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**Groupwise Weighted Parallel Interference Cancellation
for Asynchronous Multirate DS-CDMA**

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“In the technical world every *solution* has a *problem*”

Alwin van Meeteren

Abstract:

A.F. van Meeteren, *Groupwise Weighted Parallel Interference Cancellation for Asynchronous Multirate DS-CDMA*, Delft, the Netherlands, Sep. '98-Mar. '99.

This report presents the performance of Groupwise Weighted Parallel Interference Cancellation (GW-PIC) receivers within a direct sequence code division multiple access (DS-CDMA) scheme.

A multirate model, derived from a single rate model, is presented. In this multirate model several Multiuser Detection receivers have been studied. In this thesis one of them, the GW-PIC receiver is studied in an AWGN channel and in Rayleigh fading channels.

It has been shown that implementing Groupwise weighting factors into a PIC receiver can increase the BER performance of the receiver in an AWGN environment. An extensive study has been done to the influence of weighting factors in such an environment. In Rayleigh fading channels, however, the implementation weighting factors in a Groupwise way is not the best solution. Finally, the influence of delay estimation errors on the BER performance of the GW-PIC receiver has been studied in a Rayleigh fading channel.

Indexing Terms: Code Division Multiple Access (CDMA), Multirate, Multiuser Detection, Interference Cancellation, Multiple Processing Gain, Weighting Factors.

Samenvatting:

A.F. van Meeteren, *Groepsgewijze Gewogen Parallelle Interferentie Onderdrukking voor Asynchrone Multirate DS-CDMA*, Delft, Nederland, september '98-maart '99.

Dit rapport presenteert de prestatie van Groepsgewijze Gewogen Parallelle Interferentie Onderdrukking (eng: GW-PIC) ontvangers in een direct sequence code division multiple access (DS-CDMA) systeem.

Een model voor een systeem met meerdere datasnelheden is gepresenteerd,, afgeleid van een model voor een systeem met één datasnelheid. In dit "multirate" model zijn er verscheidene multiuser ontvangers bestudeerd. In dit afstudeerwerk is één van hen, de GW-PIC ontvanger bestudeerd in een AWGN kanaal en in "Rayleigh fading" kanalen.

Het is aangetoond dat het implementeren van Groepsgewijze weeg factoren in een PIC ontvanger de bit fouten kans prestatie van deze ontvanger kan verbeteren in een AWGN omgeving. Een uitgebreide studie is er verricht naar de invloed van weeg factoren in zo'n omgeving. Echter in "Rayleigh fading" kanalen is de implementatie van weeg factoren op een Groepsgewijze manier niet de beste oplossing. Uiteindelijk is er ook nog een studie verricht naar de invloed van fouten in schattingen van de vertraging van het aangekomen signaal op de bit fouten kans prestatie van de GW-PIC ontvanger in een "Rayleigh fading" kanaal.

Trefwoorden: Code Division Multiple Access (CDMA), Multirate, Multiuser Detection, Interference Cancellation, Multiple Processing Gain, Weighting Factors.

Preface

This thesis work has been done in the laboratory for Telecommunications and Traffic Control, Delft University of Technology, The Netherlands.

I like to thank my parents, my brother and, last but not least, my girlfriend for supporting me (in hard times) during my academic education.

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Delft, March 8, 1999

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List of symbols and abbreviations

Symbols:

$\mathbb{C}$	Complex Set
$\mathbb{R}$	Real Set
$\mathbf{A}$	Amplitude matrix
$\mathbf{C}$	Channel Coefficient matrix
$\mathbf{S}$	Code matrix
β	Roll-off factor
$\eta(t)$	Additive White Gaussian Noise
σ	Variance of the noise
ρ_{ij}	Crosscorellation of user i to user j
τ	Delay
ϕ_k	Carrier phase
$\Psi_{i,k}^{(n)}$	Sampled Interference Term of user i , path k
A_k	Complex Amplitude of user k .
$b_k^{(n)}$	n th transmitted bit
B	Bandwidth
c_k	Channel Attenuation factor
f_0	Coherence Bandwidth
$\mathbf{I}$	Identity Matrix
j	Complex unit, $\sqrt{-1}$
h_k	Channel Impulse Response
k	Index for users
K	Maximum number of users
L	Number of paths
M	Length of a transmitted block of symbols
n	Symbol Interval
N_c	Number of chips in code waveform
$p(t)$	rectangular pulse
$r(t)$	received signal
$\mathbf{R}$	Crosscorrelation matrix
s_k	Code waveform
T_c	Chip duration time
u_k	Received signal of user k
w	noise after detection

W_j	Weight matrix
W_1	Weighting factor high bit rate users
W_2	Weighting factor medium bit rate users
W_3	Weighting factor low bit rate users
$X_{rc}(f)$	Frequency response of the root raised cosine filter

Abbreviations:

AWGN	Additive White Gaussian Noise
BER	Bit Error Rate
BPSK	Binary Phase Shift Keying
CDMA	Code Division Multiple Access
DCS-1800	Digital Cordless System at 1800 MHz.
DECT	Digital European Cordless Telecommunication
DS-CDMA	Direct Sequence Code Division Multiple Access
GW-PIC	Groupwise Weighted Parallel Interference Cancellation
GSIC	Groupwise Successive Interference Cancellation
GSM	Global System for Mobile Communications
IC	Interference Cancellation
IMT-2000	International Mobile Telecommunications 2000
ISI	Inter Symbol Interference
ITU	International Telecommunications Union
LOS	Line Of Sight
MAI	Multiple Access Interference
MC	Multi Code
MF	Matched Filter
MPG	Multiple Processing Gain
MRC	Maximum Ratio Combining
MUD	Multi User Detection
NMT	Nordic Mobile Telecommunications
PIC	Parallel Interference Cancellation
PG	Processing Gain
PN	Pseudorandom Noise
SIC	Successive Interference Cancellation
SNR	Signal to Noise Ratio
TACS	Total Access Cellular Systems
TDMA	Time Division Multiple Access
QPSK	Quadrature Phase Shift Keying
UMTS	Universal Mobile Telecommunication System
VSF	Variable Spreading Factor
WPIC	Weighted Parallel Interference Cancellation

Operators:

$*$	Convolution
$(\cdot)^*$	Complex Conjugate
$\text{diag}(\cdot)$	Vector consisting of diagonal elements of a matrix
$\hat{(\cdot)}$	Estimate
$(\cdot)^H$	Conjugate Transpose
$\text{sgn}(\cdot)$	Signum function
$\Re(\cdot)$	Real Part

Chapter 1

Introduction.

In the 1980s analogue cellular communication systems became operational in Europe, based in national or imported standards. Examples are NMT (Nordic Mobile Telecommunications) systems at 450 MHz and 900MHz, TACS (Total Access Cellular Systems) and C-Netz. Moreover, a large variety of first generation analogue systems for cordless, paging and private mobile radio were installed.

In the 1990s Europe has introduced and will further develop second generation digital mobile telecommunication systems such as GSM, DCS1800 and DECT. These systems are able to provide several tens of millions of mobile radio users with mobile phones, pagers and private mobile radio equipment.

Today the mobile communications scene is significantly different. Mobile telecommunication has become a mass market. People are used to communicate while they are moving. On the demand side, the call for new and more sophisticated services and applications requiring higher bit rates will increase. On the supply side, network operators and service providers will be looking for opportunities to distinguish themselves from their competitors. The main developments are done by the International Telecommunication Union (ITU). These developments will eventually lead to the introduction of a new generation mobile telecommunication systems called IMT-2000, in Europe referred to as the 'Universal Mobile Telecommunication System' (UMTS).

UMTS will integrate the system concepts of digital second generation mobile systems. Users like to employ their mobile terminal anywhere. In outdoor environments ranging from rural areas to high-density city centres, but also in indoor environments. Services traditionally offered by separate systems will in UMTS be supported by one single infrastructure. Since UMTS will integrate data, video and speech transmission at high quality, a need for different data rates in the same system is welcome. Data transmission with rates varying from 10 kbps to 2 Mbps is required. To remain the quality of the transmission high, bit error rates of less than 10^{-2} are wanted for speech and bit error rates of less than 10^{-6} for data transmission are required [1]. Data rates from ranging from 144 kbps are required for high mobility users with wide area coverage and 2 Mbps for low mobility users with local coverage. To achieve this there are two multiple access methods are under attention.

One of these multiple access schemes is TDMA (Time Division Multiple access) where each user transmits in an appointed time slot. Another multiple access scheme is CDMA (Code Division Multiple Access). CDMA can be implemented in different ways. One of them is DS-CDMA (Direct Sequence Code Division Multiple Access) and will be the used scheme in this report. DS-CDMA consists of spreading the signal by modulating the carrier with a user specific wideband signature waveform [1]. All users use the full bandwidth and thus interfere with each other. The users can be separated from each other using the codes assigned to them. The received signal is correlated with this code to detect the user having that specific code. This method of detection is called Matched Filtering. A condition to be able to detect the signal belonging to the specific user is that the codes used for all the users have some sort of orthogonality between them. This means that the autocorrelation of the codes is very high and the crosscorrelation between them is very low. Since the signals won't arrive at the same moment but with different delays, it is not possible to generate codes, which are always orthogonal. Because of the lack of orthogonality between the codes when there are varying delays, the users suffer from interference caused by other users. This interference is called MAI (Multiple Access Interference) and limits the capacity of a CDMA system.

Depending on the used channel and because of the MAI it is not always possible to detect the separate users perfectly with only Matched Filtering. Other receivers have been proposed where the output of a Matched Filter is used as an estimate from the received user specific signal. This estimate is being used in these receivers to subtract it from the received wideband signal and a new Matched Filtering stage. Examples of this are SIC (Successive Interference Cancellation) and GSIC (GroupWise Serial Interference Cancellation) receivers and, if the subtraction is done in a parallel way, the PIC (Parallel Interference Cancellation) and GW-PIC (GroupWise Weighted Parallel Interference Cancellation) receivers. These receivers use the transmitted signals of all users to detect the transmitted signal for a specific user. This method is called multi-user detection (MUD) or interference cancellation (IC) [2][3][4]. The methods used in the receivers are more complex than the Matched Filtering method and therefore, it is recommended to implement this in complex situations where all the users are supposed to be detected, such as Base Stations.

Since UMTS offers the possibility to transmit data, voice and even video, there are different data rates, which vary from user to user. There are several possibilities to design a multiple data rate (multi rate) multi-user communication system [5]. One way of designing such a system is to allocate parallel communication channels to higher bit rate users. This implicates that a higher bit rate user transmits using several spreading sequences. This method is called Multi Code (MC) access. Another way of designing a multi rate system is to use a modulation alphabet with more symbols for the higher bit rate users. The method used in this work is called Multiple Processing Gain (MPG) or Variable Spreading Factor (VSF) and consists of modulation of all signals in to the same bandwidth. The high bit rate users have a smaller symbol duration time than the lower bit rate users. Modulation in to the same bandwidth implies that the high bit rate users have a lower processing gain and can transmit more symbols than lower bit rate users. Combinations of the three methods mentioned above can also be made.

A GroupWise Weighted Parallel Interference Cancellation receiver will be studied in this work. This receiver is placed in a multi rate environment, which uses the Multiple Processing Gain access method. The receiver's performance has been tested and evaluated in an AWGN channel as well as a Rayleigh fading channel. A major part of this work consists of the evaluation of the influence of the weighing factors on the performance of this receiver for high, medium and low bit rate users. After this has been done the receiver will be evaluated in comparison with a full PIC receiver and a GSIC receiver in the same environments. In Chapter 2 the system model, which is used in this work, is introduced and in Chapter 3 different types of receivers are introduced. The GW-PIC receiver operating in an AWGN channel is described in Chapter 4 and the same receiver operating in a Rayleigh fading channel is described in Chapter 5. Finally, several conclusions are drawn and recommendations are made in Chapter 6.

Previously, the study of IC receivers in a multirate environment has been done without the application of weighting factors in these IC receivers, like the GSIC receiver [6], and the PIC receiver [7]. In this work, weighting factors are applied and it can be seen that in some cases this improves the BER performance of the IC receivers very much. It is therefore an interesting study and many more interesting studies can be done extending this thesis work.

Chapter 2

Multirate CDMA system in a fading environment

In this chapter a mathematical model for the BPSK DS-CDMA system with K number of users will be considered. The system model for an asynchronous single rate CDMA system will be described in Section 2.1. This description consists of three parts. The first part describes the transmitted signal, the second describes the channel and finally the received CDMA signal will be described. After a single rate system has been described, a multirate asynchronous CDMA system will be modelled in a mathematical way.

2.1 System model for an asynchronous single rate CDMA system.

In a BPSK DS-CDMA system all the users transmit in the same bandwidth. To keep the different users separated, transmits each user with his own specific code. The bit stream of each user is multiplied by the specific code, which has a higher bitrate than the bitstream itself. The bits are thus spread into a broader bandwidth. The transmitted signal is then being sent over the channel, which contains different kinds of noise. The noise disrupts the signal and the signal is detected, making use of the same codes again. After correlation with the codes, the noise is spread over a broad bandwidth and the received signal gets the small bandwidth it had before the multiplication with the codes. The power of the noise is spread and the power of the received signal is despread. Thus the signal can be easily detected and regained (see Figure 2.1).

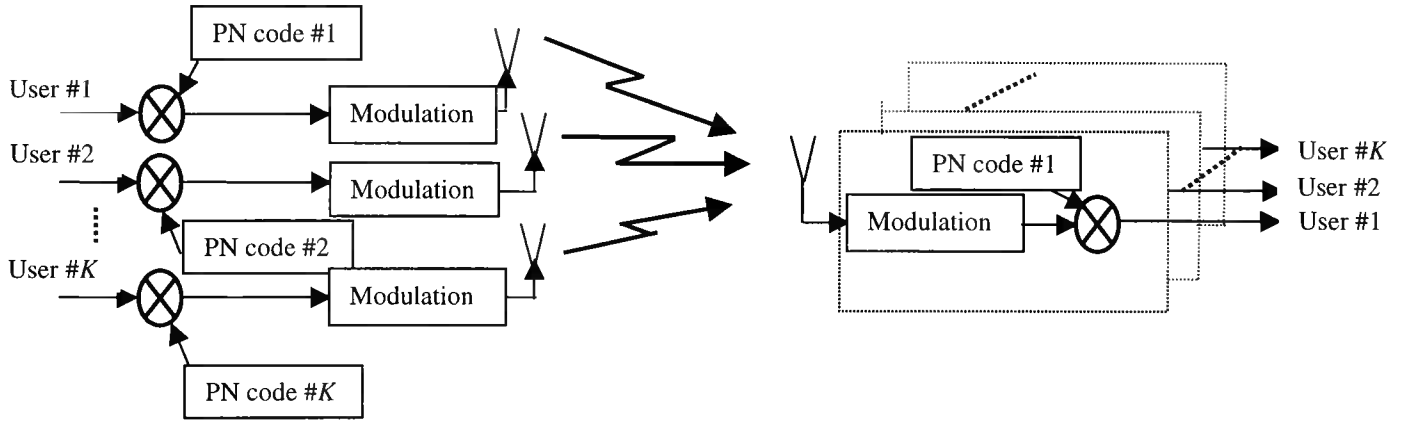


Figure 2.1: System model for a DS-CDMA system

2.1.1 Transmission of the CDMA signal.

Some basic definitions are needed to model and express the transmitted DS-CDMA signal. The data of each user k on symbol interval n , which is BPSK modulated, is defined as $b_k^{(n)} \in \{-1, 1\}$ and $n = 0, 1, \dots, M-1$ where M is the length of the transmitted block of symbols. This data is spread by multiplication by a user specific code, which is a real pseudo-random sequence of binary symbols $s_k^{(i)}$, which are called chips. From these chips a signature waveform, which is unique for each user, is constructed. In (Formula 2.1) is the mathematical model for the signature wave denoted.

$$s_k(t) = \begin{cases} \frac{1}{\sqrt{N_c T_c}} \sum_{n=0}^{N_c-1} s_k^{(n)} p(t - nT_c) & t \in [0, T) \\ 0 & \text{otherwise} \end{cases} \quad (2.1)$$

In (Formula 2.1) is N_c the number of chips, T_c the time length of a chip and T the time length of a symbol interval, while $s_k^{(i)} \in \{-1, 1\}$ and $p(t)$ is the chip waveform, which is given by Formula (2.2).

$$p(t) = \begin{cases} 1 & t \in [0, T_c) \\ 0 & \text{otherwise} \end{cases} \quad (2.2)$$

This chip waveform is a rectangular pulse with an amplitude of one. The energy of this signature waveform is normalised as in (Formula 2.3).

$$\int_0^T |s_k(t)|^2 dt = 1 \quad (2.3)$$

The bandpass spread-spectrum transmitted signal of the k th user can now be written as in (Formula 2.4).

$$u_k(t) = \sum_{n=0}^{M-1} A_k b_k^{(n)} s_k(t - nT - \Delta\tau_k) e^{j\theta_k} \quad (2.4)$$

In this complex envelope $u_k(t)$ is the transmitted spread spectrum signal, A_k stands for the carrier amplitude and θ_k is the random carrier phase. The random delay term $\Delta\tau_k$ is due to the asynchronous transmission of the symbols and is uniformly distributed in the interval $[0, T)$ [8].

The bandwidth of the message bit rate of user k is reciprocally related to the symbol interval T . After multiplication by the code the message bit rate will be related to the chip duration time T_c , which is much shorter than the symbol interval time. The signal has now a bandwidth related to T_c and will be much broader. This broad bandwidth can be seen as the original bandwidth multiplied by a certain factor, which is called processing gain. The processing gain (PG) is defined in (Formula 2.5).

$$\text{PG} = \frac{T}{T_c} \quad (2.5)$$

The processing gain can also be defined as the chip rate divided by the data rate. The larger the PG, the less sensitive the system is to noise. The power of the noise that is spread into a bandwidth is related to the PG.

2.1.2 Frequency selective slowly fading Rayleigh channel

The study of transmitted signals in a mobile telecommunication system requires a study of the physical medium through which the signal is being transmitted. The most simple model for a physical medium is one with only AWGN (Additive White Gaussian Noise). This models the thermal noise plus other noise sources unrelated to the transmitted signals. The noise power in a frequency band with bandwidth B is equal to $2\sigma^2 B$, where σ^2 represents the double sided spectral density of the noise. However, in mobile telecommunication studies this model is good to begin with but not good enough to model the system in a realistic way. In realistic situations the transmitted signal is often received coming from different propagation paths. The received signal may consist of multiple signals from the same transmitter, which followed different paths. The copies of the signal hence created can differ from each other in phase, caused by the varying time of arrival at the receiver. Adding copies with different phases in the channel output causes an amplitude variation in the received signal. This phenomenon is called *multipath fading* [9].

A realistic channel model consists of a channel model where the transmitter can not be “seen” from the receiver. Buildings, mountains or trees block the propagation paths of the signal. This means that there is no *line of sight* (LOS) component and the channel that the signal travels through is called a *Rayleigh faded channel*. The signal at the receiver’s end experiences Rayleigh fading, which has a probability density function shown in:

$$p_R(r) = \begin{cases} \frac{r}{\sigma_r^2} e^{-\frac{r^2}{2\sigma_r^2}} & r \geq 0 \\ 0 & \text{otherwise} \end{cases} \quad (2.6)$$

Where σ_r^2 is the variance of the received power and r is the envelope amplitude of the received signal [19]. The channel is modelled as a time variant linear filter with the following impulse response:

$$h_k(t) = \sum_{l=1}^{L_k} c_{k,l}(t) \delta(t - \tau_{k,l}(t)) e^{j\phi_k} \quad (2.7)$$

where L_k is the number of propagation paths, $c_{k,l}$ is the channel attenuation factor, $e^{j\phi_k}$ is the change in phase of user k and path l and $\delta(t - \tau_{k,l}(t))$ is the Dirac function, which is delayed with a time factor $\tau_{k,l}(t)$ which is the delay of the l th path of user k varying in time. The asynchronous channel model for user k can now be modelled as depicted in Figure 2.2.

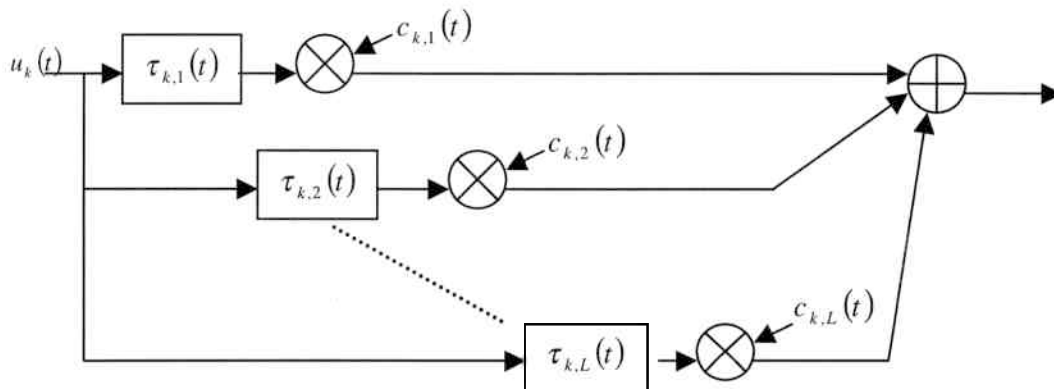


Figure 2.2: Asynchronous channel model for user k [6].

The number of paths is assumed to be equal for all users. That is the reason that in Figure 2.2 L_k is replaced by L . The phase of the impulse response is uniformly distributed over $[-\pi, \pi]$.

The channel is modelled as a *slowly fading frequency selective Rayleigh process*. The channel is *frequency selective* because the *coherence bandwidth* f_0 is smaller than the bandwidth used in spread spectrum systems like these. The coherence bandwidth is defined as the bandwidth over which the channel passes all spectral components with a linear phase [9]. In these systems a very broad bandwidth is used and the signal components falling outside f_0 will be affected differently than the signal components which are inside f_0 . The channel is *slowly fading* because the channel response remains approximately the same during a symbol interval.

If the receiver or the transmitter is in motion, the channel changes as the location of the receiver or the transmitter changes in time. In Figure 2.3 simulated realisations of Rayleigh fading channels corresponding to a velocity of 80 km/h and a symbol rate of 16 kbits/s is shown. In the figure normalised amplitude during the transmission of one thousand symbols is illustrated. It is clearly seen that the amplitude of the received signal changes as the transmitter is in motion. The amplitude changes even more rapidly when the transmitter moves faster.

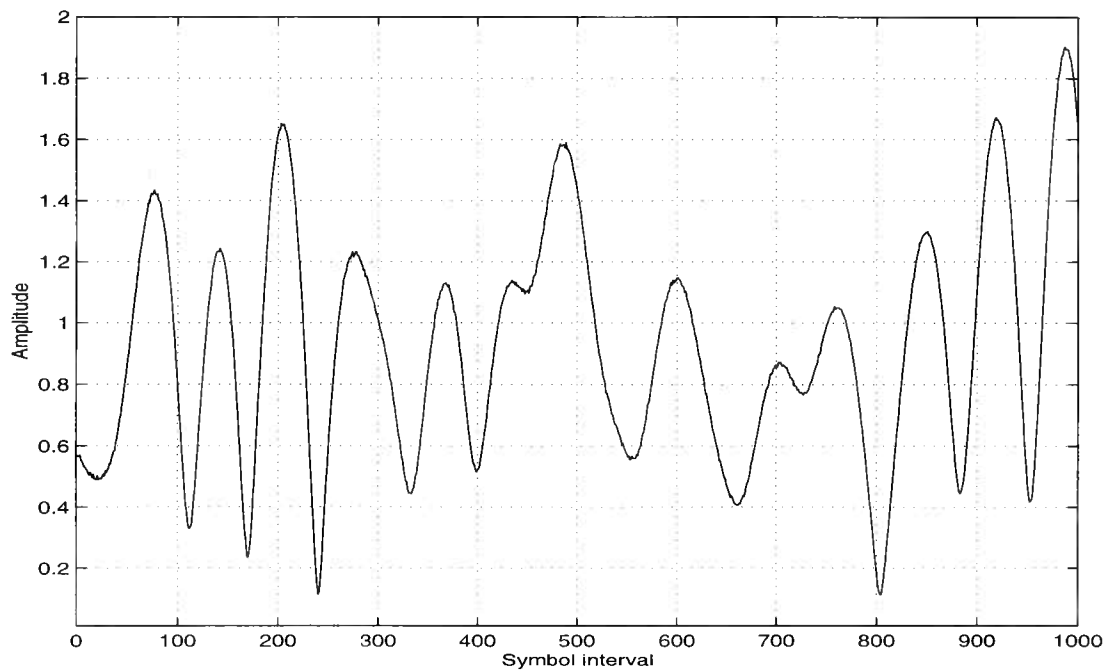


Figure 2.3: Fading process, vehicle speed is 80 km/h [8].

2.1.3 Reception of the CDMA signal

After the signal has passed through the slowly fading frequency selective Rayleigh channel, described in Section 2.1.2., the signal is received. This received signal can be seen as a convolution system with the transmitted signal (Formula 2.4) and the impulse response of the channel (Formula 2.7) [10] and AWGN $\eta(t)$. The received signal can then be written as:

$$r(t) = \sum_{k=1}^K \sum_{l=1}^L c_{k,l}(t) e^{j\phi_k} u_k(t - \tau_{k,l}) + \eta(t) \quad (2.8)$$

This formula can be extended using (Formula 2.4) and describes the received signal as denoted in:

$$r(t) = \sum_{n=0}^M \sum_{k=1}^K A_k b_k^{(n)} \sum_{l=1}^L c_{k,l}^{(n)} s_k(t - nT - \tau_{k,l} - \Delta\tau_k) + \eta(t) \quad (2.9)$$

Since in this formula there are many summations it is more convenient to describe Formula 2.9 with the use of matrices. The time discrete received signal can be represented as a column vector $\mathbf{r}$, which is given by:

$$\mathbf{r} = \mathbf{SCAb} + \mathbf{n} \in \mathbb{C}^{(M+2)N_c N_{sc} \times 1} \quad (2.10)$$

where $\mathbf{S}$ is a signature waveform or code matrix, $\mathbf{C}$ the channel coefficient and carrier phase matrix, $\mathbf{A}$ the carrier amplitude matrix $\mathbf{b}$ the binary data vector and $\mathbf{n}$ the noise vector. The number of chips is given by N_c and the number of samples per chip is given by N_{sc} . The code matrix $\mathbf{S}$ consists of the sampled waveforms $s_k(t)$ forming the columns of the matrix. The time discrete signature waveform of user k may be represented as:

$$\mathbf{s}_k = (s_k^{(0)}, s_k^{(1)}, \dots, s_k^{N_c N_{sc}-1})^T \quad (2.11)$$

Each propagation path of each user for symbol intervals $m = 0, 1, \dots, M-1$ has his own column in $\mathbf{S}$ and the delay of the first propagation path of the first user is set to zero. The maximum difference in delay between the first propagation path of the first user and the first propagation path of the last user is assumed to be smaller than one symbol interval (Formula 2.12). The reason for this assumption is that it keeps the complexity of the channel the same, while the modelling of the channel in a mathematical way is getting less complex. The delays are also assumed to satisfy the inequality noted in (Formula 2.13).

$$\tau_{K,l} + \Delta\tau_K - (\tau_{1,l} + \Delta\tau_1) < T \quad (2.12)$$

$$\tau_{1,l} + \Delta\tau_1 \prec \tau_{2,l} + \Delta\tau_2 \prec \dots \prec \tau_{K,l} + \Delta\tau_K \quad (2.13)$$

With fixed propagation delays the code matrix S may now be written as:

$$S = \begin{bmatrix} S^{(0)}(0) & 0 & 0 & \dots & 0 \\ S^{(0)}(-1) & S^{(1)}(0) & 0 & & \vdots \\ S^{(0)}(-2) & S^{(1)}(-1) & S^{(2)}(0) & & \\ \vdots & & & \ddots & \\ 0 & & & & S^{(M-1)}(0) \\ 0 & 0 & & & S^{(M-1)}(-1) \\ 0 & 0 & 0 & \dots & S^{(M-1)}(-2) \end{bmatrix} \in \mathbb{R}^{(M+2)N_s N_{sc} \times MKL} \quad (2.14)$$

where $S^{(i)}(0)$, $S^{(i)}(-1)$ and $S^{(i)}(-2)$, with $i = 0, 1, \dots, M-1$, can be written as:

$$\begin{bmatrix} S^{(i)}(0) \\ \dots \\ S^{(i)}(-1) \\ \dots \\ S^{(i)}(-2) \end{bmatrix} = \begin{bmatrix} \text{Vertical bars} & 0 \\ \dots & \dots \\ \text{Vertical bars} & \dots \\ \dots & \dots \\ \text{Vertical bars} & 0 \end{bmatrix} \quad (2.15)$$

The $(k-1)L+l$ th vertical bar (column) in (Formula 2.15) can be seen as the time discrete signature waveform of user k shifted vertically corresponding to the random time offset depending on the transmission moment $(\tau_{k,l} + \Delta\tau_k)$ and corresponding to the delay of path l . As can be seen from Formula 2.15, a particular symbol transmitted in interval n may overlap with the symbols of other users transmitted in interval $[n-2, n+2]$ and cause interference with these symbols. This is due to the uncoordinated transmission of the users [8].

The channel coefficient matrix C consists of the random carrier phases ϕ_k and of the channel coefficients $c_{k,l}^{(n)}$ (Formula 2.9) and can be written as (2.16):

$$C = \text{diag}(C^{(0)}, C^{(1)}, \dots, C^{(M-1)}) \in \mathbb{C}^{MKL \times MK} \quad (2.16)$$

where the matrix $\mathbf{C}^{(m)}$ can be written as in (Formula 2.17).

$$\mathbf{C}^{(m)} = \text{diag}(\mathbf{c}_1^{(m)}, \mathbf{c}_1^{(m)}, \dots, \mathbf{c}_L^{(m)}) \in \mathbb{C}^{KL \times K} \quad (2.17)$$

and (2.18)

$$\mathbf{c}_k^{(m)} = (c_{k,1}^{(m)} e^{j\phi_k}, c_{k,2}^{(m)} e^{j\phi_k}, \dots, c_{k,L}^{(m)} e^{j\phi_k})^T \in \mathbb{C}^L \quad (2.18)$$

The matrix $\mathbf{A}$ contains the carrier amplitudes of the users, and is written as in (Formula 2.19).

$$\mathbf{A} = \text{diag}(\mathbf{A}, \mathbf{A}, \dots, \mathbf{A}) \in \mathbb{R}^{MK \times MK} \quad (2.19)$$

$\mathbf{A}$ is a diagonal, $K \times K$ matrix as in (Formula 2.20).

$$\mathbf{A} = \text{diag}(A_1, A_2, \dots, A_K) \quad (2.20)$$

The elements of data vector $\mathbf{b}$ are the transmitted binary symbols of the different users, when they transmit a block of M symbols.

$$\mathbf{b} = (b_1^{(0)}, b_2^{(0)}, \dots, b_K^{(0)}, b_1^{(1)}, \dots, b_K^{(M-1)})^T \in \{-1, 1\}^{MK} \quad (2.21)$$

The crosscorrelation coefficient ρ_{ij} is a measure for the lack of orthogonality between two sets of codes, e.g. s_i and s_j . The crosscorrelation coefficient can be written as [11]:

$$\rho_{ij} = \langle s_i, s_j \rangle = \int_0^T s_i(t) s_j(t) dt \quad (2.22)$$

A crosscorelation matrix $\mathbf{R}$ can be defined consisting of various crosscorelation coefficients:

$$\mathbf{R} = \{\rho_{ij}\} \quad (2.23)$$

When one of the coefficients ρ_{ij} is equal to zero it means that the sequence s_i is orthogonal to sequence s_j . $\mathbf{R}$ can be denoted as [6]:

$$\mathbf{R} = \mathbf{S}^T \mathbf{S} = \begin{bmatrix} R[0] & R^H[1] & 0 & \dots & 0 \\ R[1] & R[0] & R^H[1] & & \vdots \\ 0 & R[1] & R[0] & \ddots & \\ \vdots & & \ddots & \ddots & R^H[1] \\ 0 & \dots & & R[1] & R[0] \end{bmatrix} \in \mathbb{R}^{NK \times NK} \quad (2.24)$$

When there are three users the matrices $R[0]$ and $R[1]$ become:

$$R[0] = \begin{bmatrix} 1 & \rho_{12} & \rho_{13} \\ \rho_{12} & 1 & \rho_{23} \\ \rho_{13} & \rho_{23} & 1 \end{bmatrix} \quad (2.25)$$

$$R[1] = \begin{bmatrix} 0 & \rho_{21} & \rho_{31} \\ 0 & 0 & \rho_{32} \\ 0 & 0 & 0 \end{bmatrix} \quad (2.26)$$

2.2 System model for an asynchronous multirate CDMA system

There are several possibilities to design a multiple data rate (multi rate) multi-user communication system [5]. One way of designing such a system is to allocate parallel communication channels to higher bit rate users. This implicates that a higher bit rate user transmits using several spreading sequences. This method is called Multi Code (MC) access. Another way of designing a multi rate system is to use a modulation alphabet with more symbols for the higher bit rate users. For the high bit rate users there is 16 QAM and for the low bitrate users there is QPSK available, but this is complicated to implement and this method is not very flexible. The method used in this work is called *Multiple Processing Gain* (MPG) or *Variable Spreading Factor* (VSF) and consists of modulation of all signals in to the same bandwidth. The high bit rate users have a smaller symbol duration time than the lower bit rate users. Modulation in to the same bandwidth implies that the high bit rate users have a lower processing gain and can transmit more symbols than lower bit rate users. The problem with this method is that the high bit rate users are as much spread by the codes as the lower bit rate users. They are more sensible to noise. To compensate this the higher bit rate users are given more energy than the lower bit rate users. Combinations of the three methods mentioned above can also be made. In this Section the method of MPG will be discussed.

To be able to model a multirate system a distinction between has to be made between a *physical* or *real* user and an *effective* or *virtual* user [4]. A real user transmits with a telecommunication device, e.g. a portable phone and transmits with a certain PG. A real user that transmits with the highest PG available in the system is equal to one virtual user. The symbol rate of this lowest bit rate user is assumed to be $\frac{1}{T}$, where T is the corresponding symbol interval. This interval is called *supersymbol interval*.

Other users, transmitting at a higher data rate will have symbol rates of $\frac{m}{T}$, where m is an even integer. These users are per real user equal to m virtual users. Assumed that there are K virtual users in the system during one supersymbol interval, there will be K symbols transmitted. Even when there are virtual users the system model derived in Section 2.1 will still be valid. In a MPG system the codes will be partially used in order to spread the bits according to their rates. The transmitted signal of the m th virtual user of real user k becomes:

$$u_{k,m}(t) = \sum_{n=0}^{M_k-1} A_k b_{k,m}^{(n)} s_{k,m}(t - nT - \Delta\tau_k) e^{j\theta_k} \quad (2.27)$$

This formula is almost the same as Formula 2.4 except that u_k is replaced by $u_{k,m}$, b_k by $b_{k,m}$, M by M_k and s_k by $s_{k,m}$. The reason for this replacement is that in stead of having only real users, there are now virtual users, which all transmit with their different bits and partially used codes. The parts used from the codes are depending on the PG. When a user has the highest PG the whole code is used, otherwise parts of the code are used as depicted in Figure 2.4.



Figure 2.4: Length of the codes in a MPG scheme for one real user with PG equal to the even number of chips N_c and two virtual users belonging to another real user with PG of $N_c/2$.

In the system with MPG as an access scheme in the case of one real user with processing gain equal to N_c and one real user with processing gain equal to $N_c/2$, there are the signature waveforms $s_1(t)$ and $s_2(t)$ for the user with the lowest PG and $s_3(t)$ for the user with the highest PG. These signature waveforms are denoted in (Formulae 2.28, 2.29 and 2.30).

$$s_1(t) = \frac{1}{\sqrt{N_c T_c / 2}} \sum_{n=0}^{N_c/2-1} s_{1,n} p(t - nT_c) \quad (2.28)$$

$$s_2(t) = \frac{1}{\sqrt{N_c T_c / 2}} \sum_{n=N_c}^{N_c-1} s_{1,n} p(t - nT_c - T/2) \quad (2.29)$$

$$s_3(t) = \frac{1}{\sqrt{N_c T_c}} \sum_{n=0}^{N_c-1} s_{3,n} p(t - nT_c) \quad (2.30)$$

where N_c is the number of chips and T_c is the duration of a chip. The corresponding modulator matrix is [12]:

$$S_{MPG} = \begin{bmatrix} \sqrt{2}s_1^U & 0 & s_3^U \\ 0 & \sqrt{2}s_1^L & s_3^L \end{bmatrix} \quad (2.31)$$

where:

$$s_k = (s_{k,0}, s_{k,1}, \dots, s_{k,N_c-1})^T \quad (2.32)$$

$$s_k^U = (s_{k,0}, s_{k,1}, \dots, s_{k,N_c/2-1})^T \quad (2.33)$$

$$s_k^L = (s_{k,N_c/2}, s_{k,N_c/2+1}, \dots, s_{k,N_c-1})^T \quad (2.34)$$

From the matrix S_{MPG} , written as above, the correlation matrix can be found and is denoted in Formula 2.35.

$$R_{MPG} = S_{MPG}^T S_{MPG} = \begin{bmatrix} R[0] & R^H[1] & 0 \\ R[1] & R[0] & R^H[1] \\ 0 & R[1] & R[0] \end{bmatrix} \quad (2.35)$$

In this case where the factor $\sqrt{2}$ is used to equalise the received energies, $R[0]$ and $R[1]$ become:

$$R[0] = \begin{bmatrix} 1 & 2\rho_{11}^{UL} & \sqrt{2}\rho_{13}^U \\ 2\rho_{11}^{UL} & 1 & \sqrt{2}\rho_{13}^L \\ \sqrt{2}\rho_{13}^U & \sqrt{2}\rho_{13}^L & 1 \end{bmatrix} \quad (2.36)$$

$$R[1] = \begin{bmatrix} 0 & 2\rho_{11}^{UL} & \sqrt{2}\rho_{31}^{UL} \\ 0 & 0 & \sqrt{2}\rho_{31}^{UL} \\ 0 & 0 & 0 \end{bmatrix} \quad (2.37)$$

In the synchronous case the matrices of (Formula 2.36) and (Formula 2.37) reduce to (2.38) and (2.39).

$$R[0] = \begin{bmatrix} 1 & 0 & \sqrt{2}\rho_{13}^U \\ 0 & 1 & \sqrt{2}\rho_{13}^L \\ \sqrt{2}\rho_{13}^U & \sqrt{2}\rho_{13}^L & 1 \end{bmatrix} \quad (2.38)$$

$$R[1] = \begin{bmatrix} 0 & 0 & 0 \\ 0 & 0 & 0 \\ 0 & 0 & 0 \end{bmatrix} \quad (2.39)$$

In this chapter the modelling of the DS-CDMA system has been presented. MPG can be used as method to implement different data rates into the same system. The transmission of the signal as well as the reception of the signal were discussed. In the next chapter several types of multiuser detection receivers are presented. These receivers can be used to detect the received signal and to reproduce the transmitted signal, which were mentioned in Section 2.1.3 and Section 2.1.1, respectively.

Chapter 3

Multiuser detection receivers

Unlike FDMA and TDMA multiple access techniques, in a DS-CDMA system all users use the same bandwidth at the same time. This gives rise to Multiple Access Interference (MAI). The caused MAI deteriorates the reception of the signal. To tackle this problem Multiuser Detection (MUD) schemes have been proposed [2][3][4]. These schemes use the received signal of all users to detect the transmitted data of a single user. The conventional receiver is less complex than MUD receivers but is too simple to generate an equal performance. The optimal Multiuser Detection structure has found to be too complex to use and suboptimal techniques were considered. One class of suboptimal receivers is the class of interference cancellation receivers. In this class there is a distinction to be made: Serial Interference Cancellation (SIC) and Parallel Interference Cancellation (PIC). After the consideration of the conventional receiver, the suboptimal MUD receivers will be described in this chapter.

3.1 Conventional receiver

At the transmitters end of the system the data stream of the different users was spread by a user specific signature waveform. To separate the different users at the receiver's end the received signal is correlated with the same user specific waveform (3.1).

$$y_{k,l}^{(n)} = \int_{nT+\tau_{k,l}+\Delta\tau_k}^{(n+1)T+\tau_{k,l}+\Delta\tau_k} s_k(t - nT - \tau_{k,l} - \Delta\tau_k) r(t) dt \quad (3.1)$$

$$= s_k(T - t) * r(t) \Big|_{t=(n+1)T+\tau_{k,l}+\Delta\tau_k} \quad (3.2)$$

The signature waveform is delayed according to the time offset due to the asynchronous aspect of the system ($\Delta\tau_k$) and according to the delay of the path l of user k ($\tau_{k,l}$). The output of the integrator is sampled and is denoted by $y_{k,l}^{(n)}$.

This correlator structure is equivalent to a filter with an impulse response of $s_k(T-t)$ (Formula 3.2). This is the reason that the other name for this correlator structure is *Matched Filtering* [13].

In conventional single user detection the goal is to separate the data symbols of a user from the data symbols coming from other propagation paths from the same user and the data symbols from other users. Assumed that a system has K users, L resolvable propagation paths and an AWGN noise component $\eta(t)$, then the correlator output of user k , propagation path l and symbol interval m can be written as (3.3):

$$\begin{aligned}
 y_{k,l}^{(m)} &= A_k e^{j\phi_k} c_{k,l}^{(m)} b_k^{(m)} \\
 &+ \sum_{\substack{l'=1 \\ l' \neq l}}^L \sum_{m=0}^{M-1} A_k e^{j\phi_k} c_{k,l'}^{(m)} b_k^{(m)} \int_{-\infty}^{\infty} s_k(t-mT-\tau_{k,l}-\Delta\tau_k) s_k(t-mT-\tau_{k,l'}-\Delta\tau_k) dt \\
 &+ \sum_{\substack{k'=1 \\ k' \neq k}}^K \sum_{l=0}^L \sum_{m=0}^{M-1} A_{k'} e^{j\phi_{k'}} c_{k',l}^{(m)} b_{k'}^{(m)} \int_{-\infty}^{\infty} s_{k'}(t-mT-\tau_{k',l}-\Delta\tau_{k'}) s_k(t-mT-\tau_{k,l}-\Delta\tau_k) dt \\
 &+ \int_{-\infty}^{\infty} \eta(t) s_k(t-mT-\tau_{k,l}-\Delta\tau_k) dt
 \end{aligned} \tag{3.3}$$

$$= A_k e^{j\phi_k} c_{k,l}^{(m)} b_k^{(m)} + \Psi_{k,l}^{(m)} + w_{k,l}^{(m)} \tag{3.4}$$

The output of the Matched Filter consists of four components. The first component contains the *information* of the transmitted data symbol $b_k^{(m)}$. The second component contains the influence of the other fading paths of user k . This component can be seen as the *multipath interference* component. The third component reflects the influence of other users in the DS-CDMA system. This component is the *multiple access interference* (MAI) component. Finally the last term in (3.3) contains the AWGN, which is also correlated with the signature waveform of user k . These four components are written in a less extended form in Formula (3.3). In this formula the component $\Psi_{k,l}^{(m)}$ denotes the multipath interference and the multiple access interference. The Matched Filter, having separated all the L resolvable paths, yields a vector, on symbol interval n , described as:

$$\mathbf{y}_k^{(n)} = (y_{k,1}^{(n)}, y_{k,2}^{(n)}, \dots, y_{k,L}^{(n)})^T \in \mathbb{C}^L \tag{3.5}$$

While detecting the data symbols, advantage can be taken of the L th order diversity in the output of the Matched Filter. This advantage can be taken by combining the output vector components in a coherent way. The best performance is achieved by using *Maximal Ratio Combining* (MRC). This method adds each of the Matched Filter outputs after it has been weighted by the corresponding channel coefficient. The MRC combiner relies more on the outputs that have more strength. The signals with the least corruption by the channel have the largest channel coefficients and thus have the most influence on the combined signal. This explains intuitively the optimum performance of this method. The MRC structure is depicted in Figure 3.1 and is here referred to as a RAKE Matched Filter.

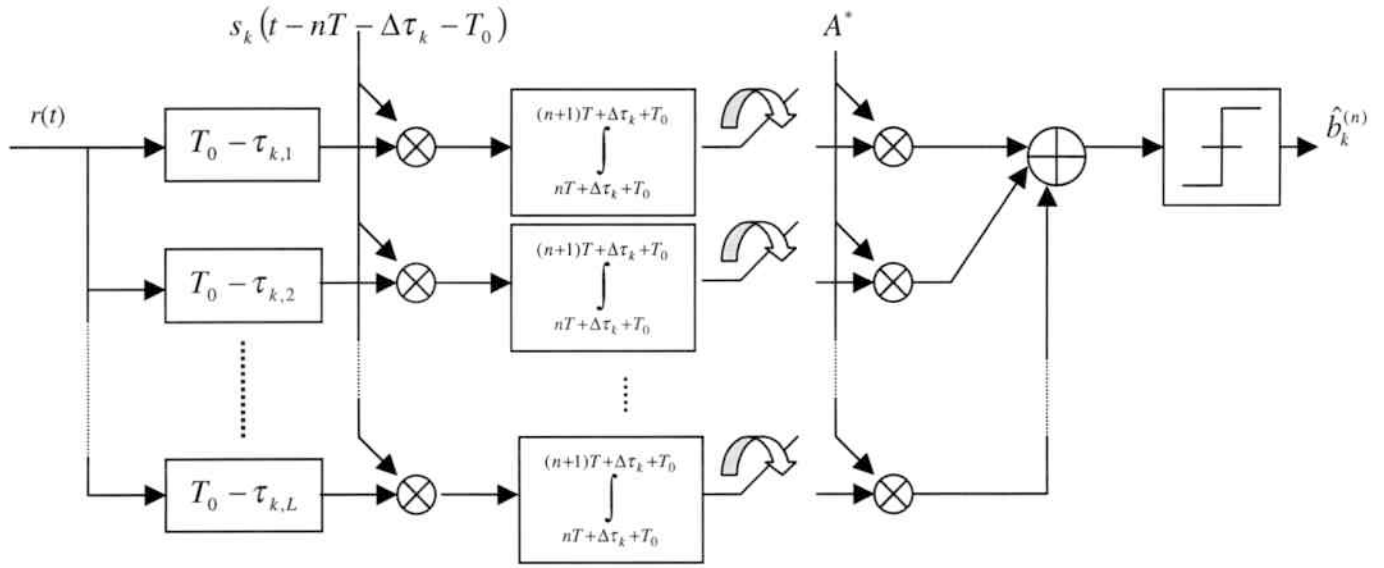


Figure 3.1: RAKE correlator for L resolvable paths and for one user k .

The symbol is detected using hard decisions. These decisions are made by the signum function, which is defined, as [10]:

$$\text{sgn}(x) = \begin{cases} 1 & x > 0 \\ 0 & x = 0 \\ -1 & x < 0 \end{cases} \quad (3.6)$$

$$\hat{b}_{k,l}^{(n)} = \text{sgn} \left[\Re \left\{ \mathbf{c}_k^{H_{(n)}} \mathbf{y}_k^{(n)} \right\} \right] \quad (3.7)$$

In matrix notation $\mathbf{y}$ becomes:

$$\mathbf{y} = \mathbf{S}^T \mathbf{S} \mathbf{C} \mathbf{A} \mathbf{b} + \mathbf{S}^T \mathbf{n} = \mathbf{R} \mathbf{C} \mathbf{A} \mathbf{b} + \mathbf{w} \quad (3.8)$$

The multiple access interference may have crucial effects on the performance of this conventional receiver. Due to the interference of strong users this receiver is less able to detect weak (low energy) users. The level of interference may be significant even with low crosscorrelations between the signature waveforms. When a transmitter is close to the receiver the signal coming from this transmitter is much stronger than a signal coming from a transmitter that is further away. The detection of a signal coming from a large distance is worse than the detection of signal coming from a transmitter that is near by. This is called the Near-far effect. To tackle this problem and other effects that may cause bad detection subtractive single rate receivers were developed.

3.2 Suboptimal Subtractive receivers

Optimal multiuser detectors were found too complex [14], which lead to a consideration of different suboptimal nonlinear techniques. All these techniques use some form of interference cancellation, which can be illustrated by writing the Matched Filter output in the form [8]:

$$y_{k,l}^{(n)} = r_{k,l}^{(n)} + \Psi_{k,l}^{(n)} + w_{k,l}^{(n)} \quad (3.9)$$

In this formula $r_{k,l}^{(n)}$ represents the desired term consisting of the information about the transmitted symbol, $\Psi_{k,l}^{(n)}$ the interference component, consisting of the multiple access interference as well as the multipath interference, and $w_{k,l}^{(n)}$ represents the noise factor. If the detector is able to estimate the interference component $\Psi_{k,l}^{(n)}$, $r_{k,l}^{(n)}$ can be estimated by canceling this interference component. This is denoted in (Formula 3.10).

$$\hat{r}_{k,l}^{(n)} = y_{k,l}^{(n)} - \hat{\Psi}_{k,l}^{(n)} \quad (3.10)$$

The transmitted symbol can be recovered using the estimate $\hat{r}_{k,l}^{(n)}$. Since the interference component of the other users and the other transmission paths is cancelled, the decision based upon the estimate of the desired term $\hat{r}_{k,l}^{(n)}$ will be more reliable. The system performance improves when interference cancellation is used.

Interference cancellation can be done in several ways. Two of the receivers belonging to the class of interference cancellation receivers have received the most attention. These two receivers are the Successive Interference Cancellation (SIC) receiver and the Parallel Interference Cancellation (PIC) receiver. These receivers are considered in Section 3.2.1 and Section 3.2.2, respectively. Section 3.2.3 describes Groupwise Weighting techniques for the PIC receiver.

3.2.1 Successive Interference Cancellation

In the group of Successive Interference Cancellation receivers the estimate of the interference is cancelled in a serial way. The order of the cancellation of the interference is from the strongest user to the weakest user. This order is chosen in such a way because the strongest user can be detected with most reliability and are likely to cause significant interference to the weaker users. After the detection of the strongest user and the cancellation of the interference the second strongest user is detected. This process continues until there are no more users left to detect. It is clear that a ranking from the strongest user to the weakest user is needed. In fading channels the ranking has to be done on-line because the received powers of the users vary from symbol to symbol.

To be able to cancel the interference of the weaker user k the symbols $(b_{k-1}^{(n-2)}, \dots, b_{k-1}^{(n+2)})$ are needed. To detect the symbols of the user $k-1$ the symbols $(b_{k-2}^{(n-4)}, \dots, b_{k-2}^{(n+4)})$ are required. This is due to the asynchronous transmission of the symbols and to the multipath spread. From this it can be concluded that the detection of a single symbol of user k requires knowledge of the received signal over a period that is proportional to the users stronger than k . This can cause a delay that is too long for practical use, especially when there are a lot of users.

One way of solving the problem of the long decision delay is to process the received signals in groups [15]. Instead of detecting the received signal user by user some signals are taken together and are treated as a group. The groups can be formed by selecting the received signals based on their powers. First the group with the strongest users is detected and their interference is cancelled from the other users. For example in single rate systems a group size of two is found to influence the performance only marginally. By taking only two users in a group the complexity of the system and with this the decision delay is severely decreased [15]. There is a trade-off between simplicity of the system on the one side and the decrease in performance on the other side. Receivers that detect the users and cancel interference serially and in a groupwise way are called Groupwise Successive Interference Cancellation (GSIC) receivers. The basic GSIC receiver is depicted in Figure 3.2.

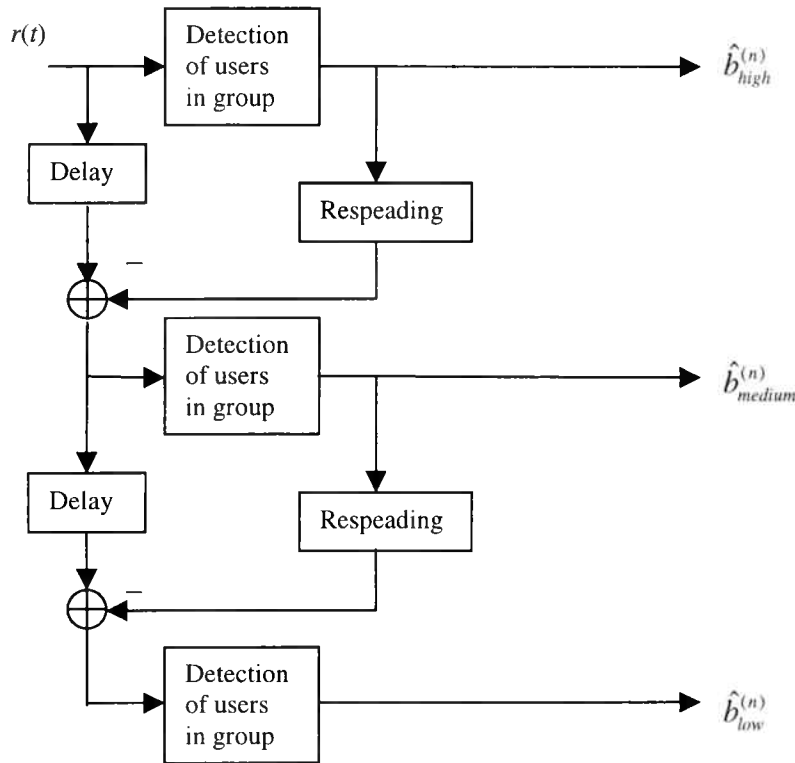


Figure 3.2: Basic Groupwise Successive Interference Cancellation [6].

3.2.2 Parallel Interference Cancellation

The Parallel Interference Cancellation (PIC) receiver estimates and subtracts the multiple access interference (MAI) in a parallel way. The cancellation can be done directly on the received signal, before Matched Filtering (wide band domain or chip level) or on the symbols after detection with Matched Filtering (narrow band domain or symbol level). Both methods are analytically equal [8] and the chip level approach is followed during implementation. In PIC not only the present symbols but also the two previous symbols and the two coming symbols can cause interference with a symbol. This is due to the asynchronous transmission of the symbols and to the multipath delays. Therefore a *processing window* with a large size, consisting of the bits $(b^{(n-2)}, \dots, b^{(n+2)})$, is used. All these symbols are stored in a register in order to be able to make a good estimate of the MAI.

To obtain the initial estimates of the signal Matched Filtering can be used. A delay from three symbol intervals is necessary to make correct Matched Filter decisions. Again it is necessary to store these decisions in a register and when this register contains five symbols a first PIC stage can be applied.

The decisions coming from the first PIC stage are again stored in the register. If the register has seven symbols the second PIC stage can be applied and so on. To be able to apply the i th PIC stage, the register has to store $3+2*i$ symbols.

3.2.3 Groupwise Weighted Parallel Interference Cancellation

In the previous two sections attempts were made to cancel the estimated interference *completely* from the received signal in order to clear this signal from interference. In some cases the interference estimate can be very poor and instead of clearing the signal from interference, interference is added to the received signal. After interference cancellation the detection might get worse by using poor estimates of the MAI than before interference cancellation. In order to tackle this problem the estimated interference can be *partially* cancelled by applying *weighting factors* [16][17][18]. The reliability of the MAI estimates is judged. From this judgement a weighting factor deciding the amount of power taken into the cancellation step can be determined. Often the reliability of the MAI estimate is higher if the signal has more power. It would therefore be logical to appoint weighting factors according to the power level of the received symbols. When these weighting factors are integrated in Parallel Interference Cancellation (PIC) receivers, the performance of PIC receivers can be improved. These receivers are known as Weighted Parallel Interference Cancellation (WPIC) receivers [8].

The i th interference cancellation stage produces preliminary decisions, which may be expressed in a matrix form as:

$$\hat{\mathbf{b}}(i) = \text{sgn} [\Re \{ \mathbf{C}^H \hat{\mathbf{y}}(i) \}] \quad (3.11)$$

In where sgn is the signum function as defined in (Formula 3.6) and $\hat{\mathbf{y}}(i)$ is the vector that represents the interference cancelled output of stage i . Because the initial stage uses only Matched Filtering and no cancellation has been done yet, $\hat{\mathbf{y}}(i)$ is set to $\mathbf{h}$ for the initial stage ($i=0$). For subsequent stages $\hat{\mathbf{y}}(i)$ can be written as:

$$\hat{\mathbf{y}}(i) = \mathbf{W}_i (\mathbf{h} - \hat{\Psi}(i)) + (\mathbf{I} - \mathbf{W}_i) \hat{\mathbf{y}}(i-1) \quad (3.12)$$

The interference estimate used by the i th stage is based on the decisions of the previous stage $i-1$ and may be written in the form:

$$\hat{\Psi}(i) = (\mathbf{R} - \mathbf{I})\mathbf{C}\mathbf{A}\hat{\mathbf{b}}(i-1) \quad (3.13)$$

The diagonal matrices $\mathbf{W}_i$, $i=1, \dots, I$, are the weight matrices that define the amount of cancellation used at stage i . These matrices consist of weight factors that may vary from symbol to symbol and have values ranging from zero (no part in the cancellation) to one (a full part in the cancellation). A closer examination of a multistage WPIC receiver reveals that the outputs of the first stage (3.14) and second stage (3.15) are [8]:

$$\begin{aligned} \hat{\mathbf{y}}(1) &= \mathbf{W}_1(\mathbf{h} - \hat{\Psi}(1)) + (\mathbf{I} - \mathbf{W}_1)\mathbf{h} \\ &= \mathbf{h} - \mathbf{W}_1\hat{\Psi}(1) \end{aligned} \quad (3.14)$$

$$\begin{aligned} \hat{\mathbf{y}}(2) &= \mathbf{h} - \mathbf{W}_1\hat{\Psi}(1) - \mathbf{W}_2(\hat{\Psi}(2) - \mathbf{W}_1\hat{\Psi}(1)) \\ &= \hat{\mathbf{y}}(1) - \mathbf{W}_2(\hat{\Psi}(2) - \mathbf{W}_1\hat{\Psi}(1)) \end{aligned} \quad (3.15)$$

Every interference cancellation stage cancels from the output of the previous stage (partially) an amount of interference, which is obtained as a difference between the new interference estimate and the interference which the previous stages have already cancelled. If the weight matrices are identity matrices the WPIC scheme reduces to a PIC scheme.

The WPIC receiver is drawn in Figure 3.3. In a single rate environment as well as in a multirate environment the power of a user in the received signal is decisive for the quality of the MAI estimate of this user. A high power user will have a better MAI estimate than a user with less power. In a multirate environment, the power differences between the different users with different data rates are big and between users with the same data rate are small. This is the reason that weighting factors could be chosen in a *Groupwise* way. This means that all the users belonging to a group of users with the same data rate, thus with about the same power, will have the same weighting factor. By implementing these patterns in a WPIC receiver, this receiver will become a *GW-PIC* (Groupwise Weighted Parallel Interference Cancellation) receiver [17].

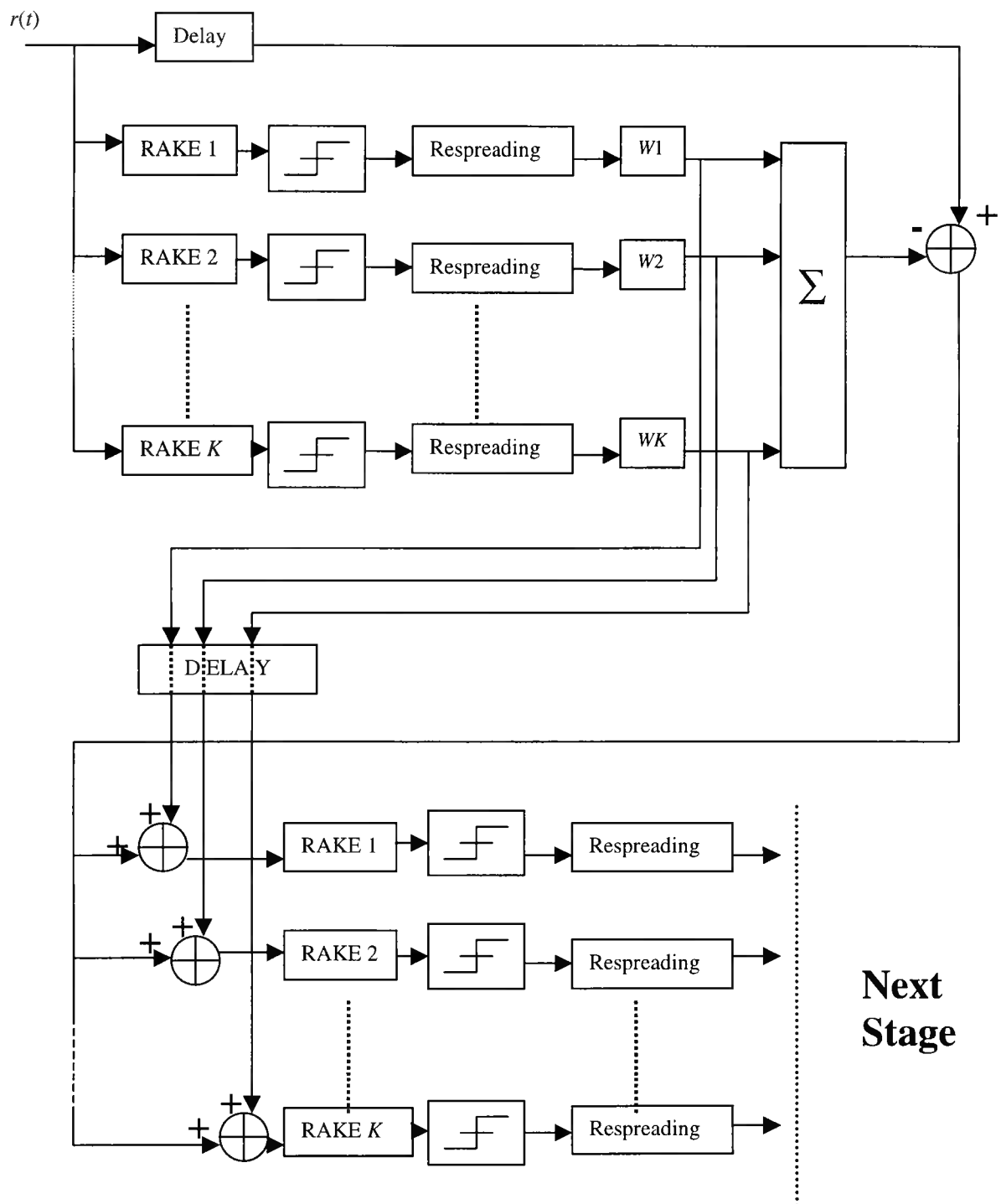


Figure 3.3: i th stage of a WPIC receiver [17].

Chapter 4

Performance of multirate receivers in AWGN channels.

In the study of unknown receivers it is custom to start the simulation in a less complex environment. In this case this environment is the additive white Gaussian noise channel or also known as AWGN. Simulating receivers in such a simple environment enables us to understand the basic performance relationships of these receivers. After conclusions are drawn on the basic performance relationships, the receiver will be placed in a more complex environment, for example a fading channel (Chapter 5). The performance of the different receivers will be evaluated using the Monte Carlo principle [19].

This chapter begins with the evaluation of the receivers in a single rate system. In this most simple system the performance of a receiver will be evaluated while changing the number of users. The results of this are shown in Section 4.2. After this part the receiver is evaluated in a multirate environment. Section 4.3 describes the results of a study of the influence of weighting factors in such an environment. The best weighting factors will be those where the BER performance of the GW-PIC receiver is best. After selecting the weighting factors the GW-PIC receiver will be compared with other receivers in a multi rate environment. Remarks will be made and conclusions will be drawn at the end of this chapter in Section 4.4.

4.1 Simulation Parameters

Although we do all simulations in an AWGN channel it is still important that the right parameters are chosen and remain unchanged during the simulations. If this is not the case then a good comparison to other receivers will not be possible. One of the most important parameters is the selected Gold code. These codes consist of randomly chosen Gold sequences of length 31 with one bit randomly chosen and added to make the total length of the used codes 32. These codes are used in the simulations to (re)spread the signal.

Once the codes are selected to (re)spread the signal, the chips are sampled with four samples per chip and 32 chips per symbol for the lowest bit rate user, 16 chips per symbol for the medium bit rate user and 8 chips per symbol for the highest bit rate user. Since the simulations in Section 4.1 are done with only low bit rate users, the chips per symbol for these users in this simulation will be 32.

In Section 4.2 there are 4 low bit rate users, 4 medium bit rate users and 2 high bit rate users considered. This means that 32 chips per symbol for the low bit rate users, 16 chips per symbol for the medium bit rate users and 8 chips per symbol for the high bit rate users are used.

One problem that has to be tackled in a digital telecommunication system is Inter Symbol Interference. This is caused by improper filtering as pulses pass through the communication channel. The pulses will be spread in time, which results in the smearing into adjacent time slots; this is depicted in figure 4.1.

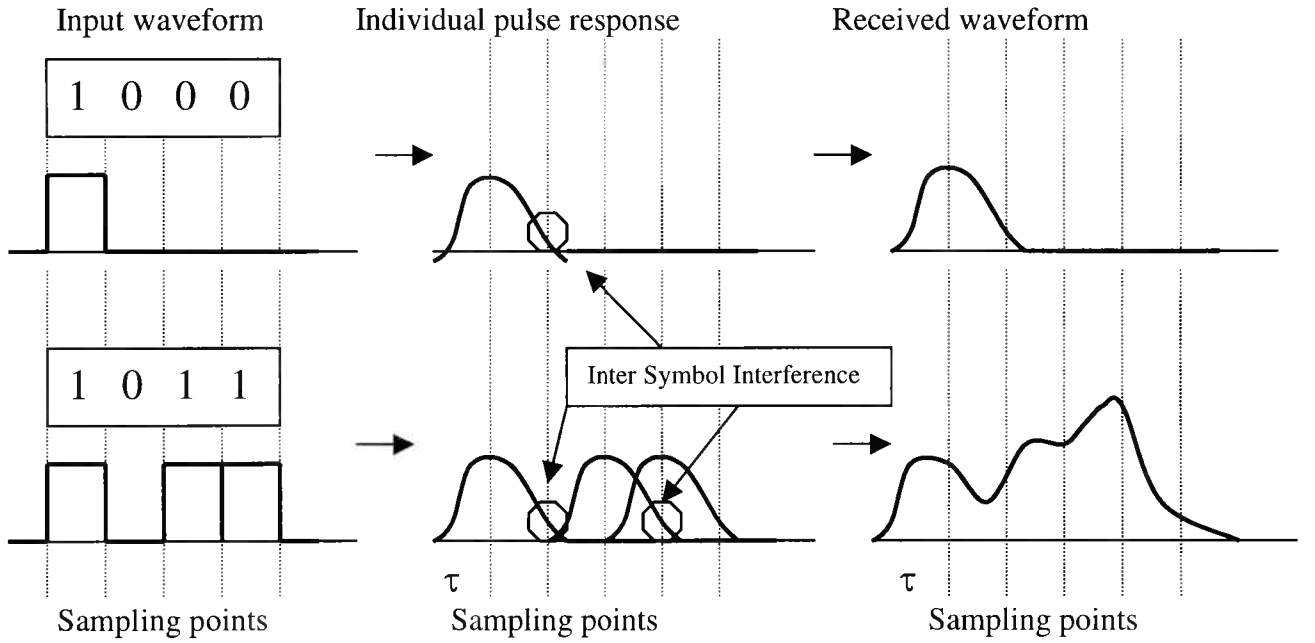


Figure 4.1: Examples of ISI on received pulses in a binary communication System[20]

As can be seen in Figure 4.1 there is an overlap of the pulses at the receiver's end of the system. This overlap causes an added interference, which makes it more difficult to extract the transmitted signal. To tackle this problem it is important to choose a frequency response that limits the ISI and is physical realizable. The idea is to find pulse shapes that go through zero at adjacent sampling points and yet have an envelope that decays much faster than $1/x$ [20]. A filter that satisfies these conditions is the raised cosine filter. The raised cosine filter has the following frequency response:

$$X_{rc}(f) = \begin{cases} T & \left(0 \leq |f| \leq \frac{1-\beta}{2T}\right) \\ \frac{T}{2} \left\{ 1 + \cos \left[\frac{\pi T}{\beta} \left(|f| - \frac{1-\beta}{2T} \right) \right] \right\} & \left(\frac{1-\beta}{2T} \leq |f| \leq \frac{1+\beta}{2T} \right) \\ 0 & \left(|f| \geq \frac{1+\beta}{2T} \right) \end{cases} \quad (4.1)$$

In these simulations the roll-off factor β has been chosen to be: $\beta = 0.75$.

The pulse form of this raised cosine pulse is depicted in Figure 4.2.

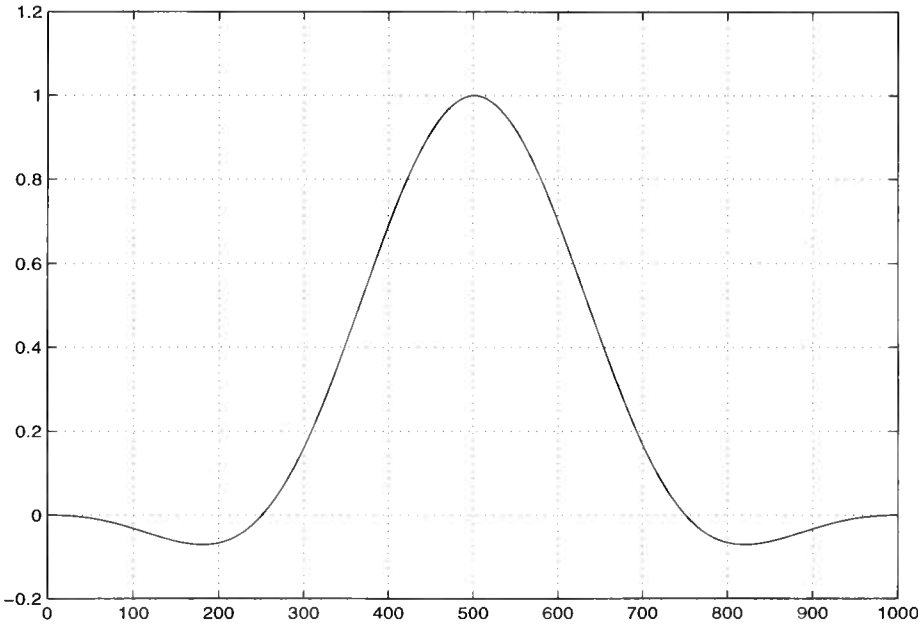


Figure 4.2: 1000 samples of the Raised cosine pulse form ($\beta = 0.75$).

The parameters mentioned in this section are summarized in Table 4.1.

Transmission mode	Asynchronous
Spreading modulation	BPSK
Processing gains	32,16,8
Basic Code length	32
Basic data rate	16 kbits/s
Code family	Gold
Number of samples/chip	4
Chip waveform/Roll-off factor	Root-raised-cosine/0.75
Channel model	AWGN
Samples per chip	4
Symbol block	500
Channel estimation	Assumed known

Table 4.1: Simulation parameters for AWGN channels

4.2 Performance in a single rate system

In this section the performance of a Matched Filter and a PIC receiver in a single rate system is investigated. In this particular system we consider only low-rate users and no medium-rate users or high-rate users. The results of this simulation in this system are depicted in Figure 4.3.

It is clear that a PIC receiver performs much better than Matched Filtering, especially for higher signal-to-noise ratios. The performance of the PIC receiver as well as the Matched Filter worsens when the number of users increases. The reason for this is that the First Estimate gets worse because of the lack of perpendicularity in the Gold codes used here. The Gold codes do not have an infinite length thus the cross-correlation between them becomes larger when we have a lot of users using these codes. This means that the signal sent over the channel will suffer from AWGN as well as the lack of cross-correlation between the codes. The First Estimate influences the PIC receiver in all the stages, which results in the PIC performance getting worse when the number of users increases.

A way to decrease the influence of the First Estimate is to weigh the influence of the First Estimate. This means that we take only a part of the First Estimate into account because, as Figure 4.3 depicted, this estimate is not always reliable [16] [18]. As a first test the weighting factors for all low rate users are chosen to be 0.4. Doing this we see that we get better results with the same AWGN channel and the same number of users. This is depicted in Figure 4.4. The conclusion that can be drawn from these results is that in a single rate system there is an overall amelioration of the receiver if weighting factors are applied.

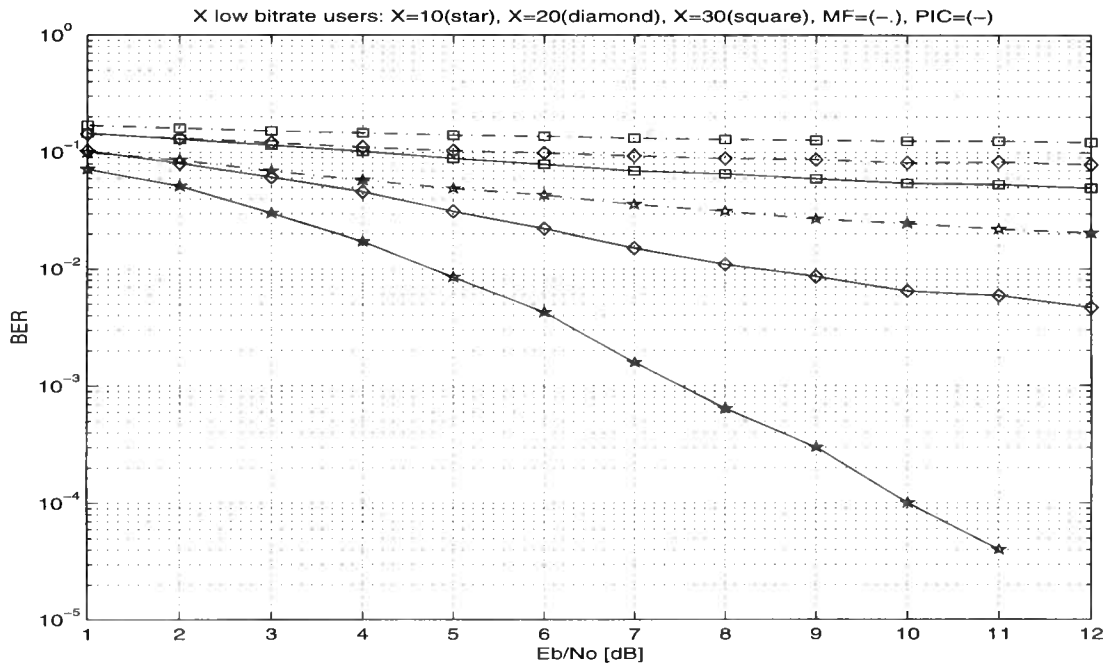


Figure 4.3: Bit error rates for a 2 stage PIC receiver and Matched Filtering with different number of users.

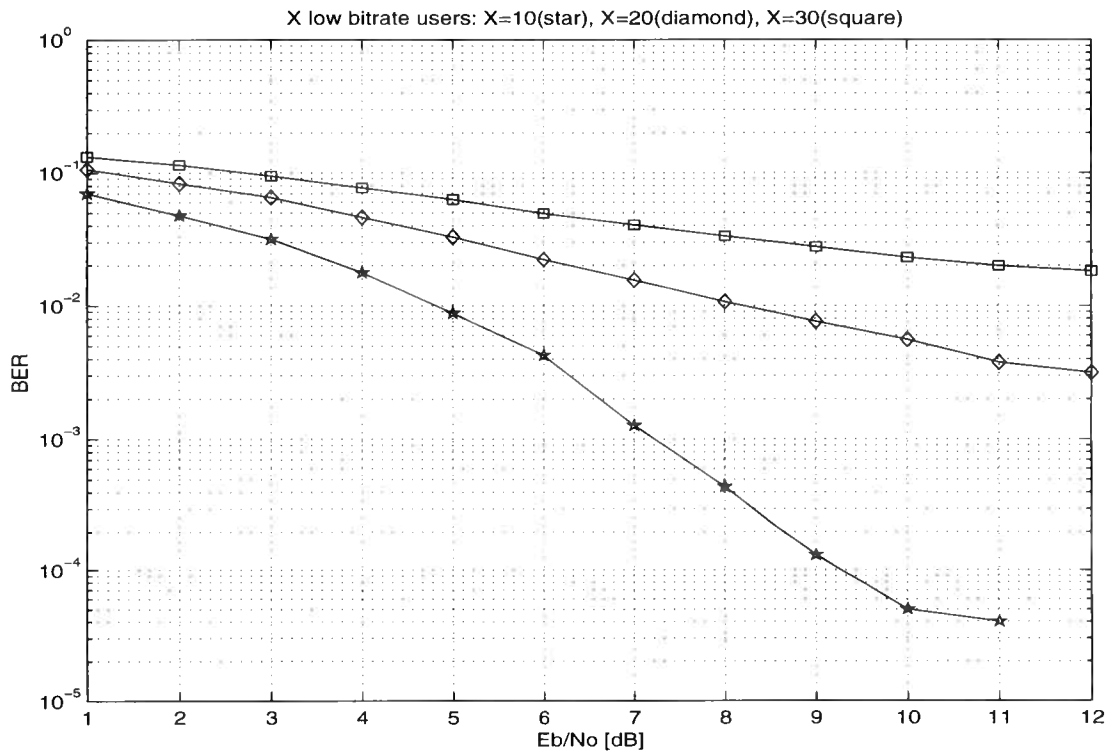


Figure 4.4: Bit error rates for a 2 stage weighted PIC receiver with a weighting factor of 0.4 in the first stage and one in the second stage with different number of users.

4.3 Performance in a multi rate system

In this section the GW-PIC is placed in a multirate environment. In this environment there are 4 low bit rate users, 4 medium bit rate users and 2 high bit rate users. All users have equal received energies, but between these users there are natural power differences. These power differences are caused by the different rate of changes in energy depending from the bit rate. The power of the highest bit rate user is the largest.

4.3.1 The performance of a Matched Filter and a PIC receiver in a multi rate system

This section begins with the study of the performance of a normal PIC receiver and a Matched Filter. The different bit error rates are plotted in Figure 4.5.

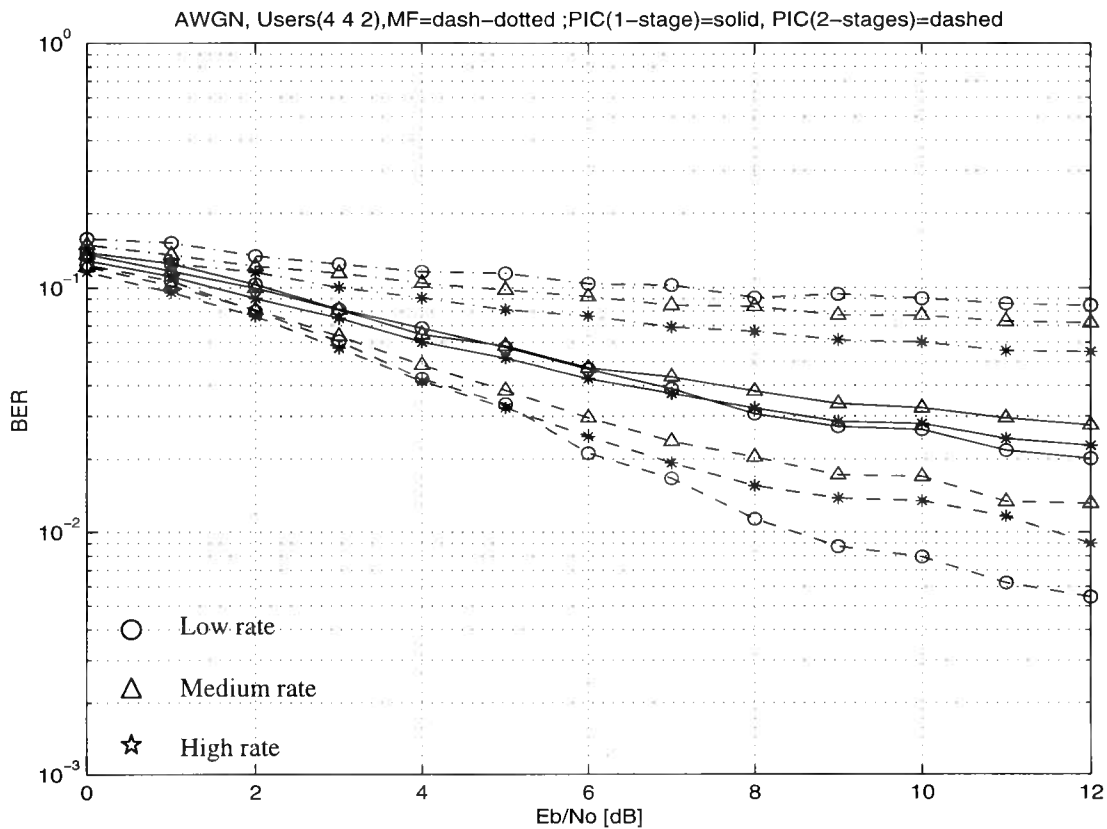


Figure 4.5: Bit error rates versus signal to noise ratio for different users with different data rates. The three graphs represent the BER after Matched Filtering (dash-dotted), after a 1 stage PIC detector (solid) and after a 2 stages PIC receiver (dashed).

It can be seen from the graphs that the PIC gives a better result than the Matched Filter. The reason for this is that a PIC detector makes use of the Matched Filter output, which is called the First Estimate for the PIC receiver. This First Estimate is being subtracted from the received signal. Doing this we clean the received signal of Multiple Access Interference (MAI). Hereafter the PIC receiver uses the Matched Filter another time to make an estimate of the transmitted signal. Because of the power differences between the groups and because of the fact that the more power a received signal has got the better it can be detected using Matched Filtering, there is a difference in performance between groups. It can be seen from the upper three graphs in Figure 4.5 that the highest bit rate user performs better than the other two and that the medium rate bit user performs better than the lowest bit rate user after Matched Filtering.

Since the PIC receiver uses Matched Filtering to make a First Estimate, the highest bit rate users have the best First Estimate followed by the medium bit rate users and finally the low bit rate users with the worst First Estimate. At the same time the estimate for the MAI will be better for the low bit rate users and medium rate users, because they make use of the good First Estimates of the highest bit rate users. Since the medium bit rate users have a better First Estimate than the low bit rate users; the best estimate for the MAI is made for the low bit rate users. This means that the high bit rate users have the advantage of having the best First Estimate and the low bit rate users the advantage of the best estimate for the MAI.

In the single rate system (Section 4.2) there was an improvement in performance when weighting factors were applied. In a multirate system there is also an improvement of the BER performance of the PIC receiver when weighting factors are applied. Simulations have been done at a signal to noise ratio of 8 dB. Keeping the signal to noise ratio at a fixed level the weighting factors in the first stage of the GW-PIC were varying from zero (only Matched Filtering after one stage) to one (Full PIC stage) and were equal for all users. The weighting factors of the second stage were set to one for all users. From Figure 4.6 it can be seen that there is an amelioration in detection of all users when certain weighting factors are applied. The influence of weighting factors in an multirate AWGN environment is striking and is interesting enough for a further investigation. This investigation will be described in the next section.

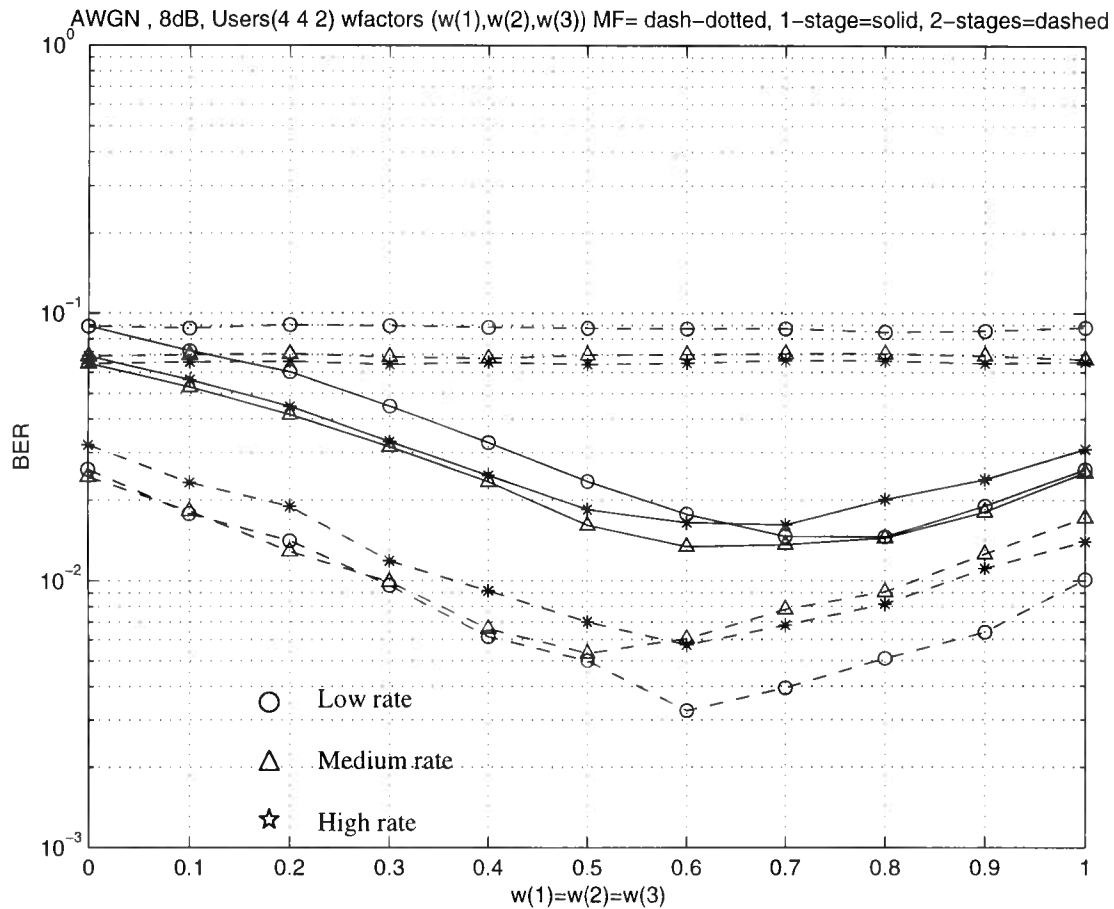


Figure 4.6: Bit error rates versus equal weighting factors in the first stage of the GW-PIC receiver for different users with different data rates and with a signal to noise ratio of 8 dB. The equal weighting factors in the first stage of the GW-PIC receiver ranging from zero to one. The weighting factors in the second stage were set to unity (full PIC).

4.3.2 The influence of weighting factors on the performance of low, medium and high bit rate users

To study the influence of weighting factors on the performance of GroupWise Weighted PIC receivers several conditions have to be respected. A first condition is to study these receivers after only one stage. The reason for this is that the influence of the weighting factors can be seen directly after one stage without a possible change in the results from the influence of a next stage. A second condition to compare the performance of the GroupWise Weighted PIC receiver is to keep the sum of the three weighting factors equal to a constant. When the weights are changed this constant should be unchanged so that a comparison can be made between different receivers. This constant represents the amount of energy that is used in the cancellation step.

When the full PIC receiver is taken for an example than this constant will be equal to three. Cancellation will be done in this case with all of the power of the First Estimates from the high bit rate users, the medium bit rate users and the low bit rate users. The weighting factors for a full PIC receiver will be 1 for the high bit rate users as well as the medium bit rate users and the low bit rate users. Since the First Estimates from the users are expected to be less accurate than later estimates it is wise to choose the weighting factors between zero and one. Choosing a weighting factor for a group equal to zero means that the estimate for this group of users is not taken in account in the following cancellation step. Choosing it one for a group of users means that the total power from this estimate is taken in account in the following cancellation step and a full cancellation will be effected. A summary of these conditions is depicted in Table 4.2.

Weighting factor high bit rate user	$0 \leq W_1^2 \leq 1$
Weighting factor medium bit rate user	$0 \leq W_2^2 \leq 1$
Weighting factor low bit rate user	$0 \leq W_3^2 \leq 1$
Sum of the Weighting factors	$W_1^2 + W_2^2 + W_3^2 = C \quad C = 1.8$

Table 4.2: Conditions for the different weighting factors.

The weighting factors W_1 , W_2 and W_3 represent the fraction of the **amplitude** that is taken into account in the cancellation step for the high bit rate users, the medium bit rate users and the low bit rate users, respectively.

The weighting factors are squared because the comparison of the different receivers with the different weighting factors can only be done correctly if the same fraction of **energy** is used in the first stage. The sum of these squared weighting factors has been chosen to be 1.8. The constant is set to 1.8 because there are a lot of combinations possible to obtain this number and it represents a realistic reliability of the correctness of the First Estimates.

The study of the influence of weighting factors begins with giving equal weighting factors to the different users. Since the squared sum of the weighting factors has to be equal to 1.8, they are equal to $\sqrt{0.6}$ for each user. This is depicted in Table 4.3.

Weighting factor high bit rate user	$W_1 = \sqrt{0.6}$
Weighting factor medium bit rate user	$W_2 = \sqrt{0.6}$
Weighting factor low bit rate user	$W_3 = \sqrt{0.6}$

Table 4.3: Weighting factors for equal used parts of the First Estimate of the three different groups

The results of this simulation are shown in Figure 4.7. This figure emphasizes that with equal weighting factors of $\sqrt{0.6}$, the overall performance of the receiver compared to a PIC receiver (Figure 4.5) could be ameliorated.

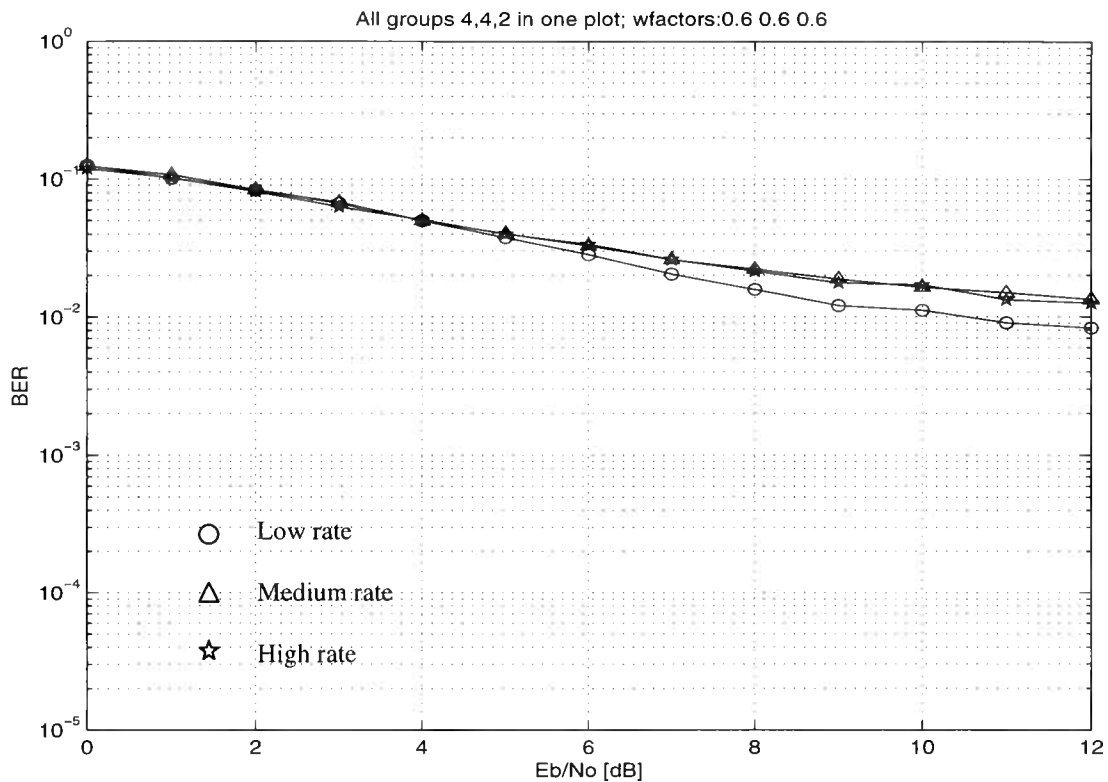


Figure 4.7: Bit error rates versus signal to noise ratio for 2 high bit rate users, 4 medium bit rate users and 4 low bit rate users.

$$W_1 = W_2 = W_3 = \sqrt{0.6}.$$

There are three weighting factors for a three rate system like this. This means that there is a degree of freedom of three. To comprehend the influence of these three weighting factors on the performance of the GroupWise Weighted PIC receiver the degree of freedom should be brought back to two. To achieve this, one weighting factor has to remain constant and the other two have to vary, while the sum of the squared weighting factors has to remain equal to 1.8. The following simulations are done with the parameters shown in Table 4.4.

Weighting factor high bit rate user	$W_1^2 = 0.8$
Weighting factor medium bit rate user	$W_2^2 = X$
Weighting factor low bit rate user	$W_3^2 = Y$
Sum of the Weighting factors	$W_1^2 + X + Y = C \quad C = 1.8$

Table 4.4: Conditions for the different weighting factors to limit the degree of freedom in the simulations.

As can be seen in Table 4.4 the weighting factors of the low bit rate users and of the medium bit rate users are varying, whereas the weighting factors of the high bit rate users is set to a constant. This constant is equal to 0.8 and the squared sum of the three weighting factors remains 1.8.

The first simulation results come from the case where the weighting factor of the medium bit rate users is equal to zero and the weighting factors of the low bit rate users is equal to 1 (Table 4.5).

$X = 0$
$Y = 1$

Table 4.5: Weighting factors of the low bit rate users set to unity and weighting factors of the medium bit rate users are set to zero.

In this special case where the First Estimate of the medium rate users is not taken into account in the cancellation step, all users are equally well detected (Figure 4.8). All users are worse detected than with a normal (full) PIC detector after one stage. This result is the outcome of the not used First Estimates of the medium bit rate users. In this special case the low bit rate users as well as the high bit rate users suffer from a worse estimate for the MAI, because the First Estimate from the medium bit rate users is left out of consideration in the cancellation step.

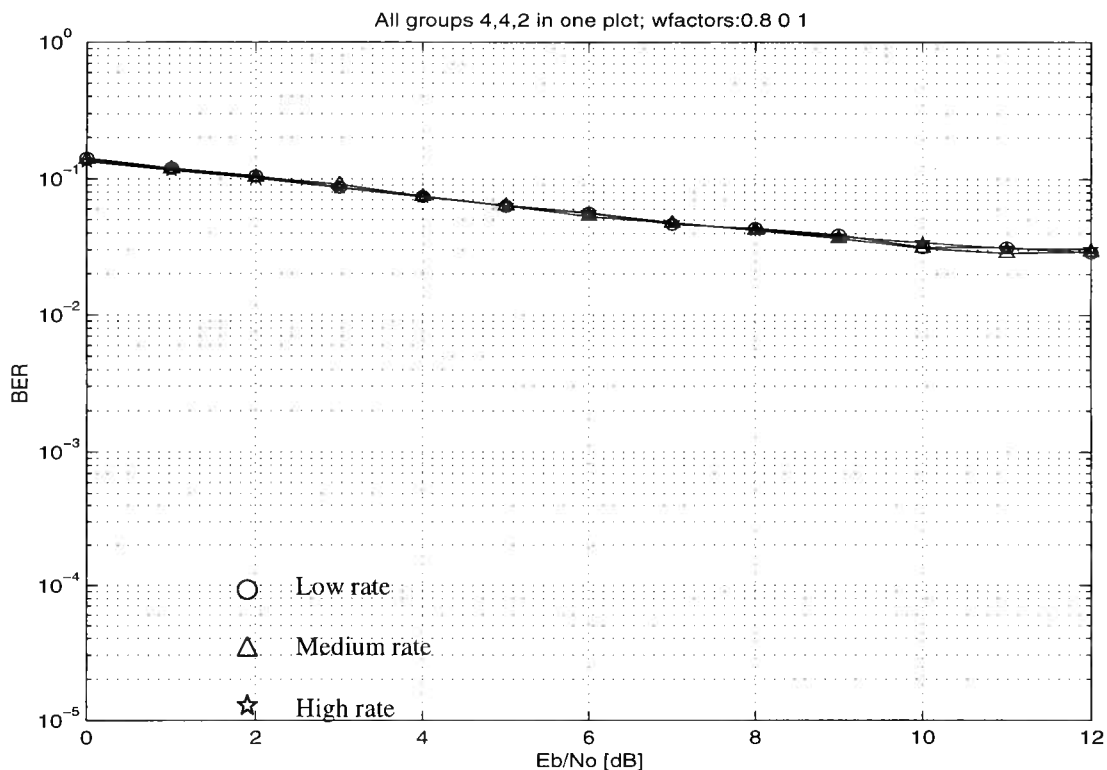


Figure 4.8: Bit error rates versus signal to noise ratio for 2 high bit rate users, 4 medium bit rate users and 4 low bit rate users.

$$W_1 = \sqrt{0.8}, W_2 = 0, W_3 = 1.$$

From Figure 4.8 can be concluded that setting a weighting factor equal to zero has not the best outcome, because information about one of the groups after Matched Filtering is not taken into account.

The next simulations are done increasing the weight of the medium bit rate users (W_2) while keeping the square of weighting factor of the high bit rate users equal to 0.8. This is done with the purpose to evaluate the effect of changing the role of the First Estimates from the low bit rate users from a big role in the cancellation step to a small role in the cancellation step. The next three evaluations have the weighting factors shown in Table 4.6, Table 4.7 and Table 4.8.

$X = 0.2$
$Y = 0.8$

Table 4.6: Squared weighting factors of the low bit rate users set to 0.8 and the squared weighting factors of the medium bit rate users are set to 0.2.

The results of the runs with this set of weighting factors are shown in Figure 4.9.

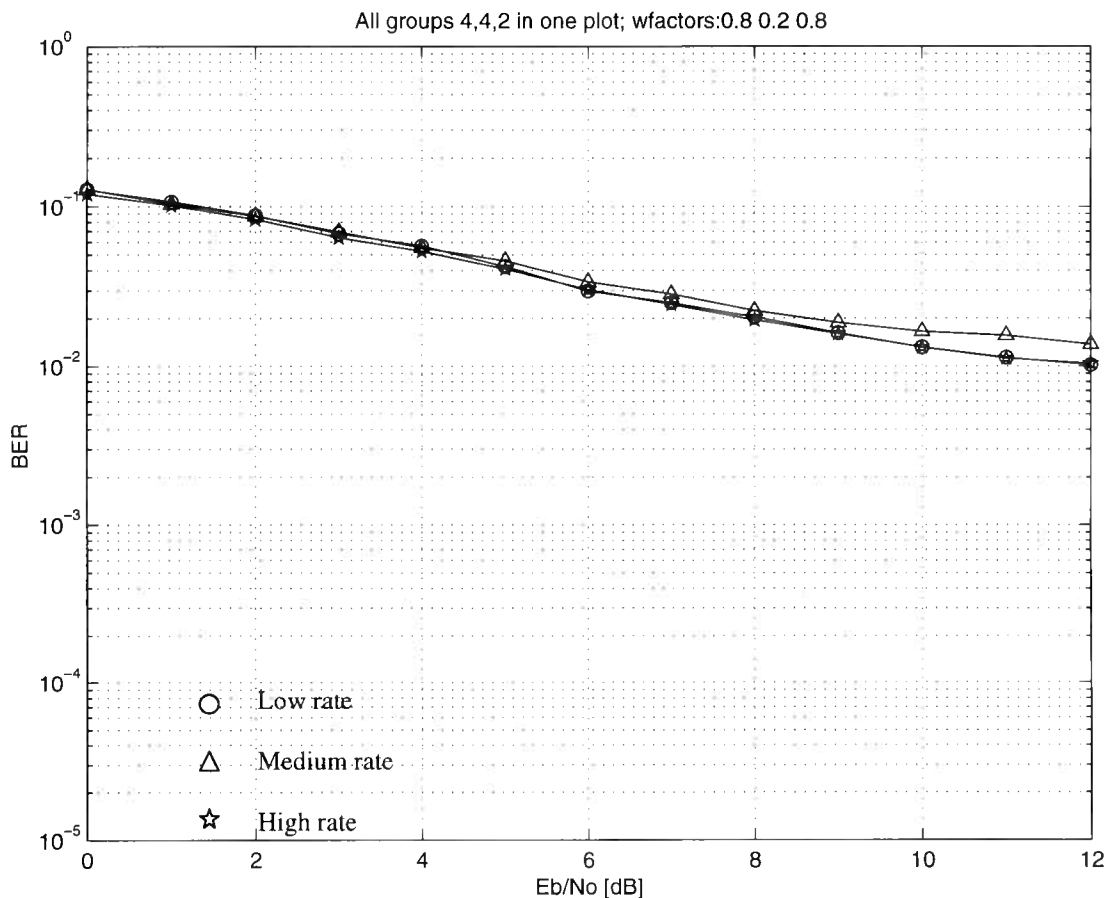


Figure 4.9: Bit error rates versus signal to noise ratio for 2 high bit rate users, 4 medium bit rate users and 4 low bit rate users.

$$W_1 = \sqrt{0.8}, W_2 = \sqrt{0.2}, W_3 = \sqrt{0.8}.$$

$X = 0.6$
$Y = 0.4$

Table 4.7: Squared weighting factors of the low bit rate users set to 0.4 and the squared weighting factors of the medium bit rate users are set to 0.6.

Results of these runs are depicted in Figure 4.10.

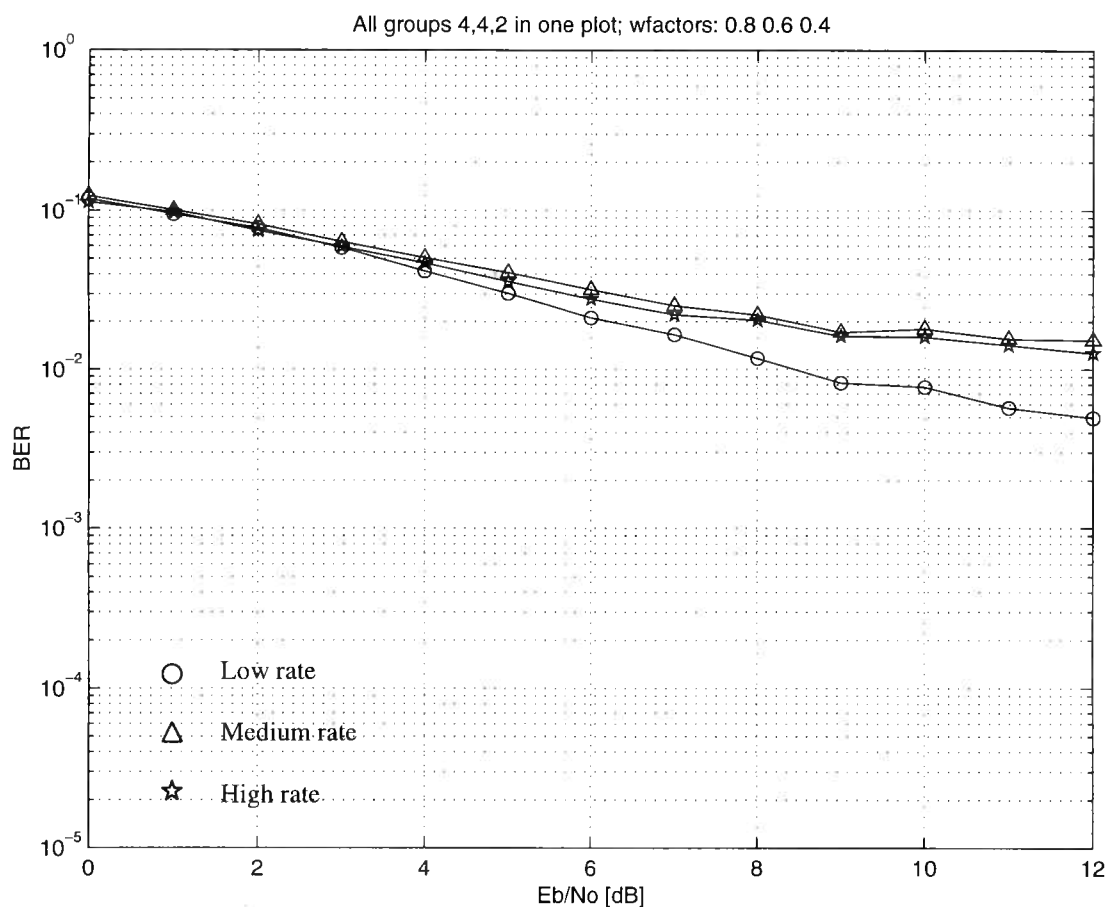


Figure 4.10: Bit error rates versus signal to noise ratio for 2 high bit rate users, 4 medium bit rate users and 4 low bit rate users.

$$W_1 = \sqrt{0.8}, W_2 = \sqrt{0.6}, W_3 = \sqrt{0.4}.$$

$X = 0.8$
$Y = 0.2$

Table 4.8: Squared weighting factors of the low bit rate users set to 0.2 and the squared weighting factors of the medium bit rate users are set to 0.8.

These results are printed in Figure 4.11.

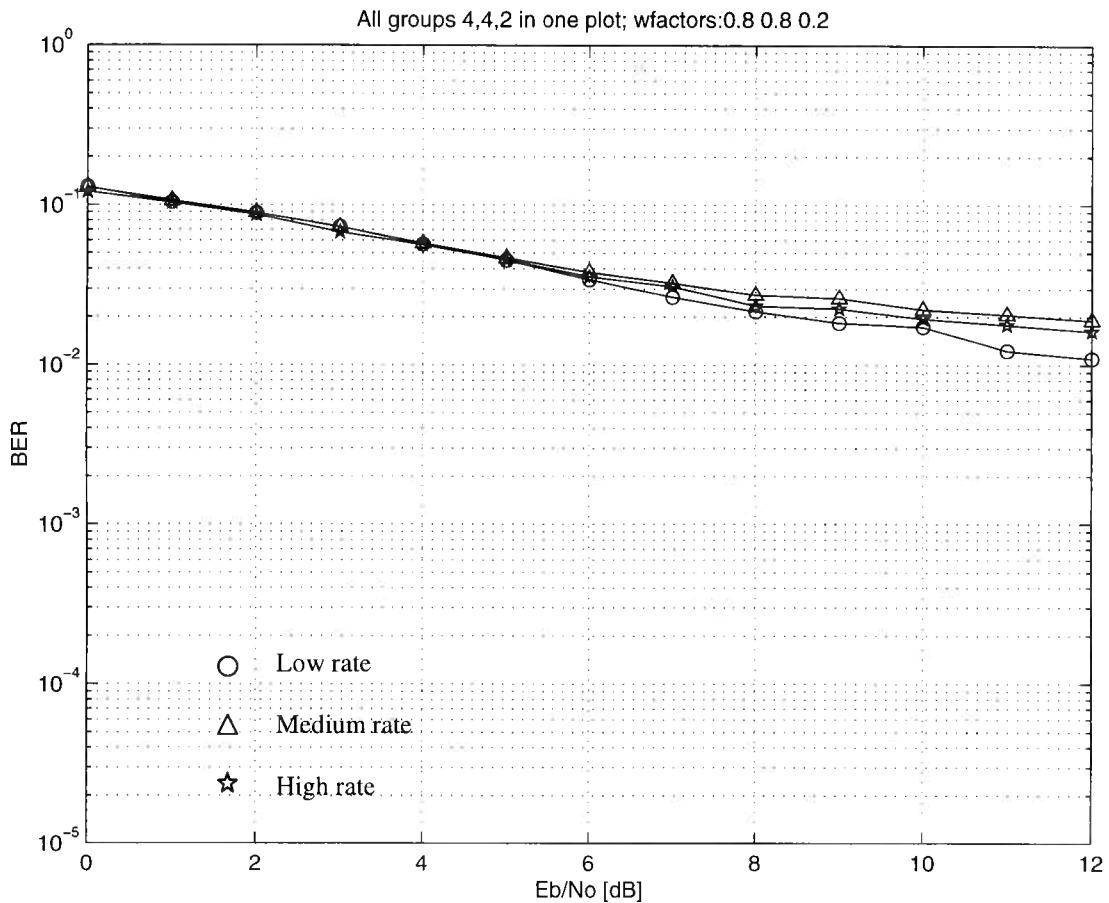


Figure 4.11: Bit error rates versus signal to noise ratio for 2 high bit rate users, 4 medium bit rate users and 4 low bit rate users.

$$W_1 = \sqrt{0.8}, W_2 = \sqrt{0.8}, W_3 = \sqrt{0.2}$$

Evaluating these three pictures (Figure 4.9, Figure 4.10 and Figure 4.11), it may be concluded that the performance of the weighted PIC receiver for the low bit rate users gets significantly better compared to the full PIC detector after one stage. For the medium and the high bit rate users there is not a significant improvement on the receivers' performance when weighting factors are applied. The best results has the weighted PIC receiver with squared weighting factors of 0.8 for the high bit rate users, 0.6 for the medium bit rate users and 0.4 for the low bit rate users.

From this a suspicion has been raised that if a good performance of the weighted PIC receiver in detecting the low bit rate users in multirate environment is desirable, there is another condition that weighting factors should satisfy: the weighting factors should be as equally distributed as possible between zero as a minimum value and one as a maximum value. Simulations with weighting factors shown in Table 4.9 and Table 4.10 are done to verify this.

Weighting factor high bit rate user	$W_1^2 = 1$
Weighting factor medium bit rate user	$W_2^2 = 0.6$
Weighting factor low bit rate user	$W_3^2 = 0.2$
Sum of the Weighting factors	$W_1^2 + X + Y = C \quad C = 1.8$

Table 4.9: Different weighting factors for a good detection of the low bit rate users.

Results of the simulations run with the weighting factors from Table 4.9 are shown in Figure 4.12.

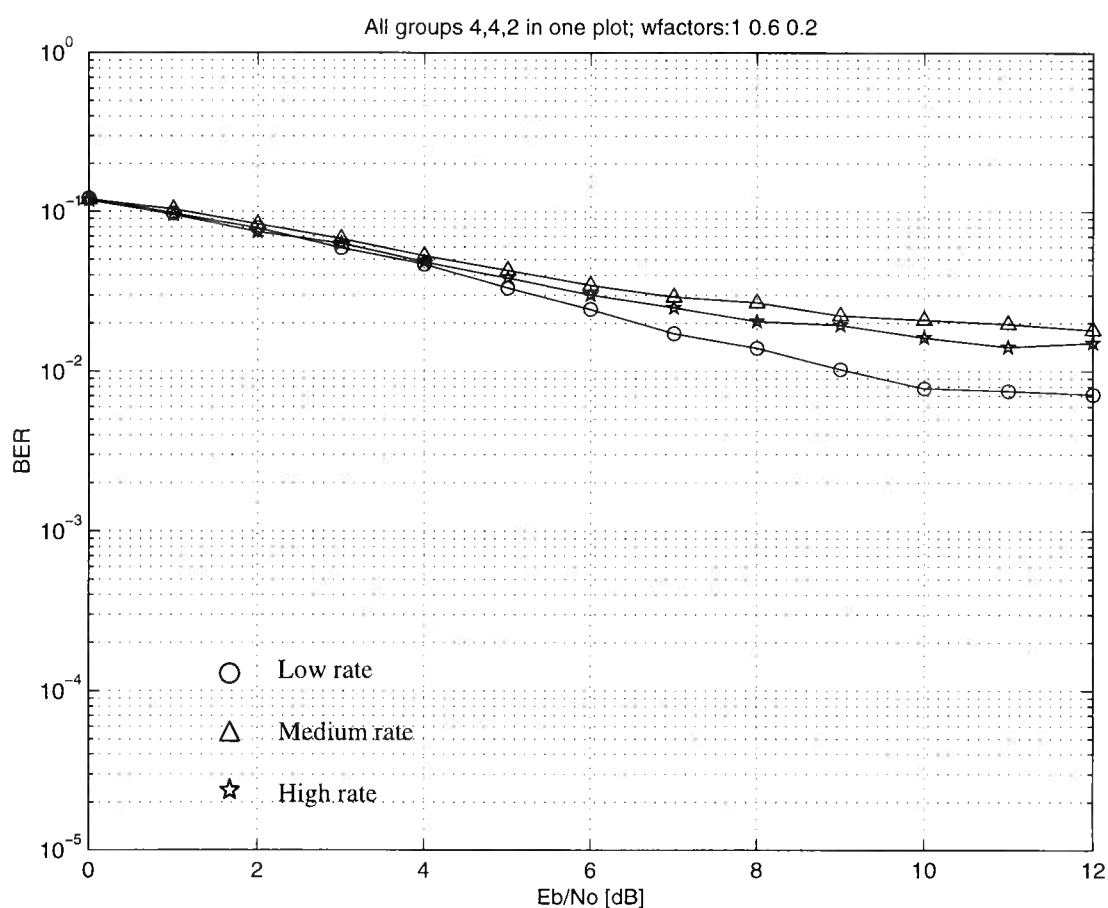


Figure 4.12: Bit error rates versus signal to noise ratio for 2 high bit rate users, 4 medium bit rate users and 4 low bit rate users.

$$W_1 = 1, W_2 = \sqrt{0.6} \quad W_3 = \sqrt{0.2}.$$

Weighting factor high bit rate user	$W_1^2 = 0.9$
Weighting factor medium bit rate user	$W_2^2 = 0.7$
Weighting factor low bit rate user	$W_3^2 = 0.2$
Sum of the Weighting factors	$W_1^2 + X + Y = C \quad C = 1.8$

Table 4.10: Different weighting factors for a good detection of the low bit rate users.

Results of the simulations run with the weighting factors from Table 4.10 are shown in Figure 4.13.

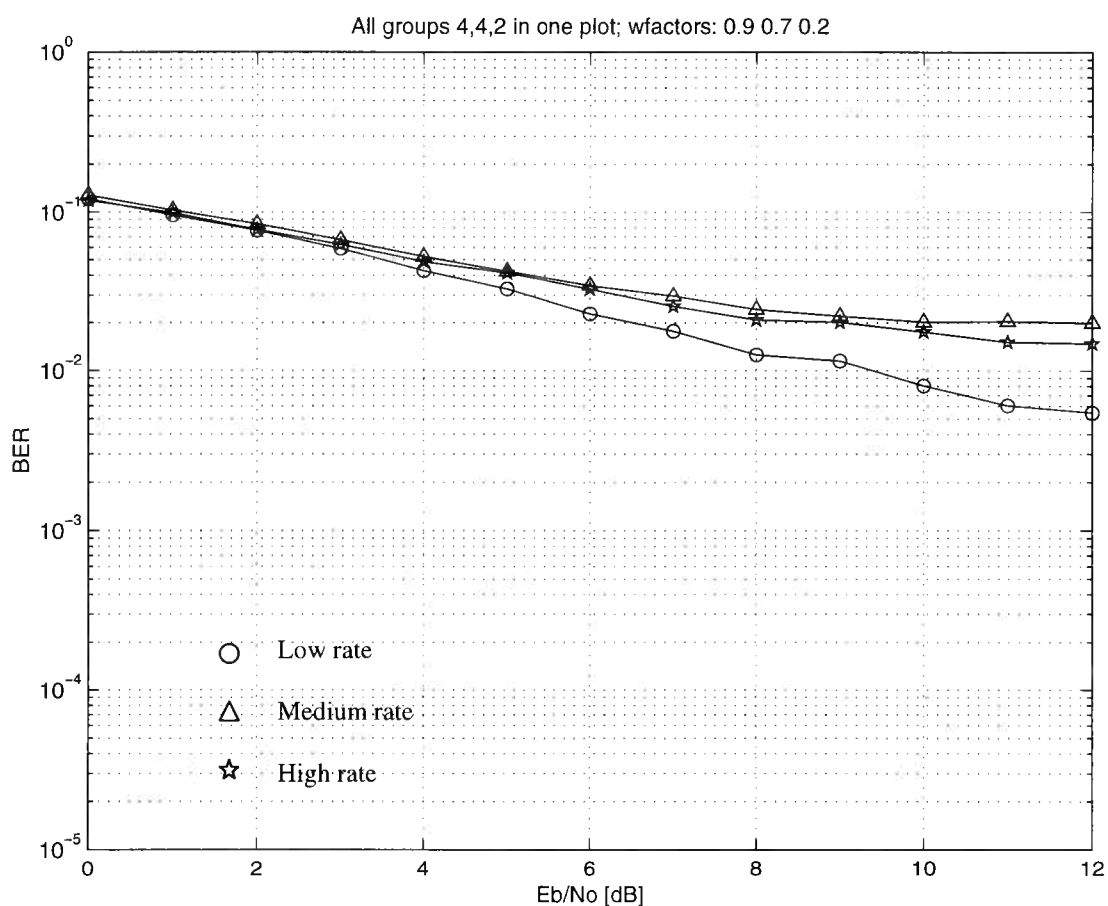


Figure 4.13: Bit error rates versus signal to noise ratio for 2 high bit rate users, 4 medium bit rate users and 4 low bit rate users.

$$W_1 = \sqrt{0.9}, W_2 = \sqrt{0.7}, W_3 = \sqrt{0.2}.$$

Evaluation of Figure 4.12 and Figure 4.13 has proven the suspicion to be true. An equal distribution of the weighting factors between zero as the bottom limit and one as the top limit, with the weighting factors of the high bit rate users as the largest and those of the low bit rate users as the smallest, gives the best results for the detection of the low bit rate users in a multi rate environment with a GroupWise Weighted PIC receiver.

Notice that in the case of Table 4.9 the performance of the weighted PIC receiver in detecting the low bit rate users is worse than the conditions for the weighting factors written down in Table 4.7 and Table 4.10. This is the effect of using the full First Estimate in the cancellation step. The First Estimate of the high bit rate users is the most reliable compared to other users but is not exactly correct. Setting a weighting factor for the high bit rate user equal to one is not an ideal case.

A summary of the conclusions drawn in this section is given in Table 4.11. In this table the conditions for the weighting factors are shown to get the best detection of the low bit rate users. A weighted PIC receiver can not improve the detection of the high bit rate users or the medium bit rate users very much compared to a full PIC receiver.

Description	Mathematical meaning
The weighting factors of the high bit rate users are larger than the weighing factors of the other two users. The weighting factors of the medium bit rate users are larger than the weighing factors of the low bit rate users.	$W_1^2 > W_2^2 > W_3^2$
Setting a weighting factor equal to zero is to be avoided.	$W_1^2 \neq 0$ $W_2^2 \neq 0$ $W_3^2 \neq 0$
Setting a weighting factor equal to one is to be avoided.	$W_1^2 \neq 1$ $W_2^2 \neq 1$ $W_3^2 \neq 1$
Conditions following from the above	$I > W_1^2 > W_2^2 > W_3^2 > 0$
The difference between the applied weighting factors should be almost equal.	$W_1^2 - W_2^2 \approx W_2^2 - W_3^2 \approx d$ $d = (W_1^2 - W_3^2)/2$

Table 4.11: Conditions for the weighing factors to get the best detection of low bit rate users with a GroupWise Weighted PIC receiver.

With Table 4.11 the section about the influence of weighting factors on the performance of the GW-PIC detector has been finished. In the sections coming after this section a GW-PIC receiver shall be examined. The weighing factors for this receiver are chosen on the best performance of this receiver in detecting low bit rate users. So far, although optimal weighting factors are not yet found, the best performance in detecting the low bit rate users has the GW-PIC receiver with weighting factors for the first stage shown in Table 4.12. In Section 5.3 of this work the GW-PIC receiver will have these weighting factors for the first stage followed by a full PIC stage.

Weighting factor high bit rate users in the first stage	$W_1^2 = 0.8$
Weighting factor medium bit rate users in the first stage	$W_2^2 = 0.6$
Weighting factor low bit rate users in the first stage	$W_3^2 = 0.4$

Weighting factor high bit rate users in the second stage	$W_1^2 = 1$
Weighting factor medium bit rate users in the second stage	$W_2^2 = 1$
Weighting factor low bit rate user in the second stage	$W_3^2 = 1$

Table 4.12: Selected weighting factors on the basis of best performance of the GW-PIC after one stage in detecting low bit rate users extended to a two stage GW-PIC receiver.

4.3.3 Comparison of receiver performance of the GW-PIC to other receivers

In Section 4.3.2 it has been made clear that the GW-PIC receiver detects low bit rate users the best when the weighting factors of Table 4.12 are applied.

After a second stage of full PIC, both the GW-PIC receiver and the PIC receiver have the best detection of the low bit rate users. The increase in BER performance after two stages, when the weighting factors from Table 4.12 are applied, is visualized in Figure 4.14.

Finally, a basic GSIC receiver has been placed in the multirate environment of 4 low bit rate users, 4 medium bit rate users and 2 high bit rate users, this is depicted in Figure 4.15. The conclusion may be drawn that the GW-PIC receiver detects the different users better than a full PIC receiver and a basic GSIC receiver.

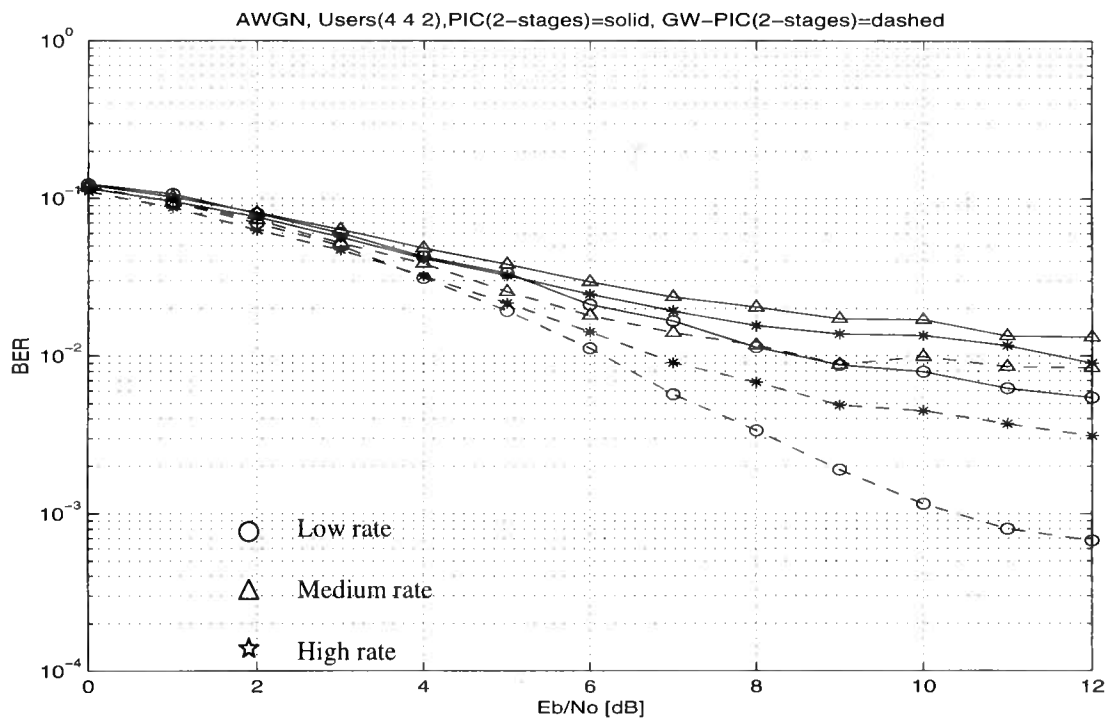


Figure 4.14: BER of two stage PIC receiver and a two stage GW-PIC receiver in a three processing gain system with 2 high, 4 medium and 4 low bit rate users. Weighting factors that are used in this figure are mentioned in Table 4.12.

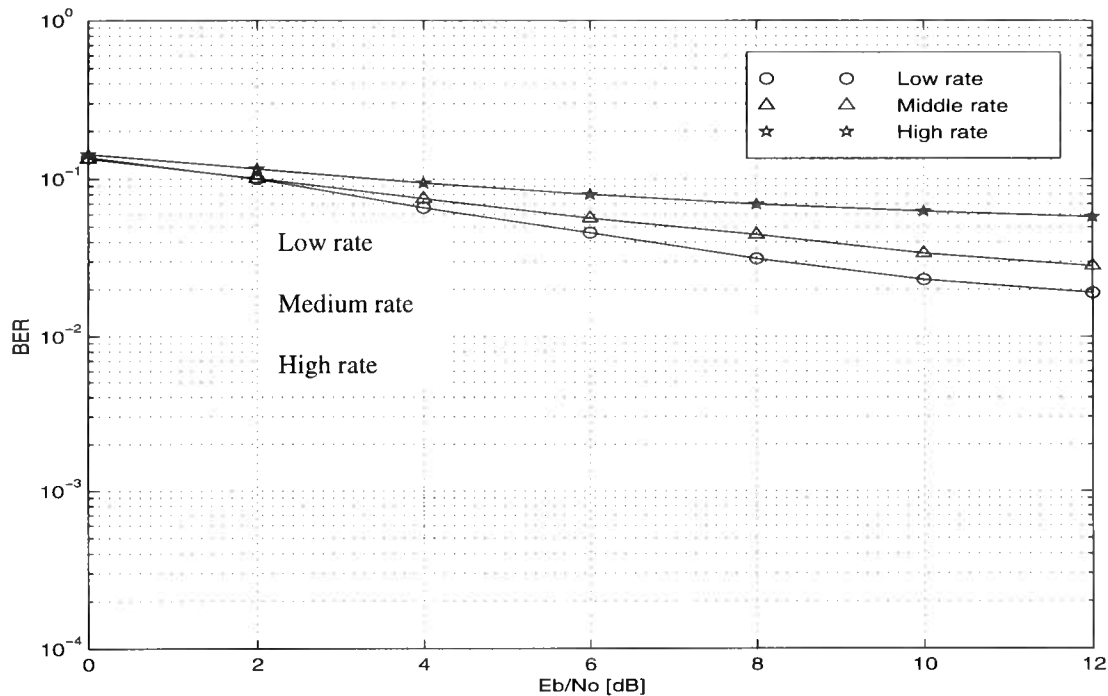


Figure 4.16: BER of basic GSIC receiver in a three processing gain system with 4 low, 4 medium and 2 high bit rate users [6].

4.4 Discussion

In an AWGN channel there are several conclusions to be drawn. In a single rate as well as in a multirate system the performance of a PIC receiver can be improved when weighting factors are applied. The performance of the GW-PIC receiver in a multirate environment is strongly dependent of the chosen weighting factors. The GW-PIC can get better BER performances than the basic GSIC receiver or the full PIC receiver in a multirate AWGN environment. In this work there was no study made of the optimization of the weighting factors, only the influence of weighting factors has been investigated and several conclusions were drawn. This study can be used to further investigate the GW-PIC and to optimize the weighting factors to get even better results in the AWGN environment.

Chapter 5

Performance of multirate receivers in fading channels

In Chapter 4 the GW-PIC receiver has been studied in an AWGN channel. In mobile telecommunication studies this model is good to begin with but not good enough to model the system in a realistic way. In realistic situations the transmitted signal is often received coming from different propagation paths. The received signal may consist of multiple signals from the same transmitter, which followed different paths. The copies of the signal hence created can differ from each other in phase, caused by the varying time of arrival at the receiver. Adding copies with different phases in the channel output causes an amplitude variation in the received signal. This phenomenon is called *multipath fading* [9].

In the first section of this chapter the simulation parameters used in fading channels will be described. In Section 5.2 the BER performance of the GW-PIC in a single path and in a two path multirate Rayleigh fading channel will be presented. This chapter continues with the analysis of the robustness of the GW-PIC receiver to fixed delay errors. Conclusions will be drawn in the last section of this chapter.

5.1 Simulation parameters

The channel is modelled as a *slowly fading frequency selective Rayleigh process* (see Section 2.1.2). The simulation parameters in fading channels are the same as the ones used in the AWGN channel of Chapter 4. The same Gold codes are used together with the same number of samples per chip, meaning four samples per chip. In Section 4.1 of this Chapter there are 4 low bit rate users, 4 medium bit rate users and 2 high bit rate users in a single path and a two path Rayleigh fading channel. The low bit rate users transmit at a data rate of 16 kbits/s, the medium bit rate users at twice this rate and the high bit rate users at four times this rate, i.e. 64 kbits/s. The transmissions take place at a frequency of 2 GHz. and a vehicle speed of 80 km/h.

In the section that describes the performance of the GW-PIC receiver with fixed delay errors the most complicated channel, i.e. a two path Rayleigh fading channel, is chosen. The root raised cosine pulse with a roll-off factor of 0.75 is used to model the sent pulses as in Chapter 4.

Without loss of generality the assumption is made that the maximum difference in delay between the first propagation path of the first user and the first propagation path of the last user is smaller than one symbol interval of a low bit rate user (Formula 2.12). The *multipath spread* is thus assumed to be smaller than the symbol interval of the lowest bit rate user. The power profile of the two paths is the same, which means that the signal power is equally divided over the two paths. The simulation parameters are shown in Table 5.1

Transmission mode	Asynchronous
Spreading modulation	BPSK
Processing gains	32,16,8
Basic Code length	32
Basic data rate	16 kbits/s
Code family	Gold
Number of samples/chip	4
Chip waveform/Roll-off factor	Root-raised-cosine/0.75
Channel model	1-path Rayleigh fading channel 2-path Rayleigh fading channel
Mobile speed	80 km/h
Multipath spread	$T_m < T_s$
Samples per chip	4
Symbol block	500
Channel estimation	Assumed known

Table 5.1: Simulation parameters for fading channels

5.2 Performance in a Multiple Processing Gain system

This section begins with the evaluation of the BER performance of the GW-PIC receiver in a one path Rayleigh fading channel. This means that the only the changes in the amplitude of the received channel is considered. The movement of the transmitting source (see Figure 2.3) causes this varying amplitude. In a one path Rayleigh fading channel, the received signal is not added to a copy caused by another path of the same user. This is only the case for a two (or more) path Rayleigh fading channel, which means that there is one (or more) extra path. The amplitude is dependent of the phase, which is correlated, with the moment of arrival at the receiver's end.

Figure 5.1 depicts the influence of equal weighting factors on the BER performance of the GW-PIC receiver at a signal to noise ratio of 8 dB. It can be seen from this figure that the full PIC receiver detects the received signal with the least errors.

After the received signal passes through a full PIC stage almost no amelioration in BER performance can be obtained by adding another full PIC stage. The BER performance after one stage in a single path multirate Rayleigh fading channel can be improved applying weighting factors, but the equal weighting factors for the best detection after one stage are equal to one (full PIC). The performance of the GW-PIC detector is bound within two limits. The upper bound is obtained after only RAKE Matched Filtering (all weighting factors equal to zero) and the lower bound is obtained after a full PIC stage (all weighting factors equal to one).

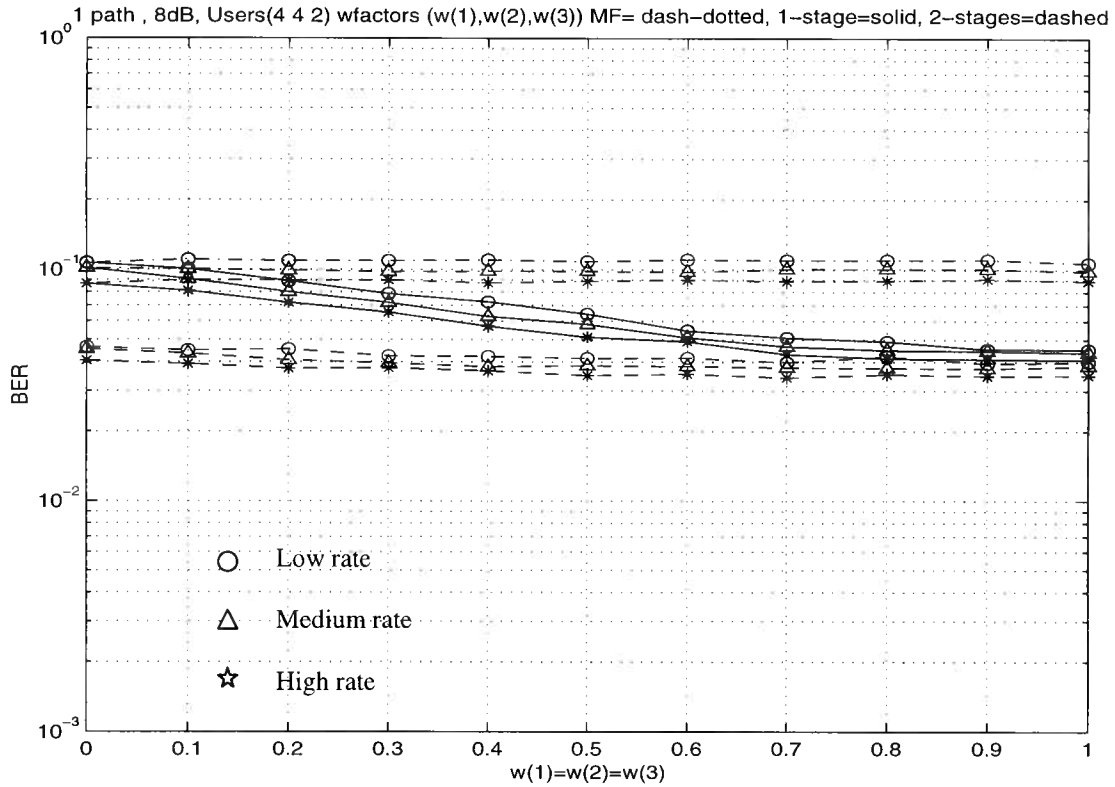


Figure 5.1: Bit error rates versus equal weighting factors at a signal to noise ratio of 8 dB in a single path system for 2 high, 4 medium and 4 low bit rate users. The equal weighting factors in the first stage of the GW-PIC receiver ranging from zero to one. The weighting factors in the second stage were set to unity (full PIC).

Figure 5.2 is the same as Figure 5.1 except that there are now two resolvable paths for the RAKE Matched Filter. The same trend as in Figure 5.1 can be seen. The system performance is the best after a full PIC stage and the worst after only RAKE Matched Filtering. The performance of the receiver in a two-path scenario is slightly better than in a single path scenario. It is less likely that the received signals coming from two different propagation paths are in a deep fade at the same time.

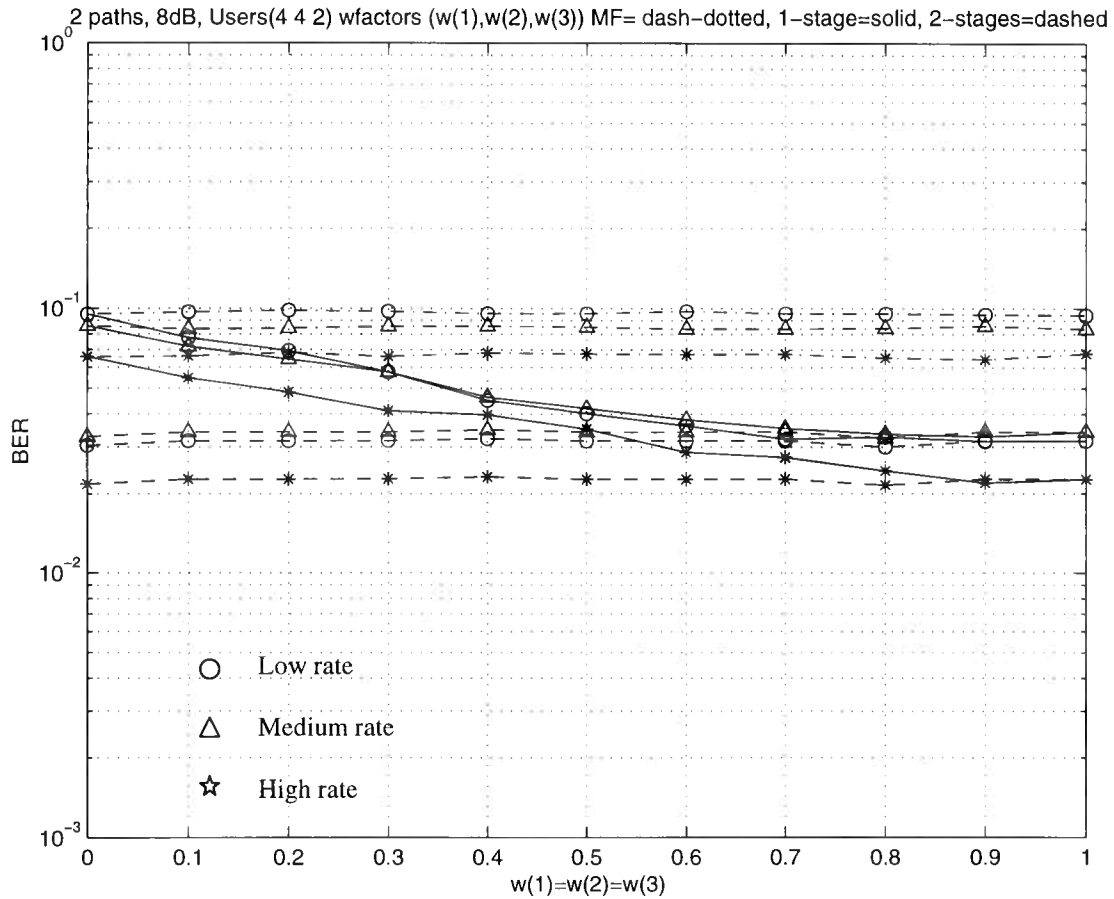


Figure 5.2: Bit error rates versus equal weighting factors at a signal to noise ratio of 8 dB in a two path system for 2 high, 4 medium and 4 low bit rate users. The equal weighting factors in the first stage of the GW-PIC receiver ranging from zero to one. The weighting factors in the second stage were set to unity (full PIC).

In Figure 5.2 (as in Figure 5.1) there is an upper limit and a lower limit of BER performance of the GW-PIC receiver. The upper bound is attained after only RAKE Matched Filtering and is slightly lower than in the single path scenario. The lower bound is obtained after a full PIC stage.

Figure 5.3 depicts the GW-PIC receiver in a two path Rayleigh fading channel with a fixed weighting factor of 0.7 for all users in the first stage. The signal to noise ratio varies from 4 to 14 dB. It can be seen that after one stage the PIC receiver (weighting factors equal to one) performs better than the GW-PIC with weighting factors equal to 0.7.

From these figures it can be concluded that applying (equal) weighting factors in a groupwise way is not the best solution in fading channels. The received powers of the different users differ from symbol to symbol. The weighting factors should be chosen according to these varying powers in order to get a better BER performance.

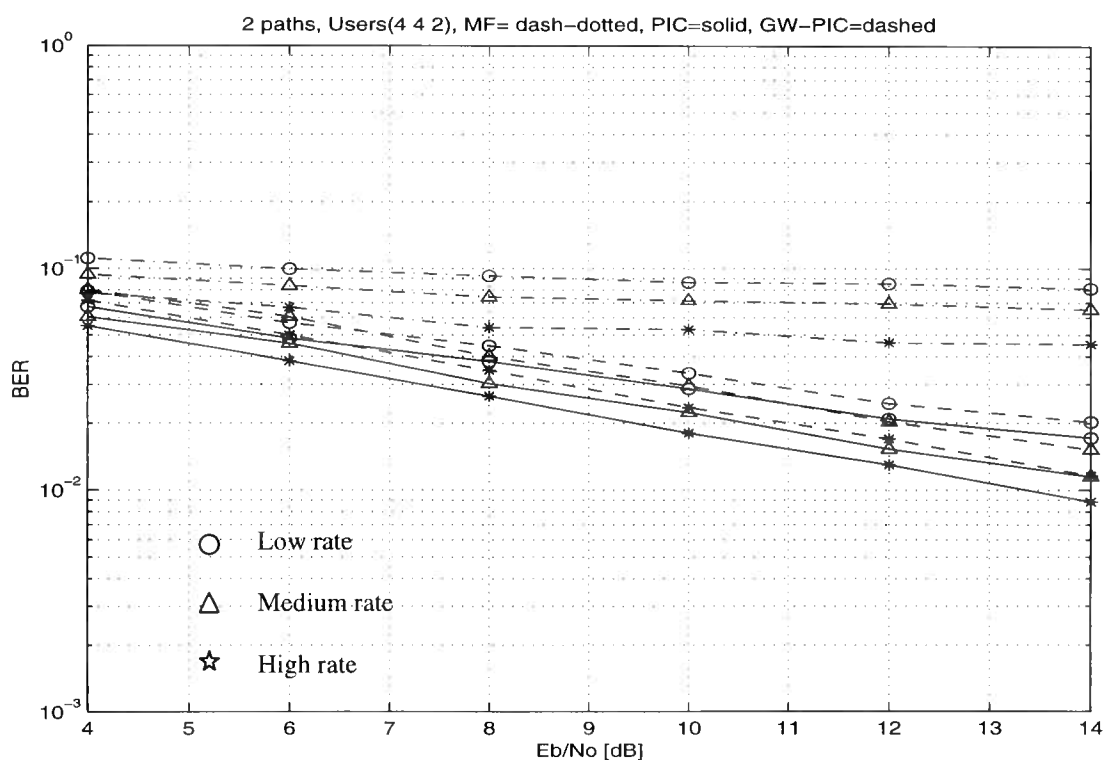


Figure 5.3: BER Performance to signal to noise ratio of the RAKE Matched Filter (dash-dotted), 1 stage GW-PIC receiver with equal weighting factors of 0.7 (dashed) and a 1 stage PIC receiver (solid) in a two paths Rayleigh fading channel with 2 high, 4 medium and 4 low bit rate users.

Compared to the basic GSIC in a two paths Rayleigh fading channel the GW-PIC performs better. The BER performance of the basic GSIC is depicted in Figure 5.4.

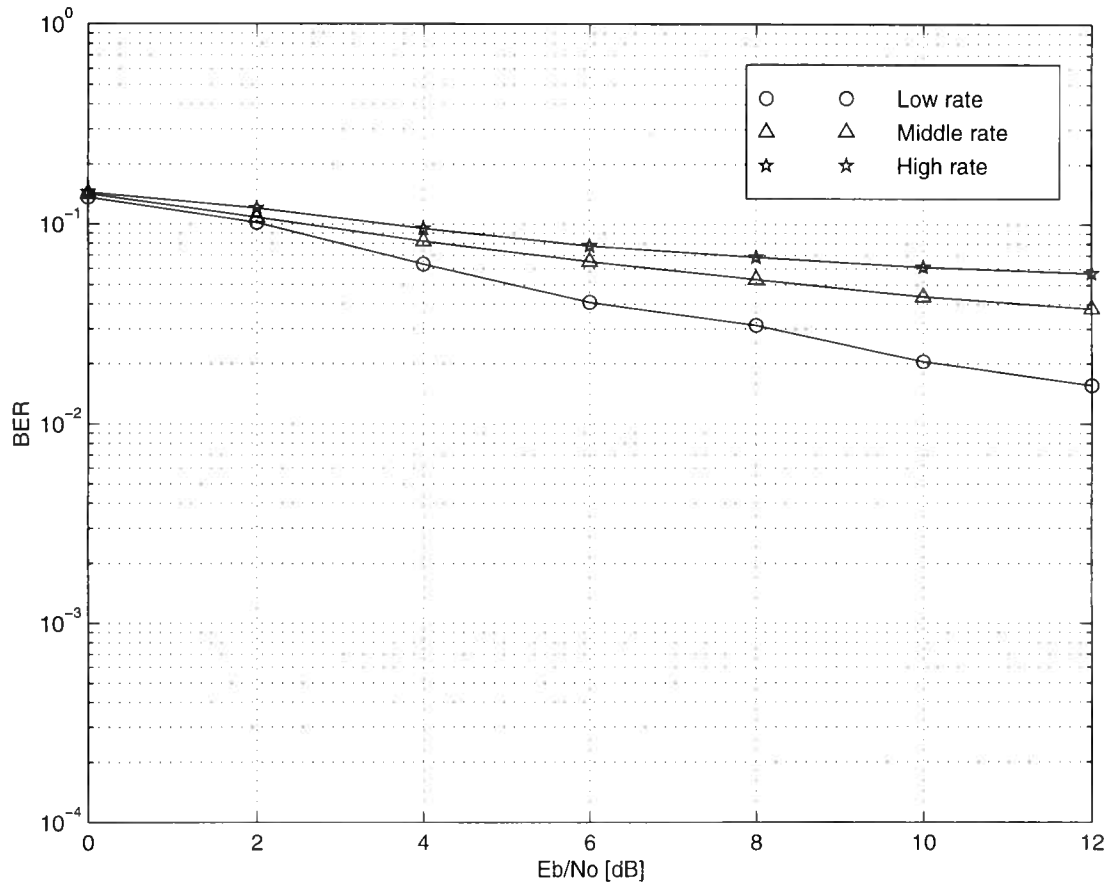


Figure 5.4: BER Performance to signal to noise ratio of the basic GSIC receiver in a two paths Rayleigh fading channel with 2 high, 4 medium and 4 low bit rate users [6].

5.3 Performance in a known channel with fixed delay estimation errors

In the previous section it was assumed that the RAKE correlator, used for the Initial Estimate, knows the delays of the different users. In this section there will be a fixed delay estimation error introduced. The RAKE receiver correlates the received signals with a Gold code that is shifted in time according to the delay estimation error. The result of the mismatched correlation is a decrease of energy at the output of the RAKE correlator. This means that the quality of the First Estimates as well as the quality of the MAI estimates decrease. The performance of the GW-PIC receiver will be worse when a fixed delay error is introduced. This is depicted in Figure 5.5. In this figure the weighting factors in the first stage of the GW-PIC are equal to $\sqrt{0.8}$, $\sqrt{0.6}$, $\sqrt{0.4}$ for the high bit rate users, the medium bit rate users and the low bit rate users, respectively. The weighting factors in the second stage of the GW-PIC are set to unity (full PIC).

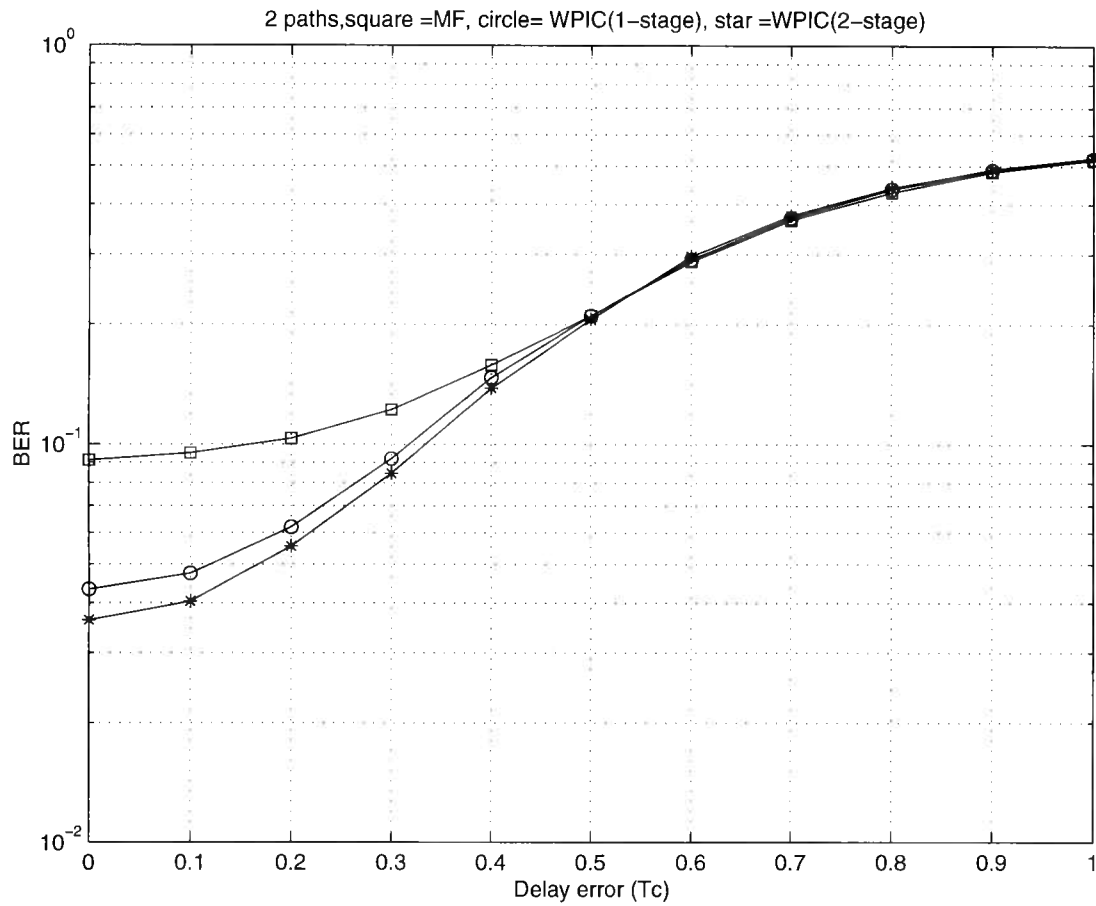


Figure 5.5: BER performance against fixed delay errors measured in tenths of chip duration time in a two path Rayleigh fading channel with 2 high, 4 medium and 4 low bit rate users and at a signal to noise ratio of 8 dB. RAKE Matched Filtering is square, 1 stage GW-PIC is circle and a 2 stage GW-PIC with weighting factors noted in Table 4.12 is star

From Figure 5.5 it can be seen that the RAKE Matched Filter as well as the GW-PIC receiver are relatively robust against delay estimation errors up to one tenth of a chip duration time. If the delay estimation error is larger than a tenth of a chip duration time then the BER Performance of the GW-PIC receiver degrades rapidly (see also [21]).

5.4 Discussion

Several conclusions can be drawn from the results of the simulations done in fading channels. The first conclusion is that the BER performance of the RAKE Matched Filter is slightly better in a two path Rayleigh fading channel than in a single path Rayleigh fading channel. This probability that the signals coming from two different propagation paths undergo a deep fade in their power at the same moment is smaller. The overall performance of the GW-PIC in a two path Rayleigh fading channel is better than the performance of this receiver in a single path Rayleigh fading channel.

A second conclusion that can be drawn is that the GW-PIC with equal weighting factors can not perform better than a full PIC receiver in fading channels. All the received powers in a fading channel fluctuate. The GW-PIC treats all the estimates coming from these signals with fluctuating powers in the same way in the cancellation of the MAI. The weighting factors should be chosen according to the quality of the estimates. The quality of the estimates fluctuates from symbol to symbol and the amount of power taken into the cancellation step should also fluctuate from symbol to symbol. In the GW-PIC receiver this is not the case.

The high bit rate users are the best detected in fading channels by the RAKE Matched Filter as well as by the (GW-) PIC receiver (see also [7]). These users generate the best MAI estimates. It can be seen that if these estimates are taken fully in account in the cancellation (the full PIC case) the BER performance is best. Adding another PIC stage slightly improves the performance. The bad First Estimates from the low bit rate users influence the performance in all stages of the (GW-) PIC receiver.

A tolerable delay estimation error has been found to be smaller than one tenth of a chip duration time. When the delay estimation error is increasing the BER performance of the (GW-) PIC receiver deteriorates linearly in a logarithmic scale until a delay estimation error of six tenths of a chip duration time. When the delay estimation error is larger than six tenths of a chip duration time the BER performance of the GW-PIC receiver saturates.

Chapter 6

Conclusions and Recommendations

6.1 Conclusions

This report describes the BER performance of multirate GW-PIC receivers in both AWGN channels and Rayleigh fading channels. In the AWGN channels an extensive study has been done to the influence of weighting factors. Several conclusions on weighting factors can be drawn from this work.

From the study on weighting factors in an AWGN channel it may be concluded that the BER performance of the PIC receiver can be improved by applying weighting factors. In a multirate environment the weighting factors have a large influence on the detection of the different groups. Choosing a set of weighting factors influences the BER performance of the GW-PIC receiver averaged over all users as well as the BER performance of the GW-PIC receiver for each individual group. An extensive study has been done to the effect of weighting factors on the detection of low bit rate users. From this study it was concluded that the performance of the GW-PIC receiver in detecting low bit rate users is best, when the weighting factors are equally divided over the interval $]1,0[$, with the highest weighting factors for the high bit rate users and the lowest weighting factors for the low bit rate users. These conclusions could be used as directions for optimisation of the BER performance of the GW-PIC. Another conclusion drawn from the studies in an AWGN channel is that the GW-PIC receiver can get a better performance than the (full) PIC receiver and than the basic GSIC receiver when certain weighing factors are used.

Analysis of the GW-PIC in a slowly fading frequency selective channel showed that the BER performance of the GW-PIC receiver with equal weighting factors is not better than the BER performance of the full PIC receiver. The more the equal weighting factors are approaching one (full) PIC, the better the performance of the GW-PIC in fading channels gets. The PIC receiver has the best BER performance in fading channels compared to the GW-PIC (with equal weighting factors smaller than one) and compared to the basic GSIC receiver. The conclusion on the performance of the GW-PIC with a fixed delay estimation error is that a fixed delay estimation error of less than one tenth of a chip duration time has little influence on the BER performance of the GW-PIC receiver.

6.2 Recommendations

In these simulations a GW-PIC receiver was studied in different channels. In AWGN channels the GW-PIC has a better performance than the (full) PIC receiver has when certain weighting factors are used. It is interesting to optimise these weighting factors to get an even better result with the GW-PIC receiver in an AWGN channel. This thesis can be used for directions to achieve an optimal performance of MUD receivers in an AWGN channel. However, in fading channels the BER performance of the PIC receiver is better than the BER performance of the GW-PIC receiver with equal weighting factors. This is due to the received power that varies from symbol to symbol. It would therefore be interesting to apply weighting factors not in a fixed groupwise way but in a varying way depending on the power level of the received symbols. This can be achieved by estimating the received power of the symbol that has to be detected and appoint a weighting factor chosen from a field of values ranging from zero to one. An algorithm has to be written to select the weighting factor.

Other combinations than weighting factors with PIC receivers could be made. For example, implementing weighting factors in serial structures like the basic GSIC or receiver derived from the basic GSIC could ameliorate the BER performance of these receivers.

In these simulations the only study that has been done on the delay estimation errors is one with fixed delay estimation errors. A study on the BER performance with varying delay estimation errors would be interesting because this is closer to reality.

The performance of GW-PIC receivers with channel speeds other than 80 km/h could be studied. With a lower channel speed the variation of the received power is less strong and the weighting factors could ameliorate the system performance compared to a full PIC receiver.

Finally, different modulation techniques could be used and the influence of this on the system performance needs to be studied further. The UMTS proposal as described within FRAMES [22] uses QPSK.

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Appendix A

Extended Abstract of VTC 1999-Fall Paper

A.F. van Meeteren, T. Ojanperä, H. Nikookar, R. Prasad, "Groupwise Weighted Parallel Interference Cancellation for Asynchronous Multirate DS-CDMA", VTC-Fall, Amsterdam, Netherlands, September 1999

Groupwise Weighted Parallel Interference Cancellation for Asynchronous Multirate DS-CDMA

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Extended Abstract:

In the development of third mobile telecommunication systems, at ITU level called IMT-2000 (International Mobile Telecommunications by 2000), attention for the realisation of mobile systems, which are able to integrate speech, data and video transmission is increasing day-by-day. To realise the integration of these services in one system, it will be necessary to provide multiple data rates. Wideband services with data rates up to 384 kbits/s for wide area coverage and up to 2 Mbits/s for indoor environments are required [1].

There are several ways to implement multiple data rates, i.e. multirate, communication for DS-CDMA (Direct Sequence Code Division Multiple Access). One of them is the use of a MPG (Multiple Processing Gain) scheme, which spreads the signals depending of their data rate. A group of low data rate users is spread more than a group of high data rate users. This is realised by using the entire spreading code for low data rate users and only part of the spreading code for high data rate users [2]. With this scheme all the users, independent of their data rate, have the same bandwidth. Since the group of high data rate users is less spread, this group is more sensitive to external interference sources. To compensate this, the parts of the codes used to spread the high data rate users are given the same energy as the codes used for the low data rate. However, the data rates of the high data rate users are higher than the data rates of the low bit rate user. When the signals are received with equal energies, the received power of the high data rate users is higher than the received power of the low data rate users [3].

Within this CDMA system a user is using the same channel as other users, causing interference with them. This is called *Multiple Access Interference* (MAI). The caused MAI deteriorates the reception of the signal. To tackle this problem multiuser detection schemes have been proposed [4,5]. The optimal multiuser detection structure has found to be too complex to use and suboptimal techniques were considered. One class of suboptimal receivers is the class of interference cancellation receivers. In this class there is a distinction to be made: Serial Interference Cancellation (SIC) and Parallel Interference Cancellation (PIC). Both structures use temporal estimates of the MAI to reduce the MAI in the received signal by subtracting these estimates from it. The difference between a SIC receiver and a PIC receiver is that a SIC receiver subtracts the estimates of the MAI in a serial way [3] and a PIC receiver subtracts the MAI in a parallel way. A PIC receiver is therefore much faster than a SIC receiver.

A drawback of reducing the MAI in the received signal with the use of a PIC receiver is that the PIC receiver subtracts the MAI estimates *entirely* from the received signal. The MAI estimates are often not completely correct and by subtracting these estimates from the received signal, the detection might even get worse. In this paper a *weighting factor* is given to the reliability of the estimated MAI. This factor decides the amount of power from the estimates used in the subtractive part of the receiver. A good MAI estimate can be seen as a reliable estimate and can be given a big role in the subtraction from the received signal. The opposite can be done for bad MAI estimates.

The power of a user in the received signal is decisive for the quality of the MAI estimate of this user. A high power user will have a better MAI estimate than a user with less power. The power differences between the different users with different data rates are big and between users with the same data rate are small. This is the reason that weighting factors could be chosen in a *Groupwise* way. This means that all the users belonging to a group of users with the same data rate, thus about the same power, will have the same weighting factor. By implementing these patterns in a PIC receiver, this receiver will become a *GW-PIC* (Groupwise Weighted Parallel Interference Cancellation) receiver.

In this work the effect of (Groupwise) weighting factors on the BER (Bit Error Rate) performance of PIC detection in an asynchronous multirate environment is investigated. A comparison of the GW-PIC receiver with a PIC receiver in an asynchronous singlerate AWGN (Additive White Gaussian Noise) environment is made. After interpretation of these results, the GW-PIC receiver is placed in an asynchronous multirate AWGN environment and the influence of weighting factors in such an environment is extensively studied. This study has been done in a AWGN environment and then has been extended to a single path and a two paths asynchronous multirate Rayleigh fading channel, in which the GW-PIC receiver with equal weighting factors has been considered. Finally, the impact of delay estimation errors has also been investigated in the paper.

If this extended abstract is accepted, then in the final version of the paper the BER simulation results of the GW-PIC with different weighting factors will be presented and a comparison of the results will be provided.

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