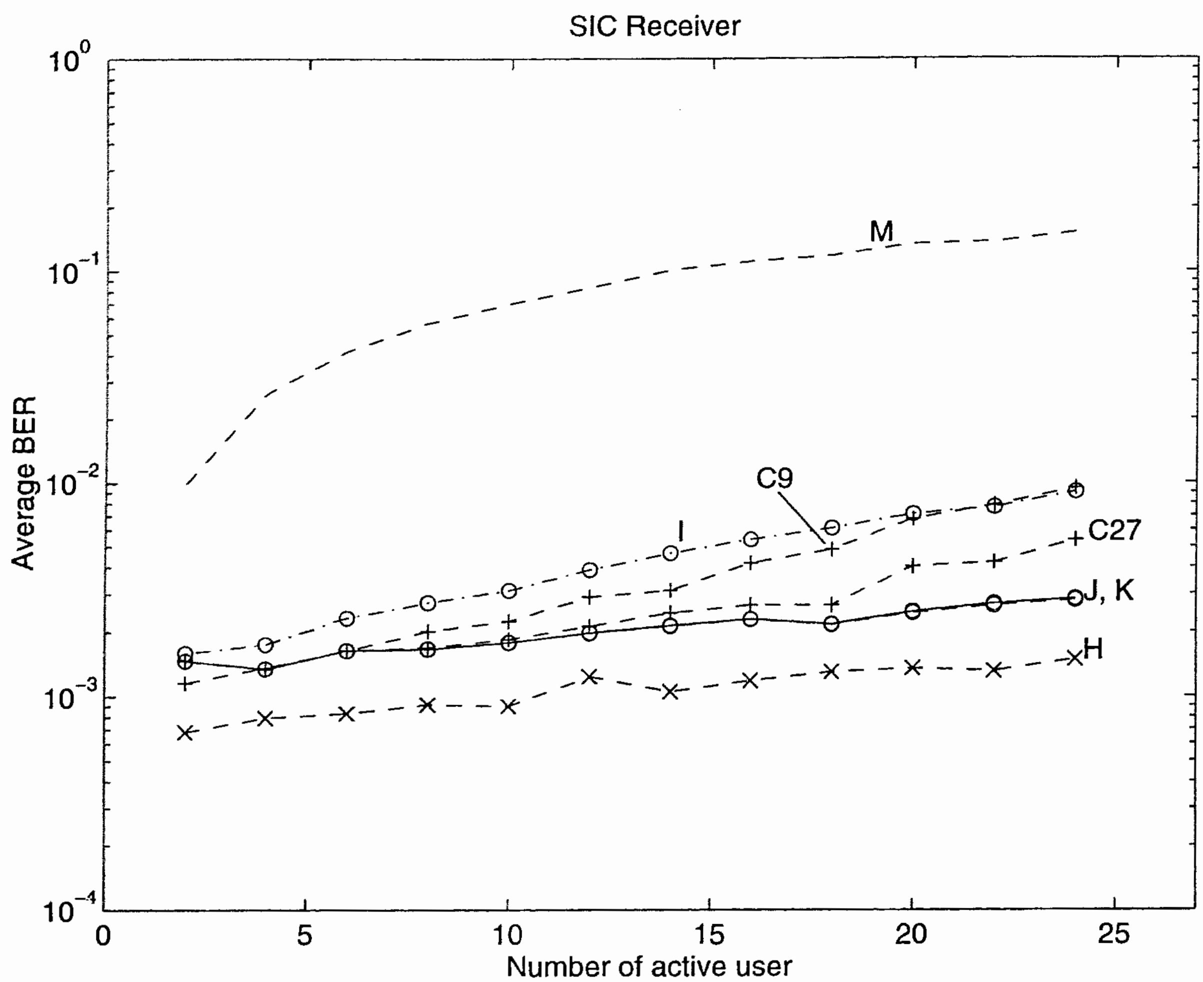

Figure 4.11: One pilot w/ guard intervals, Hybrid Rec., BER vs. user, SNR 25 dB

Parameter for Simulation	
Parameter	**Value**
Spreading Factor	31
Despread SNR/dB	15
Pilot Block	One Pilot with Guard Intervals
Interpolator length	11
Data-Frame Length	27
Number of User	variable
$f_d T$	0.01
τ_{max}/T	1.0

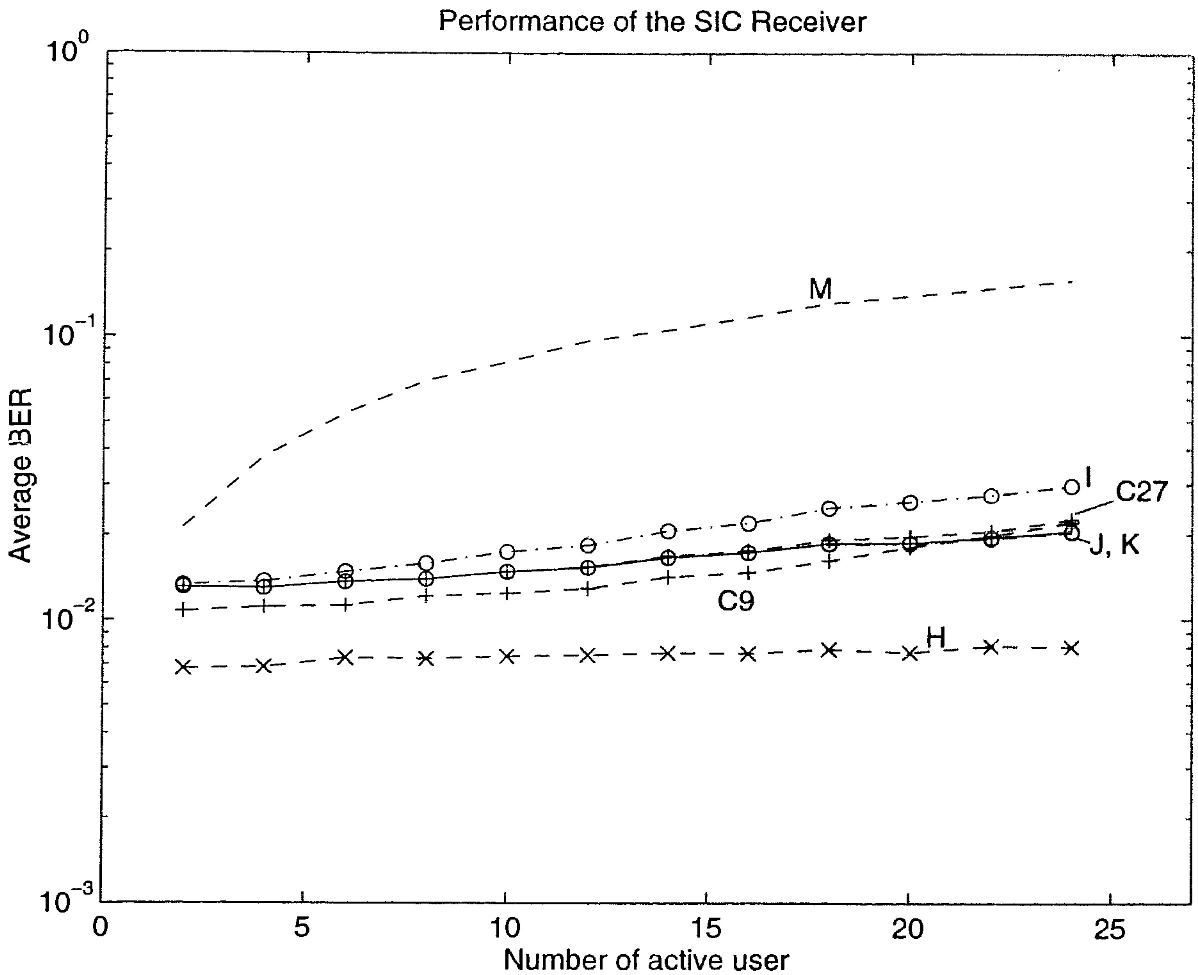

Figure 4.12: One pilot w/ guard intervals, Hybrid Rec., BER vs. user, SNR 15 dB

Parameter for Simulation	
Parameter	**Value**
Spreading Factor	31
Despread SNR/dB	15
Pilot Block	One Pilot with Guard Intervals
Interpolator length	11
Data-Frame Length	27
Number of User	variable
$f_d T$	0.01
τ_{max}/T	1.0

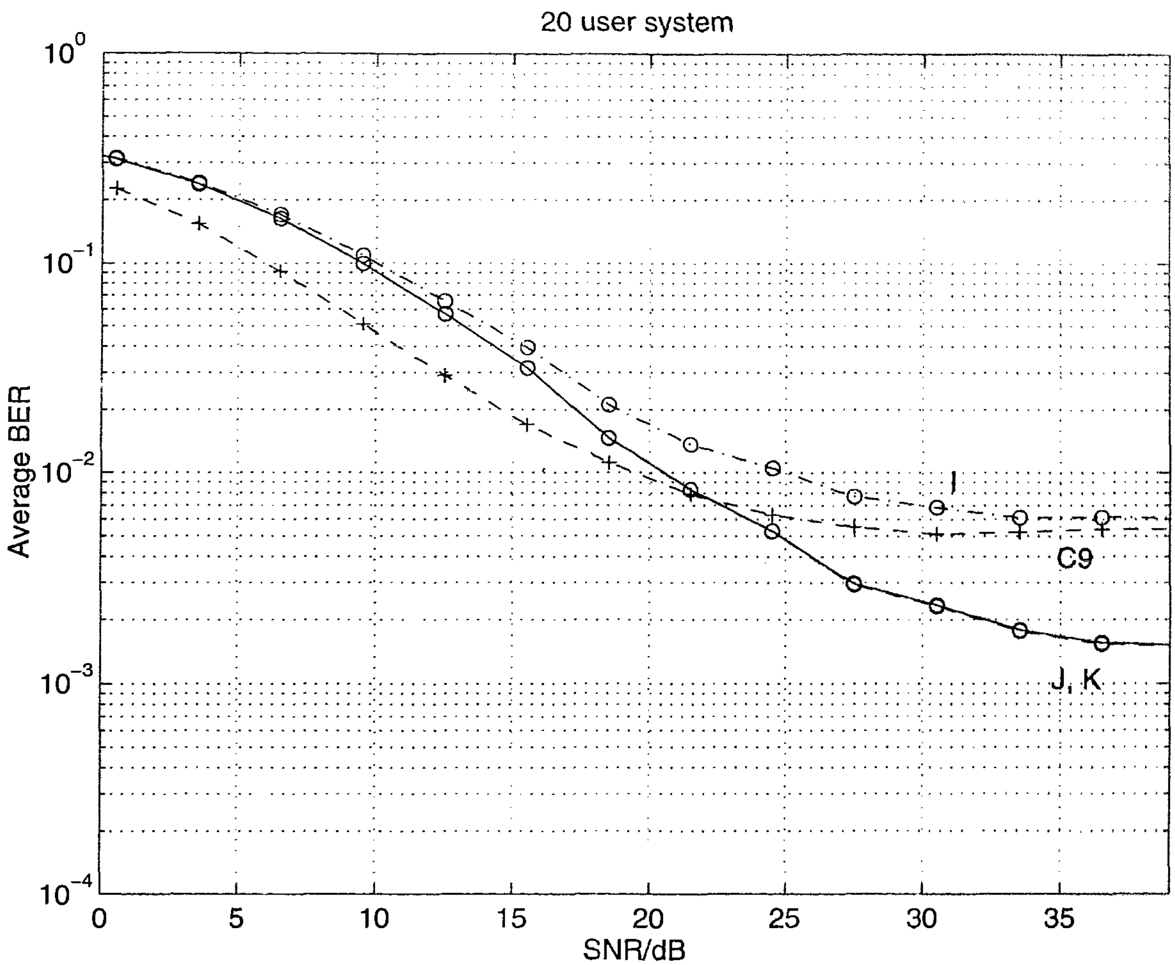

Figure 4.13: Three pilot frame, Hybrid Rec., BER vs. SNR, 20 User System

Parameter for Simulation	
Parameter	**Value**
Spreading Factor	31
Despread SNR/dB	variable
Pilot Block	Three pilot symbols
Interpolator length	11
Data-Frame Length	27
Number of User	20
$f_d T$	0.01
τ_{max}/T	1.0

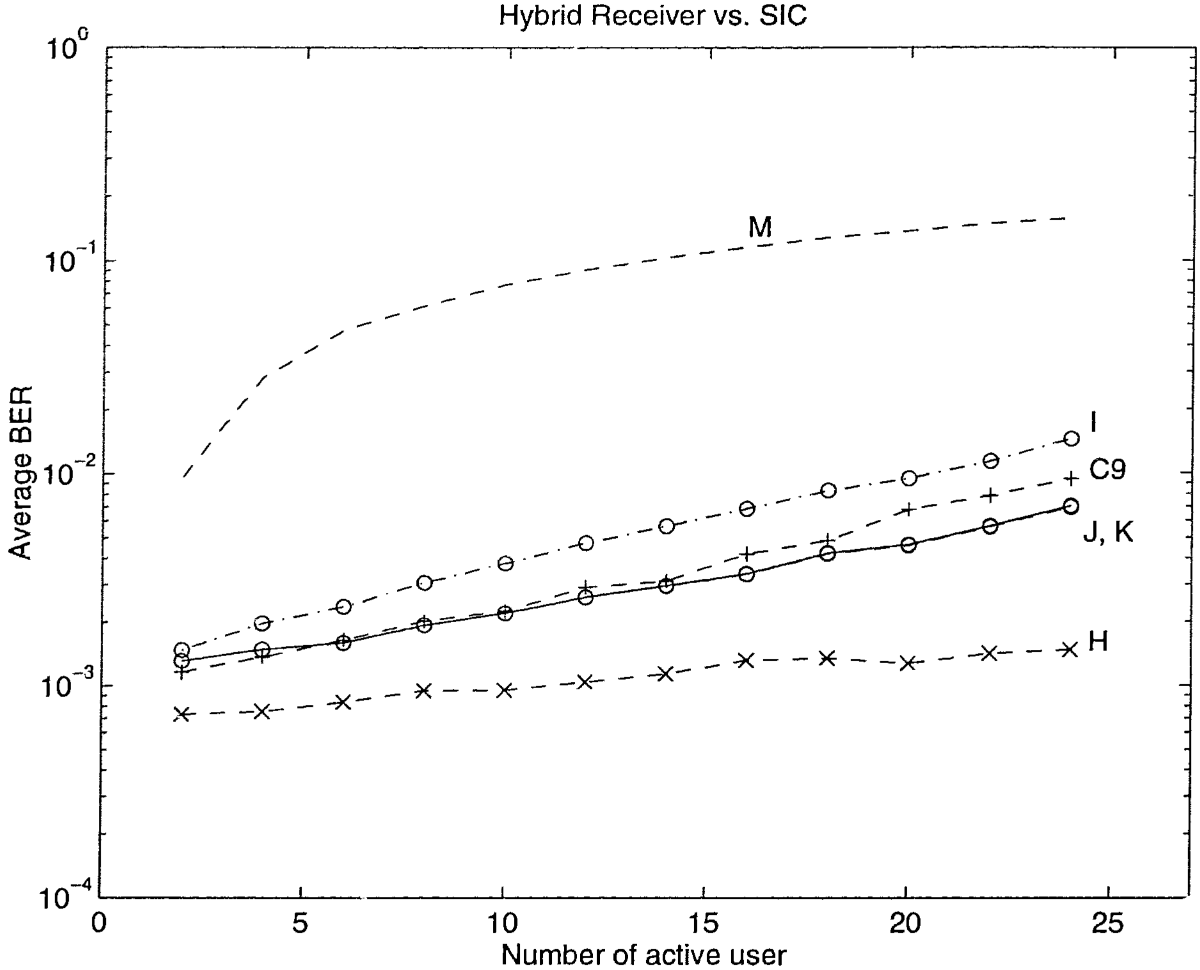

Figure 4.14: Three pilot frame, Hybrid Rec., BER vs. user, SNR 25 dB

Parameter for Simulation	
Parameter	**Value**
Spreading Factor	31
Despread SNR/dB	25
Pilot Block	Three pilot symbols
Interpolator length	11
Data-Frame Length	27
Number of User	variable
$f_d T$	0.01
τ_{max}/T	1.0

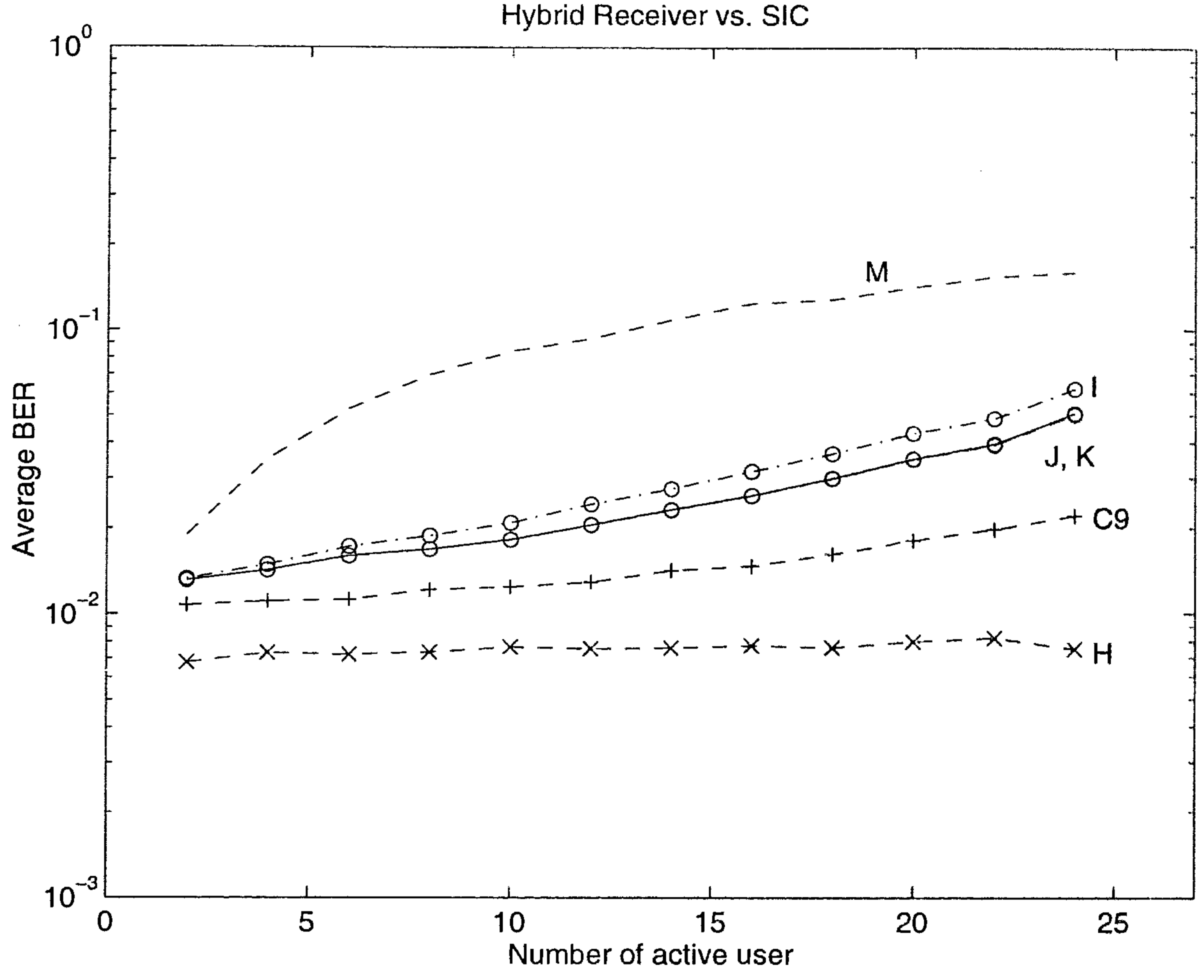

Figure 4.15: Three pilot frame, Hybrid Rec., BER vs. user, SNR 15 dB

Parameter for Simulation	
Parameter	**Value**
Spreading Factor	31
Despread SNR/dB	15
Pilot Block	Three pilot symbols
Interpolator length	11
Data-Frame Length	27
Number of User	variable
$f_d T$	0.01
τ_{max}/T	1.0

shorter integration time in the receiver. The double SIC schemes A, B, C and D will also perform on these transmissions, however, the pilots were in this case transmitted with the same power as the data. The hybrid receiver I, J and K can only operate on this frame structure in cells with a smaller time shift than T. Two different maximum time delays have been considered. A small cell (cell radius 3.9 km) with a $\tau_{max} = 0.25T$ and a medium size (radius 7.8 km) cell with $\tau_{max} = 0.5T$ have been simulated. The pilots were transmitted with 1.333 and 2 times the power used for transmitting the data, respectively.

Figure 4.16 shows results for a 20 user system with a maximum time delay of $\tau_{max} = 0.25T$. It can be seen that the hybrid receiver I and J offer a good solution for this scenario. In comparison with the SIC receiver A, B, C and D, the hybrid receiver reduces the error floor and enables a capacity increase in the system. It can also be observed that all SIC receiver improve their performance as compared to the system with a maximum time delay of T considered in section 4.3.1.

For a varying number of user in the system and a channel SNR of 25 dB the hybrid approach decreases the BER for a higher loaded cell. For a BER of $2 \cdot 10^{-3}$ the hybrid receiver can serve 20 user whereas the SIC receiver C accomodates only 15 user. For a SNR of 15 dB the hybrid receiver shows a similar performance as the SIC schemes.

A time shift of $\tau_{max} = 0.5T$ degrades the performance of the hybrid schemes and they experience a worse BER in the low SNR region (figure 4.19). Still, in the low noise region the approach is beneficial as compared to the SIC schemes.

Figure 4.20 shows results for a system with a different number of users. The hybrid receiver shows now an almost similar performance as compared to the three-stage SIC C. If the channel SNR increases to 15 dB the performance of the hybrid receiver is worse than the comparison systems B, C and D.

4.5 Discussion

We presented in this chapter simulation results for different successive interference cancelling receiver and two main concepts were investigated.

The double SIC scheme was implemented in a different number of stages and a hybrid receiver was presented that combined a decorrelator and an SIC data detector. Three different data transmission frame structures have been proposed that enable the use of a hybrid receiver. Among the three, the one pilot frame with preceding and following guard interval seems to be the most promising approach for both receiver concepts. Using the hybrid receiver on this frame structure offers an additional capacity increase.

Whether one of the proposed frame structures and receivers should be used in a practical application depends on the expected system parameters and other issues like delay, cell size, SNR and more. It could be seen that for low SNR in the system the more favourable approach would be to use a short frame length to decrease the estimation error and BER. In the high SNR region the multiuser access interference limits the performance and a receiver placing more emphasis on removing this interference would be the right choice. The hybrid receiver could be chosen for this kind of channel SNR. It offers in addition a complexity that is in most cases lower than the comparable double SIC scheme.

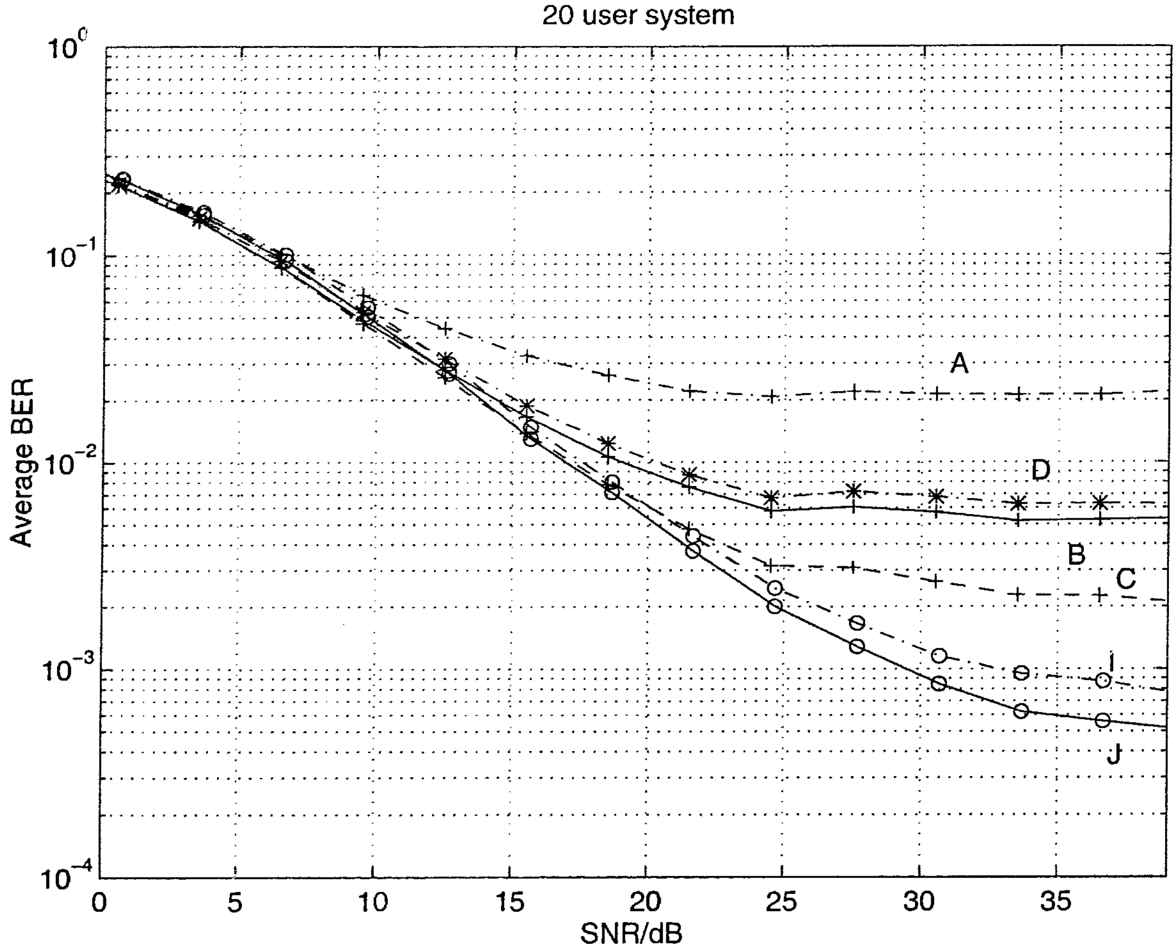

Figure 4.16: One pilot w/ higher power, BER vs. SNR, 20 user system, $\tau_{max} = 0.25T$

Parameter for Simulation	
Parameter	**Value**
Spreading Factor	31
Despread SNR/dB	variable
Pilot Block	One Pilot with 1.33 times higher power
Interpolator length	11
Data-Frame Length	9
Number of User	20
$f_d T$	0.01
τ_{max}/T	0.25

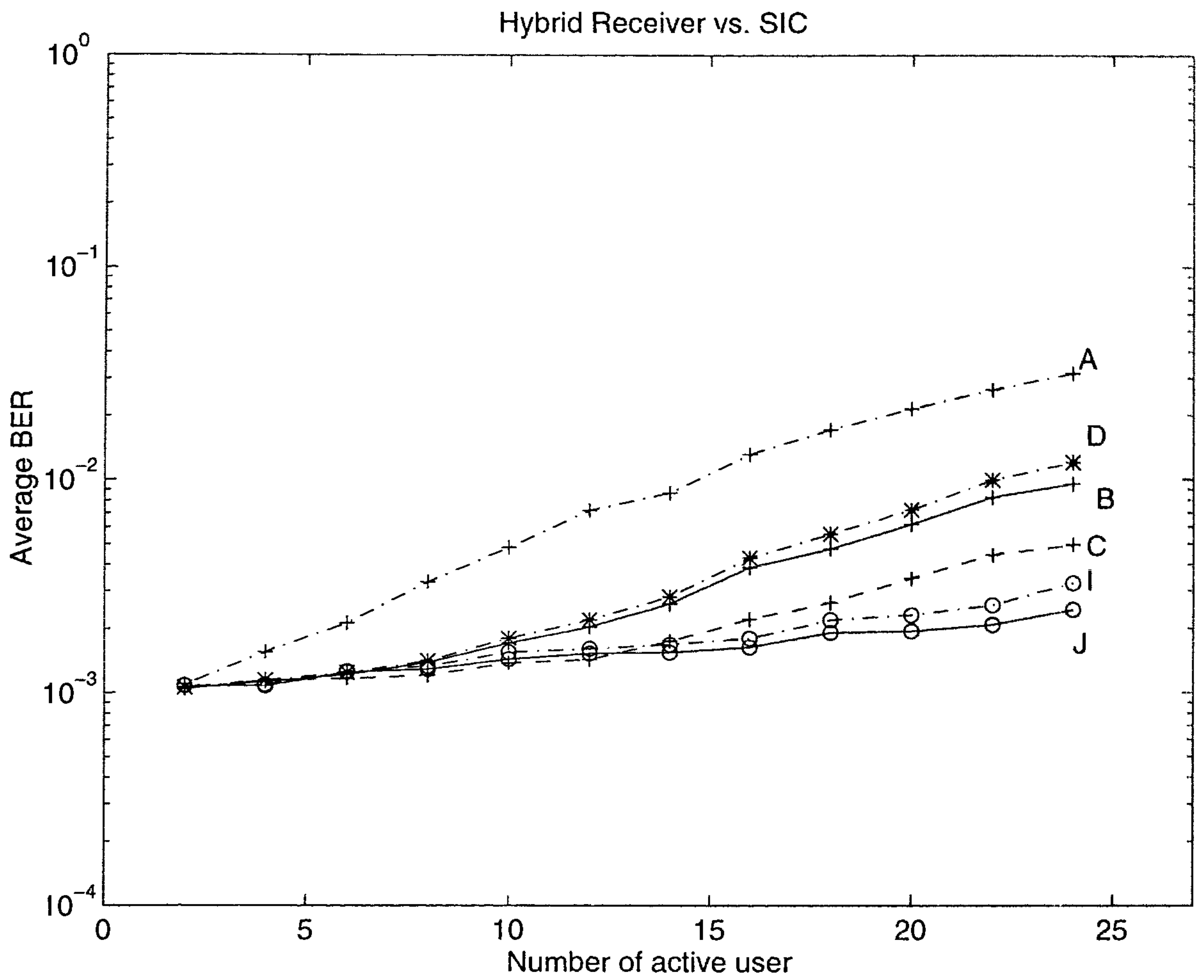

Figure 4.17: One pilot w/ higher power, BER vs. user, SNR 25 dB, $\tau_{max} = 0.25T$

Parameter for Simulation	
Parameter	**Value**
Spreading Factor	31
Despread SNR/dB	25
Pilot Block	One Pilot with 1.33 times higher power
Interpolator length	11
Data-Frame Length	9
Number of User	variable
$f_d T$	0.01
τ_{max}/T	0.25

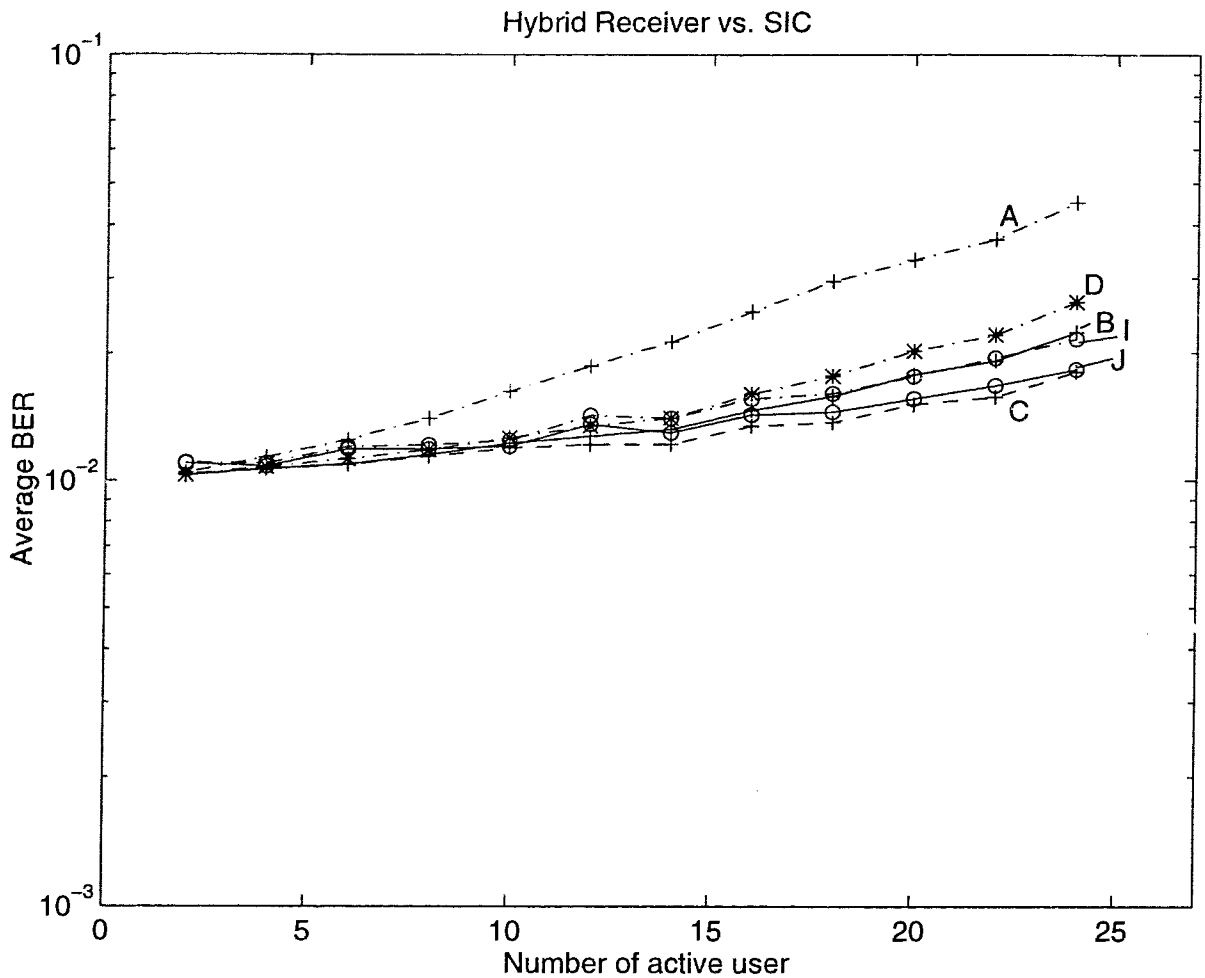

Figure 4.18: One pilot w/ higher power, BER vs. user, SNR 15 dB, $\tau_{max} = 0.25T$

Parameter for Simulation	
Parameter	**Value**
Spreading Factor	31
Despread SNR/dB	15
Pilot Block	One Pilot with 1.33 times higher power
Interpolator length	11
Data-Frame Length	9
Number of User	variable
$f_d T$	0.01
τ_{max}/T	0.25

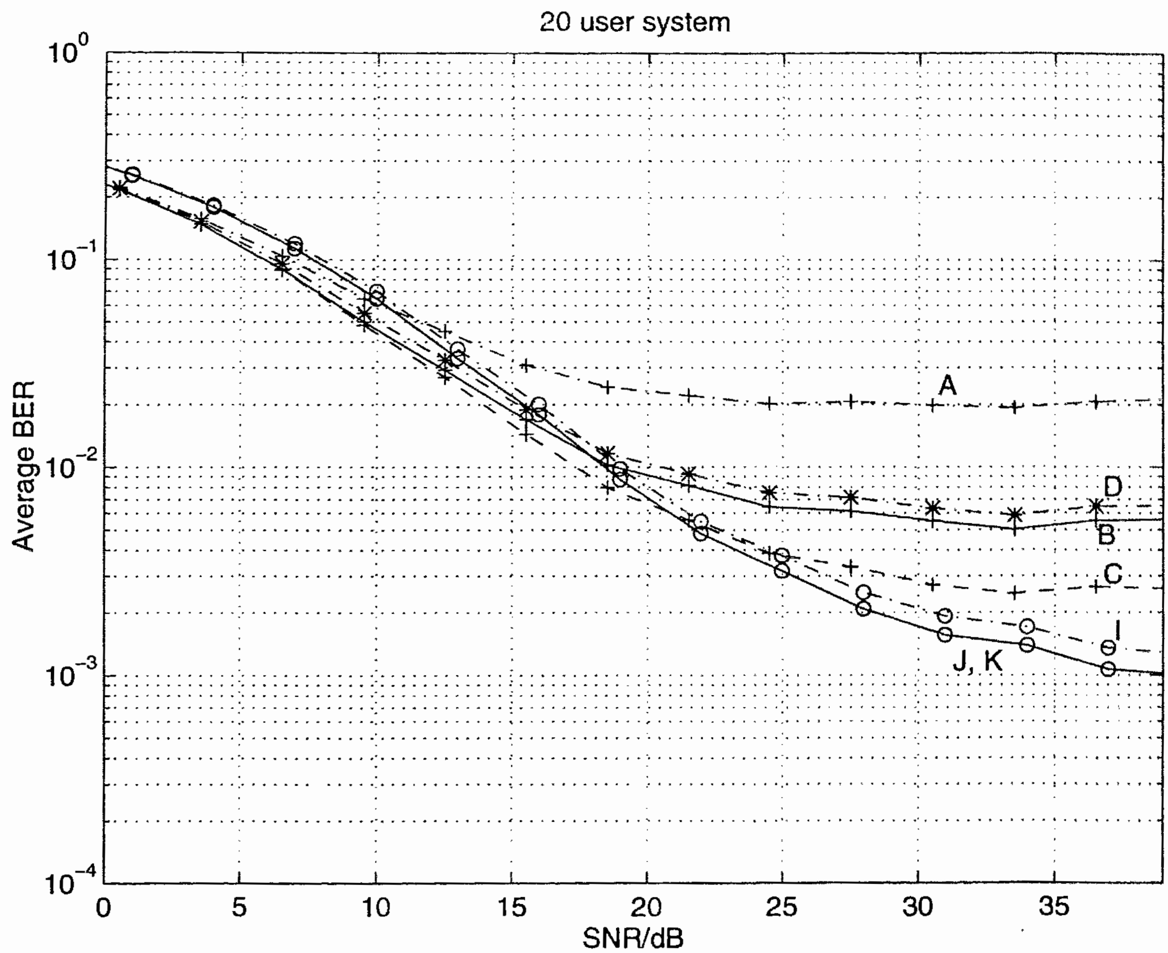

Figure 4.19: One pilot w/ higher power, BER vs. SNR , 20 User System, $\tau_{max} = 0.5T$

Parameter for Simulation	
Parameter	**Value**
Spreading Factor	31
Despread SNR/dB	variable
Pilot Block	One Pilot with double transmitted power
Interpolator length	11
Data-Frame Length	9
Number of User	20
$f_d T$	0.01
τ_{max}/T	0.5

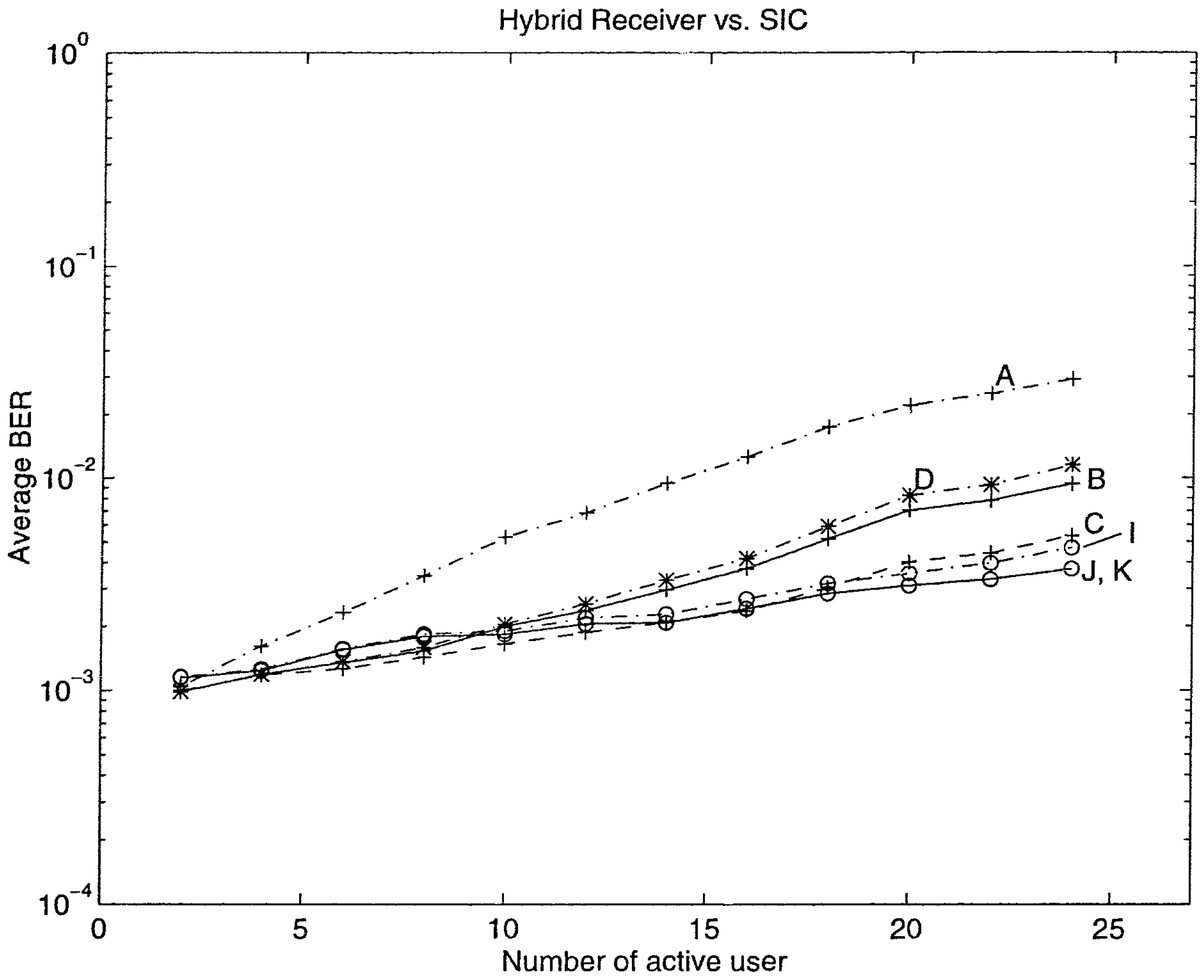

Figure 4.20: One pilot w/ higher power, BER vs. user, SNR 25 dB, $\tau_{max} = 0.5T$

Parameter for Simulation	
Parameter	**Value**
Spreading Factor	31
Despread SNR/dB	25
Pilot Block	One Pilot with double transmitted power
Interpolator length	11
Data-Frame Length	9
Number of User	variable
$f_d T$	0.01
τ_{max}/T	0.5

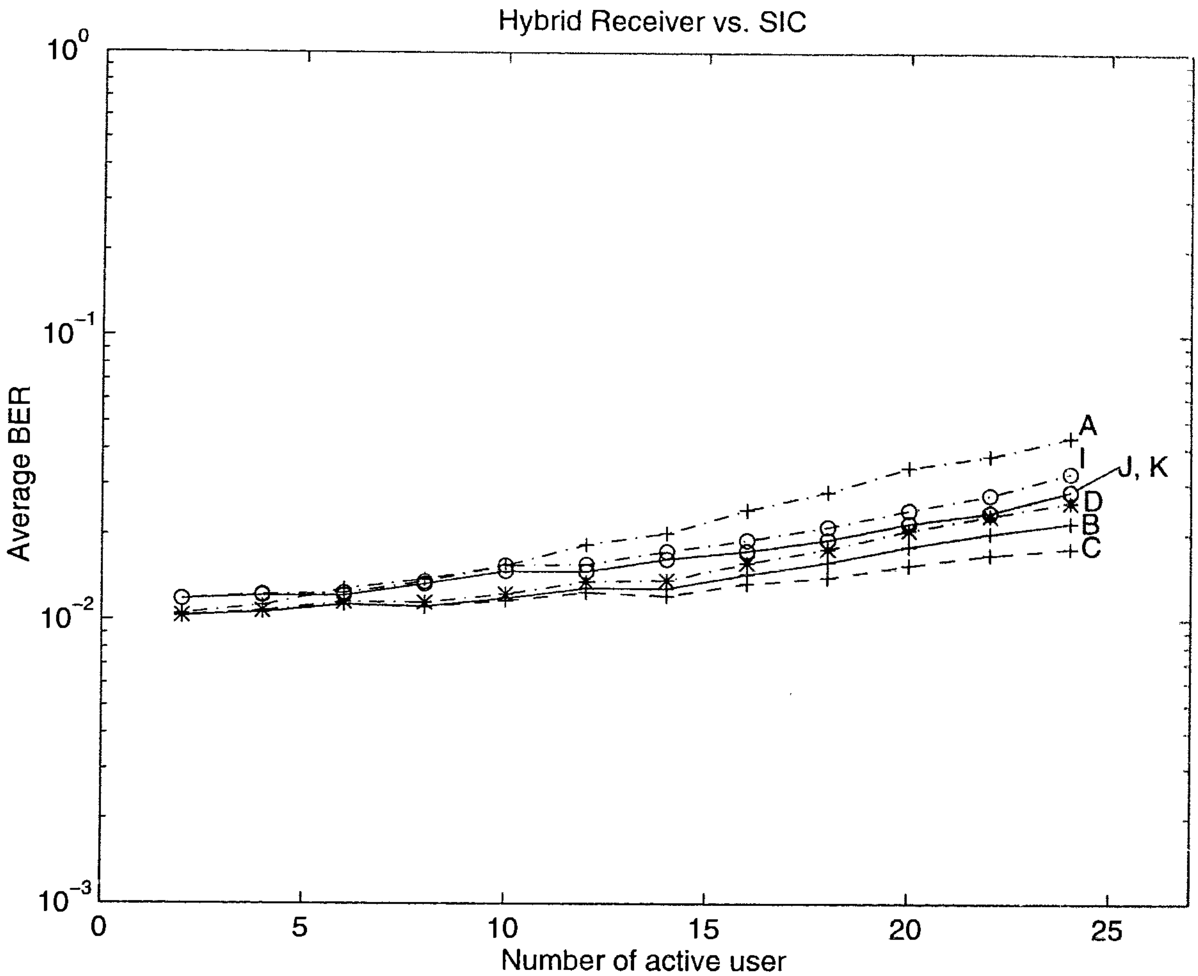

Figure 4.21: One pilot w/ higher power, BER vs. user, SNR 15 dB, $\tau_{max} = 0.5T$

Parameter for Simulation	
Parameter	**Value**
Spreading Factor	31
Despread SNR/dB	15
Pilot Block	One Pilot with double transmitted power
Interpolator length	11
Data-Frame Length	9
Number of User	variable
$f_d T$	0.01
τ_{max}/T	0.5

Chapter 5

Conclusions

5.1 Conclusions

This thesis has focused on pilot symbol assisted successive interference cancelling receiver operating in a Raleigh fading channel.

The main concepts that have been investigated were identical in the way the data detection was performed—a successive interference cancellation scheme using previously obtained channel estimates for the detection and regeneration of the signals was employed.

Based on this receiver model two different multiuser access interference suppression schemes were used for the pilot symbol intervals. The double SIC receiver also used a SIC concept for this purpose and the regeneration of the pilot symbol transmissions was done by using the correlator outputs as channel estimates. The performance of this receiver was investigated on two different frame structures. It was found that in a quasi-synchronous system, the use of the frame structure with one pilot symbol preceeded and followed by a guard interval was a good choice under all noise conditions. For a SNR higher than 17 dB, using this frame structure was beneficial over a frame without guard intervals and of comparable length. Moreover, the error floor was reduced and the system capacity was improved. For a SNR lower than 17 dB a slight degradation of the system performance was observed.

The hybrid receiver approached the problem of MAI in the pilot symbol intervals by using a decorrelator that could completely decouple the different users' signals. To keep the complexity of the decorrelator at a moderate level, different pilot insertion schemes have been proposed that enable the use of a synchronous decorrelator with a matrix of order $K \cdot K$ for operation in quasi-synchronous cells.

For quasi-synchronous systems with a maximum time delay of T, a frame with three consecutive transmitted pilot symbols and the frame with one pilot symbol plus guard intervals were investigated in combination with the hybrid receiver. It was found that the three-pilot frame in combination with the hybrid receiver yielded to a decreased performance in the low SNR region as compared to the double SIC scheme. Benefits could only be achieved for SNR higher than 27 dB. The one pilot frame with guard intervals is clearly a better choice for use with the hybrid receiver. It was shown that the hybrid approach could reduce the residual error floor and the system capacity as compared to the double SIC receiver when both systems were operating under similar conditions. For high noise level the hybrid scheme was at least as good as the double SIC receiver.

The use of the hybrid receiver in medium and small cell sites with a maximum delay of $0.5T$ and $0.25T$, respectively, and only one inserted pilot symbol per frame was possible by shortening the integration time during the pilot symbol reception.

It could be seen that for the small cell size with a maximum delay of $0.25T$ for the received signals, the hybrid receiver improves the error floor in the system and for SNR higher than 15 dB the system capacity can be increased. Using the same concept in a medium size cell with a maximum delay of $0.5T$ yielded to some performance decrease in the high noise region. However, the error floor in the system could still be decreased by the use of the hybrid approach.

In summary, the presented hybrid receiver in combination with the frame structure with one pilot symbol and guard intervals is a very good choice. It can be used for synchronous cells up to cells where the maximum time delay is smaller or equal to one symbol time T. The system capacity can be improved in the low noise region and there is no performance degradation for lower SNR. The updating of the inverse of the crosscorrelation matrix that is needed for this receiver can be efficiently done by an order recursive algorithm. The receiver can therefore be regarded as a good compromise scheme for a quasi-synchronous DS/CDMA system.

5.2 Future Work

An important topic that should be considered is the frequency selective mobile channel and the performance of a Rake hybrid receiver operating in this environment. Multi-cell interference could be integrated into the system model and their influence on the BER could be investigated. Further work should be done on the partial analysis of the system to verify the simulation results.

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