Adaptive DSP Algorithms for UMTS:
Blind Adaptive MMSE and PIC Multiuser Detection

Adaptive Wireless Networking (AWGN)

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Abstract

A study of the application of blind adaptive Minimum Mean Square Error (MMSE) and Parallel Interference Cancellation (PIC) multiuser detection techniques to Wideband Code Division Multiple Access (WCDMA), the physical layer of Universal Mobile Telecommunication System (UMTS), has been performed as part of the Freeband Adaptive Wireless Networking project. This study was started with an analysis of Code Division Multiple Access (CDMA) and conventional CDMA detection. After that blind adaptive MMSE and PIC detection have been analyzed for general CDMA systems. Then the differences between WCDMA and general CDMA were analyzed and the results have been used to determine how blind adaptive MMSE and PIC can be implemented in WCDMA systems.

Blind adaptive MMSE has been implemented in WCDMASim, a WCDMA simulator and some preliminary simulation results obtained with this simulator are presented. These simulation results do not yet show the performance that was expected of blind adaptive MMSE detection based on simulation results obtained in previous research. The cause for these unexpected results is not yet known and will be the subject of further research.

Implementation of PIC detection in WCDMASim was found to require changes to the architecture of the WCDMASim simulator. Implementation of these changes and solving the problems with blind adaptive MMSE detection are considered for future work.
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Acronyms

AWGN     Additive White Gaussian Noise
BER      Bit-Error-Rate
CDMA     Code Division Multiple Access
DPCCH    Dedicated Physical Control Channel
DPDCH    Dedicated Physical Data Channel
DPDCHs   Dedicated Physical Data Channels
DSP      Digital Signal Processing
FDD      Frequency Division Duplex
FPFAs    Field Programmable Function Arrays
FPGAs    Field Programmable Gate Arrays
GPPs     General Purpose Processors
NRZ      Non-Return-to-Zero
MAI      Multiple Access Interference
MIC      Multistage Interference Cancellation
MMSE     Minimum Mean Square Error
MOE      Mean Output Energy
MSE      Mean Square Error
OVSF     Orthogonal Variable Spreading Factor
PIC      Parallel Interference Cancellation
SIC      Successive Interference Cancellation
SNR      Signal to Noise Ratio
SNRs     Signal to Noise Ratios
TFCI     Transport Format Combination Indicator
TPC      Transmission Power Control
FBI      Feedback Information
UMTS     Universal Mobile Telecommunication System
WCDMA    Wideband Code Division Multiple Access
Chapter 1

Introduction

The Freeband Adaptive Wireless Networking (AWGN) project aims to develop methods and technologies that can be used to design efficient adaptable and reconfigurable base stations and terminals for third and fourth generation wireless communication systems, as for example UMTS. In order to achieve this, the project consists of two activities.

The first activity is the mapping of (different) communication systems on a platform of heterogeneous reconfigurable architectures. The platform is heterogenous in the sense that Digital Signal Processing (DSP) is performed on General Purpose Processors (GPPs), Field Programmable Gate Arrays (FPGAs) and Field Programmable Function Arrays (FPFAs). These architectures are all reconfigurable in the sense that they can be reprogrammed in-system with new DSP algorithms.

The second activity is the study and development of adaptive DSP algorithms for UMTS. These algorithms are adaptive in the sense that they allow the communication system to adapt to changing environmental conditions (e.g., changing numbers of users in a cell or varying noise figures due to reflections or user movements) as well as changing user demands.

This report is the first report of the ‘adaptive DSP algorithms for UMTS activity’. The focus of the research described in this report is on multiuser detection algorithms which allow a UMTS base station to adapt to the number of active users in a cell and reduce the influence of Multiple Access Interference (MAI).

MAI, the interference that the signal of a user experiences caused by the signals of the other users on a channel, is inherent in cellular wireless Code Division Multiple Access (CDMA) communications systems. Conventional CDMA receivers treat MAI as if it were additive random noise. In multiuser detection information about the signals of the other users is used to improve detection of the signal of each individual user. In previous research ([2], [3]) blind adaptive Minimum Mean Square Error (MMSE) multiuser detection and Parallel Interference Cancellation (PIC) multiuser detection
Chapter 1. Introduction

were found to be two of the more promising multiuser detection techniques for CDMA communications systems. In the CDMA model that was used in this previous research however, a lot of the effects that are present in an actual cellular wireless CDMA communications system, as for example fading and multipath, were ignored. Therefore in this report, blind adaptive MMSE and PIC multiuser detection will be studied using an accurate model of the uplink of Wideband Code Division Multiple Access (WCDMA), the physical layer of the Universal Mobile Telecommunication System (UMTS) communications system. The goals of this study are to:

1. Determine if the blind adaptive MMSE and PIC multiuser detection techniques can be applied to WCDMA.

2. Implement blind adaptive MMSE and PIC multiuser detection in a WCDMA simulator.

3. Determine the performance of blind adaptive MMSE and PIC multiuser detection in WCDMA through simulations and assess the effects of detector parameter choices.

In a later phase of the project the obtained results can be used to study the implementation related issues of blind adaptive MMSE and PIC multiuser detection. This serves as a preparation for the implementation of these detection techniques in a test-bed. Eventually this will lead to an implementation of blind adaptive MMSE and PIC multiuser detection for UMTS on a heterogeneous reconfigurable architecture.

Chapter 2 of this report starts with a general description of CDMA. After that the CDMA system models that are frequently used in literature describing multiuser detection are given. These CDMA system models were also used in the mentioned previous research. Finally Chapter 2 contains a description of conventional CDMA detection. The models described in Chapter 2 are used to describe blind adaptive MMSE detection in Chapter 3 and PIC detection in Chapter 4. In Chapter 5 a mathematical model for WCDMA is developed. This WCDMA model is used to describe blind adaptive MMSE detection for WCDMA in Chapter 6 and PIC detection for WCDMA in Chapter 7. Chapter 8 describes the simulator that is used to study the performance of blind adaptive MMSE and PIC detection. This chapter also describes the extensions to the simulator that have been implemented to support simulation of these two detection techniques. In Chapter 9 some preliminary simulation results are presented. Finally in Chapter 10, the conclusions and recommendations that can be drawn from the research described in this report are given. Chapter 10 also gives some directions for future research.
Chapter 2

CDMA and Conventional CDMA Detection

In this chapter first a general description of CDMA is given. After that continuous- and discrete-time CDMA system models are given that are frequently used in literature describing multiuser detection. These system models will be used to describe blind adaptive MMSE and PIC detection in the next two chapters. At the end of this chapter conventional CDMA detection is discussed briefly.

2.1 CDMA

CDMA uses direct-sequence spread-spectrum techniques to achieve efficient multiple access communications. In a direct sequence spread spectrum transmitter each bit of a binary nonreturn-to-zero information signal is modulated by one period of a binary nonreturn-to-zero pseudo-random sequence to generate the transmitted signal, see Figure 2.1. This pseudo-random sequence is also referred to as signature sequence, signature waveform, or in an older terminology code, explaining the term code-division multiple access. The pseudo-random sequence is composed of elementary pulses of duration $T_c$ commonly referred to as chips. Because the duration of a chip of the pseudo-random sequence is usually a lot shorter than the duration $T$ of a bit of the information signal the modulated signal will be a wide-band signal with nearly the same spectrum as the pseudo-random sequence. The bandwidth expansion ratio $T/T_c$ is also known as the spreading gain. In a CDMA system the particular signature sequence of a user identifies the particular point-to-point channel corresponding to that user.

CDMA receivers employ the signature sequence of a user as the key to recover the transmitted information. Detection of the transmitted data is accomplished with a correlation demodulator driven by a synchronized replica of the signature waveform used at the transmitter, see Figure 2.2.
Chapter 2. CDMA and Conventional CDMA Detection

Figure 2.1: Direct-sequence spread-spectrum transmitter.

The by design low correlation between the signature waveforms of the different users gives the CDMA system its multiple access properties.

In an ideal CDMA system orthogonal signature sequences would be used. A CDMA system using orthogonal signature sequences will cancel out all Multiple Access Interference (MAI) and yield single user performance. However, fully orthogonal systems are not practical for two reasons. First, for a given limited number of chips there only exist a limited number of orthogonal signature sequences. Secondly, and more importantly, if the users are not transmitting synchronously, the signature sequences are out of phase and lose their orthogonal property. CDMA systems normally use shift-register sequences or combinations of shift-register sequences for their signature se-

Figure 2.2: Direct-sequence spread-spectrum receiver.

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2.1. CDMA

sequences. It is not possible to obtain signature sequences for any pair of users that are orthogonal for all time offsets using this method.

Not having completely orthogonal signature sequences would not cause a dramatic performance decrease, if the received power of the interfering signal is smaller than the received power of the signal of the desired user. However, if the interfering signals are much stronger than the signal of the desired user, proper detection becomes impossible. This scenario occurs often in practical systems. For example consider the link from a cellular phone to a base station. If one cellular phone is transmitting from a position close to the base station and another cellular phone is transmitting from a position further away from the base station, then the received power of the signal of the first cellular phone will be higher than that of the distant cellular phone, assuming that their transmit powers are equal. Thus, the detection of the distant user will result in a severe increase in bit errors. This situation is generally called the near-far problem.

A natural solution to the near-far problem is the use of power control. The power control system compares the received power levels. It uses a control channel to transmit power status information to the cellular phones. The cellular phone will adapt its transmitting power in such a way that the received power at the base station is equal for all users. Power control alleviates the near-far problem at the expense of receiver complexity and decreased usable bandwidth.

The length is another property of the signature sequences that has to be taken into consideration. Traditionally signature sequences with a length $N$ were used in CDMA so that $NT_c = T$, thus the period of the signature sequence is equal to the duration of a bit, so each bit is modulated by the entire signature sequence. This kind of signature sequences are so called short codes. The disadvantage of using signature sequences with a relatively short length is that the crosscorrelations between the signature sequences vary relatively strongly from each other. Since the crosscorrelations between the signature sequence of a channel and the signature sequence of the other channels determines the amount of interference on the channel, varying crosscorrelations will result in a bit-error-rate performance that varies between channels. To solve this so called long codes can be used.

Long codes are signature sequences with a period that is much longer than the duration of a bit, for example signature sequences with a period of $2^{22}T_c$. When these signature sequences are used to modulate a bit stream each bit will be modulated by a different section of the signature sequence. The amount of interference for a particular bit therefor depends on the crosscorrelations between the section of the signature sequence that is used to modulate that bit and the sections of the signature sequences that are used to modulate the bits that are transmitted at the same time in the other channels. This results in an interference on each channel that varies from bit to bit, but averaged over the bits there will be the same amount of
interference on each channel. Therefore the bit-error-rate performance on all
the channels will be the same.

2.2 Continuous-time Synchronous CDMA Model

The basic continuous-time synchronous $K$-user CDMA model describes the
received signal of a CDMA system in which the bit streams of $K$ synchronous
users antipodally modulate $K$ signature waveforms which are transmitted
over an Additive White Gaussian Noise (AWGN) channel. Both the bit
streams and the signature waveforms are represented by Non-Return-to-
Zero (NRZ) signals. The received signal for one symbol period in such a
system can be expressed in complex envelope notation as:

$$r(t) = \sum_{k=1}^{K} A_k b_k c_k(t) + \sigma n(t). \quad (2.1)$$

Where the following notation is used:

- $c_k(t)$ is the real valued deterministic signature waveform assigned to
  the $k$th user, normalized to have unit energy
  $$||c_k||^2 = \int_{0}^{T} c_k^2(t) dt \triangleq 1. \quad (2.2)$$
  The signature waveforms are assumed to be zero outside the interval
  $[0,T]$, and therefore, in an AWGN channel, there is no inter symbol
  interference.

- $A_k$ is the received amplitude of the $k$th user’s signal. $A_k^2$ is referred to
  as the energy of the $k$th user.

- $b_k \in \{-1, +1\}$ is a bit transmitted by the $k$th user.

- $n(t)$ is white Gaussian noise with zero mean and unit variance. It mod-
  els thermal noise plus other noise sources unrelated to the transmitted
  signals. According to (2.1) the noise power in a frequency band with
  bandwidth $B$ is $2\sigma^2 B$. In the literature, the one-sided noise spectral
  density $2\sigma^2$ is frequently denoted by $N_0$.

2.3 Discrete-time CDMA Model and Conventional
Detection

CDMA detectors commonly have a front-end whose objective is to obtain a
discrete-time representation of the received continuous-time waveform $r(t)$. 

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2.3. Discrete-time CDMA Model and Conventional Detection

One way of converting the received waveform into a discrete-time representation is to pass it through a matched filter bank, see Figure 2.3, each matched to the signature waveform of a different user. The outputs of the matched filters are then sampled at the end of each bit period.

The output of the matched filter for a user \( k \) for synchronous CDMA can be expressed as:

\[
Z_{mfk} = \int_0^T r(t)c_k(t)dt = A_k b_k \int_0^T c_k(t)c_k(t)dt + \sum_{j \neq k} A_j b_j \int_0^T c_j(t)c_k(t)dt + \sigma \int_0^T n(t)c_k(t)dt
\]  

(2.3)

By using the fact that \( c_k(t) \) is normalized to have unit energy the matched filter outputs can be expressed as

\[
Z_{mfk} = A_k b_k + \sum_{j \neq k} A_j b_j \rho_{jk} + n_k,
\]  

(2.5)

where \( \rho_{jk} = \langle c_j, c_k \rangle = \int_0^T c_j(t)c_k(t)dt \)

(2.6)

and

\[
n_k = \sigma \int_0^T n(t)c_k(t)dt
\]  

(2.7)

is a Gaussian random variable with zero mean and variance equal to \( \sigma^2 \).

So the matched filter output consists of a desired signal component, a MAI component and a noise component.

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The decisions of the conventional CDMA detector for the bits transmitted by user $k$ are based on the sign of the output of the matched filter bank for user $k$ at the end of each bit period. This detector will perform well when the cross correlations $\rho_{jk}$ between the signature sequence of the desired user and the signature sequences of the other users in the system are a lot smaller than the autocorrelation of the signature sequence of the desired user. The amplitudes $A$ of the signals of all the users in the system have to be about equal as well for good performance. The performance of the conventional CDMA detector will deteriorate however when the number of active users in the system increases, because then the multiple access interference component can still become as large as the desired signal component.
Chapter 3

Blind Adaptive MMSE Detection

In this chapter the blind adaptive MMSE detector will be described and mathematically analyzed. This analysis is a summary of the analysis of blind adaptive MMSE detection in literature ([4], [5] and [6]). First a general notation for linear multiuser detectors, the class of multiuser detectors to which the blind adaptive MMSE detector belongs, will be developed. After that it will be shown that the blind adaptive MMSE detector, which minimizes the mean output energy, also minimizes the mean square error. Finally, an adaptive algorithm will be derived for the implementation of the blind adaptive MMSE detector. For convenience in this chapter it will be assumed that the desired user is user 1, but the same reasoning can of course be applied to all users in the system.

3.1 Linear Multiuser Detectors

The blind adaptive MMSE detector is an example of a linear multiuser detector. Linear multiuser detectors apply a linear transformation to the outputs of the matched filter bank to produce a new set of outputs, which hopefully provide better performance when used for estimation. Other examples of linear multiuser detectors are the conventional, decorrelating and MMSE detector [4].

Since matched filtering is also a linear operation, the matched filter bank followed by a linear transformation used in linear multiuser detection can be seen as a matched filter bank with modified sequences. So the signature sequence $c$ is replaced by a modified signature sequence $m$. A linear multiuser detector for user 1 can be characterized by the modified sequence $m_1$, which is the sum of two orthogonal components. One of these components is the signature sequence of user 1, $c_1$. The other component is denoted as
Chapter 3. Blind Adaptive MMSE Detection

$x_1$ and will be referred to as the $x$ sequence, so

$$m_1 = c_1 + x_1,$$  \hspace{1cm} (3.1)

with $m_1, c_1, x_1 \in \mathbb{R}^N$, where $N$ is the number of chips per symbol and

$$\langle c_1, x_1 \rangle = 0.$$  \hspace{1cm} (3.2)

Since $x_1$ is orthogonal to $c_1$, any $x_1$ can be chosen to minimize the correlation between the multiple access interference and $m_1$, while the correlation with user 1 remains constant. Thus

$$\langle c_1, m_1 \rangle = ||c_1||^2 \triangleq 1.$$  \hspace{1cm} (3.3)

The linear detector makes its decision for user 1 based on the sign of the output of the matched filter with modified sequence for user 1, so

$$\hat{b}_1 = \text{sgn}(\langle r, m_1 \rangle).$$  \hspace{1cm} (3.4)

Every linear multiuser detector can be written in this form, so it is a canonical representation for linear multiuser detectors [4].

The output of the matched filter with modified sequence for user 1 is equal to

$$Z_1 = \langle r, m_1 \rangle = A_1 b_1 \langle c_1, c_1 + x_1 \rangle + \sum_{k=2}^{K} A_k b_k \langle c_k, c_1 + x_1 \rangle + \sigma \langle n, c_1 + x_1 \rangle.$$  \hspace{1cm} (3.5)

Using (3.3) and the fact that $\langle c_1, c_k \rangle = \rho_{1k}$ gives

$$Z_1 = A_1 b_1 + \sum_{k=2}^{K} A_k b_k (\rho_{1k} + \langle c_k, x_1 \rangle) + \sigma \langle n, c_1 + x_1 \rangle.$$  \hspace{1cm} (3.6)

### 3.2 Minimizing Mean Output Energy

The blind adaptive MMSE detector in fact minimizes the Mean Output Energy (MOE), in contrary to what its name implies. In this section it will be shown that by minimizing the Mean Output Energy, the Mean Square Error (MSE) is also minimized.

The Mean Output Energy of a linear multiuser detector for user 1 is defined as:

$$\text{MOE} \triangleq E[\langle r, m_1 \rangle^2].$$  \hspace{1cm} (3.7)
3.2. Minimizing Mean Output Energy

The trivial solution to minimizing this equation is setting $m_1 = 0$. However, since $m_1$ is defined as the sum of $c_1$ and $x_1$, this solution is eliminated. It can be expected intuitively that minimizing the output energy of the linear detector is a sensible approach. This is because the energy at the output of the detector can be written as the sum of the energy due to the desired signal plus the energy due to the interference (background noise and multiple access interference). Any $x_1$, as long as $x_1$ is orthogonal to $c_1$, can be chosen to minimize the interference, but it will not influence the energy of the desired signal. The $x_1$ that minimizes the Mean Output Energy also minimizes the Mean Square Error as the following reasoning shows.

The Mean Output Energy and the Mean Square Error of the linear detector for user 1 can be written as, respectively,

\begin{equation}
\text{MOE}(x_1) = E[(\langle r, c_1 + x_1 \rangle)^2] \tag{3.8}
\end{equation}

and

\begin{align}
\text{MSE}(x_1) &= E[(A_1 b_1 - \langle r, c_1 + x_1 \rangle)^2] \\
&= E[(A_1 b_1)^2 - 2A_1 b_1 \langle r, c_1 + x_1 \rangle + (\langle r, c_1 + x_1 \rangle)^2]. \tag{3.9}
\end{align}

Then the fact is used that the received signal $r$ can be decomposed in a desired signal portion $A_1 b_1 c_1$ and a residue term $R$. The residue consists of multiple access interference and white Gaussian noise.

\begin{align}
\text{MSE}(x_1) &= E[(A_1 b_1)^2] - E[2A_1 b_1 \langle A_1 b_1 c_1 + R, c_1 + x_1 \rangle] + \\
&= E[\langle r, c_1 + x_1 \rangle)^2]. \tag{3.10}
\end{align}

The last term of equation 3.10 is equal to the mean output energy $\text{MOE}(x_1)$. Since $b_1 \in \{-1, +1\}$, $(A_1 b_1)^2 = A_1^2$ and $E[(A_1 b_1)^2] = A_1^2$, equation 3.10 can be written as

\begin{align}
\text{MSE}(x_1) &= A_1^2 - E[2A_1 b_1 \langle c_1, c_1 + x_1 \rangle] - \\
&= E[2A_1 b_1 \langle R, c_1 + x_1 \rangle] + \text{MOE}(x_1). \tag{3.11}
\end{align}

Further it is assumed that bit $b_1$ is independent of the other bits $b_k, k \neq 1$ and that $b_1$ is independent of the white Gaussian noise. From these two assumptions, it follows that $b_1$ is uncorrelated with the multiple access interference. Finally, by also using the obvious fact that $b_1$ is independent of $m_1$ the term $E[2A_1 b_1 \langle R, c_1 + x_1 \rangle]$ can be written as $E[2A_1 b_1]E[(R, c_1 + x_1)]$, so

\begin{align}
\text{MSE}(x_1) &= A_1^2 - E[2(A_1 b_1)^2 \langle c_1, c_1 + x_1 \rangle] - \\
&= E[2A_1 b_1]E[(R, c_1 + x_1)] + \text{MOE}(x_1) \tag{3.12}
\end{align}
Chapter 3. Blind Adaptive MMSE Detection

Since $E[2A_1b_1] = 0$, because $b_1$ is equally likely to be 1 or -1, the term $E[2A_1b_1E[(R, c_1 + c_1)]]$ can be eliminated, resulting in

$$\text{MSE}(x_1) = A_1^2 - 2A_1^2\langle c_1, c_1 + x_1 \rangle + \text{MOE}(x_1)$$

$$= A_1^2 - 2A_1^2\|c_1\|^2 + \text{MOE}(x_1)$$

$$= \text{MOE}(x_1) - A_1^2. \quad (3.13)$$

So the mean square error and the mean output energy differ by only a constant and the arguments that minimize both functions are the same. To implement the MSE function in an algorithm, knowledge of the data bits for user 1 is needed. For an implementation of the MOE function this knowledge is not needed, which means that an algorithm based on the MOE function does not require training sequences. It can be shown that the mean output function MOE($x_1$) is strictly convex over the set of signals orthogonal to $c_1$. Therefore, the output energy has no local minima other than the unique global minimum. With this property, the stochastic gradient descent method can be used to adaptively implement the blind adaptive MMSE detector [5].

### 3.3 Stochastic Gradient Descent Method

The stochastic gradient descent method is based on the gradient decent method. The gradient descent method is used to find the parameter $\theta_{\text{min}}$ that minimizes the following function:

$$\Xi(\theta) = E[g(X, \theta)]. \quad (3.14)$$

Where $X$ is a random variable and $g(\cdot)$ is a function. If the function $\Xi$ is convex, then for any initial condition $\theta_0$, the gradient descent algorithm converges to the minimum of $\Xi$. The algorithm follows the direction of steepest descent (i.e., the direction opposite to the gradient $\nabla \Xi$):

$$\theta_i = \theta_{i-1} - \mu \nabla \Xi(\theta_{i-1}), \quad (3.15)$$

where the subscript $i$ is used to indicate the iteration number of the algorithm. If the step size $\mu$ is arbitrarily small, then eventually $\theta_i$ will be as close to $\theta_{\text{min}}$ as desired. In practice, the step size can be progressively decreased as the algorithm converges. According to (3.14) the probability distribution of $X$ has to be known in order to compute the gradient. Though this knowledge may be available, the use of these distributions can be avoided by using the stochastic gradient descent method.

In the stochastic version of the algorithm the unknown term $\nabla \Xi(\theta_{i-1}) = \nabla E[g(X, \theta_{i-1})]$ is replaced by the unaveraged $\nabla g(X_i, \theta_{i-1})$. With $X_i$ the realization of the random variable $X$ for iteration $i$. This results in the following stochastic gradient descent algorithm:

$$\theta_i = \theta_{i-1} - \mu \nabla g(X_i, \theta_{i-1}). \quad (3.16)$$

where the subscript $i$ again is used to indicate the iteration number.
3.4 Adaptive Implementation

The stochastic gradient decent method can be used to find the $x$ sequence $x_{\text{opt}}$ that minimizes the Mean Output Energy. The MOE function, given as $\text{MOE}(x_1) = E[(r, c_1 + x_1)^2]$, is than the equivalent of the $\Xi(\theta)$ function, the $x$ sequence is the equivalent of $\theta$ and the received signal $r$ is the equivalent of $X$. To minimize the Mean Output Energy the $x$ sequence is adapted each bit period using the stochastic gradient descent algorithm (3.16). This means that one iteration of the stochastic gradient descent algorithm is performed for each bit period. Since subscripts are already used to indicate users, the iteration number is indicated with an index $[i]$. So the stochastic gradient descent algorithm for adaptation of the $x$ sequence can be written as

$$x_1[i] = x_1[i - 1] - \mu \nabla((r[i], c_1 + x_1[i - 1]))^2. \quad (3.17)$$

Here $r[i]$ indicates the received signal for the bit period of the $i$th bit in the bit stream. $x[i-1]$ is the value of the $x$ sequence obtained from the previous received signal $r[i-1]$ during the previous iteration of the algorithm. Note that $c_1$ has to remain the same for all bit periods, which implies the use of a short code CDMA system.

Equation (3.17) requires the gradient of $(r[i], c_1 + x_1[i - 1])^2$ for the $i$th bit interval with respect to $x_1$. This gradient is

$$\nabla_{x_1}((r[i], c_1 + x_1[i - 1]))^2 = 2r[i]c_1 + x_1[i - 1])r[i]. \quad (3.18)$$

Equation (3.18) states that the gradient of the mean output energy is equal to a scaled version of the received signal $r[i]$. After all, $2r[i]c_1 + x_1[i - 1])$, is only a constant.

At this stage, the gradient descent algorithm for the $x$ sequence can be modified so that it satisfies the orthogonality condition $\langle c_1, x_1 \rangle = 0$. This is done by replacing the received signal $r[i]$, by the component of $r[i]$ which is orthogonal to $c_1$. The orthogonal component $r_{ort}[i]$ is written as:

$$r_{ort}[i] = r[i] - \langle r[i], c_1 \rangle c_1. \quad (3.19)$$

By using the stochastic gradient decent algorithm for the $x$ sequence (3.17), the expression for the gradient (3.18) and the component of the received signal orthogonal to $c_1$ (3.19) the algorithm for minimizing the mean output energy can be found:

$$x_1[i] = x_1[i - 1] - 2\mu (r[i], c_1 + x_1[i - 1])r_{ort}[i]$$

$$x_1[i - 1] - 2\mu (r[i], c_1 + x_1[i - 1]) \langle r[i] - (r[i], c_1)c_1 \rangle. \quad (3.20)$$

1$r[i]$ is a vector of $N$ samples of the received signal, sampled at the chip times, for the $i$th bit period. Where $N$ is the number of chips per symbol.
Chapter 3. Blind Adaptive MMSE Detection

The expression \( \langle r[i], c_1 \rangle \) in equation (3.20) is a normal matched filter operation for user 1, as used in the matched filter bank described in Section 2.3. The expression \( \langle r[i], c_1 + x_1[i - 1] \rangle \) in equation (3.20) is a matched filter for user 1 with modified sequence (3.5). Since the \( x_1 \) sequence of this filter is adapted by the stochastic gradient descent rule this filter is referred to as the adaptive filter. The output of the matched filter for user 1 for the \( i \)th bit period is written as:

\[
Z_{mf1}[i] = \langle r[i], c_1 \rangle. \tag{3.21}
\]

Analogously, the output of the adaptive filter for user 1 for the \( i \)th bit period is written as:

\[
Z_1[i] = \langle r[i], c_1 + x_1[i - 1] \rangle = Z_{mf}[i] + \langle r[i], x_1[i - 1] \rangle. \tag{3.22}
\]

Substituting (3.21) and (3.22) in (3.20), the adaptation rule for the \( x \) sequence can be written as:

\[
x_1[i] = x_1[i - 1] - 2\mu Z_1[i](r[i] - Z_{mf1}[i]c_1). \tag{3.23}
\]

The output of the adaptive filter \( Z_1[i] \) is used as the decision statistic of the blind adaptive MMSE detector for user 1:

\[
\hat{b}_1[i] = sgn(Z_1[i]) = sgn(Z_{mf}[i] + \langle r[i], x_1[i - 1] \rangle). \tag{3.24}
\]

The output energy of the adaptive filter will be minimal when the \( x_1 \) sequence has converged to \( x_{1,\text{opt}} \), the \( x \) sequence that minimizes the mean output energy for user 1. In Section 3.2 it was shown that the mean square error is than also minimized. So when the \( x_1 \) sequence has converged to \( x_{1,\text{opt}} \) the decision statistic for user 1 of the blind adaptive MMSE detector is equivalent to the decision statistic for user 1 of the MMSE detector. Figure 3.1 gives a graphical representation of the implementation of the blind adaptive MMSE detector.

The natural choice for initialization of the \( x \) sequence is \( x_1[0] = 0 \). Whether the algorithm is stable depends on the value for the step size \( \mu \). Equations to determine the maximum value of \( \mu \) for which the algorithm is still stable are given in [5]. In previous research [3] however the values obtained with these equations turned out to result in an unstable algorithm. In practice the value for the step size \( \mu \) is usually determined by trial-and-error. A smaller step size will result in a longer adaptation time. On the other hand, a smaller step size will also result in an \( x_1[i] \) which is closer to \( x_{1,\text{opt}} \), where \( x_{1,\text{opt}} \) is the orthogonal sequence component that results in a global minimum of the mean output energy. So the best value for \( \mu \) is a trade-off between adaptation time and accuracy.
3.4. Adaptive Implementation

Figure 3.1: Blind adaptive MMSE detector.
Chapter 4

PIC Detection

In this chapter the Parallel Interference Cancellation (PIC) detector will be analyzed. The chapter starts with a short discussion of decision-driven detectors, the class of detectors to which the PIC detector belongs. After that the PIC detector is introduced. Then the different ways in which Parallel Interference Cancellation can be implemented are discussed. Finally further refinements of the implementations of the PIC detector are given, based on mathematical analysis of these PIC implementations in literature.

4.1 Decision-driven Detectors

The idea behind decision-driven detectors is simple: if a decision has been made about an interfering user’s bit, then the interfering signal caused by this user can be reconstructed and subtracted from the received signal. This will cancel the MAI caused by this user, provided that the decision was correct; otherwise it will double the contribution to the interference of this user.

The simplest decision driven detector is the Successive Interference Cancellation (SIC) detector, which consists of a bank of matched filters followed by a sorter, see Figure 4.1. In a lot of implementations the sorter sorts the signals of the different users in order of decreasing received power. However, this is not necessarily best since it fails to take into account the crosscorrelations among users. A sensible alternative is to order users according to the signal-to-noise ratios computable using

\[
E \left[ \left( \int_0^T r(t)c_k(t)dt \right)^2 \right] = \sigma^2 + A_k^2 + \sum_{j \neq k} A_j^2 \rho_{jk}^2,
\]

(4.1)

which can be estimated easily from the matched filter outputs. Interference cancellation for a user \( k \) is achieved by subtracting from the matched filter outputs of user \( k \) an estimate of the MAI generated by users with higher
received powers than that of user $k$. For the synchronous case this results in the following decision rule:

$$\hat{b}_k = \text{sgn} \left( Z_{m_k} - \sum_{j=1}^{k-1} A_j \rho_{jk} \hat{b}_j \right),$$  

(4.2)

where it is assumed that the users are numbered in order of decreasing received power, so the user with the highest received power has the lowest number.

The SIC detector has a number of disadvantages. First of all, the SIC detector may actually double the interference, when it makes a wrong decision. Once such an error is made it is likely that the detector will accumulate more errors, because the other decisions are dependent on this decision. Since the demodulation of the first user will never be done in absence of multiple access interference, wrong decisions are more likely in the case of a large number of users. So performance may decrease rapidly when the number of users increases and the performance of the detector for a particular user is greatly affected by the order in which the users are cancelled. For the same reason, the SIC detector operates best in an unbalanced power environment. This is not a very likely environment in actual cellular wireless CDMA communications systems, because those systems usually employ some form of power control. Another problem of the SIC detector is that the signals of the different users are demodulated sequentially, which causes the delay in demodulating the signals to grow with the number of users.
4.2 PIC Detector

The PIC detector is a detector that also belongs to the class of decision-driven detectors. It alleviates the problem of the SIC detector that the performance of successive cancellation for a particular user is greatly affected by the order in which the users are cancelled.

The PIC detector is based on a technique that employs multiple iterations in detecting the data bits and cancelling the interference, therefore it is also referred to as Multistage Interference Cancellation (MIC) detector. In its simplest form the PIC detector consists of two stages. The first stage consists of a conventional bank of matched filters. The second stage performs, for each user, reconstruction and subtraction of an estimate of the interference from all other users, see Figure 4.2. So the second stage in fact performs successive cancellation for all users. This leads to the following decision rule:

$$\hat{b}_k = \text{sgn} \left( y_k - \sum_{j \neq k} A_j p_{jk} \hat{b}_j \right).$$  \hspace{1cm} (4.3)

The motivation for such a detector is given by comparing with the SIC detector. In the SIC detector, the demodulation of user $k$ is carried out in the presence of the multiple access interference of all users that have a weaker received power. After all, the SIC detector demodulates the received signals in order of decreasing received power. By using the PIC detector,
the interference caused by all interfering users can be estimated through the bank of matched filters and is then cancelled out. Another advantage of the PIC detector over the SIC detector is that the PIC detector demodulates the signals of the individual users in parallel, therefore the delay in demodulating signals in the PIC detector does not increase with the number of users.

The PIC detector requires more knowledge of the received signal than the conventional CDMA detector. Since it has to reconstruct the transmitted signal of all the users, it requires information about: the timing of the desired user, the timing of the interfering users, the received amplitudes and the signature waveforms of the desired and interfering users. The timing and signature waveforms of all the users also have to be known in a base station that uses conventional CDMA detection, since it has to detect all the users. So PIC slightly increases the processing requirements for a base station, because it will have to implement some form of amplitude estimation.

4.3 PIC Implementation

There are two implementations for PIC detection[7]. The implementation that was used to describe Parallel Interference Cancellation detection in the previous section is the so-called narrowband implementation. In this implementation interference is cancelled from the narrowband outputs (matched filter outputs or outputs of previous stages) by using the estimates of the data symbols and channel gains as well as the known cross-correlations between users. This can be clearly seen from Figure 4.2 and equation (4.3).

The decision rule of the two-stage narrowband PIC detector for synchronous CDMA given in the previous section can be easily extended to a decision rule for a multi-stage detector, called a $S$-stage narrowband PIC detector. For stage $s + 1$ the decision rule of the $S$-stage narrowband PIC detector can be expressed as

$$\hat{b}_{k}^{(s+1)} = \text{sgn}\left(Z_{k}^{(s+1)}\right)$$  \hspace{1cm} (4.4)

with

$$Z_{k}^{(s+1)} = Z_{mf_{k}} - \sum_{j \neq k} A_{j} \rho_{jk} \hat{b}_{j}^{(s)}$$  \hspace{1cm} (4.5)

as the decision statistic for stages $s > 1$ and

$$Z_{k}^{(1)} = Z_{mf_{k}}$$  \hspace{1cm} (4.6)

as the decision statistic for stage $s = 1$. A schematic representation of an $S$-stage narrowband PIC detector is given in figure 4.3.

The second implementation for PIC detection requires the estimation, regeneration, and cancellation of the signal of each interferer from the received signal of each of the desired users, see Figure 4.4. Since the wideband signals of all the users have to be regenerated this implementation is referred
4.3. PIC Implementation

Figure 4.3: S stage narrowband PIC detector.
Figure 4.4: S stage wideband PIC detector.
4.4. Amplitude Estimation

to as the wideband implementation. The decision rule for stage \( s + 1 \) of the \( S \)-stage wideband PIC detector can be expressed as

\[
\hat{b}_k^{(s+1)} = \text{sgn} \left( Z_k^{(s+1)} \right)
\]  

(4.7)

with

\[
Z_k^{(s+1)} = \int_0^T \hat{r}_k^{(s)}(t)c_k(t)dt
\]  

(4.8)

as the decision statistic, where the received signal \( \hat{r}_k^{(s)} \) for user \( k \) at stage \( s \) is estimated according to

\[
\hat{r}_k^{(s)}(t) = r(t) - \sum_{j=1}^K \hat{u}_j^{(s)}(t)
\]  

(4.9)

and the signal \( \hat{u}_j^{(s)}(t) \) corresponds to the reconstructed signal of user \( j \) at stage \( s \). This signal is reconstructed according to

\[
\hat{u}_j^{(s)}(t) = A_j \hat{b}_j^{(s)}c_j(t).
\]  

(4.10)

At stage \( s = 1 \) there has not been any interference cancellation, therefore

\[
\hat{r}_k^{(1)}(t) = r(t).
\]  

(4.11)

Both implementations of the PIC detector have the same theoretical bit-error-rate performance [7]. For short-code CDMA systems the narrowband implementation requires less computations than the wideband implementation, because it avoids regeneration of the wideband signal. For long-code systems this advantage disappears because in that case the crosscorrelations have to be recalculated for each symbol anyway. In both cases the narrowband implementation requires more memory because the crosscorrelations \( \rho_{jk} \) between the signature sequences of all the users have to be stored.

4.4 Amplitude Estimation

The equations for the decision rules of the narrowband implementation (4.5) and the wideband implementation (4.10) show that both PIC detector implementations require knowledge of the received amplitudes of the signals of all the users. Since this information is not directly available at the receiver, the received amplitudes have to be estimated. A common way to do this is to use the matched filter outputs or outputs of a previous stage, which are both referred to as soft decisions, as a joint estimation of the detected bits and the received signal amplitudes. This avoids the use of separate channel estimation algorithms that increase the overall complexity of the receiver.
In this case the decision statistic of the narrowband PIC detector for stages \( s > 1 \) (4.5) can be rewritten as

\[
Z_k^{(s+1)} = Z_{mk} - \sum_{j \neq k} Z_j^{(s)} \rho_{jk}
\]  

(4.12)

and the equation for the reconstructed signal of the wideband PIC detector for stages \( s > 1 \) (4.10) can be rewritten as

\[
\hat{u}_j^{(s)}(t) = Z_j^{(s)} c_j(t).
\]  

(4.13)

Unfortunately the amplitude estimates derived from the outputs of the matched filter bank are, especially in the first stages, not very accurate, because they still contain a lot of interference. Therefore enhanced performance can be achieved at the cost of increased receiver complexity by using separate channel estimation algorithms.

### 4.5 Partial Cancellation PIC

In Correal et al. [8] it is shown that straightforward implementation of wideband PIC based on complete subtraction of the interference estimates results in a biased decision statistic. The bias has its strongest effect in the first stage of interference cancellation, in the subsequent stages its effect diminishes. However if the bias leads to incorrect cancellation at the first stage the effects of these errors may be observed at the next stages. In Potman [3] it is shown that the decision statistic of a straightforward implementation of narrowband PIC, based on complete subtraction of the interference, is also biased.

In Correal et al. [8] a simple method to mitigate the effect of the bias and improve the performance of parallel multistage interference cancellation is proposed. This method is based on multiplying the amplitude estimates with a partial-cancellation factor \( 0 \leq C_K^{(s)} \leq 1 \) that varies with the stage of cancellation \( s \) and system load \( K \). This multiplication has to be performed before the amplitude estimates are used to subtract the interference in case of the narrowband implementation, or before the amplitude estimates are used to reconstruct the signal in case of the wideband implementation. This can be interpreted as modifying equation (4.5) and (4.9) to include a partial-cancellation factor \( C_K^{(s)} \) resulting in respectively

\[
Z_k^{(s+1)} = y_k - \sum_{j \neq k} C_K^{(s)} A_j \rho_{jk} \hat{b}_j^{(s)}
\]  

(4.14)

for the decision statistic in case of narrowband PIC and

\[
\hat{r}_j^{(s)}(t) = r(t) - C_K^{(s)} \sum_{j=1}^K \hat{u}_j^{(s)}(t)
\]  

(4.15)
4.5. Partial Cancellation PIC

for the estimated received signal for user $k$ in case of wideband PIC.
Chapter 5

WCDMA Uplink Model

In this chapter a model of the WCDMA uplink is developed. The chapter starts with a general description of the WCDMA uplink in Section 5.1. After that, in Section 5.2, a general mathematical model of the WCDMA uplink for asynchronous multipath fading channels is given.

5.1 WCDMA Uplink

The WCDMA uplink is quite different from the basic CDMA system described in Chapter 2. Figure 5.1 shows the spreading and modulation in the WCDMA uplink. Instead of just one information stream a user can transmit up to six simultaneous information streams in frames over six so called Dedicated Physical Data Channels (DPDCHs). In addition to the six DPDCHs each user also has one Dedicated Physical Control Channel (DPCCH) available. The WCDMA uplink frame has a duration of 10 ms and is split into 15 slots. Each slot of the DPCCH contains pilot bits, Transport Format Combination Indicator (TFCI) bits, Transmission Power Control (TPC) bits and Feedback Information (FBI) bits [9].

The information stream(s) that a user wants to transmit over the DPDCHs are spread to the WCDMA chip-rate of 3.84 Mcps with an Orthogonal Variable Spreading Factor (OVSF) channelization code ($c_k,1 \ldots c_k,6$ in Figure 5.1). The use of a variable spreading factor makes it possible to support different data-rates on the DPDCHs. For the WCDMA uplink DPDCHs spreading factors ranging from 4 to 256 are supported. The DPCCH is always spread with an OVSF channelization code with spreading factor 256 ($c_k,c$ in Figure 5.1).

The spread DPCCH and DPDCHs are than I-Q/code multiplexed and the chips of the resulting signal are scrambled with a long or short complex scrambling code ($c_k[u]$ in Figure 5.1). So WCDMA can be considered a long or a short code CDMA system depending on the type of scrambling codes that are used. The long scrambling codes are Gold codes with 25...
degree generator polynomials truncated to the 10ms WCDMA uplink frame length, resulting in 38400 chips with 3.84 Mcps. The short scrambling codes have been chosen from the extended S(2) code family and have a length of 256 chips. The complex scrambling sequence is formed, in the case of short codes by combining two real valued codes and in the case of long codes by combining a single sequence with a delayed version of itself.

The complex-valued chip sequence that is the result of the scrambling process is split into its real and imaginary parts. Both parts are filtered with a root-raised cosine pulse-shaping filter with roll-off factor 0.22 and are than fed to the I and Q branches of the carrier modulator. A more detailed description of the UMTS physical layer can be found in [9] and the UMTS physical layer specification, of which [10] is the overview document.

5.2 General WCDMA Uplink Model

Based on the above description and using a slightly modified version of the notation used in the UMTS physical layer specification [10] the complex symbol $g_k[i]$ corresponding with the $i$th bits of the spread and I-Q/code multiplexed DPDCHs and DPCCH of user $k$ can be mathematically expressed as:

$$g_k[i] = \sum_{d=1,3...}^{D} \beta_d b_{k,d}[i] c_{k,d} + j \left( \beta_c b_{k,c}[i] c_{k,c} + \sum_{d=2,4...}^{D} \beta_d b_{k,d}[i] c_{k,d} \right)$$

(5.1)

where
5.2. General WCDMA Uplink Model

- $D$ is the number of DPDCHs used by user $K$,
- $\beta_d$ is the gain factor for the DPDCHs,
- $b_{k,d}^i$ is the $i$th bit on the $d$th Dedicated Physical Data Channel (DPDCH) of user $k$,
- $c_{k,d}^c$ is the real-valued channelization code for the $d$th DPDCH of user $k$,
- $\beta_c$ is the gain factor for the DPCCH,
- $b_{k,c}^i$ is the $i$th bit on the DPCCH of user $k$,
- and $c_{k,c}^c$ is the channelization code for the DPCCH of user $k$.

The transmitted signal of user $k$ is than equal to:

$$s_k(t) = A_k p(t) c_k^c[i] g_k[i]$$ (5.2)

where

- $A_k$ is the amplitude for user $k$,
- $p(t)$ is the transmitter pulse shape,
- and $c_k^c[i]$ is the complex-valued long or short scrambling code for user $k$ used to scramble complex symbol $g_k[i]$.

To generate the pulse shape $p(t)$ a root-raised cosine pulse shaping filter with impulse response

$$RC_0(t) = \frac{\sin \left( \frac{\pi}{T_c} (1 - \alpha) \right) + 4\alpha \frac{\pi}{T_c} \cos \left( \frac{\pi}{T_c} (1 + \alpha) \right)}{\pi \frac{T_c}{T_c} \left( 1 - \left( \frac{4\alpha}{T_c} \right)^2 \right)}, \quad (5.3)$$

is used, where the roll-off factor $\alpha$ is equal to 0.22 and the chip duration $T_C$ is equal to $\frac{1}{\text{chip-rate}}$.

It is assumed that there are $K$ users active in the cell that the model covers and that each of these users experiences a Rayleigh fading multipath channel. This results in the following received signal $r(t)$ at the base station:

$$r(t) = \sum_{k=1}^{K} \sum_{i=1}^{M} \sum_{l=1}^{L_k} h_{k,l}(t) s_k(t - iT - \tau_{k,l}) + \sigma_n(t)$$ (5.4)

where

- $K$ is the number of active users in the cell that the base station covers,
Chapter 5. WCDMA Uplink Model

- \( M \) is the number of complex symbols in a frame,
- \( L_k \) is the number of multipaths for user \( k \),
- \( h_{k,l}(t) \) is the time dependent complex fading coefficient for multipath \( l \) of user \( k \),
- \( \tau_{k,l} \) is the delay of multipath \( l \) of user \( k \),
- \( \sigma \) is the standard deviation of the noise,
- and \( n(t) \) is white Gaussian noise with zero mean and unit variance.

The blind adaptive MMSE and PIC multiuser detectors that have been derived for a general CDMA system in the previous two chapters will be derived again for WCDMA in the next two chapters. The consequences for these detection techniques of the more realistic channel that is assumed in the WCDMA model will be indicated in the next two chapters as well.
Chapter 6

Blind Adaptive MMSE Detection for WCDMA

In this chapter blind adaptive MMSE detection for WCDMA is described. First a simplified version of the general WCDMA uplink model of the previous chapter is derived in Section 6.1. This simplified model is used to describe blind adaptive MMSE detection for synchronous AWGN channel WCDMA in Section 6.2. Finally section 6.3 describes generalization to asynchronous WCDMA systems and other channel models.

6.1 Synchronous AWGN Short Code WCDMA Uplink Model

To derive an implementation of blind adaptive MMSE detection for WCDMA it is useful to reduce the WCDMA model of the previous chapter to a WCDMA model for a synchronous WCDMA system transmitting over an AWGN channel, since a similar model was used in the derivation in Chapter 3. So it is assumed that there is no multipath and no fading and that all users are synchronous. In that case the received signal (5.4) reduces to:

\[ r(t) = \sum_{k=1}^{K} s_k(t) + \sigma n(t) \]  

(6.1)

When short scrambling codes are used the transmitted signal for user \( k \) (5.2) can be written as:

\[ s_k(t) = A_k p(t) c^*_k g_k[i] \]  

(6.2)

since for short scrambling codes \( c^*_k[1] = c^*_k[2] = \ldots = c^*_k[M] = c^*_k \). To further simplify the model it is also assumed that all users only use one DPDCH and that \( \beta_d = \beta_c = 1 \). The expression for the complex symbol (5.1) then reduces to:

\[ g_k[i] = b^d_{k,1}[i] c^*_k + j b^e_{k,1}[i] c^e_k \]  

(6.3)
So the received signal can be written as:

\[ r[i] = \sum_{k=1}^{K} A_k p(t) c_k^c \left( b_k[i] c_{k,1}^c + j b_k[i] c_{k,c}^c \right) + \sigma n[i] \]

\[ = \sum_{k=1}^{K} A_k p(t) c_k^c c_{k,1}^c b_k[i] + j A_k p(t) c_k^c c_{k,c}^c b_k[i] + \sigma n[i] \] (6.4)

The product of the channelization code and the spreading code will be referred to as the \textit{combined code}, denoted by \( cc_k, d \), where the subscript \( k \) is used to indicate the user and the subscript \( d \) is used to indicate the DPDCH to which the channelization code corresponds. So \( cc_{1,1} \) refers to the combined code of DPDCH 1 of user 1 and \( cc_{1,c} \) refers to the combined code of DPCCH of user 1. For convenience it will be assumed that the desired user is user 1. The received signal can then be written as:

\[ r[i] = A_1 p(t) cc_{1,1} b_1[i] + j A_1 p(t) cc_{1,c} b_1[i] + R, \] (6.5)

where the residue term \( R \) consists of multiple access interference and white Gaussian noise:

\[ R = \sum_{k=2}^{K} A_k p(t) cc_{k,1} b_k[i] + j A_k p(t) cc_{k,c} b_k[i] + \sigma n[i] \] (6.6)

### 6.2 Blind Adaptive MMSE for Synchronous AWGN WCDMA

Recall that the blind adaptive MMSE detector is fully described by a matched filter (3.21), an adaptive filter (3.22) and an update-rule for the adaptive filter coefficients (3.23). Substituting the received signal of the synchronous WCDMA model (6.5) into the equation for the matched filter output (3.21) and replacing the spreading sequence in that equation by the combined code for DPDCH 1 of user 1 results in the following expression for the output of the matched filter for DPDCH 1 of user 1:

\[ Z_{m,f1,1}[i] = \langle A_1 p(t) cc_{1,1} b_1^d[i] + j A_1 p(t) cc_{1,c} b_1^d[i], R, cc_{1,1} \rangle \\
= \langle A_1 p(t) cc_{1,1} b_1^d[i], cc_{1,1} \rangle + \langle j A_1 p(t) cc_{1,c} b_1^d[i], cc_{1,1} \rangle + \langle R, cc_{1,1} \rangle + \langle R, cc_{1,1} \rangle \]

\[ = A_1 p(t) b_1^d[i] \langle cc_{1,1}, cc_{1,1} \rangle + j A_1 p(t) b_1^d[i] \langle cc_{1,c}, cc_{1,1} \rangle + \langle R, cc_{1,1} \rangle \] (6.7)

Where the inner product with itself of the combined code for DPDCH 1 of user 1 is:

\[ \langle cc_{1,1}, cc_{1,1} \rangle = \sum_{n=1}^{N} c_1^c[n] c_1^c[n] c_1^c[n] c_1^c[n] \] (6.8)
6.2. Blind Adaptive MMSE for Synchronous AWGN WCDMA

where the overbar is used to indicate the complex conjugate. This can be expanded to:

\[
\langle cc_{1,1}, cc_{1,1} \rangle = \sum_{n=1}^{N} (\Re(c_1^*[n]) + j\Im(c_1^*[i]))(\Re(c_1^*[n]) - j\Im(c_1^*[n]))c_1^*[n]^2
\]

\[
= \sum_{n=1}^{N} (\Re(c_1^*[n])^2 + \Im(c_1^*[i])^2)c_1^*[n]^2
\]

\[
= 2N \quad (6.9)
\]

with \(N\) the length in chips of the combined code. The last step follows from the fact that \(\Re(c_1^*[n]) \in \{-1, 1\}\), \(\Im(c_1^*[i]) \in \{-1, 1\}\) and \(c_1^*[n] \in \{-1, 1\}\). In Section 2.2 the signature sequences were assumed to be normalized to have unit energy. Equation 6.9 shows that this is not the case in WCDMA. When the output of the matched filter is used in conventional CDMA detection this is not of influence, because the sign of the matched filter output is not changed. However, when the output of the matched filter is used to estimate the transmitted signal the scaling caused by the signature sequences has to be taken into account, as will be shown later in this section.

Substituting the received signal of the synchronous WCDMA model (6.5) into the equation for the adaptive filter output (3.22) and replacing the spreading sequence in that equation by the combined code for DPDCH 1 of user 1 results in the following expression for the output of the adaptive filter for DPDCH 1 of user 1:

\[
Z_{1,1}[i] = \langle A_{1p}(t)cc_{1,1}b_1^*[i] + jA_{1p}(t)cc_{1,c}b_1^*[i] + R, cc_{1,1} + x_{1,1} \rangle
\]

\[
= \langle A_{1p}(t)cc_{1,1}b_1^*[i], cc_{1,1} + x_{1,1} \rangle + j\langle A_{1p}(t)cc_{1,c}b_1^*[i], cc_{1,1} + x_{1,1} \rangle + \langle R, cc_{1,1} + x_{1,1} \rangle
\]

\[
= A_{1p}(t)b_1^*[i] \langle cc_{1,1}, cc_{1,1} + x_{1,1} \rangle + jA_{1p}(t)b_1^*[i] \langle cc_{1,c}, cc_{1,1} + x_{1,1} \rangle + \langle R, cc_{1,1} + x_{1,1} \rangle
\]

\[
= A_{1p}(t)b_1^*[i] \langle \langle cc_{1,1}, cc_{1,1} \rangle + \langle cc_{1,1}, x_{1,1} \rangle \rangle + jA_{1p}(t)b_1^*[i] \langle \langle cc_{1,c}, cc_{1,1} \rangle + \langle cc_{1,c}, x_{1,1} \rangle \rangle + \langle R, cc_{1,1} \rangle + \langle R, x_{1,1} \rangle \quad (6.10)
\]

The inner product \(\langle cc_{1,1}, cc_{1,1} \rangle\) is again equal to \(2N\). The inner product \(\langle cc_{1,1}, x_{1,1} \rangle\) is equal to zero because \(x_{1,1}\) is orthogonal to \(cc_{1,1}\) by definition. The blind adaptive MMSE detector makes its decisions for the received bits based on the sign of the adaptive filter outputs, as indicated by equation (3.24). Since DPDCH 1 is mapped to the real part of the transmitted complex symbol, decisions for the received bits on DPDCH 1 are made based on the sign of the real part of the \(Z_{11}\) matched filter outputs.
The update rule for the filter coefficients of the adaptive filter for DPDCH 1 of user 1 is found by replacing $x_1$ by $x_{1,1}$, $Z_1$ by $Z_{1,1}$, $Z_{m1}[i]$ by $Z_{m1,1}[i]$ and $c_1$ by $cc_{1,1}$ in equation (3.23):

$$x_{1,1}[i] = x_{1,1}[i - 1] - 2\mu Z_{1,1}[i](r[i] - Z_{m1,1}[i]cc_{1,1}).$$

(6.11)

In the term $(r[i] - Z_{m1,1}[i]cc_{1,1})$ in the equation above a reconstructed version of the transmitted signal of user 1 is removed from the received signal, so the remaining signal consists only of interference and noise. Substituting the found expressions for the received signal (6.5) and the matched filter output (6.7) into this term results in the following equation:

$$r[i] - Z_{m1,1}[i]cc_{1,1} = A_1p(t)cc_{1,1}b_d^d[i] + jA_1p(t)cc_{1,1}b_1^d[i] + R - \langle A_1p(t)b_d^d[i]cc_{1,1,1,1}1,\rangle + \langle A_1p(t)b_1^d[i]cc_{1,1,1,1}1,\rangle + A_1p(t)cc_{1,1}b_d^d[i] + jA_1p(t)cc_{1,1}b_1^d[i] + R - A_1p(t)b_d^d[i]cc_{1,1}1,\rangle + jA_1p(t)b_1^d[i]cc_{1,1}1,\rangle + \langle R,cc_{1,1,1,1}1,\rangle$$

(6.12)

Since the inner product $\langle cc_{1,1},cc_{1,1}1,\rangle$ is equal to $2N$, the reconstructed version of the received signal of user 1 is $2N$ times larger than the actual component of user 1 in the received signal. Therefor the matched filter outputs have to be scaled with a factor $\frac{1}{2N}$ when they are used in the update rule for the adaptive filter coefficients. This results in the following final equation for the update rule for the filter coefficients of the adaptive filter for DPDCH 1 of user 1:

$$x_{1,1}[i] = x_{1,1}[i - 1] - 2\mu Z_{1,1}[i](r[i] - Z_{m1,1}[i]cc_{1,1})$$

(6.13)

The value of the $Z_{1,1}$ is also scaled by a factor $2N$ but this can be easily compensated for by adjusting the step-size $\mu$.

### 6.3 Asynchronous WCDMA Systems and Other Channel Models

Equations (6.7), (6.10) and (6.13) describe a blind adaptive MMSE detector for DPDCH 1 of user 1 in a synchronous WCDMA system. Detectors for the other DPDCHs and the DPCCH of user 1 can be described by replacing the channelization code used in those equations by the appropriate channelization code for the DPDCH that has to be detected. Detectors for the DPDCHs and DPCCH of other users can be described by replacing the
6.3. Asynchronous WCDMA Systems and Other Channel Models

scrambling code used in equations (6.7), (6.10) and (6.13) by the scrambling code of the desired user.

Extension to asynchronous systems is also straightforward, just as the conventional detector the blind adaptive MMSE detector can be synchronized with the received signal through the use of pilot symbols. Since the blind adaptive MMSE detector has the same input and output parameters as the conventional CDMA detector it can be used in a rake receiver structure to take advantage of the multipath diversity that is present in multipath channels. Although the performance of the blind adaptive MMSE detector will be negatively influenced by fading, just as the performance of the conventional detector, from a functional point of view no special measures are necessary to use the blind adaptive MMSE detector on fading channels.
Chapter 7

PIC Detection for WCDMA

In this chapter PIC detection for WCDMA is described. First a simplified version of the general WCDMA uplink model of Chapter 5 is derived in Section 7.1. This simplified model is used to describe PIC detection for synchronous AWGN channel WCDMA in Section 7.2. Finally section 7.3 describes generalization to asynchronous WCDMA systems and other channel models.

7.1 Synchronous AWGN WCDMA Uplink Model

To derive an implementation of PIC detection for WCDMA it is useful to reduce the WCDMA model of Chapter 5 to a WCDMA model for a synchronous WCDMA system transmitting over an AWGN channel, since a similar model was used in the derivation in Chapter 4. In contrary to the model derived in Section 6.1, the model derived in this section will not be limited to short code WCDMA systems since PIC detection can also be implemented for long code CDMA systems.

When it is assumed that there is no multipath and no fading and that all users are synchronous, the received signal (5.4) reduces to:

\[ r(t) = \sum_{k=1}^{K} s_k(t) + \sigma n(t) \]  

The transmitted signal of user \( k \) (5.2) in that case is equal to:

\[ s_k(t) = A_k p(t) c_k^d[i] g_k[i] \]  

To further simplify the model it is also assumed that all users only use one DPDCH and that \( \beta_d = \beta_c = 1 \). The expression for the complex symbol (5.1) than reduces to:

\[ g_k[i] = b_{k,1}^d[i] c_{k,1}^d + j b_{k}^c[i] c_{k,c}^c \]
So the received signal can also be written as:

\[
    r[i] = \sum_{k=1}^{K} A_k p(t) c_k^i[i] \left( b_k^d[i] c_{k,1}^i + j b_k^c[i] c_{k,c}^i \right) + \sigma n[i]
\]

(7.4)

When it is assumed that the desired user is user 1, the received signal can be written as:

\[
    r[i] = A_1 p(t) c_1^i[i] c_{1,1}^i b_1^d[i] + j A_1 p(t) c_1^i[i] c_{1,c}^i b_1^c[i] + R,
\]

(7.5)

where the residue term R consists of multiple access interference and white Gaussian noise:

\[
    R = \sum_{k=2}^{K} A_k p(t) c_k^i[i] c_{k,1}^i b_k^d[i] + j A_k p(t) c_k^i[i] c_{k,c}^i b_k^c[i] + \sigma n[i].
\]

(7.6)

The main difference with the model in Section 6.1 is that in the model in this section the scrambling sequences depend on the indices of the transmitted bits, indicated by the use of an index \([i]i\) in \(c_k^i[i]\). This models the fact that a different section of the scrambling code is used to scramble each data bit in long code WCDMA. In short code WCDMA the dependency of the scrambling sequences on the indices of the transmitted bits disappears and the model can be reduced to the model in Section 6.1.

### 7.2 PIC for Synchronous AWGN WCDMA

In this section a PIC detector for DPDCH 1 of user 1 that works in short-code as well as in long-code synchronous WCDMA system transmitting over an AWGN channel is described. As was stated in Section 4.3, for long-code CDMA, the narrowband and wideband implementation of the PIC detector have the same theoretical bit-error-rate performance and computational complexity. Since the wideband implementation requires less memory, the wideband implementation will be used in this section.

The wideband PIC detector using soft decision amplitude estimation and partial cancellation is described by a decision rule (4.7), a decision statistic (4.8), an estimate of the received signal (4.15) and a reconstructed transmitted signal (4.13). To arrive at the decision statistic for DPDCH 1 of user 1 the single spreading sequence in equation (4.8) has to be replaced with the combination of scrambling code and channelization code used in WCDMA:

\[
    Z_{1,1} = \int_0^T z_{1,1}^{(s)}(t) c_1^i(t) c_{1,1}^i(t)
\]

(7.7)
For an estimate of the received signal of DPDCH 1 of user 1 the reconstructed transmitted signal of the other users in the system is subtracted from the received signal, just as in equation (4.15). In addition the reconstructed signal of the DPCCH of user 1 has to be subtracted from the received signal as well. This results in the following estimate of the received signal of DPDCH 1 of user 1:

\[
\hat{r}_{1,1}(t) = r(t) - C_1^{(s)} \left( jZ_{1,c}^{(s)} c_1^c(t)c_1^c(t) + \sum_{j=2}^{K} \hat{u}_j^{(s)}(t) \right)
\]

(7.8)

To obtain an equation for the reconstructed transmitted signal of another user in the system the fact that the transmitted signal of a user consists of a DPDCH signal and a DPCCH signal has to be taken into account. Also the single spreading sequence in equation (4.13) has to be replaced with the combination of scrambling code and channelization code. This results in the following equation for the reconstructed signal:

\[
\hat{u}_j^{(s)}(t) = Z_{j,c}^{(s)} c_j^c(t)c_j^c(t) + jZ_{j,c}^{(s)} c_j^c(t)c_j^c(t).
\]

(7.9)

### 7.3 Asynchronous WCDMA Systems and Other Channel Models

Equations (7.7), (7.8) and (7.9) describe a PIC detector for DPDCH 1 of user 1 in a synchronous short- or long-code WCDMA system. Just as in the case of the blind adaptive MMSE detector these equations can be generalized to the other DPDCHs and the DPCCH of user 1 by replacing the channelization code used in those equations by the appropriate channelization code for the DPDCH that has to be detected. Detectors for the DPDCHs and DPCCH of other users can be described by replacing the scrambling code used in equations (7.7), (7.8) and (7.9) by the scrambling code of the desired user.

Extension to asynchronous systems is also straightforward, just as the conventional detector, the PIC detector can be synchronized with the received signal through the use of pilot symbols. To take advantage of the multipath diversity that is present in multipath channels, the single matched filters for each user in each stage of the PIC detector can be replaced by a Rake structure of matched filters. This will however drastically increase the processing power required by the detector. Although the performance of the PIC detector will be negatively influenced by fading, just as the performance of the conventional detector, from a functional point of view no special measures are necessary to use the PIC detector on fading channels.
Chapter 8

WCDMASim

In this chapter WCDMASim, the simulator that is used to study blind adaptive MMSE and PIC detection for WCDMA, will be described. The chapter starts with a general description of WCDMASim. Since the uplink is of most interest for studying multiuser detection, the WCDMASim uplink simulator will be described in more detail in Section 8.2. Section 8.3 describes the extensions required for simulating blind adaptive MMSE and PIC multiuser detection and other extensions that were made to WCDMASim.

8.1 WCDMASim

WCDMASim is an accurate simulator of the WCDMA physical layer developed by the Mobile and Portable Radio Research Group (MPRG) of the Virginia Polytechnic Institute and State University (Virginia Tech). WCDMASim strictly adheres to the WCDMA physical layer standards and simulates the WCDMA uplink and downlink over a multiple access specular multipath Rayleigh fading channel. It uses a conventional Rake receiver with equal gain combining. WCDMASim is a Matlab program with some of the computationally intensive functions implemented as Matlab mex files, written in C.

8.2 WCDMASim Uplink Simulator

The WCDMASim uplink simulator consists of four main blocks: an initialization phase block, a desired signal block, a MAI signals block and a Rake receiver block, as indicated by Figure 8.1. The initialization phase block is only executed once, before the actual simulation is started. The other blocks are executed for each iteration of the simulation loop. Since WCDMASim is a frame based simulator, one iteration of the simulation loop processes one WCDMA frame.
Chapter 8. WCDMASim

### Initialization Phase

- Set simulation parameters (including scrambling sequence)
- Generate Record of Fading Signals (desired mobile only)
- Generate symbols for DPCCH (Control Channel)
- Generate bits for DPDCH
- Apply OVSF Spreading to DPDCH and DPCCH
- Apply Scrambling Sequence
- Apply Pulse Shaping Filter
- Apply Specular Multipath Channel

**This Portion Iterates on a Frame-by-Frame Basis**

**desired signal**

- Generate symbols for DPDCH and DPCCH
- Apply OVSF Spreading
- Apply Scrambling Sequence
- Apply Pulse Shaping Filter
- Apply Specular Multipath Channel

**MAI Signals**

- Generate symbols for DPDCH and DPCCH
- Apply OVSF Spreading
- Apply Scrambling Sequence
- Apply Pulse Shaping Filter
- Apply Specular Multipath Channel

**RAKE Receiver**

- Matched Filtering and Decimation to Chip Rate
- Descramble Signal
- Despread Signal
- Estimate Channel Phase Offset
- Correct for Channel Phase Offset
- Combine RAKE Finger Outputs
- Detect Data Bits
- Compute Error

---

**Figure 8.1: WCDMASim uplink simulator structure [1].**

### 8.2.1 Initialization Phase Block

In the initialization phase block first some simulation parameters, as for example noise variance and doppler frequency, are calculated from the simulation parameters specified by the user of the simulator. Then the channelization codes and the scrambling code for the desired user are generated. The pulse shape is generated in this block as well, by taking the desired number of samples from a truncated version of the root raised cosine impulse response specified in the UMTS physical layer specification. Then the pilot symbols for the desired user are generated and the bits of these pilot symbols are inserted in the DPCCH frame of the desired user. Two frames of the DPDCHs for the desired user, containing randomly generated bits, are generated in this block as well. The DPCCH frame and the DPDCH frame(s) are then spread with the appropriate OVSF code, I-Q/code multiplexed, scrambled and pulse shape filtered to generate the transmitted signal for two frames of the desired user. After that the transmitted signals for the current and previous frame of the interfering users are generated. The DPDCH and DPCCH frames that are used to generate these signals both contain random data bits.

The initialization phase block of the WCDMASim uplink simulator is also used to precalculate the fading coefficients of the specular multipath fading channel model that is used. The multipath profile of the channel can be specified by the user of the simulator by giving delay/relative power...
8.2. WCDMASim Uplink Simulator

pairs. For each multipath a series of fading coefficients is calculated by multiplying its relative power with a series of complex numbers whose amplitude has Rayleigh distribution and whose phase is uniformly distributed from $-\pi$ to $\pi$. Usually, enough fading coefficients for an entire simulation are precalculated in the initialization phase block. Only in very long simulations additional fading coefficients will have to be calculated when the actual simulation is already started. After the fading coefficients are calculated WCDMASim enters the simulation loop.

8.2.2 Desired Signal Block

In the simulation loop first the desired signal block is executed. In this block first an additional frame for each DPDCH, containing random bits, is generated. These frames are spread with the OVSF code, I-Q/code multiplexed together with the DPCCH of the desired user, scrambled and pulse shape filtered. So now three frames of the transmitted signal of the desired user have been generated. These three frames are needed to be able to select the appropriate delayed parts of the transmitted signal to create the multipath signal later on. The apply specular multipath channel sub-block of the desired signal block takes these three signal frames and processes them through a specular multipath fading channel. This means that the sub-block first concatenates the three frames together into a super-frame. For each multipath the signal samples that correspond with the delay of that multipath are extracted from the super-frame. These samples are multiplied with the fade coefficients and the resulting values for all the multipaths are added to generate the multipath signal for the desired user. Since calculating fade coefficients for every sample of the transmitted signal frame is very computationally intensive, the simulator applies the fading coefficients to the transmitted signal samples in a piecewise constant fashion. In practice this means that a new fading coefficient is picked once every 400 chips.

8.2.3 MAI Signal Block

The MAI signal block is executed after the desired signal block. The MAI block performs the same functions for the interfering users as the desired signal block did for the desired user. However, to reduce the simulation times, application of the specular multipath fading channel is optional for the interfering users. The output of the MAI signal block is the (multipath) signal for the interfering users. To generate the received signal the simulator adds the multipath signal of the desired user, the multipath signal of the interfering users and a noise signal whose variance is determined from the signal-to-noise-ratio specified for the simulation.
8.2.4 Rake Receiver Block

The rake receiver block takes the received signal as input. Since the rake receiver in the simulator is assumed to be a ‘perfect’ rake receiver, it has exact knowledge of all the multipath delays of the channel. This knowledge is used to generate a delayed version of the received signal for each multipath delay. Each delayed version of the received signal is first passed through the receive filter, which filters the signal with the root raised cosine pulse shape and decimates the signal to the chip rate. After that each delayed version of the received signal is descrambled and despread. The pilot symbols on the despread DPCCH of the delayed received signal are then used to estimate the phase offset of the delayed received signal. These phase offset estimates are used to correct the bit estimates on the DPDCCHs and the DPCCH of the delayed received signal for their phase offset. The phase corrected bit estimates on the DPDCCHs and DPCCH of all the delayed received signals are then equal gain combined. Finally hard decisions for the bits on the DPDCCHs and the DPCCH are made based on the equal gain combiner outputs.

8.3 Extending WCDMASim

WCDMASim has to be extended in order to be able to simulate blind adaptive MMSE and PIC multiuser detection. Subsection 8.3.1 describes the implementation of the blind adaptive MMSE detector in WCDMASim and the changes to the simulator the implementation of this detector required. Subsection 8.3.2 describes the implementation of the PIC detector and the changes to WCDMASim that were required to implement this detector. Finally Subsection 8.3.3 describes the additional changes, not related to a particular detector implementation, that were made to WCDMASim.

8.3.1 Implementing Blind Adaptive MMSE Detection

Blind adaptive MMSE detection requires a short code CDMA system, as indicated in Section 3.4. The WCDMA specification fulfills this requirement by allowing short scrambling sequences in the uplink as indicated in Section 5.1. Although short codes were originally not implemented in WCDMASim, adding short code generation to WCDMASim is relatively straightforward. The Frequency Division Duplex (FDD) spreading and modulation document [11] of the UMTS physical layer specification describes how both the short- and the long UMTS scrambling sequences can be generated using shift registers. The code to generate short scrambling sequences could therefore be derived relatively easily from the long scrambling sequence generation code that was already present in WCDMASim. Figure 8.2 shows the shift register configuration to generate the UMTS short scrambling sequences.
8.3. Extending WCDMASim

Figure 8.2: Short scrambling sequence generation

Adaptive Wireless Networking (AWGN)
The blind adaptive MMSE detector implementation has to be added to the rake receiver block of WCDMASim. The blind adaptive MMSE detector has the same input and output parameters as the conventional CDMA detector, therefore its implementation does not require any changes to the input parameters of the rake receiver block. In the rake receiver block some changes are required. Since WCDMASim is a frame based simulator its conventional detector implementation operates on an entire WCDMA frame at once. So in this implementation, first a delayed version of an entire WCDMA frame is generated with the delay of the multipath that is currently processed. This delayed frame is then pulse-shape filtered, descrambled and despread. After that the phase offset of the received signal is determined from the pilot symbols in the DPDCH that has just been despread. Finally the entire frame of DPDCHs and DPCCH bits are corrected for this phase offset. The blind adaptive MMSE detector however has to update its adaptive filter coefficients after each detected bit and uses these updated coefficients for the detection of the next bit. Therefore in the blind adaptive MMSE detector implementation in the Rake receiver block the descrambling and despreads of an entire frame that is used in the conventional detector is replaced with a loop over all the symbols in a frame. Inside this loop equations (6.7), (6.10) and (6.7) are implemented.

At the time this report is written the blind adaptive MMSE detector implementation in WCDMASim is still being debugged.

8.3.2 Implementing PIC Detection

Implementation of the PIC detector does require changes to the input parameters of the Rake receiver block, since the PIC detector, in addition to the desired user, also has to detect the interfering users to perform the cancellation. Therefore the PIC detector needs information about the timing of the interfering users and it needs to know the scrambling and channelization codes of the interfering users. This information has to be passed to the Rake receiver block. In order to be able to do this, changes to the MAI signal block are required as well, because in the original WCDMASim implementation the channelization and scrambling codes that were used to generate the MAI are not stored. To implement the actual PIC detector equations equations (7.7), (7.8) and (7.9) will have to be implemented in the Rake receiver block of WCDMASim.

At the time this report is written not all the changes described above have already been implemented in WCDMASim.
8.3. Extending WCDMASim

8.3.3 Additional Changes

The code of the original WCDMASim uses a mix of different coding styles, resulting in slightly different variable names used for representing the same data in different parts of the simulator. This can be quite confusing to someone not familiar with the code who has to make changes or additions to the simulator. Therefor parts of WCDMASim were rewritten to make the used coding style more consistent before any other changes to WCDMASim were made.

To aid the implementation and debugging of blind adaptive MMSE and PIC multiuser detection, the possibility to plot the signals that are present in different parts of the simulator has been added to WCDMASim. This makes it possible to visualize the DPDCH and DPCCH data streams, the channelization and scrambling codes, the uplink frames, the transmitted signal, the rayleigh faded transmitted signal, the MAI signal, the noise and finally the signal that arrives at the receiver.
Chapter 9

Simulation Results

At the time this report is written the blind adaptive MMSE detector and the PIC detector are only partially implemented in WCDMASim. Therefore this chapter only contains some preliminary simulation results.

9.1 Data Stream to Received Signal

Figures 9.1 and 9.2 show the signals that are present in different parts of WCDMASim, during simulation of a WCDMA system. The system that is being simulated has 5 active users and uses short scrambling sequences. The desired user uses a spreading factor of 64 for a single DPDCH and moves with 50 km/h. The simulated channel is a single path channel with a Signal to Noise Ratio (SNR) of 10dB.

Figure 9.1(a) shows the first 8 bits on the DPDCH and the first two bits on the DPCCH of the desired user. The DPDCH and DPCCH are spread with the DPDCH and the DPCCH channelization codes of Figure 9.1(b). The spread DPDCH and the DPCCH are mapped to the inphase channel, respectively the quadrature channel and then multiplied with the complex scrambling code of Figure 9.1(c). This results in a complex uplink frame of which the first 512 chips are shown in Figure 9.1(d). The uplink frame is filtered with the pulse shaping filter to generate the transmitted signal of Figure 9.1(e). The velocity of the desired user is used to calculate the Rayleigh fading coefficients. The transmitted signal is multiplied with these Rayleigh fading coefficients to obtain the Rayleigh faded signal of Figure 9.1(f). Figure 9.2(a) shows the output of the MAI Signal Block of the simulator and in Figure 9.2(b) the noise is shown. These two signals are added to the Rayleigh faded signal to obtain the received signal that is shown in Figure 9.2(c).
Figure 9.1: From data streams to received signal in the WCDMASim simulator.
Figure 9.2: From data streams to received signal in the WCDMASim simulator (continued).
Figure 9.3: Bit-error-rate comparison between a long and a short code WCDMA system.
9.2 Long and Short Scramble Code Comparison

Figure 9.3 shows a comparison between the Bit-Error-Rate (BER) of a WCDMA system using long scramble codes and the BER of a system using short scramble codes for a number of Signal to Noise Ratios (SNRs). The system that is being simulated has 4 active users. The desired user uses a spreading factor of 32 for a single DPDCH and moves with 5 km/h. In the simulation a multipath channel is used. To obtain the BER of the system for a specific SNR 140 WCDMA frames are simulated. Since one WCDMA frame contains 38400 chips and the used spreading factor is 32 this corresponds with 168000 simulated data bits.

Figure 9.3 shows that in this particular case there is hardly any BER performance difference between a long and a short scramble code WCDMA system. However, as was stated in Section 2.1, the cross correlation between short sequences and thus the experienced interference and BER depend on the particular short sequences. So to make a statement about the generality of this result more research will have to be performed on this particular issue.

9.3 Conventional and Blind Adaptive MMSE Detector Comparison

Figure 9.4 shows a comparison between the BER of a conventional detector and a blind adaptive MMSE detector for a number of SNR. The WCDMA system that is being simulated has 11 active users and uses short scrambling sequences. The desired user uses a spreading factor of 32 and a single DPDCH. The desired user does not move and there is no multipath, so an AWGN channel is used. To obtain the BER of the system for a specific SNR 1500 WCDMA frames are simulated. Since one WCDMA frame contains 38400 chips and the used spreading factor is 32 this corresponds with 1800000 simulated data bits. This means that the BER values for SNR values of 8dB and larger are not very reliable, because these BER values are based on the occurrence of only a few error events.

In figure 9.4 it can be seen that the blind adaptive MMSE detector has about the same BER performance as the conventional detector for the same SNR. This result was not expected, since blind adaptive MMSE detection performed a lot better than conventional detection in simulations for a general CDMA system performed in previous research [2] and theoretically should also perform a lot better in WCDMA. At the moment the blind adaptive MMSE detector implementation in WCDMASim has not been tested enough to make a statement about the cause of this unexpected result. Therefor additional simulations of blind adaptive MMSE detection using WCDMASim will have to be performed in future research.
Chapter 9. Simulation Results

Figure 9.4: Bit-error-rate comparison between conventional and blind adaptive MMSE detection.
Chapter 10

Conclusions and Recommendations

10.1 Conclusions

In this report two multiuser detection algorithms, blind adaptive MMSE detection and PIC detection have been analyzed for suitability for application in UMTS communications systems. These detection algorithms were first described for general CDMA communications systems. After that the differences between general CDMA and WCDMA, the specific type of CDMA that is used in the physical layer of UMTS, have been described. Also, compared to previous research [3], a more realistic model of the communications channel has been introduced. This was followed by a description of the application of blind adaptive MMSE and PIC detection algorithms to WCDMA. After that, the changes to WCDMASim, a WCDMA simulator, required for simulating blind adaptive MMSE and PIC detection were described. Finally some preliminary simulation results that were obtained using the modified WCDMASim were shown. The challenges and problems that were encountered while obtaining these simulation results have been described.

Although not straightforward, application of the blind adaptive MMSE and PIC multiuser detection algorithms to WCDMA was found to be possible within the limitations set by the UMTS specification. Implementation of blind adaptive MMSE and PIC multiuser detection in the WCDMASim simulator however turned out to be more cumbersome.

Adding the blind adaptive MMSE detector (and the short scrambling code generation it requires) to WCDMASim was relatively easy. But the blind adaptive MMSE requires configuration of some detector parameters. This turned out to be quite hard because of long simulation times. For example, simulation of a WCDMA system with 11 users, a spreading factor of
32, one DPDCH per user and a simple AWGN channel without multipath or fading takes 15 seconds for 10 WCDMA frames on a Pentium 4 2.4 GHz. This results in a simulation time of 1.25 ms per simulated bit. So the simulation of 1 million bits takes 20 minutes. These long simulation times made experimenting with different detector parameters and general debugging of detector implementations difficult.

The current implementation of blind adaptive MMSE detection in WCDMASim does not yet show the improvements in bit-error-rate performance that were expected based on simulation results obtained for other CDMA communications systems in previous research [2]. It is not yet clear if this is caused by incorrectly chosen detector parameters, bugs in the implementation of the detector in the simulator, or the fact that the influence of MAI on the performance of conventional WCDMA detection is negligible. In the latter case, not only blind adaptive MMSE detection, but no multiuser detection technique in general will show a performance improvement.

Implementation of the PIC detector requires some changes to the overall architecture of the WCDMASim simulator. Although these changes are certainly possible, they require a redesign of the architecture of WCDMASim and have therefore not yet all been implemented in WCDMASim.

So looking back at the goals that were set in Chapter 1 the conclusions from the research described in this report can be summarized as follows:

1. The application of blind adaptive MMSE and PIC multiuser detection algorithms to WCDMA is possible within the limitations set by the UMTS specification.

2. Blind adaptive MMSE detection has been implemented in WCDMASim. The WCDMASim simulator requires changes to its software architecture for implementation of PIC detection.

3. The preliminary simulation results of blind adaptive MMSE detection do not yet show a BER performance improvement compared to conventional WCDMA detection. PIC detection has not yet been simulated due to the changes to the simulator it requires.

After having obtained more experience with WCDMASim, this simulator turned out to be not as suitable for the project as was expected when it was initially chosen. The code of the simulator was inconsistent because of the different coding styles that were used by the different people that apparently developed the simulator. Also the code of the simulator could have been modularized in a better way. These problems have already partially been solved by rewriting parts of the code of the simulator to make the coding style more consistent. But the lack of modularity still remains, which makes applying changes or additions to the simulator difficult.
10.2 Recommendations and Future Work

To determine the suitability of blind adaptive MMSE detection for WCDMA reliable simulations of this detection technique will have to be performed. This still requires a number of additional research steps:

1. The derivation of blind adaptive MMSE detection for WCDMA will have to be re-evaluated to make sure it does not contain any mistakes.

2. The implementation of blind adaptive MMSE detection in the simulator has to be checked for consistency with the derivation.

3. In Honig et al [5] equations are given to analytically determine the optimum detector parameters for the blind adaptive MMSE detector. Although some problems with these equations were found in previous research [3] it may be useful to study if they are applicable in case of WCDMA.

4. Additional simulations will have to be performed to gain knowledge about the behavior of blind adaptive MMSE detection in WCDMA.

The BER performance of PIC detection cannot be determined using analytical techniques [4]. So in order to determine the performance of PIC in WCDMA, simulation results for PIC in WCDMA will have to be found in literature, or this detection technique will have to be implemented in a WCDMA simulator and simulated. Implementation of PIC detection in a simulator is also useful to obtain practical experience with this particular detection technique.

However because major changes to the architecture of WCDMASim are required to implement PIC detection and WCDMASim was found to have some other problems, as for example long simulation times and bad modularity, developing a WCDMA simulator that is more suited for the requirements of multiuser detection may be considered at this point.
Bibliography


