Code Acquisition Techniques for CDMA-Based Mobile Networks

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Submitted for the Degree of
Doctor of Philosophy
from the
University of Surrey

UniS

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August 2003
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Summary

The initial code acquisition techniques for direct sequence code division multiple access (DS/CDMA) communication networks are investigated in this thesis.

Conventional methods of code acquisition, which are basically based on the auto correlation and cross correlation properties of spreading codes, fail in the presence of multiple access interference (MAI) and the near-far effect. This fact motivates the study for interference resistant acquisition algorithms in the hostile channel environment. Training-based acquisition is investigated and the effect of training sequence structure on acquisition performance is discussed. A new training sequence architecture is proposed which results in a shorter acquisition time.

Demands for high bit rate services and needs for more efficient exploitation of resources lead to the study of acquisition algorithms that do not need the preamble or training sequences. In this context, blind adaptive algorithms for code acquisition are investigated. The mismatch problem of blind algorithms is addressed and a novel method of mismatch problem handling for Constraint Minimum Output Energy (C-MOE) is proposed. The algorithm results in good acquisition performance under different channel conditions and system loadings.

The idea of joint acquisition and demodulation of data, where the outcome of the acquisition mode is an interference suppressor filter, is also discussed. It is shown that in this class of receivers, a one-step constraint acquisition process is not sufficient for handling both the mismatch problem and exploiting the multi-path diversity. Therefore, a novel receiver is proposed which is able to handle the mismatch problem as well as the channel diversity. This receiver is based on a two-step constraint minimum output energy algorithm and comparatively provides a good acquisition and demodulation performance.

**Keywords:** Code acquisition, interference cancellation, joint acquisition and demodulation
Dedicated To

Maryam & Negar
Acknowledgments

I thank my supervisors, Prof. R. Tafazolli and Prof. B. Evans, without whose help this thesis could not have been finished.
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Acronyms

3G  Third Generation Mobile Systems
ACQ  Acquisition
AICH  Acquisition Indicator Channel
ASIC  Application Specific Integrated Circuit
AWGN  Additive White Gaussian Noise
BCH  Broadcast Channel
BER  Bit Error Rate
B-MOE  Basic Minimum Output Energy
BPSK  Binary Phase Shift Keying
CDMA  Code Division Multiple Access
CFAR  Constant False Alarm Rate
C-MOE  Constraint Minimum Output Energy
DL  Downlink
DOA  Direction Of Arrival
DS  Direct Sequence
DS/CDMA  Direct Sequence CDMA
DS/SS  Direct Sequence Spread Spectrum
Eb/No  Energy of Bit/(2*Noise Variance)
E-JAD  Enhanced JAD
ETSI  European Telecommunication Standards Institute
FA  False Alarm
FDD  Frequency Division Duplex
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<th>Description</th>
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<tr>
<td>FDMA</td>
<td>Frequency Division Multiple Access</td>
</tr>
<tr>
<td>FH/CDMA</td>
<td>Frequency Hopping CDMA</td>
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<tr>
<td>FIR</td>
<td>Finite Impulse Response</td>
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<tr>
<td>F-N</td>
<td>Frobenius Norm</td>
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<tr>
<td>GWSSUS</td>
<td>Gaussian Wide Sense Stationary Uncorrelated Scattering</td>
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<tr>
<td>ICI</td>
<td>Inter Chip Interference</td>
</tr>
<tr>
<td>IS95</td>
<td>Industry Standard 95</td>
</tr>
<tr>
<td>ISI</td>
<td>Inter Symbol Interference</td>
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<tr>
<td>ITU</td>
<td>International Telecommunication Union</td>
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<tr>
<td>JAD</td>
<td>Joint Acquisition and Demodulation</td>
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<tr>
<td>LFSR</td>
<td>Linear Feedback Shift Register</td>
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<tr>
<td>LMMSE</td>
<td>Linear Minimum Mean Square Error</td>
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<tr>
<td>LMS</td>
<td>Least Mean Squares</td>
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<tr>
<td>LOS</td>
<td>Line of Sight</td>
</tr>
<tr>
<td>LS</td>
<td>Least Squares</td>
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<td>MAI</td>
<td>Multiple Access Interference</td>
</tr>
<tr>
<td>MC-CDMA</td>
<td>Multi Carrier CDMA</td>
</tr>
<tr>
<td>MHz</td>
<td>Mega Hertz</td>
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<tr>
<td>ML</td>
<td>Maximum Likelihood</td>
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<tr>
<td>MOE</td>
<td>Minimum Output Energy</td>
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<tr>
<td>MSE</td>
<td>Mean Square Error</td>
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<td>M-Sequence</td>
<td>Maximal Length Sequence</td>
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<tr>
<td>MUSIC</td>
<td>Multiple Signal Classification</td>
</tr>
<tr>
<td>MVDR</td>
<td>Minimum Variance Distortion-less Response</td>
</tr>
<tr>
<td>NLOS</td>
<td>None Line Of Sight</td>
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<tr>
<td>OVSF</td>
<td>Orthogonal Variable Spreading Factor</td>
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<td>PDF</td>
<td>Probability Density Function</td>
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<td>PE</td>
<td>Polynomial Expansion</td>
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<td>PMAI(offset)</td>
<td>Power of MAI Relative to the Desired User</td>
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<td>PN</td>
<td>Pseudo Noise</td>
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<td>QoS</td>
<td>Quality of Service</td>
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<td>RACH</td>
<td>Random Access Channel</td>
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<tr>
<td>RLS</td>
<td>Recursive Least Square</td>
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<td>SD/CA</td>
<td>Signal Detection / Code Acquisition</td>
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<td>SIR</td>
<td>Signal to Interference Ratio</td>
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<td>SNR</td>
<td>Signal to Noise Ratio</td>
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<td>SPRT</td>
<td>Sequential Probability Ratio Test</td>
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<td>TDD</td>
<td>Time Division Duplex</td>
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<td>TDMA</td>
<td>Time Division Multiple Access</td>
</tr>
<tr>
<td>UL</td>
<td>Uplink</td>
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<tr>
<td>UMTS</td>
<td>Universal Mobile Telecommunication System</td>
</tr>
<tr>
<td>UTRA</td>
<td>UMTS Terrestrial Radio Access</td>
</tr>
<tr>
<td>VLSI</td>
<td>Very Large Scale Integrated Circuit</td>
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<td>WCDMA</td>
<td>Wideband CDMA</td>
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Notations

\( a[n] \)  
\( j \)th  element of the spreading sequence of \( k \)th  user

\( \overline{\alpha}(\theta_{a,k}) \)  
Steering vector relating to the \( k \)th  direction of arrival

\( \overline{A}_0 \)  
Multi-path code signature matrix

\( b_{k,i} \)  
\( i \)th  symbol of \( k \)th  user

\( b_{k,n} \)  
\( n \)th  bit of \( k \)th  user

\( b' \)  
Estimation of the desired symbol

\( \overline{b}, \tilde{b} \)  
Vector of information symbols and its estimation, respectively

\( B(l,z) \)  
Path function (in a Markov chain) relating to the \( l \)th  path

\( c(t) \)  
Time invariant channel impulse response

\( c(t_0,l) \)  
Time variant channel impulse response

\( c_0 \)  
Speed of light

\( c_{k,p} \)  
Channel coefficient of the \( k \)th  user in \( p \)th  path

\( c_l \)  
Complex channel coefficient of the desired user in \( l \)th  path

\( c_{p,k}(t) \)  
Time invariant channel pulse response relating to the \( k \)th  user

\( \overline{c} \)  
Multi-path channel coefficient vector

\( \overline{c}_{op} \)  
Estimation of the channel coefficient vector \( \overline{c} \)

\( C \)  
Number of total acquisition states in a Markov chain

\( E(n,j,k) \)  
Event function relating to detection of the correct cell where \( n \)th  cell is the correct one and "j" miss and "k" false alarm have been counted before detecting the \( n \)th  cell

\( f(D) \)  
Characteristic polynomial function of a code generator
Notations

\( f_c \)  
Carrier frequency (Hz)

\( f_{\text{max}} \)  
Maximum Doppler frequency

\( \tilde{f} \)  
Receiver FIR filter vector

\( \tilde{f}_{\text{mse-op}} \)  
Optimum receiver vector based on the MSE criterion

\( \tilde{f}_{\text{moe}} \)  
Receiver FIR filter, calculated based on the MOE criterion

\( \tilde{f}_p \)  
Beam-former filter

\( g_0(D) \)  
General polynomial function

\( G \)  
Processing gain

\( H_0 \)  
False acquisition-state hypothesis

\( H_1 \)  
Correct acquisition-state hypothesis

\( H_0(z) \)  
Transfer function for advancing from incorrect cell to another incorrect cell

\( H_d(z) \)  
Transfer function for advancing from correct cell to acquisition state

\( H_c(z) \)  
Transfer function for advancing from correct cell to incorrect cell

\( I_0(.) \)  
Zero order Bessel function of the first kind

\( I_d(.) \)  
Intra-cell interference

\( I_c(.) \)  
Inter-cell interference

\( K_c \)  
Rician factor

\( l \)  
Distance between the base station and the mobile terminal

\( L \)  
Number of transmission paths

\( L_e \)  
Number of elements in an antenna array structure

\( L_p \)  
Path loss in dB

\( L_d \)  
Power loss caused by delay \( \delta T_c \), \( \delta \in [0,1) \)

\( M \)  
Ratio of the spreading code period to the symbol time duration
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<td>$n_d(t)$</td>
<td>Additive white Gaussian noise</td>
</tr>
<tr>
<td>$n_{kt}$</td>
<td>Integer number of chips delay relating to the $k_{th}$ user's signal in $t_{th}$ path</td>
</tr>
<tr>
<td>$N$</td>
<td>Period of a code sequence (chips)</td>
</tr>
<tr>
<td>$N_l$</td>
<td>Uncertainty region caused by distance $l$ (in chips)</td>
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<td>$P$</td>
<td>Sampling rate at the receiver side</td>
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<td>$P(n,j,k)$</td>
<td>Probability function related to $E(n,j,k)$</td>
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<td>$P_{acq\text{-}error}$</td>
<td>Probability of acquisition error</td>
</tr>
<tr>
<td>$P_d$</td>
<td>Probability of detection</td>
</tr>
<tr>
<td>$P_f$</td>
<td>Probability of acquisition failure</td>
</tr>
<tr>
<td>$P_{fa}$</td>
<td>Probability of false alarm</td>
</tr>
<tr>
<td>$P_{fa,u}$</td>
<td>Upper bound of the probability of false alarm</td>
</tr>
<tr>
<td>$P_{ij}$</td>
<td>Probability of transition from state “i” to state “j” on a Markov chain</td>
</tr>
<tr>
<td>$P_m$</td>
<td>Probability of miss</td>
</tr>
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<td>$P_{op}$</td>
<td>Probability of error of the optimum receiver (Single user, AWGN channel)</td>
</tr>
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<td>$P_{om}$</td>
<td>Probability of error of the optimum receiver (Multiple users, AWGN channel, synchronous case)</td>
</tr>
<tr>
<td>$P_{om}^{as}$</td>
<td>Probability of error of the optimum receiver (Multiple users, AWGN channel, asynchronous case)</td>
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<tr>
<td>$P_{rd}(\cdot)$</td>
<td>Rayleigh PDF</td>
</tr>
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<td>$P_{Rd}(\cdot)$</td>
<td>Rician PDF</td>
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<tr>
<td>$P_s(\cdot)$</td>
<td>PDF of shadowing</td>
</tr>
<tr>
<td>$P_{Tc}(t)$</td>
<td>Chip shape waveform of duration $T_c$</td>
</tr>
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<td>$r_c(t)$</td>
<td>Auto-correlation function of a Rayleigh fading process</td>
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<td>$r_k(t)$</td>
<td>Received signal of $K_{th}$ user</td>
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<tr>
<td>$r_o(t)$</td>
<td>Received signal of the desired user</td>
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<td>$\bar{F}_{rd}$</td>
<td>Cross correlation vector between received vector $\bar{F}$ and data sequence $d$</td>
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<td>$\bar{r}_n$</td>
<td>Sampled received vector falling in the $n_{th}$ observation window</td>
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<tr>
<td>$\bar{r}_n^{(k)}$</td>
<td>Sampled received vector of $k_{th}$ user in the $n_{th}$ observation window</td>
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<td>$R_{aa}(k)$</td>
<td>Auto correlation function of sequence $a$</td>
</tr>
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<td>$R_{aa'}(k)$</td>
<td>Cross correlation function of sequences $a$ and $a'$</td>
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<tr>
<td>$\bar{R}$</td>
<td>Autocorrelation matrix of the received signal vector $\bar{F}$</td>
</tr>
<tr>
<td>$\hat{R}$</td>
<td>Estimated autocorrelation matrix of the received signal vector $\bar{F}$</td>
</tr>
<tr>
<td>$\bar{R}_0$</td>
<td>General Autocorrelation matrix</td>
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<td>$S_c(.)$</td>
<td>Jakes power spectral density</td>
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<tr>
<td>$S_G(.)$</td>
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<tr>
<td>$s_k(t)$</td>
<td>Continuous time signature waveform, spreading sequence of $k_{th}$ user</td>
</tr>
<tr>
<td>$t[n]$</td>
<td>$n_{th}$ bit of the training sequence</td>
</tr>
<tr>
<td>$T$</td>
<td>Time duration of information symbol</td>
</tr>
<tr>
<td>$T(n,j,k)$</td>
<td>Acquisition time relating to the $E(n,j,k)$</td>
</tr>
<tr>
<td>$T_{seq,av}$</td>
<td>Average acquisition time</td>
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<tr>
<td>$T_c$</td>
<td>Time duration of code chips</td>
</tr>
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<td>$T_d$</td>
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<tr>
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<td>Average dwell time</td>
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<td>Time interval between two random access slots</td>
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<td>$\tilde{u}_k^0, \tilde{u}_k^{-1}, \tilde{u}_k^{+1}$</td>
<td>$k$th user signature waveform related to the current, previous and the next symbols, which are falling into the observation window</td>
</tr>
<tr>
<td>$\hat{u}_0$</td>
<td>Estimate of the desired user signature waveform</td>
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<tr>
<td>$\overline{u}_{\text{wrong}}$</td>
<td>A pre-known wrong hypothesis</td>
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<td>Carrier frequency (Radian/s)</td>
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<tr>
<td>$\alpha$</td>
<td>Decision threshold</td>
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<tr>
<td>$\sigma_T$</td>
<td>Constant coefficient used for the Taylor series expansion of $R$</td>
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<td>$\delta_{k,l}$</td>
<td>A fraction of a chip delay relating to the $k$th user's signal in $l$th path</td>
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<td>$\Delta T$</td>
<td>Time uncertainty interval</td>
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<td>$\Delta \Omega$</td>
<td>Frequency uncertainty interval</td>
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<tr>
<td>$\theta_{a,k}$</td>
<td>Direction of arrival of $k$th user</td>
</tr>
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<td>$\theta_k$</td>
<td>Random phase of $k$th user</td>
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<tr>
<td>$\lambda_{v,i}$</td>
<td>i$\text{th}$ eigen-value of the correlation matrix</td>
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<td>$\lambda_{v,\text{max}}$</td>
<td>Maximum eigen-value of the correlation matrix</td>
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<tr>
<td>$\lambda_k$</td>
<td>Power control error of $k$th user</td>
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<td>$\lambda_{\text{moe}}$</td>
<td>Lagrangian multiplier used for the C-MOE criterion</td>
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<tr>
<td>$\mu$</td>
<td>Step size in a LMS algorithm</td>
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<td>$\mu_d$</td>
<td>Mean of a Gaussian random variable</td>
</tr>
<tr>
<td>$\nu$</td>
<td>Penalty factor in the C-MOE algorithm</td>
</tr>
</tbody>
</table>
\( \xi_{k,0}, \xi_{k,m} \) Shadowing random variables of \( k \) user observed from desired user cell and its own cell respectively.

\( \rho_{jk} \) Cross correlation between \( j \)th and \( k \)th users

\( \rho_{j,\delta}^{l} \) & \( \rho_{k,l}^{r} \) Left hand and right hand sides, partial cross correlations between \( j \)th and \( k \)th users

\( \sigma \) Standard deviation of a Gaussian random variable

\( \sigma_{a}^{2} \) Variance of acquisition time

\( T_{o} \) Delay of the desired user in comparison with the sampling time at the receiver

\( \tau_{e} \) Relative delay (Delay of \( k \) user in comparison with \( j \)th user)

\( \phi_{k} \) Voice activity of \( k \)th user

\( \Omega_{i} \) Probability of starting the cell search process from \( i \)th cell

\( \nabla(.) \) Gradient

\( (.)^{*} \) Complex Conjugate

\( (.)^{T} \) Matrix/Vector transpose

\( (.)^{H} \) Complex conjugate transpose

\( \| . \| \) Euclidian norm

\( \text{trace}(.) \) Sum of the main diagonal elements of a matrix

\( \langle \bar{x}, \bar{y} \rangle \) Inner product of two vectors \( \bar{x} \) and \( \bar{y} \)

\( (.) \otimes (.) \) Convolution

\( [.] \) The nearest integer value, smaller than the operand

\( U_{n=0}^{k}(X_{n}) \) Union of events \( X_{n}, n \in [0,k] \)

\( T_{\delta}^{n}(\bar{x}) \) Left hand shift. Shifts vector \( \bar{x} \), \( n \) times
$T_n^r(\bar{x})$  
Right hand shift. Shifts vector $\bar{x}$, $n$ times

In this thesis the following applies:

Matrixes are denoted using uppercase letters with a bar (e.g. $\bar{A}$), or with a hat (e.g. $\hat{A}$).

Vectors are denoted using lowercase letters with a bar (e.g. $\bar{a}$), or with a hat (e.g. $\hat{a}$).

All other variables and parameters are scalar.
1 Introduction

1.1 Wireless network

The wireless communication idea goes back to 1895 by G. Marconi who used radio waves over long distance. However its rapid growth started after Bell laboratories, in 1960, proposed the cellular concept. Communication networks are now faced with explosive growth in number of subscribers, as well as increasing demands for higher bit rate services. As a result, the ability to support larger frequency bandwidth and to provide better quality of services (QoS) to the end user becomes more prominent. Furthermore, the limitation on available frequency bandwidth is already a problem in metropolitan areas, therefore the demand for new bandwidth efficient multiple access schemes, as well as advanced receiver architectures, is essential in providing low cost communications solutions for next generation systems.

Frequency division multiple access (FDMA), and time division multiple access (TDMA), are two traditional ways of allocating frequency spectrum resources to separate users. The FDMA and TDMA rely on partitioning the users in the available frequency and time domains and hence the maximum number of active users depends on the available slots of frequency and time.
However, in code division multiple access technology, known as CDMA, all users transmit their information bits at the same time and occupy the same frequency band using different spreading codes. The capacity of network in CDMA mainly depends on the number of available orthogonal or nearly orthogonal signature sequences (spreading codes) that are used for separating the users' transmissions.

There are two most commonly used spread spectrum techniques that provide the code division multiple access schemes. These are known as direct sequence CDMA (DS/CDMA) and frequency hopping CDMA (FH/CDMA) [Pete95] [Pras98] [Ojan98-a].

Generally speaking in a DS/CDMA the stream of information bits are multiplied by a user specific high rate spreading code to generate the spread spectrum signal. Alternatively, FH/CDMA scheme spreads the information bits by changing their transmission frequency regularly according to a user specific spreading code.

The FH scheme has been mostly used for military applications, while the DS/CDMA has attracted most interest in commercial systems. IS-95 was the first commercial system to adopt CDMA as an air-interface in North America; this latter scheme (DS/CDMA) is considered in this thesis.

Based on the standardisation process conducted by the International Telecommunication Union (ITU) and its European co-ordinator, the European Telecommunication Standards Institute (ETSI), the Wideband CDMA (WCDMA) has been selected for the UMTS, Universal Mobile Telephone System, Terrestrial Radio Access (UTRA) for services which are using frequency division duplex (FDD), where TDMA-CDMA was selected for time division duplex (TDD) services [Ojan98-b].

1.2 CDMA-receiver structure

The type of multiple access scheme and characteristics of mobile wireless channel have a direct influence on the design and parameters of the receiver.

The transmission channel in mobile wireless network is an open area. The WCDMA
transmitted signal arrives to the receiver through different transmission paths with different delays, phases, attenuations and different direction of arrivals. It also suffers from Doppler frequency shifts, which depends on the relative velocity of transmitter and receiver. Hence the transmitted signal is spread in time, space and frequency. The combination of the above problems and the fact that many users are sending their information at the same time and in the same frequency band with unique power profiles, have suggested the need for a new receiver design that can suppress multiple access interference, and that can exploit the multi-path characteristic of channel. Figure 1.1 shows an uplink (mobile to base station) multi-path channel.

1.3 Code synchronisation; the motivation

The previous section has defined the key requirements of a CDMA receiver design. To remove the multiple access interference, MAI, and inter symbol interference (ISI), the spreading codes should have a very low cross-correlation and very sharp auto-
correlation characteristic. Exploiting these characteristics on the other hand needs a good level of synchronisation between the transmitted code sequence and its locally generated replica at the receiver.

This suggests that code synchronisation is an essential part of any CDMA receiver. It is also discussed in the next chapter that the code acquisition process is more sensitive to MAI, ISI and the near-far effect than the demodulation process is.

This motivates the study of interference resistant and near-far resistant code delay acquisition algorithms. In addition, the concept of joint signal acquisition and demodulation, where the output of the acquisition mode is an interference resistant demodulator, can also provide an efficient solution.

Code synchronisation process is usually implemented in two stages: Code Acquisition (coarse alignment) and Tracking (fine alignment).

First stage is used to bring the delay offset between the incoming spread signal and the locally generated replica of spreading code to within the pull-in range of the tracking loop. This range is usually less than one chip time duration. The second stage is initiated to minimise the delay offset and to track the slow variations of channel such as instantaneous power and phase of the propagation paths, delays etc... [Pete95].

Interference resistant coarse alignment schemes are the subject of this thesis. Their extension to interference resistant receivers for joint CDMA signal acquisition and demodulation is also investigated.

1.4 Outline of the thesis and novel achievements

Adaptive interference suppression algorithms for joint code time delay acquisition and demodulation of direct sequence CDMA signal are studied in this thesis.

It is explained that the acquisition capacity of CDMA networks is more sensitive to multiple access interference and the near-far effect than demodulation, and thus interference resistant algorithms for the acquisition step will be more essential than the
demodulation process.

Conventional acquisition algorithms show a poor performance in highly loaded networks and especially in the presence of strong near-far effect. Also the overhead introduced by power control schemes, which is essential for conventional receivers, could be excessive for the future high data rate services. All of these factors strongly motivate the study of interference resistance acquisition algorithms.

On the other hand, the MAI suppressor algorithms have already attracted great deal of interest for the demodulation process. These algorithms introduce a considerable level of complexity to the receiver structure, thus the idea of joint signal acquisition and demodulation (JAD), where the output of acquisition mode is an interference resistant demodulator, is an interesting approach.

The training-based JAD scheme is studied in this thesis. Particularly the structure of the training sequence and its effect on overall acquisition performance is investigated. It is shown that by using a novel training sequence structure, faster acquisition time without increasing complexity can be achieved.

In packet radio networks, code acquisition or time synchronisation may be necessary for every packet session. The acquisition algorithms that are based on preamble structure or training sequences can impose a large amount of overhead to the network and reduce the efficiency of scheme. In this context, blind interference resistant code acquisition algorithms are studied. In the blind acquisition approach the only available information about desired user is its spreading code and no training information is used.

The minimum output energy (MOE) is a linear algorithm, which is suitable for applying the idea of joint acquisition and demodulation of data. This algorithm is studied in presence of multi-path fading, strong multiple access interference and unknown modulated information bits. Both uplink and downlink scenarios will be considered. For this class of receivers and in a multi-path mobile environment, a new set of important problems will be raised. One of the most critical issues is the mismatch between the best receiver hypothesis and the actual desired user signature waveform. In practice, the mismatch can significantly degrade the performance of the receiver from both acquisition
and a demodulation perspective.

A novel and practical algorithm for handling the mismatch is proposed that significantly improves the performance of the acquisition process.

It is also explained that the final receiver, which is the outcome of the acquisition mode, in practice, cannot exploit the diversity of channel effectively. The reason behind this is studied and a proper method of channel estimation is applied. By applying the method of mismatch handling used for the acquisition process, a new formula for calculation of the final demodulator filter is proposed. The channel estimation step only uses the available information at the output of acquisition mode; it does not need any new collection of data and is based on the new MOE optimisation process for refinement of the receiver.

The thesis is organised as follows:

Chapter 2 studies the principles of the acquisition process for CDMA multiple access technology. Conventional acquisition algorithms and their related concepts are introduced. As an important and practical conventional acquisition approach, the maximum likelihood scheme is studied.

The concept of acquisition capacity as well as the generalised transfer function for modelling the acquisition process is explained. Also, the limitation of conventional receivers for both acquisition and demodulation is addressed in this section.

Chapter 3 introduces the adaptive receiver systems and the mobile channel model. This model is to capture the interesting characteristics of the channel according to the time delay acquisition. Some of these characteristics are chip level and symbol level uncertainty as well as inter-symbol interference (ISI), and inter-chip interference (ICI).

It is also explained that for an adaptive JAD receiver, the structure of interference should be stationary. As the result, the class of short spreading codes is introduced and considered in all part of this thesis.

Chapter 4 focuses on interference resistance linear acquisition algorithms and their related issues that should be solved in a practical system.
Training algorithm and MOE are two most interesting approaches, which are studied in the presence of a mobile fading channel. Novel training structure as well as novel method of mismatch handling are introduced for these two methods respectively.

Proper construction of acquisition hypothesis as well as the effect of multi-path diversity on acquisition performance is also addressed. Finally, a time delay analysis is carried out, which is based on the general transfer function introduced in chapter 2 and the simulation results of chapter 4.

Chapter 5 studies the performance of the final JAD receiver and addresses the BER performance improvements.

Proper channel estimation without increasing the complexity, and compatible with the structure of the acquisition process is introduced in this section, to exploit the diversity of channel, which cannot be captured during the acquisition mode efficiently.

Also by combining the idea of mismatch handling proposed in chapter 4, and output of the channel estimation mode, a novel enhanced JAD receiver (E-JAD) structure is proposed. Performance of the E-JAD and effectiveness of channel estimation algorithm is studied in detail for an uplink and downlink multi-path channel, where simulation results show a significant improvement in comparison to the fundamental JAD, basic MOE (B-MOE) for demodulation (with perfect knowledge of multi-path delay) and conventional demodulator (the Rake receiver with perfect knowledge of delays and channel coefficients).

The novel achievements of this research are listed as follows:

- Adaptive algorithms for joint acquisition and demodulation of data were studied for the first time in multi-path channel environments.
- Effect of Training sequence structure on joint signal acquisition and demodulation performance for an LS based JAD algorithm was studied, and a new training structure, which results in a faster acquisition time is proposed.
- Blind constraint minimum output energy version of JAD algorithm was studied and the effect of mismatch, in a multi-path channel was considered. A novel and
practical method of adaptive handling of mismatch in a mobile multi-path channel is proposed.

- Performances of the modified Training and the proposed blind adaptive C-MOE version of JAD were evaluated for both uplink and downlink of a mobile Rayleigh fading channel. Evaluation also compares the effect of flat and frequency selective fading channel conditions on acquisition performance. MAI suppression, near-far resistance, acquisition window size and effect of chip level uncertainty on acquisition performance were also considered for performance evaluations.

- Problem of effective multi-path combining based on JAD algorithm, which has a prominent influence on initial BER performance of the receiver (but not on acquisition performance), was addressed for the first time. To solve this problem, a new structure of JAD receiver, which benefits from a suitable channel estimation and receiver refinement, is proposed. The integrated channel estimation does not need new collection of the received signal and uses the correlation matrix and information, available after the acquisition mode.

- By combining the idea of mismatch handling (based on adaptive calculation of fictitious noise) used in the acquisition step, with the receiver refinement algorithm, a novel JAD receiver structure is proposed. In addition to the structure itself, a novel formula for calculation of the final demodulation vector is introduced as well.

- BER performance of the proposed receiver was extensively evaluated in a multi-path Rayleigh mobile channel for both uplink and downlink scenarios. This was done in the presence of MAI and the near-far effect. Also, the effect of the acquisition performance on the BER performance was considered by evaluating the results using different acquisition window sizes.
In direct sequence code division multiple access (DS/CDMA) communication networks, the information-bearing signal of each user is multiplied by a unique code sequence. This code sequence is to provide orthogonality between different users and is chosen to have special properties, which facilitate demodulation of the transmitted signal by the intended receiver. Additionally, it should also make demodulation of the transmitted signal by an unintended receiver as difficult as possible. Since the spreading signal has a bandwidth much larger than the data bandwidth, the spread-spectrum transmission bandwidth is dominated by the spreading signal and actually is independent of the data signal.

Direct sequence (DS) modulation spreads the power of the original signal over a much wider spectrum simply by multiplying the spreading sequence by the original information bits in the time domain. This results in a lower spectral density and generally satisfies the two following conditions that are attributed to spread spectrum techniques [Pete95].

1) The transmission bandwidth must be much larger than the information bandwidth.
2) The transmission bandwidth should be independent of the information signal.

In the demodulation process, a conventional DS/CDMA receiver will correlate the
received signal with a locally generated replica of the user spreading code. Based on the characteristics of spreading codes, which will be elaborated later, the correlating process de-spreads the energy of desired user into a smaller bandwidth (i.e. data bandwidth). Furthermore, the energy of the undesired users remains spread in the frequency domain, and hence the desired user can be detected effectively. This process assumes that the receiver has perfect knowledge of the desired user spreading code. In addition, successful detection of the desired signal depends on the available information surrounding the correct time delay of desired signal or alternatively the phase of the spreading code.

In practical systems, this information is not available before call set-up, and must be estimated so as to support coherent demodulation.

Estimation of desired signal time delay generally known as coarse synchronisation; this process becomes more challenging if we consider the time varying characteristics of the transmission medium in mobile networks, and the presence of interfering users that transmit in the same frequency, and time interval. These interfering users, usually referred to as multiple access interference, can be much stronger than the desired one, affecting the performance of the detector [Pete95] [Vite95].

Also, due to the time varying nature of the mobile channel, the coarse synchronisation must be followed by a tracking system, typically known as fine alignment.

Different code structures have been proposed and used for different CDMA systems. Existing IS-95B standard, which is based on synchronous narrowband CDMA radio access, uses maximum length sequences, known as M-Sequences, to provide multiple access technique. On the other hand wide band CDMA, WCDMA, for universal mobile telecommunication services, UMTS, rely on asynchronous network scenario using a bandwidth of 5 MHz.

Thus, for UMTS communication network, instead of using different phase shifts of the same M-sequences for each cell, (which provides a good auto-correlation characteristics) different codes with good cross-correlation and auto-correlation
characteristics has been proposed. Most common families of these codes are Gold and Kasami codes.

The code structure further impacts on how the code synchronisation, cell acquisition and handover synchronisation are performed [Dina98].

2.1 Basic acquisition principles

Generally, both the phase and frequency of the received spread spectrum signal is unknown to the receiver. Although the uncertainty region of both parameters is continuous, however it is usually segmented into the discrete parts, generally referred to as cells. During the acquisition process each cell should be checked individually.

Figure 2.1 shows an example of two-dimensional segmentation of the uncertainty region, where each square represents a cell in time and frequency domain.

![Figure 2.1: Two-dimensional uncertainty region](image)

$\Delta T$ and $\Delta \Omega$ in Figure 2.1 show the uncertainty domains of the received signal associated with the time delay and carrier frequency respectively. The maximum size of each time domain’s cell can be one chip time duration or less. Block diagram of a basic acquisition system is shown in Figure 2.2 [Pete95]. In this general method of acquisition, the received signal is multiplied by a locally generated replica of its pseudo noise (PN) code sequence. An energy detector calculates the energy of the disspread signal in a specific frequency band. Of course this PN sequence should be the same as the one used in the transmitter for spreading the information signal. By
shifting the carrier frequency and code timing of the local PN generator, we can move within the cell space defined in the Figure 2.1, and using a specified decision rule, the decision device can decide on the correct/incorrect cells. Sometimes the correct and incorrect decision cells are referred to as hypothesis 1 (H1) and hypothesis 0 (H0). In the correct cell, the de-spreading process is done effectively and the energy of the received signal is compressed into the desired band of frequency and maximum energy is detected. In this thesis, we look at the time delay acquisition and always assume that we have perfect knowledge of the carrier frequency.

![Figure 2.2: Block diagram of a basic time delay search system](image)

**2.2 Parameter definition for a general acquisition system**

Functionality of the acquisition system in Figure 2.2, which was explained in previous section, shows the importance of the auto-correlation and cross-correlation characteristics of spreading codes. Before studying these correlation functions characteristics, some important parameters related to the context of the acquisition process are listed as follows.

a) Probability of detection ($P_d$): this value represents the probability of selecting a cell, as a correct cell while it is actually the correct one.

b) Probability of missing ($P_m$): this value represents the probability of missing the correct cell during complete search over the entire uncertainty area. It is
clear that $p_n = 1 - p_d$.

c) Probability of false alarm ($P_{fa}$): this value states the probability of selecting a wrong cell instead of the correct one (false alarm) and has a severe effect on overall acquisition time. Rejecting a falsely selected cell, generally, is more time consuming than missing the correct one during a complete search of the uncertainty region.

d) Overall acquisition time ($T_{acq}$): overall acquisition time represents the average time of acquisition. A practical acquisition system tries to keep this parameter as low as possible.

Good autocorrelation property of spreading codes is essential to maximise the probability of detection, $P_d$, while low cross correlation is essential for effectively rejecting the multiple access interference. Therefore in a conventional acquisition block, the cross correlation characteristics of codes have a significant impact on $P_{fa}$ and consequently on the overall acquisition time $T_{acq}$.

2.3 Spreading codes and their characteristics

It was already mentioned that in CDMA multiple access systems, for ease of both generation of codes and synchronization processes, the spreading waveform should be a pseudorandom sequence. This means that it can be generated as a deterministic signal that statistically satisfies the requirements of a random sequence [Vite95]. Generation and characteristics of four types of pseudorandom sequences are recalled here.

**Maximal Length Sequences:** Maximal length sequences, M-sequences, are the largest code sequences that can be generated by a linear shift register such as one presented in Figure 2.3. The generating function is given by equation (2.1). In equation (2.1) $f(D)$ is called the characteristic polynomial of the linear feedback shift register, LFSR, and specify the main characteristics of the code generator. The polynomial $g_0(D)$ depends on the initial condition of the shift register and determines the phase
shift of the generated sequence.

$$G(D) = a_0 + a_1D + a_2D^2 + \cdots = \sum_{i=0}^{\infty} a_iD^i = \frac{g(D)}{f(D)}$$  \hspace{1cm} (2.1)$$

Such a characteristic polynomial that can generate the maximal length sequence is called the primitive polynomial. Primitive polynomials exist for all degree $n > 1$. Tables of primitive polynomials can be found in [Fan96] and [Stah73] for $n \leq 40$.

As was explained above, the autocorrelation of spreading codes is very important according to the probability of false alarm. Autocorrelation function of M-sequences is a two-value parameter that can be shown as:

$$R_{nn}(k) = \frac{1}{N} \sum_{n=1}^{N} a_n a_{n+k} = \begin{cases} 1 & k = 0 \\ \frac{1}{N} & k \neq 0 \end{cases}$$  \hspace{1cm} (2.2)$$

where $a_n \in \{1,-1\}$ and $N = 2^n - 1$ is the period of M-sequence. The cross-correlation of two different codes are of similar importance. Unfortunately the M-sequences are not immune to cross-correlation and may have large cross-correlation values. Welch introduced a lower bound on this value for a sequence of period $N$ in a set of $Q$ different sequences [Welc74] as:

$$R_{am}(k) \geq N \sqrt{\frac{Q-1}{NQ-1}} = \sqrt{N} \quad k \in [0,N-1]$$  \hspace{1cm} (2.3)$$

**Gold Sequences:** Gold codes are quite important because of their proper cross-correlation characteristics and the large number of codes that one Gold generator can supply. The cross-correlation between these sequences are uniform and bounded [Gold67][Gold68]. Gold sequences have a three valued cross-correlation as:
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\[ R_{\text{acq}}(k) \in \{-t(n), -1, t(n) - 2\} \] where \( n \) is the number of delay elements in Figure 2.3 and:

\[
t(n) = \begin{cases} 
1 + 2^{(n+1)/2} & \text{for } n \text{ odd} \\
1 + 2^{(n+2)/2} & \text{for } n \text{ even}
\end{cases}
\] \quad \{2.4\}

Two special M-sequences \( a, a' \) can be used to generate Gold codes. These sequences should satisfy some rules [Dina98] and are called the preferred pairs. Figure 2.4 [Dina98] shows a Gold generator of period \( N=31 \). By adding (in module 2) the output of one of the preferred sequence generators with different phases of the output of the second generator, different members of the Gold code family are obtained. In practice different phases of the second pair are achieved by loading the related shift register with different initial states. For codes of period \( N, N+2 \) family members are available.

**Kasami Sequences:** Kasami code sequences are known to have very low cross-correlation [Fan96]. Basically, adding an M-sequence \( a \), with different shifts of another M-sequence \( a' \) or M-sequences \( (a', a'') \) will generate the small or large set of Kasami codes. Sequences \( a', a'' \) are formed by appropriately decimation of sequence \( a \) [Dina98].

![Figure 2.4: A general Gold code generator [Dina98]](image)

In this way the set of \( \{a, a'\} \) generates the small set of Kasami codes with \( Q = 2^{n/2} \) family members of period \( N = 2^n - 1 \) where \( n \) is the period of \( a \) and should be even.
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The auto correlation and cross correlation of these sequences, take on the values from the set \{-1, -(2^{n/2} + 1), 2^{n/2} - 1\}. Comparing with the Welch lower bound, it can be seen that the small set of Kasami sequences is optimal.

On the other hand the set of \{a, a', a''\} generates a large set of Kasami codes. Such a set of Kasami codes consists of sequences of period \(N = 2^n - 1\), for \(n\) even, and contains both the Gold and small set of Kasami sequences. Cross correlation and auto correlation of members of large Kasami set take on the values from the set \{-1, -1 ± 2^{n/2}, -1 ± 2^{n/2+1}\}. Comparing with the Welch lower bound it is seen that the set of long Kasami codes is not optimal.

**Orthogonal codes:** Fixed length and variable length orthogonal codes, such as Walsh Hadamard and OVSF codes (orthogonal variable spreading factor codes) are also available. These codes only provide the orthogonality between synchronous channels. For detailed specification and their usage in practical systems, e.g. in downlink and uplink of UMTS (Universal Mobile Telecommunication System) see reference [Holm02].

2.4 Classification of code acquisition schemes

The process of code acquisition and delay estimation can be classified from different point of views. These are defined in the following subsection.

2.4.1 Serial cell search

In a serial cell search, the receiver evaluates the cells of uncertainty region (e.g. cells in Figure 2.1) one by one and by comparing the output statistics with a predefined threshold, decides whether it is a correct or false cell (H1/H0). This method has a low complexity but results in a long average acquisition time, \(T_{\text{avg}}\).

Three important parameters can be defined: mean and variance of acquisition time, as well as its cumulative probability density function.

Mean and variance of acquisition time can be calculated by relating an event function
E(n,j,k) leading to a correct synchronisation with each possible correct cell. In event function \( E(n,j,k) \), "n" indicates the position of correct cell (assuming that search starts from cell number 1), "j" represents the number of missed detections and k is the number of evaluated false alarms related to this events. Then an acquisition time function \( T(n,j,k) \) and probability function \( P(n,j,k) \) can be calculated for each event. These acquisition time function, related probability and final average acquisition time were calculated in [Pete95] and are presented respectively as follow:

\[
T(n,j,k) = nT_i + jCT_i + kT_f
\]

\[
P(n,j,k) = \frac{1}{C} P_d (1-P_d)^{jC+n} \left( \begin{array}{c} jC+n \\ k \end{array} \right) P_d^k (1-P_d)^{(jC+n)-k}
\]

\[
T_{acq,av} = \sum_{n,j,k} T(n,j,k)P(n,j,k) = (C-1)T_{d,av} + \frac{2-P_d}{2P_d} + \frac{T_i}{P_d}
\]

where \( T_{d,av} = T_i + T_fP_d \) and C is the total number of cells. \( T_i \) denotes the evaluation time of every cell and depends on the evaluation strategy. In this calculation it was assumed that all cells have equal probability of being correct, and there is no uncertainty in frequency domain. Acquisition time is generally a random variable. Sometimes the variance of this variable is also important for investigating the reliability of acquisition system design. For the above acquisition system, the variance was also calculated in [Pete95] and [Chen77] and is shown by Equation 2.8.

Dicarlo [Dica77] and [Dica88] have also calculated the cumulative probability density function of acquisition time for the fixed integration time serial search strategy. However we don’t pursue these concepts here.

\[
\sigma_{acq}^2 = E(T_{acq}^2) - (T_{acq,av})^2 = \frac{C^2-1}{12} \frac{(C-1)^2}{P_d} + \frac{(C-1)^2}{P_d^2} T_{d,av}^2 \\
+ 2(C-1) \frac{1-P_d}{P_d^2} T_i^2 + 2(C-1) \frac{1-P_d}{P_d^2} T_f T_{fa} P_{fa}
\]
where the operator $E(.)$ is the expectation operator.

### 2.4.2 Parallel cell search

Parallel cell search is optimum in the sense that it provides minimum possible acquisition time with a given probability of detection, $P_d$. But this scheme requires high hardware complexity because it uses the same architecture for all cells of the uncertainty region, simultaneously (e.g. a bank of matched filters). In most cases the maximum output shows the correct cell and therefore, this acquisition strategy uses the maximum-likelihood estimation.

The complexity consideration is now partially suppressed by advances in digital signal processing techniques and very large-scale-integration (VLSI) technologies. These advances have made low-cost implementations feasible and practical. For example in [Gaud98] a non-coherent parallel SD/CA (signal detection/code acquisition) algorithm, which is suitable for implementation in an application-specific integrated circuit (ASIC) was proposed, and its performance was investigated. Some other practical issues were addressed in reference [Fanu96]. References [Rick97], [Sour92], and [Rick94] provide a good background about parallel acquisition for the reverse link of mobile CDMA systems. In these references the acquisition performance of parallel systems is investigated under some practical assumptions and the acquisition capacity is addressed based on simulation results.

### 2.4.3 Hybrid cell search

Hybrid schemes have been developed to test a group of cells (less than total cells) simultaneously and to provide tradeoffs between hardware complexity and acquisition time. Different hybrid scenarios are reported in the literature. In [Zhua96] a non-coherent hybrid parallel PN code acquisition is proposed and the effect of MAI on performance is analysed. It can be understood that the scheme provides the flexibility in trade-off between the mean acquisition time and system complexity, in
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the absence of MAI. Figure 2.5 demonstrates the concept in [Zhua96].

![Block Diagram of Non-coherent Partial Parallel PN Code Acquisition](image)

**Figure 2.5:** Non-coherent partial parallel PN code acquisition block diagram.

### 2.4.4 Fixed and variable dwell time

Dwell time ($T_d$), is the time that the acquisition algorithm needs to finish one step of the evaluation of every cell. Average dwell time in correct cell is $T_{d,av} = T_i$, while its average in incorrect cells can be represented as $T_{d,av} = T_i + T_{fr} P_f$.

The evaluation of each cell (H0/H1) can be performed in one step (single dwell time) or in some sequential steps (multiple dwell times). The idea behind the multiple dwell times is that there is always one correct cell and many incorrect cells. The receiver uses a short duration of time for evaluation (e.g. integration) in the first step, and only if it detects the current cell as the correct (H1) one, it will evaluate the cell again with longer evaluation time. This is to avoid false alarm that its rejection always takes relatively long period of time. In this way the overall dwell time will be variable and
depends on the condition of the received signal.

Variable dwell time also can be achieved through a sequential detection method. A sequential probability ratio test system (SPRT), which actually implements a variable dwell time detection has been studied in [Pete95] [Chaw94] and [Ward65]. Studies surrounding mean acquisition time of serial search using multiple dwell times, and adaptive threshold setting for double dwell architecture can be found in [Sims97] [Vejl00] respectively.

A comparison study of a two-dwell time acquisition system, which uses different threshold settings (e.g. fixed threshold, constant false alarm rate (CFAR) criterion and optimum threshold), is presented in [Iina97]. Reference [Wang00] introduces a novel method for evaluation of mean acquisition time of a non-coherent serial search acquisition system based on variable dwell time. In reference [Wang00], two cases of multiple-dwell and sequential linear tests, which can provide variable overall dwell time, are investigated. Sequential probability ratio test (SPRT), which is actually a member of variable dwell time detectors, has been proven to be optimum in the sense that it yields the minimum average detection time for a specified \( P_d \) and \( P_{fa} \) [Pete95] [Wald48]. Figure 2.6 shows a typical logic flow diagram for a multiple-dwell detector.

**Figure 2.6: A multiple dwell time acquisition system**

In Figure 2.6, each integrator corresponds to one energy detector and one decision device as in Figure 2.2 but with different integration time \( T_i \) and different threshold \( v \).
If the output statistic is lower than the threshold in a correct phase, a miss happens and the system rejects the current phase and repeats the test for a new code phase or equivalently a new cell. However, if the current cell was detected as the correct cell, then the integration is repeated with longer integration time. In this block diagram, tracking mode starts after three successful integration steps. Block \((T_4, v_4)\) shows the tracking mode, where the cell is checked repeatedly until the correct phase is missed. In this case, control is transferred to the block \((T_5, v_5)\) for a more precise decision. If this block confirms the loss of correct cell, then the cell search is started again, otherwise the current cell is still the correct one and control is passed to block \((T_4, v_4)\).

### 2.5 MAI and conventional receivers

Conventional CDMA systems deal with multiple access interference (MAI), just as an additional source of AWGN noise. These receivers either ignore the near far problem or try to limit it with power control. Near-far problem occurs when CDMA receiver is to detect a weak-desired signal in presence of strong interfering signals, which is a common phenomenon in mobile environment. Standard single user techniques such as the matched filter, active correlators and methods described in this chapter, can be classified as conventional receivers.

#### 2.5.1 BER performance of conventional receivers

It is well known that in an AWGN channel the optimum receiver for single-user case using BPSK modulation is obtained by applying the following decision rule:

\[
b' = \text{sgn}\left( \int y(t)s(t)dt \right)
\]

where \(b' \in \{1, -1\}\) is the estimated bit and \(y(t)\) and \(s(t)\) show the received continuous signal and the desired user signature waveform respectively. Obviously the above optimum receiver assumes that the signature waveform and delay of desired user are known perfectly. The decision rule is generally implemented using an active or
passive matched filter. The bit error rate, BER, performance of this receiver can be shown as: $P_{\text{err}} = Q(\sqrt{\text{SNR}})^1$.

When there are other users in the cell, the optimum receiver performance is limited by MAI as well as near-far problem. To see these issues in more detail, BER of the optimum receiver in presence of MAI is studied in the following paragraphs.

**Synchronous case:**

When there are “K” synchronous multiple access users in the cell (e.g. in downlink scenario) the probability of error, $P^x_k$, for $k_{\text{th}}$ user can be written as [Verd98]

$$
P^x_k = \frac{1}{2^{K-1}} \sum \cdots \sum \cdots \sum \cdots Q\left(\frac{A_k}{\sigma} + \sum_{j \neq k} A_j \rho_{jk}\right)
$$

The main assumptions in equation (2.10) are the equal probable BPSK modulation in AWGN channel and independent information bit streams. In equation (2.10) $A_k$, $\sigma$ and $\rho_{jk}$ are the amplitude of $k_{\text{th}}$ user, standard deviation of noise and cross correlation between signature waveforms of $j_{\text{th}}$ and $k_{\text{th}}$ users respectively. $e_j$ is the information bit of $j_{\text{th}}$ user. An interesting characteristic is the performance of the receiver when the noise level goes to zero. In this case the $P^x_k$ will vanish if and only if the argument of each of the Q-functions is positive, that is, if

$$
A_k > \sum_{j \neq k} |A_j| \rho_{jk} \tag{2.11}
$$

Condition (2.11) for error-free decisions in the absence of background noise is commonly referred to as the open-eye condition [Verd98].

**Asynchronous case:**

The main difference between synchronous and asynchronous cases in terms of BER calculation is the number of effective interferences. In asynchronous case there are actually two adjacent bits of every interfering user that contribute to detection of one bit of the desired user. The probability of error and open-eye condition for $k_{\text{th}}$ user in

$\text{Q}(x) = \int_{-\infty}^{\infty} \frac{1}{\sqrt{2\pi}} e^{-x^2} \, dx$. 

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the asynchronous case is presented by equations (2.12) and (2.13) respectively [Verd98]:

\[
P^a_k = \frac{1}{4k-1} \sum_{e_i,d_i \in \{-1,1\}} \sum_{\text{\scriptsize{}}_{\text{\scriptsize{}}}_{\text{\scriptsize{}}}_{\text{\scriptsize{}}} e_j,d_j \in \{-1,1\}} \sum_{\text{\scriptsize{}}_{\text{\scriptsize{}}}_{\text{\scriptsize{}}}_{\text{\scriptsize{}}}} A_k \sigma + \sum_{j \neq k} \frac{A_j (e_j \rho'_{jk} + d_j \rho'_{kj})}{\sigma} \tag{2.12}
\]

\[
A_k > \sum_{j \neq k} A_j (|\rho'_{jk}| + |\rho'_{kj}|) \tag{2.13}
\]

where \( \rho'_{jk} \) and \( \rho'_{kj} \) show the left and right hand partial autocorrelation functions between signature waveforms of \( k_{th} \) and \( j_{th} \) users. These functions are defined in equation (2.14). In equation (2.14), \( \tau_{j,k} \) represents the relative delay of \( k_{th} \) user in comparison with \( j_{th} \) user. \( T \) is the time duration of information bits. \( e_j \) and \( d_j \) show two consecutive bits of \( j_{th} \) user that contribute in detection of one bit of desired user (\( k_{th} \) user in this case) and \( j < k \).

\[
\rho'_{jk} = \int_{\tau_{j,k}}^T s_j(t)s_k(t-\tau_{j,k})dt
\]

\[
\rho'_{kj} = \int_{0}^{\tau_{j,k}} s_j(t)s_k(t+T-\tau_{j,k})dt \tag{2.14}
\]

2.5.1.1 Numerical example

To illustrate the behaviour of conventional receivers in the presence of multiple access interference, a synchronous CDMA system using Gold codes of duration 31 is considered. We consider a synchronous system, so that we can apply the pre-known values of cross correlations, which in this case are -0.29, -0.03 and 0.22. In the case of asynchronous system the behaviour will be worst because it is obvious that \( |\rho_{jk}| \leq |\rho'_{jk}| + |\rho'_{kj}| \) and thus if the condition (2.11) is not valid then the condition (2.13) cannot be valid as well. The cross-correlations of interfering users were selected randomly and results were averaged over several runs. Figure 2.7 shows the BER of the single user receiver for the specified conditions.

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Figure 2.7: BER of single user receiver vs. SNR for two different numbers of multiple-access interfering users. (Equal power scenario)

Figure 2.8: BER of single user receiver vs. SNR for two different numbers of multiple access interference users and near-far ratio.
In this example, all signals are equal power; the BER floor and effect of MAI are also seen. Figure 2.8 illustrates the same results, but in the presence of the near-far effect, where signal of interfering users are 5 dB and 10 dB stronger than the desired user at the receiver side.

In both figures, $P_{\text{offset}}$ denotes the difference between the power of MAI users and desire user in dB, assuming that all MAI users have the same power.

### 2.5.2 Acquisition-based capacity of conventional receivers

Acquisition based capacity is defined as the maximum number of simultaneous transmissions supported by the network, while providing acceptable acquisition performance. For conventional acquisition systems this is a function of acquisition window length. Acquisition window length of the conventional receivers is defined as the number of received signal's samples that are used to make one output acquisition statistic. Then the maximum size of acquisition window is the period of spreading code, $N$.

Usually bit error rate (BER), or signal to noise ratio (SNR), is considered as a measure of performance in the analysis of DS/SS schemes while the processing gain ($G$), or bandwidth expansion is referred as a measure of complexity.

The BER based capacity is known to be linearly dependent on the processing gain or complexity. This is based on the assumption that the knowledge of the propagation delay of the desired user's signature waveforms is available at the receiver. The following discussion, introduced in reference [Madh93], shows that if the acquisition window size, $N$, is a linear function of processing gain, $G$ (which is generally an acceptable assumption) then the acquisition based capacity is lower than the BER based capacity.

In [Madh93] an asymptotic analysis of acquisition capacity for DS/SS systems has been introduced. This analysis is based on using a passive matched filter for acquisition and is the subject of this section. In this study, the acquisition window size, which is the length of a matched filter, defines the complexity of the acquisition
process.

Assume that the desired user is sending only its signature sequence \( \{ a_j \} \) where there are no modulating information bits, and that the spreading sequences of other users are modulated by random bits. Thus the \( k \text{th} \) interfering user's signal can be modelled as independent random sequence \( \{ x_k \} \). The desired user and \( k \text{th} \) interfering user signal is shown by equations \( 2.15 \) and \( 2.16 \) respectively.

\[
\begin{align*}
r_o(t) & = \sum_{j=0}^{\infty} a_j p_{T_c}(t - jT_c - \tau_0 T_c) + n_0(t) \\
r_k(t) & = \sum_{j=0}^{\infty} x_{kj} p_{T_c}(t - jT_c - \tau_k T_c)
\end{align*}
\]

where \( p_{T_c}(\cdot) \) represents a rectangular pulse of duration \( T_c \). \( a_j \) and \( x_{kj} \) denote the \( j \text{th} \) signature sequence of the desired user and random sequence of \( k \text{th} \) user, respectively. Delay \( \tau_0 \) was restricted to \( [0 \ldots N-1] \). \( n_0(t) \) is the additive noise added to the desired user signal. For interfering users with random chips, it is sufficient to assume that \( \tau_i \) is restricted to the range \( 0 \leq \tau_i \leq 1 \). Now we can apply a chip-matched filter at the front end of the receiver to gather the statistics, as shown in equation \( 2.17 \). For simplicity it was assumed that \( \tau_0 \) is an integer number and that the receiver can acquire the carrier frequency and the phase of the target transmission perfectly.

\[
Z_j = \int_{\frac{\tau_0}{T_c}}^{(j+1)\frac{T_c}{T_c}} r(t) dt = a_{j-\tau_0} + X_j
\]

where \( r(t) = r_o(t) + \sum_k r_k(t) \) and the additive interference \( X_j \) is given by:

\[
X_j = \sum_{k=1}^{K} [(1-\tau_k)x_{kj} + \tau_k x_{k,j-1}] + n_j
\]

All delays are measured relative to the sampling time of the receiver. Now the acquisition problem consists of estimating \( \tau_0 \) based on the sequence of statistics \( Z_j \). A discrete-time filter, which is matched to a section of the desired signature sequence
of length $N$, calculates the output $W_n$ from input sequence $Z_n$. This output sequence is used to detect the correct delay.

Acquisition may be achieved by detecting the discrete time, when the matched filter output crosses a threshold. An alternative scheme is to collect the output of matched filter for a time interval, and to select the discrete delay time associated with the maximum sample as the correct delay of the strongest path.

In reference [Madh93], the probability of acquisition failure is defined as the probability of threshold crossing at an incorrect delay, or if no threshold crossing happens at all. A false alarm happens if the threshold is exceeded before the correct time shift, and a miss occurs if the threshold is not exceeded at the correct time. To simplify the calculations it is also assumed that the desired user contribute to the output of the matched filter only at the correct delay. The matched filter output at the other time delays is only constructed by the interfering users’ signal and noise.

Assume that $Y_n$ shows the interference and noise section of the matched filter output, $W_n$, at time ‘$n$’. Thus $W_n = Y_n, n < \tau_0$ and $W_n = Y_n + N, n = \tau_0$

Based on the above assumptions the probability of false alarm is given by

$$P_f = P[U_n^{\tau_0-1}\{ Y_n / N > \alpha \}]$$

(2.19)

where, $U_n^{\tau_0-1}\{ X_n \}$ shows the union of events $X_n$. The probability of miss is given by

$$P_m = P[Y_n / N < \alpha - 1], n = \tau_0$$

(2.20)

The parameter $\alpha$ is the threshold level and its selection provides a trade-off between $P_f$ and $P_m$. It should be selected based on the PDF of $W_n$, size of uncertainty region and effect of false alarm and the miss events on the overall acquisition time of the systems.

For more details about different methods of threshold setting for a conventional acquisition system, see [Sims97][Vejl00].

The probability of acquisition failure is defined [Madh93] as the probability of the union of the events considered in equations (2.19) and (2.20). It is bounded as:
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\[ P_{fa} \leq P_f \leq P_{fa} + P_m. \]

An evaluation of capacity consists of finding the maximum permissible number of interfering signals "K" as a function of the acquisition window length N, subject to the constraint that the probability of acquisition failure tends to zero as \( N \to \infty \).

The capacity is mainly determined by the probability of false alarm. For a synchronous and equal power case, the matched filter input cased by interfering users, is a sequence of independent and identical distributed symmetric binomial random variables with parameter K. The upper and lower bounds on the false alarm probability can be presented as [Madh93]:

\[ P_{fa} \leq \tau_0 \frac{P(Y_0/N > \alpha)}{\frac{1}{2} \sum_{n=0}^{N-1} \sum_{m=0}^{N-1} P(Y_m/N > \alpha, Y_n/N > \alpha)} \]  \( (2.21) \)

The matched filter output at a specific time delay, e.g. \( Y_0 \), is the sum of NK independent and identically distributed symmetric Bernoulli random variables. By using the central limit theorem and for large values of NK, the distribution function of the random variable \((NK)^{-1/2}Y_0\) can be modelled as a standard Gaussian function. Thus for large values of NK and after averaging over delay variable \( \tau_0 \), the upper bound for probability of false alarm is calculated as:

\[ P_{fa,u} = \left(\frac{N-1}{2}\right)Q\left(\alpha(N/k)^{1/2}\right) \]  \( (2.22) \)

\( \tau_0 \) is uniformly distributed in the range of \([0 \ldots N-1]\). By approximating the equation \( (2.22) \), the maximum allowable K as a function of N, which allows \( P_{fa,u} \) tend to zero as \( N \to \infty \), was calculated in [Madh93]. This was shown as \( K \leq 0.5\alpha^2(N/\ln N) \) where \( 0 \leq \alpha \leq 1 \). According to the acquisition based capacity concept, the most important point in this inequality is the non-linear relation between maximum number of users, K, and the matched filter window size N, which is in the form of \( (N/\ln N) \).

It was also addressed in the same reference, that for an asynchronous case, the lower
bound on capacity remains unchanged. It is clear that by increasing $\alpha$ from 0 to 1, the asymptotic acquisition based capacity for synchronous case tends to $0.5(N/\ln N)$. However this also increases the probability of miss. An asymptotic analysis of the $P_{as}$ in [Madh93] shows that the required condition is, $\alpha \to 1$ in such a way that $[(1-\alpha)^2 \ln N] \to \infty$.

These lead to the following conclusions:

For large number of interfering users, the acquisition window size that gives a good acquisition performance becomes relatively large. As a result, the matched filter scheme becomes impractical, since its size is currently limited by both cost and technology. Acquisition preamble increases the transmission overhead, reducing overall system throughput particularly in packet-based systems. This drawback motivates the need to study alternative acquisition schemes. The most important point is that the acquisition-based capacity (upper bound) is a non-linear $(N/\ln N)$ function of complexity. This is in contrast to the BER based capacity, which is generally a linear function of processing gain. Consequently the acquisition performance is more limited by multiple access interference than BER performance. If, for a target BER performance, the receiver needs sophisticated algorithms to combat multiple access interference, then the same applies for time delay acquisition as well. Alternatively these problems could be solved by tight control of network timing, and reducing the dimension of time delay uncertainty. However in this thesis we investigate the problem, only from points of view of the interference resistance acquisition schemes.

Figure 2.9 illustrates the upper bound of the probability of false alarm calculated from equation (2.22) ($\alpha=1$). The upper bound is plotted for large values of $KN$, which is essential for validity of the Gaussian assumption. Also, it is worth noting that equation (2.22) is only useful for low probability of false alarms, where the probability of common false alarm events in different cells can be neglected (see equation (2.21)). One can observe from Figure 2.9 that by increasing the number of users, for example from 6 to 21, and for a fixed upper bound probability of false alarm 0.05, how dramatically the size of matched filter should be increased.
2.6 Maximum likelihood code acquisition

The maximum likelihood code acquisition system is classified as a common conventional code acquisition method. It is practical for current mobile communication systems as in IS-95 and UMTS. In these systems the acquisition process is basically based on slotted mode preamble search in the reverse link (mobile terminal to the base station) and pilot search in the forward link (base station to the mobile terminal).

This section specifies the method as well as statistical characteristics of appropriate statistical decision variables. The performance of maximum likelihood, ML code acquisition technique for slotted-mode in a frequency selective Rayleigh fading channel was studied in [Rick97] and [Park98]. In this context the H1 region is defined as a group of H1 cells, which is the case assumes the multi-path fading condition is present.

In this scenario, the mobile terminal transmits an un-modulated PN sequence (the preamble) in a short duration of time. The preamble has been aligned with a pilot signal transmitted in the downlink. Thus the uncertainty about the phase of the received preamble at the receiver side depends on the maximum distance between the base station
and the edges of the cell.

For random access slots, and as a figure of merit for acquisition algorithm, the probability of detection is more desirable than the mean acquisition time. Thus the ML technique, which is optimum from the probability of detection point of view, is considered. However, it is well known that for the forward link, where the mean acquisition time is more important, serial search algorithms provide better performance. Figure 2.5 illustrates the block diagram of a conventional ML detector. The decision rule can be based on comparing the largest correlation output with a threshold or simply just selecting the largest correlation output without any comparison. Of course the distribution of decision statistics will be slightly different.

In this section, ML selection without any threshold setting will be considered. In reference [Park98] a parallel search system similar to Figure 2.5, which is based on threshold setting is analysed. Reference [Zhua96] shows a hybrid technique where uncertainty region is divided into M blocks of N phases while M different phases are searched simultaneously. However there is always a trade-off between hardware complexity and processing time. Figure 2.10 illustrates one branch of non-coherent detector shown in Figure 2.5. In Figure 2.10, \( a_{j-r} \) represents the chip sequence of locally generated spreading code, where "r" represents the delay of \( r_{th} \) branch relative to the first branch of the detector (branch "0").

\[
\begin{align*}
\sqrt{2}\cos (w_t) & \rightarrow \sqrt{2}\sin (w_t) \\
\mathbf{H}_r^I(w) & \rightarrow \mathbf{H}_r^Q(w) \\
1 = jT_c & \rightarrow \alpha_{j-r} \\
\sum_{j=0}^{N-1} & \rightarrow (.)^2 \\
\sum_{j=0}^{N-1} & \rightarrow \gamma_r \\
\end{align*}
\]

Figure 2.10: In-phase / Quadrature phase non-coherent detector

(One of the branches in figure 2.5)
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Figure 2.11 presents the structure of desired signal in slotted mode preamble search. In this figure \( T_{\text{pre}} \), \( T_m \) and \( T_w \) denote the preamble time, message time and window time between every two random access slots respectively. \( T_r \) is the reset time for a miss detection. Although there is a penalty time for false alarm events, however in this structure it is assumed that any false alarm is recovered before the next random access slot arrives and thus the penalty time and reset time are the same.

![Figure 2.11: A general example of access slot structure](image)

It is worth noting that generally \( T_o \) is a random variable and thus the reset time \( T_r \). On the other hand, time duration \( T_p \), which denotes the processing time, should be smaller than the preamble time. Most conventional time delay acquisition algorithms apply the verification mode to confirm whether the \( H_1 \) decision in the search mode was true [Rick97]. However only a single dwell system is considered here, and it is assumed that some error handling techniques, e.g. cyclic redundancy check, verify the decision before the next slot arrives. The number of search phases in the reverse link, is determined by the maximum distance between the transmitter and receiver as:

\[
N_t = \frac{2l}{c_0 T_c}
\]

where \( l \), \( c_0 \) and \( T_c \) denote the distance between the base station and mobile terminal, speed of light and chip duration, respectively.

In the reverse link, there are \( K \) users per cell, which their transmissions arrive with independent random carrier phases and independent time delays. Before considering the
effects of fading, shadowing and power control error, the complex signal of the $k$th user can be written as:

$$ s_k(t) = \sqrt{2} \exp(j(w t + \theta_k)) \sum_{l=-\infty}^{\infty} b_{k,l} P_{t_k}(t - lT_c - \tau_k) $$

(2.23)

$\theta_k$ and $\tau_k$ represent the random carrier phase and time delay of $k$th user respectively, where $k=1...K$. $P_{t_k}(t)$ is the chip shape waveform. The $b_{k,i}$ shows the $i$th chip of $k$th user.

The period of chip sequence generally assumed to be larger than the acquisition time interval. It was shown [Park98], [Rick97] that the performance of long sequences is significantly better than the performance of short sequences. We assume un-modulated spreading sequence for desired user and random modulating bit sequences for other users. This assumption results in a deterministic sequence $b_{d,i}$ for the desired one, and random sequence $b_{r,i}$ with equal probability of +1 and -1 for the remaining users. In this case, we can also assume that the delay of all interfering users compared to the desired one, is restricted to the duration of one chip, and the delay of the first path of desired user is zero. This means, with reference to the desired user signal, the system is chip synchronous. In chapter 4, it is shown that chip asynchronous effect can be modelled as loss of power. For BPSK modulation maximum loss of 3 dB would be expected.

The fading processes for different users are modelled, as independent processes. This is a reasonable assumption if we assume that the distance between every two users is more than one wavelength of the carrier frequency. Finally the received signal can be presented as:

$$ r(t) = r_0(t) + I_i(t) + I_o(t) + n_o(t) $$

where $r_0(t)$ is the desired signal. $I_i(t)$ and $I_o(t)$ represent the intra-cell and inter-cell interference respectively and $n_o(t)$ is the additive white Gaussian noise. By considering the effect of power control error $\lambda$, voice activity $\varphi$ and multi-path effect, these terms can be rewritten as:

$$ r_0(t) = \lambda \sqrt{P_0} \sum_{p=0}^{L-1} c_{1,p} s_1(t - pT_c) $$

(2.24)
\begin{align*}
I_i(t) & = \sum_{k=2}^{K} \varphi_k \lambda_k \sqrt{P} \sum_{p=0}^{L-1} c_{k,p} s_k(t - pT_c) \\
I_o(t) & = \sum_{k=R+1}^{6K} \varphi_k \lambda_k \sqrt{P} \left( \frac{\xi_{k,m}}{\gamma_k^o} \right)^2 10^{\log a_k - \log a_{k,m}} \sum_{p=0}^{L-1} c_{k,p} s_k(t - pT_c) 
\end{align*}

$L$ is the maximum number of paths, $c_{k,p}$ is the channel coefficient of $k$th user in $p$th path. $P$ and $P_o$ are power of interfering users and desired user respectively. $\xi_{k,0}$ and $\xi_{k,m}$ show the shadowing random variables of $k$th user related to the desired user cell and $k$th user cell respectively (see Figure 2.12). Power control error of $k$th user, $\lambda_k$, is modelled as a log-normally distributed random variable with standard deviations $\sigma_1$ and $\sigma_1$ for desired and interferer users, respectively. However this error is constant over acquisition observation interval. The reason behind choosing different variances for power control error is that the desired user, which is in acquisition mode, cannot benefit from closed loop power control and its power control error statistics is different from other users. Voice activity is also a random variable and its variance is defined by $E[\varphi_1^2]$.

The shadowing terms for the $k$th user are modelled as log normally distributed random variables. Thus $\xi_{k,0}$ and $\xi_{k,m}$ are Gaussian random variables with standard deviation $\sigma_1$. The shadowing variables are generally correlated but for simplicity they are considered uncorrelated.

Figure 2.12 shows the geometry of desired user cell and its first layer of surrounding cells.

![Figure 2.12: The layout of the cell of interest and the first layer of interfering cells](image-url)
Chapter 2: Time delay acquisition principles

After base band conversion and assuming the ideal shape for chip pulse (no inter chip interference) the in phase and quadrature phase sequence of signal component can be presented by:

$$S(iT_c) = \lambda_1 \sqrt{P_0} \sum_{p=0}^{N-1} [c_{r,p} \cos(\theta_l) + c_{i,p} \sin(\theta_l)] b_{l-p}$$  \hspace{1cm} (2.27)

Sampling rate in (2.27) is the same as chip rate (see Figure 2.10). By passing the signal of equation (2.27) through parallel matched filters, the following statistics in each branch is obtained, where it was assumed that the desired signal contributes only at the output of the first “L” branches.

$$S(m) = \begin{cases} \lambda_1 \sqrt{P_0} T_c \sum_{j=0}^{N-1} [c_{r,m}(jT_c) \cos(\theta_l) + c_{i,m}(jT_c) \sin(\theta_l)] & \text{if } m \leq L \\ 0 & \text{if } m > L \end{cases}$$  \hspace{1cm} (2.28)

where $c_{r,m}$ and $c_{i,m}$ represent the real and imaginary parts of fading coefficients of the $m_{th}$ path of the first user. In reference [Rick97] these statistics have been derived for in cell and other cell interference as well.

In summary, the output of each branch of the receiver can be written as:

$$\gamma_m^2 = (S_c(m) + I_{i,c}(m) + I_{o,c}(m) + n_c(m))^2 + (S_s(m) + I_{i,s}(m) + I_{o,s}(m) + n_s(m))^2$$  \hspace{1cm} (2.29)

where the indexes “c” and “s” represent the in-phase (cosine) and quadrature phase (Sine) lines of the $m_{th}$ branch. We see that all the in-phases and quadrature phases, as well as the in-cell and out-of-cell interference terms in equation (2.29) are independent random variables. By applying the central limit theorem, it is straightforward to see that in the right hand side of equation (2.29), the square of two independent Gaussian random variables are added together. Therefore, for a known power control error value and fading coefficients of the desired user (this assumption results in a deterministic term for the desired user’s signal), $\gamma_m^2$ is a chi-square distributed random variable with two degree of freedom [Papo91]. Based on the assumptions that were already made, the signal part in equation (2.29) exists, only while “m” belongs to the $H_1$ region, and in the other cases this term vanishes. False alarm happens if the output of one of the branches where
m > L becomes bigger than the output of the branches belong to the H1 region, m ≤ L. Finally, the probability of false alarm is calculated as equation (2.30). The detail of these calculations can be found in reference [Rick97]. For calculating equation (2.30) the power control error of desired user, \( \lambda_1 \), is still assumed constant value.

\[
P_f(\lambda_1) = (N_r - L) \int_0^\infty x \exp(-\frac{x^2}{2}) \cdot \left[ 1 - Q\left(2\lambda_1^2E I 1 + K_r \frac{1}{\sigma_0^2} \sqrt{\frac{x^2}{\sigma_i^2}}\right) \right] \\
\times \prod_{i=1}^{M-L-1} \left[ 1 - \exp(-\frac{x^2}{2\sigma_i^2}) \left(1 - \exp(-\frac{x^2}{2})\right)^{M-L-1} \right] dx \tag{2.30}
\]

Definitions of the parameters in the above equation are as follows.

\( N_r \): Number of parallel branches or equivalently the number of chip phase uncertainty.

\( L \): Number of multi-path (separated by one chip delay)

\( \lambda_1 \): Power control error for desired user. \( \lambda_1 = 1 \) shows perfect power control according to the path loss and shadowing only.

\( E \): Desired user bit energy

\( I \): Total power of interference (intra cell and inter cell interference) plus power of noise. Accuracy of this parameter depends on the accuracy of models that were considered for parameters such as voice activity, cell sectoring, shadowing, interfering users and chip shaping. The reference [Rick97] provides a good vision for these parameters but here we just limit ourselves to the overall vision.

\( K_r \): For Rician channel model, where there is a line of sight path, \( K_r \) denotes the ratio of direct line's power to total diffuse power. For a Rayleigh channel \( K_r \) is zero.
\( \sigma_j^2 \): Some indication of the desired user energy to the total interference and noise power over the \( j \)th path. This parameter is given by [Rick97] as

\[
\sigma_j^2 = 1 + \lambda \frac{E}{I} \Delta \beta_j,
\]

where parameter \( \Delta \) represents the effect of averaging over observation window N, on the variance of fading. In many practical cases, where the acquisition observation window is very short in comparison with the coherency time of the channel, this parameter can be set to one.

Also, the parameter \( \beta_j \) represents the variance of fading in the \( j \)th path, where it was assumed that the total variance of fading over all multi paths is unit. Clearly \( \beta_j \) depends on the power delay profile of the channel.

### 2.6.1 Interference resistant delay acquisition algorithms

Recently some methods of code acquisition have been proposed that try to suppress the effect of interference by using training sequences or exploiting the structure of interference etc. These algorithms can be classified as interference resistant and near-far resistant methods and are subject of this thesis. Most of these methods are trying to adjust the coefficients of a finite impulse response, FIR, filter to minimise the effect of interference at the output. FIR filters can be adapted by using linear algorithms such as least mean square, LMS, recursive least square, RLS [Tarh98] [Miya97] [Mura97] or non-linear algorithms like MUSIC (Multiple Signal Classification) [Bens96][Stro96]. These methods can be implemented using training sequences [Madh98] or through blind approaches, where the only available information at the receiver side, regarding to the desired user, is its spreading code [Derr98][Madh97]. A detail literature survey of these methods will come in the next chapter.

### 2.7 Generalised transfer function

Serial, parallel or hybrid cell-searches can be modelled as a Markov chain. A transition from one state to another, during evaluation of every cell (e.g. transition between blocks of Figure 2.6 or transition between different cells of Figure 2.1)
happens with specific probability and takes a specific duration of time.

These quantities are related to the method of evaluation, system parameters as well as communication environment characteristics. Figure 2.13 shows the chain for a two dwells cell-search scenario, where cell evaluation starts from state number one.

If, after phase evaluation in state "1", the current cell is detected as correct one, control is passed to the state "2". $P_{12}$ shows the probability of this transition and $T_{11}$ is the corresponding time. $P_{10} = 1 - P_{12}$ is the probability of miss in state "1" if we assume that the cell, which is under test is actually the correct one. In this figure, state "3" represents the tracking state where $P_{23}$ is the probability of detection in state "2" that leads the system to the tracking mode. Blocks "0" and "4" represent boundary states and their related probabilities, $P_{00}$ and $P_{44}$ are equal to one. Other probabilities and times are understood by comparing Figure 2.6 and Figure 2.13. Of course two different state diagrams should be calculated related to the false and true cells.

In the following diagram, a path-factor is associated to each path (e.g. $P_{12z^7}$) and a path-function $B(\ell, z)$ is defined as the product of all path-factors on a specific path $\ell$, between two states. Path-function, clearly, is a function of $\ell$ and $z$.

A transition function $H_z$ between two states is defined as sum of all possible path-functions between these two states.

![Figure 2.13: State transition diagram for a specific code phase](image-url)
Thus the overall transfer function can be written as [Pete95]:

\[ H(z) = \sum_{l \in L} B(l, z) \]  

(2.31)

In (2.31) \( L \) shows the number of possible paths between two specified states and \( B \) is the related path transfer function. The average transition time between two states \( "i" \) and \( "j" \) can be presented as:

\[ T_{i,j} = \sum_{l \in L} \frac{d}{dz} (B(l, z))_{z=i} \]

(2.32)

Please note that if the transition diagram is designed to shows the evaluation of a correct cell, then \( T_{i,j} \) is the average time required to evaluate the correct cell and corresponds to \( T_i \), where \( "i" \) and \( "j" \) are start and tracking states respectively. On the other hand if the diagram is designed to show a false cell and the probability values and associated times are set accordingly, then the same \( T_{i,j} \), where \( "i" \) and \( "j" \) are the start and boundary states, shows the average time necessary to reject an incorrect cell.

Also the probability of detection, \( P_d \), can be easily shown as:

\[ P_d = \left( \sum_{l \in L} B(l, z) \right)|_{z=1} \]

(2.33)

where the transfer function is calculated between the start and the tracking states in a correct phase cell. By using this method, the average acquisition time of a linear cell search, such as one explained in section 2.6, can be easily investigated.

This concept has been generalised by Dicarlo, Weber and Polydoros [Dica77], [Poly83], [Poly84] and is applicable to any method that can be shown by a general diagram like Figure 2.11.

Figure 2.14 illustrates a state transition diagram representing the entire direct-sequence code synchronisation process. For a CDMA receiver, having “C” cells in code phase uncertainty region, there is “C+2” states in the transition diagram, where “C-1” of these states correspond to “C-1” incorrect code phases and one state represents the correct phase (state “C”). Also one state shows the acquisition state (state “Acq”) and one state is dedicated to starting point. Receiver starts from a
randomly selected code phase with probability of $\Omega_i$. The synchronisation/acquisition block uses one of the detection techniques to evaluate the current cell. The transfer function between each pair of cells is calculated as explained before.

It is easy to investigate that the transfer function $H_i(z)$ between $i_{th}$ cell and acquisition state can be represented as follows.

$$H_i(z) = \Omega_i H_x(z)H_0(z)^{-i} \frac{1}{1 - H_y(z)(H_0(z))^{e-i}}$$  \hspace{1cm} (2.34)

The desired function $H(z)$ is sum of $H_i(z)$ and can be summarized as equation (2.35).

$$H(z) = \sum_{i=1}^{C} H_i(z) = H_x(z) \frac{1}{1 - H_y(z)(H_0(z))^{e-i}} \sum_{i=1}^{C} \Omega_i[H_0(z)]^{-i}$$  \hspace{1cm} (2.35)

Having found the function $H(z)$, the average acquisition time is calculated as

$$T_{acq,av} = \left[\frac{d}{dz}H(z)\right]_{z=1}$$  \hspace{1cm} (2.36)

These results are applicable to any search system, which can be represented by a state-transition diagram like what is illustrated in Figure 2.14.

![Figure 2.14: The general transition diagram for acquisition process](image-url)
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The transfer functions in the above figure are defined as:

\[ H_q(z) = \text{Transfer function from incorrect cell to another incorrect cell.} \]

\[ H_d(z) = \text{Transfer function from correct cell to acquisition state.} \]

\[ H_e(z) = \text{Transfer function from correct cell to incorrect cell.} \]

2.8 Summary

The principals of time delay acquisition for DS/CDMA signal have been studied in this chapter. The key role of time acquisition and effective parameters on its performance were addressed. Different methodologies for coarse alignment were considered and a common strategy of code acquisition based on access slot structure, known as parallel Maximum likelihood acquisition, was presented in more detail.

Limitation of conventional receivers according to the both acquisition based capacity and BER based capacity was studied and it was shown that code acquisition capacity is more sensitive to MAI and the near-far effect than BER based capacity. This motivates the study for interference suppression receivers for joint acquisition and demodulation of CDMA signal (JAD), approach. In a JAD algorithm, the output of acquisition mode, which should be an interference resistant process, is an interference suppression demodulator.

Also, the concept of general transition diagram for calculating the average acquisition time was elaborated in this section. This transition diagram will be later used for time delay analysis of the studied approaches.
Chapter 3

3 Adaptive interference suppression receivers in fading channel

In the previous chapter it was discussed that the conventional methods of CDMA signal reception, neglect multiple access interference (MAI), and near-far effect. As a result, both the BER capacity and the acquisition capacity of CDMA communication systems is effectively limited.

To overcome these issues, the current CDMA standard (IS-95), and third generation universal mobile telecommunication systems, known as 3G standard for UMTS, use closed loop power control, which has several drawbacks; the overhead associated with the closed loop power control may become too much for the future packet-based CDMA systems. Furthermore, in ad-hoc wireless networks with time varying topologies it could be difficult to satisfy the closed loop power control requirements due to need for proper coordination between transmitter and receiver. Therefore this has motivated study into interference resistant algorithms for joint acquisition and demodulation of data.
3.1 Overview of multiuser receivers

The effective way of reception of CDMA signal in terms of tackling the near-far problem and MAI issue is based on multi-user detection. Maximal likelihood multi-user receiver is an optimal receiver proposed by Verdu [Verd86], which requires joint channel and data symbol estimation, and consequently is far too complex to be implemented in a real system. Other sub-optimal receivers such as [Lupa89], [Vara90], [Duel93] and [Xie90] have lower complexity, typically linear in relation to the number of active users, but all of them are centralized. In these schemes, the receiver front end is a bank of filters, which are matched to the signature waveforms of active users. The matched filter outputs collectively provide sufficient statistics for making joint decision about all the users’ symbols.

Multiuser detection receivers can be implemented as standard post-combining or pre-combining receivers.

In a post-combining receiver, the multiuser detector is implemented after multi-path combining while in a pre-combining receiver the multiuser detection is implements before multi-path combining.

Figure 3.1 illustrates general post-combining and pre-combining receiver structures.

![Figure 3.1: (a) Post-combining (b) Pre-combining interference suppression receivers.](image)

On the other hand whenever we are interested in reception of only one user (desired user), centralised multiuser detectors will be too complex in terms of both...
implementation and processing of filters outputs. By using adaptive implementation of multiuser detectors for a certain class of DS-CDMA signals, which is based on short spreading codes, single user interference suppression receivers can be obtained.

The following: [Abdu94] [Madh94] [Mill95] and [Rapa94] have done extensive work, which are all based on Minimum Mean Square Error (MMSE) criterion. All these methods use initial training sequence for initial adaptation and assume that knowledge about the coarse timing of the desired user is available.

However, using a training sequence has its own disadvantage that leads to the concept of “blind adaptive receivers for joint acquisition and demodulation”. The following provides scenarios where blind acquisition techniques could have useful application:

- **Sudden channel variations**: in mobile environments with fast channel variations, caused by transition of interference pattern or while multi-path components experience rapid and independent deep fades, connection failure becomes very probable. Recovery from tracking failure may need a high level of preamble overhead for attaining new coarse delay estimation and receiver adaptation.

- **Packet-based CDMA networks**: in packet based CDMA networks with high mobility, it is possible that time delay and channel conditions change effectively between two successive groups of packets. It might be impossible for tracking methods to follow these stepwise transitions. In addition, the preamble section, which is necessary for adaptation or delay estimation of every single packet group, results in huge overhead.

- **Highly frequency selective channels**: where there are a large number of uncorrelated multi-paths, even in the downlink and in presence of perfect power control, the received signal of the desired user suffers from the near-far effect. This near-far effect is caused by the interference signals coming from different paths with different delays.

The linear minimum mean square error (LMMSE) receiver, is one option that can be
implemented adaptively. Similar to other interference suppression receivers in fading channels, an LMMSE post-combining receiver actually tries to minimize the cost function $E(|\tilde{b} - \hat{b}|^2)$, where vector $\tilde{b}$ and $\hat{b}$ represent the data bit vector of users number "1" to "k", and their estimation respectively. Estimated vector $\hat{b}$ can be provided using either a training sequence or detected bits, which are available at the output of receiver. Post-combining LMMSE receiver in fading channel depends on the complex coefficients of the channel [Latv98] and should be changed while channel coefficients are changing. This suggests that the adaptive version of the post-combining LMMSE receiver has convergence problems in fast fading channels.

By changing the optimisation criterion to minimise the $E(|\overline{h} - \hat{h}|^2)$ instead of $E(|\tilde{b} - \hat{b}|^2)$, the pre-combining LMMSE receiver is achieved. Vector $\overline{h}$ represents the active users symbol vector, coming from different paths and affected by fading channel coefficients, and $\hat{h}$ is the local estimation of vector $\overline{h}$. In this scenario, the effect of multi-path channel is taken into account in the optimisation criterion, and the final receiver depends on the average power profiles of the channel, and not on instantaneous values of the channel coefficients [Latv98].

### 3.2 Joint time delay acquisition and demodulation of data

The acquisition capacity and its effect on the overall capacity of a system were discussed in chapter 2. It's clear that acquisition process suffer from the same issues as demodulation process, and thus the ability of joint acquisition and demodulation of CDMA signal by adaptive interference suppression receivers is an interesting concepts that needs to be investigated in a mobile multi-path fading environment.

Joint acquisition and demodulation, means that the output of the acquisition mode should provide an interference resistant receiver vector, which is ready to be used for detection of, transmitted symbols. This receiver should also be near far resistant. On the other hand, the acquisition process itself should be resistant against the MAI and the near far-effect.
In this thesis we are interested in an interference suppression single user receiver, where the desired signal is subject to synchronisation and demodulation. Thus the idea of joint acquisition and demodulation of CDMA signal will be studied hereafter.

3.3 Adaptive pre-combining LMMSE single user receiver.

In a pre-combining single user receiver, the optimisation criterion can be implemented for each path of the desired user, separately. It means that the criterion for the $i_{th}$ path is:

$$\text{MSE} = E\left\{ |h_i(n) - h_i'(n)|^2 \right\}$$

where $h_i(n) = c_i b(n)$, $c_i$ is the complex channel coefficient for $i_{th}$ path of the desired user and $b(n)$ is the $n_{th}$ symbol. Please note that the effects of path loss and fading have been considered in $c_i$. It is clear that we need one adaptive filter for each Rake finger. Also the receiver should know the multi-path delays and spreading code of the desired user. However in a joint acquisition and demodulation scenario we have to lift the assumption relative to the known multi-path delays. Figure 3.2 illustrates a general pre-combining LMMSE receiver where, $L$ is the number of resolvable paths according to the channel model.

The FIR filter is a taped-delay line filter specified by its coefficient vector $\tilde{f}$ where $\tilde{f} = [f_0, f_1, \ldots, f_{W-1}]^T$ and "W" is the number of filter taps, or the length of the received signal block $\tilde{r}(n)$ that contributes to every loop of the adaptation algorithm. The operator $(\cdot)^T$ denotes the transpose operation. Using the MSE criterion (3.1), it is well known that the optimum filter vector $\tilde{f}$ can be calculated as [Hayk85]:

$$\tilde{f}_{\text{opt}} = \tilde{R}_{\tilde{r}\tilde{r}}^{-1} \tilde{r}_{\text{in}}$$

where $\tilde{R}_{\tilde{r}\tilde{r}} = E[\tilde{r}(n)\tilde{r}^H(n)] \in C^{W\times W}$ is the autocorrelation matrix of sampled vector $\tilde{r}(n) \in C^W$. $\tilde{r}_{\text{in}} = E[\tilde{r}(n)d_i'(n)]$ is the cross correlation between received vector $\tilde{r}(n)$ and the desired response $d(n)$ over $i_{th}$ path. The operators $(\cdot)^H$ and $(\cdot)^*$ denote the Hermitian
transposition and complex conjugation respectively. In figure 3.2 we have $d_i(n) = \tilde{h}_i(n) = c_i b(n)$, where “$i$” is the channel path index and “$n$” is the iteration index. Also $\tilde{r}(n)$ is represented as $\tilde{r}(n) = [\tilde{r}[n W], \tilde{r}[n W + 1], \cdots, \tilde{r}[n W + W - 1]]^T$.

In an adaptive calculation of the filter vector $\tilde{f}_{\text{mse-opt}}$, the filter weights are updated from the initial conditions, by moving in the negative direction on the error surface of the gradient vector. This is called the steepest descent and is shown as:

$$\tilde{f}_i(n+1) = \tilde{f}_i(n) - \frac{\mu}{2} \nabla e_i(n) \quad (3.3)$$

where the gradient of MSE is calculated as:

$$\nabla e_i(n) = -2\tilde{r}_{e_i} + 2\tilde{R}_{r} \tilde{f}_i(n).$$

The steepest descent algorithm is based on perfect knowledge of the gradient vector, which in turn assumes that we have knowledge of the expectation of the auto-correlation matrix and cross-correlation vector. In practical systems, there is a need to estimate these values based on instantaneous values of the received signal.

The least mean squares (LMS) algorithm uses the approximation in equation (3.4) resulting in (3.5).

$$\nabla e_i(n) = -2\tilde{r}(n) d_i^*(n) + 2\tilde{r}(n) \tilde{r}^H(n) \tilde{f}_i(n) \quad (3.4)$$

$$\tilde{f}_i(n+1) = \tilde{f}_i(n) + \mu e_i^*(n) \tilde{r}(n) \quad (3.5)$$

Figure 3.2: An adaptive pre-combining LMMSE receiver structure.
3.4 Adaptive post-combining LMMSE single user receiver.

In a single user adaptive post-combining LMMSE receiver, the interference suppression algorithms are applied after channel combining. Therefore only one interference suppression filter is necessary and the MSE criterion is changed from equation (3-1) to

\[
MSE = E[|b(n) - b'(n)|^2]
\] (3.6)

Figure 3.3 illustrates the block diagram of the post-combining receiver. Also, in the forthcoming sections we will see that in contrast to the pre-combining receivers, post-combining receivers can be used when there is no knowledge of desired user time delay and thus are applicable for joint acquisition and demodulation of CDMA signals.

The optimum filter coefficients, and the different adaptive algorithms follow the same relations as defined in (3.3) and (3.5). However the error \( e(n) \) is now defined as \( e(n) = b(n) - b'(n) \).

Two other important parameters of an LMMSE algorithm are step size, \( \mu \), and the final MSE. Large step size increases the convergence speed, but also increases the final MSE, in addition to the probability of algorithm instability. Alternatively, a small step size requires a long training bit sequence to obtain a reasonable level of convergence.

![Figure 3.3: An adaptive post-combining LMMSE receiver structure](image-url)
To guarantee the stability of LMS algorithm, \( \mu \) should be selected to satisfy the condition \( 0 < \mu < \frac{2}{\lambda_{e_{\text{max}}}} \), where \( \lambda_{e_{\text{max}}} \) is the largest eigen-value of correlation matrix, \( \overline{R}_n \). In mobile channel, \( \lambda_{e_{\text{max}}} \) is time varying. A tighter limit on \( \mu \) to guarantee the stability, is \( 0 < \mu < \frac{2}{\sum \lambda_{e_{i}}} \) where \( \sum \lambda_{e_{i}} = \text{trace}(\overline{R}_n) \). The MSE after \( n \) iteration is given as:

\[
\text{MSE}(n) = \text{MSE}_{\text{opt}} + \text{MSE}_{\text{ex}}(n)
\]

where \( \text{MSE}_{\text{opt}} = 1 - \tau^* \tilde{\tau} \) and \( \text{MSE}_{\text{ex}}(n) = \text{trace}(\overline{R}_n E[(\tilde{f}_{\text{opt}} - \overline{f}(n))(\tilde{f}_{\text{opt}} - \overline{f}(n))^H]) \).

The relation between, \( \text{MSE}_{\text{opt}} \) and the final error \( (n \to \infty) \), \( \text{MSE}_{\text{err}} \), is as follows [Hayk85]

\[
\text{MSE}_{\text{opt}} = \frac{\text{MSE}_{\text{err}}}{1 + \sum_{i=1}^{\infty} \mu \lambda_{e_{i}}/(2 - \mu \lambda_{e_{i}})} \tag{3.8}
\]

The effect of step size on the steady state mean square error is clear from equation (3.8). In equation (3.8) \( \lambda_{e_{i}} \) denotes the \( i \)th eigen-value of auto correlation matrix \( \overline{R}_n \).

The LMMSE receiver adaptation process can be implemented by using a known training sequence or by blind receiver adaptation. Moreover, both methods can be implemented without pre-knowledge of the delay spread of the channel, to achieve the interference suppression receiver through the acquisition mode. These concepts will be introduced and investigated in detail during the next chapter. The LMMSE algorithm and its derivations belong to the family of linear adaptation algorithms.

In a mobile multiuser channel, an adaptive algorithm converges if the statistics of multiple access interference don't change, at least during the adaptation process. This assumption is satisfied, if we assume short spreading codes. The following sections speak about the characteristics of mobile channel, and the structure of the CDMA signal based on the short spreading code assumption.
3.5 Channel model

Power fluctuations and signal spread are two important phenomenons that radio signal experience during propagation through the mobile channel. The received signal power consists of three main components:

1-Path loss: the mean received signal power is a function of distance, “d”, between transmitter and receiver. This loss is a function of d^n, where “n” is dependent on the type of environment. For free-space n = 2, increasing the number of obstructions as well as the operating frequency, will increase the path loss and the value of “n”. In the following signal model for the acquisition purpose, this effect is represented just by a constant multiplicative factor, because acquisition is a short-term scenario and the path loss can be assumed to be invariant over this time interval.

For different environments and carrier frequencies, practical path loss models are available in literatures that mainly are used to calculate the coverage range for specific services. As an example, the Okumura-Hata propagation model for an urban macro cell with base station antenna height at 30 m, mobile antenna height at 1.5 m and carrier frequency 1950 MHz, [Holm02] [Saun99] proposed:

$$L_p = 137.4 + 35.2 \log_{10}(d) \text{ dB}$$

where L_p is the path loss in dB and “d” is the distance in km. For suburban area, the same model with 8 dB correction factor is assumed [Holm02]. This results in the following path loss:

$$L_p = 129.4 + 35.2 \log_{10}(d) \text{ dB}$$

2-slow fading: slow fading or shadowing is caused when the signal is obstructed by terrain configuration and man-made structures around the receiver. A lognormal probability distribution function (PDF) is mostly accepted to model the shadowing, with a mean as a function of path loss and its standard deviation, typically in the range of 5-12 dB.

The PDF of the slow fading process and the related power spectral density function can be shown as follows [Patz94][Kran90]:
\[ P_s(r) = e^{\pi^2 \sigma_s^2} \quad \text{(3.11)} \]

\[ S_G(f) = \frac{1}{\sqrt{2\pi \sigma_s^2}} e^{-\frac{f^2}{2\sigma_s^2}} \quad \text{(3.12)} \]

In equation (3.11), \( P_s(r) \) shows the probability density function of a Gaussian process with zero mean and unit variance. The parameters “m” and “s” are set to the mean and variance of the final lognormal process. \( S_G(f) \) denotes the power spectral density of the Gaussian process \( p_s(r) \) and \( \sigma_s \) is its standard deviation.

3- fast fading: fast fading is caused when multiple copies of the signal arrive at the receiver from different paths, and with different amplitudes and phases. The resultant signal may add up constructively or destructively, which can cause a wide range of power fluctuations within a few wavelengths. According to the study of acquisition performance, this is the most important effect of channel that should be considered in signal modelling and simulation results. The effect of fast fading over each resolvable path is generally represented by a complex coefficient, whose real and imaginary parts are two uncorrelated real normal processes with zero mean. Thus the envelope of this process has a Rayleigh distribution, when there is no line of sight (LOS) component, and its phase is a uniformly distributed random variable in domain \([0...2\pi]\). In presence of LOS component, the PDF of the amplitude of the fading process has a Ricean distribution. The equation (3.13) and (3.14) represent the Rayleigh and Ricean PDF respectively [Holm02].

\[ \begin{align*}
P_R(r) &= \begin{cases} 
\frac{r}{\sigma^2} \exp\left(-\frac{r^2}{2\sigma^2}\right), & r \geq 0 \\
0, & r < 0 
\end{cases} \\
\end{align*} \quad \text{(3.13)} \]

\[ \begin{align*}
P_R(r) &= \begin{cases} 
\frac{r}{\sigma^2} \exp\left(-\frac{r^2 + b_0^2}{2\sigma^2}\right) I_0(b_0 r / \sigma^2), & r \geq 0 \\
0, & r < 0 
\end{cases} \\
\end{align*} \quad \text{(3.14)} \]

In equations (3.13) and (3.14), \( \sigma^2 \) and \( b_0^2 \) denote the average power of scattered component and LOS component respectively. \( I_0(x) \) is the zero order Bessel function.
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of the first kind.

In addition, we need the power spectral density of the complex Gaussian process to be able to model the time variant characteristics of the channel, which is caused by mobility of receiver/transmitter. The Jakes power spectral density function [Jake74] is widely accepted to represent the desired spectral of the Rayleigh channel (NLOS) and is given as follows:

$$S_c(f) = \begin{cases} \frac{1}{\pi f_{d_{\text{max}}}} \sqrt{1 - \left(\frac{f}{f_{d_{\text{max}}}}\right)^2} & |f| < f_{d_{\text{max}}} \\ 0 & |f| \geq f_{d_{\text{max}}} \end{cases}$$

Inverse Fourier transform of equation {3.15} gives the corresponding autocorrelation function of the fast fading process as:

$$r_c(t) = J_0(2\pi f_{d_{\text{max}}})$$

Signal spread is also classified in three categories.

1- **Delay spread**: impulse response of multi-path propagation channel is an expanded signal over time. The delay spread is usually represented by a power delay profile and depends on the type of propagation environment (i.e. macro cell, micro cell or Pico cell) [Holm02].

2- **Angle spread**: due to the multi-path (caused by reflection, diffraction and scattering), different copies of the transmitted signal arrive at the receiver from different angles. This is known as the angular spread of channel. Generally speaking, angular spread is more significant at the mobile terminal in comparison to the base station.

3- **Frequency spread (Doppler shifts)**: when the mobile terminal or its surrounding objects are moving, arriving signals from different paths see different relative velocities and consequently, their carrier frequencies are shifted according to their relative speeds. Therefore a transmitted signal tone, will have a frequency bandwidth ($f_{d_{\text{max}}}$), at the receiver side, where $f_{d_{\text{max}}}$ denotes the maximum frequency spread of the channel.
Although the mobile channel can be represented by different types of models, i.e. geometrical or statistical models [Jake74], in this thesis we only use the GWSSUS (Gaussian Wide Sense Stationary Uncorrelated Scattering) model, which is quite acceptable for hilly or urban environments.

![Figure 3.4: Spreading of the signal in time, angle and frequency](image)

This model assumes that the channel consists of a few distinct dominant paths, while the signal arriving from each path is a superposition of many scattered signals. Different paths see independent random attenuations, phase changes and time delays. According to each scattering object, the delays are not resolvable and it means that the signal coming from different scattering objects, in comparison to the channel bandwidth, is narrowband. Therefore the overall signal, coming from one scattering object, will observe flat fading, while the channel itself is a frequency selective fading channel.

It is well known that for a GWSSUS channel, the time invariant channel impulse response can be presented as:

$$ c(t) = \sum_{l=1}^{L} c_l \delta(t - \tau_l) $$  \hspace{1cm} (3.17)

where \( L \) is the maximum number of resolvable paths and \( c_l \) and \( \tau_l \) represent the random complex coefficients of fading and time delay of the \( l \)th path, respectively. Average power of \( c_l \) and time delay \( \tau_l \) are defined by discrete form of channel power delay profile.

Considering a mobile environment, path coefficients \( c_l \) are actually time variant. This
means that for a longer period of time, the channel should be considered as a linear time variant filter [Proa01]:

\[ c(t_0, t) = \sum_{i=1}^{\ell} c_i(t) \delta(t_0 - \tau_i(t)) \] (3.18)

### 3.6 Signal model

For the purpose of this thesis, we consider an asynchronous DS/CDMA system with "K" simultaneous antipodal users over a multi-path GWSSUS channel. To be able to follow the interference suppression approaches for time delay acquisition, we only consider short spreading codes, which means that the period of spreading codes equals the time duration of one or a few modulating symbols. In other words, we can write \( MT = N T_c \), where \( T, T_c \) and \( N \) denote symbol's time duration, chip's time duration and number of chips in one period of the spreading code, respectively. \( M \) is a small integer indicating the number of information symbols that are covered by one period of the spreading code. The exact value of \( M \) depends on the available technology for implementing the tap delay line FIR filters. Also, from the above explanation, it is easy to see that the ratio \( N/M = T/T_c \) gets an integer value. Spreading code waveform of \( k_{\text{th}} \) user can be stated as:

\[
S_k(t) = \sum_{G=0}^{N-1} \sum_{j=0}^{M-1} a_{k}^{[j]} U_{j}^{G}(t - jT_c) = \sum_{G=0}^{M-1} S_{G}^{k}(t)
\] (3.19)

where \( a_{k}^{[j]} \) is the \( j \)th element of spreading sequence for \( k_{\text{th}} \) user, \( p_{\tau_c}(t) \) is a chip shaping waveform to satisfy the Nyquist criterion for no ICI (Inter Chip Interference) and usually is selected from the family of raised cosine pulses. Therefore the transmitted signal of \( k_{\text{th}} \) user can be presented as:

\[
t_{k}(t) = \sum_{G=0}^{M-1} b_{k}^{[G]} S_{G}^{k}(t - nT)
\] (3.20)

In equation (3.20), \( T \) is the time period of information symbols, \( b_{k}^{[G]} \) is the \( G \)th symbol of \( k_{\text{th}} \) transmission and \( a_{k}^{[j]} \) is its amplitude. The received signal of \( k_{\text{th}} \) user is the result
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of convolution between the transmitted signal and the channel impulse response.

\[ r_k(t) = ts_k(t) \otimes c(t_0, t) + n_s(t) \] \hspace{1cm} (3.21)

where \( n_s(t) \) represents the additive white Gaussian noise and "\( \otimes \)" denotes the convolution operator. When the processing time, in comparison to the coherence time of the channel, is small, (this case is applicable for code acquisition process) the time-invariant model of channel is sufficient, and \( r_k(t) \) is stated as:

\[ r_k(t) = \sum_{l=1}^{L} a'_k c_{s,k}(t) \sum_{m=0}^{M-1} b_{k,(m+G)} s_{c}^k (t - nT - \tau_{k,l}) + n_s(t) \] \hspace{1cm} (3.22)

In equation (3.22), \( T_{eq} \) is the acquisition time, which generally is a short period of time at the start of each data session. According to the BER performance evaluation, \( c_{s,k} \) is time-variant and should be tracked using a tracking algorithm. The instantaneous amount of \( c_{s,k} \), in comparison to its average power, changes very rapidly. The average power of \( c_{s,k} \) is defined by the statistical characteristics of the channel. When there is a power control mechanism, the power of transmitted signal, and thus \( a'_k \), is changing during this time, and should be considered especially for evaluating the BER performance.

Moreover, to attain a distinct perspective of the received signal, let us assume, without loss of generality, that \( M=1 \). This means that the time duration of the spreading code is the same as the symbol duration. By combining equations (3.20) and (3.22), and rearranging the order of summations, we can represent the received signal of \( k_{th} \) user as:

\[ r_k(t) = \sum_{a=0}^{nT<T_{eq}} \sum_{j=1}^{N-1} d_{k,a} a'_k c_{p,a} (t - jT_e - nT) + n_s(t) \] \hspace{1cm} (3.23)

where \( c_{p,a} = \sum_{l=1}^{L} c_{s,l} p_n(t - \tau_{k,l}) \) is the channel pulse response according to the user \( k \) and \( d_{k,a} = b_{k,a} a'_k \) shows the effect of \( n_{ui} \) bit of \( k_{th} \) user and its related power.

At the receiver side, the signal is passed through a chip matched filter and is sampled,
generally at time steps $n'T_c$, where $n' = 0, 1/p, 2/p, \ldots, (T_{\text{chip}}/T_c)-1/p$. Parameter "p"

denotes the sampling rate (e.g. p=1 indicates that the sampling rate is the same as the
chip rate). For the case of rectangular chip pulses, the effect of the chip matched filter
will be simply an integration over duration of $T_c/p$. In the case of a synchronised
network at the chip level, (just for simplicity of following notation) a sampling device
replaces the integration. In this case, the discrete version of received signal can be
shown as:

$$r_k(n'T_c) = \sum_{n=0}^{\lfloor n'/T_c \rfloor} \sum_{j=1}^{\lfloor n'/T_c \rfloor} d_{k,n} a_k[j] c_p(j) \((n'-j)T_c-nT) + n_o(t)$$  \hspace{1cm} (3.24)

In practical scenario, $c_p(t)$ or pulse response of channel has a limited duration $L_t T_c$
(assume that it can be presented by its samples of distance $T_c$) and it is
straightforward to represent the combined received signal of all users as:

$$r(n'T_c) = \sum_{k=1}^{K} \sum_{n=0}^{\lfloor n'/T_c \rfloor} \sum_{j=1}^{\lfloor n'/T_c \rfloor} d_{k,n} a_k[j] c_p(j) \((n'-j)T_c-nT) + n_o(t)$$  \hspace{1cm} (3.25)

The above presentation of the received signal shows the effect of inter-chip
interference (ICI) as well as inter-symbol interference (ISI), and multiple access
interference (MAI). The operator $\lfloor \cdot \rfloor$ represents the nearest integer value, which is
smaller than its operand.

The receiver collects the transmitted samples in blocks of length "W". Based on the
idea of joint delay acquisition and data demodulation, the output of acquisition mode
should actually be an interference suppression vector, which is used to extract the
energy of desired user. Then the receiver filter should be able to cover, at least one
complete period of the data symbol. However, using periodic short spreading codes
means that our uncertainty about the start time/sample of the code sequence can be
reduced to the period of the spreading codes, N. Therefore, just for simplicity, we
assume that the period of the signature sequence is the same as the symbol duration.
In this case, a receiver filter of length $(2\times N+L)\times p$ samples is sufficient to cover
one complete symbol as well as its delayed versions, even if there is uncertainty
surrounding the start time/sample of the symbol, where L is the delay spread of the
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channel. Longer receiving blocks can be used when the period of the spreading codes is more than a symbol duration $T$, i.e. $M > 1$. Also for the purpose of simulation results, in the next sections, we always assume the sampling rate to be the same as the chip rate (i.e. $p = 1$).

Furthermore, the received sampled vector $\tilde{r}_n$ during the $n_m$ observation interval (for $M = 1$ and $p = 1$), is given by:

$$\tilde{r}_n = (r[nN], r[nN + 1], ..., r[nN + 2N + L - 1])^T,$$

which covers as many as $2 \times N + L$ samples of the received signal.

### 3.7 Summary

Adaptive pre and post combining receivers were studied in this chapter. It was discussed that the post combining receivers can be used, when there is no knowledge of the desired user time delay, and therefore this concept can be applicable to joint CDMA signal acquisition and demodulation.

The multi-path mobile channel model and the asynchronous nature of the CDMA signal model, which are suitable for modelling the important characteristics of the acquisition process (such as chip and symbol level uncertainty), was derived. It was discussed that the concept of short spreading codes, is a necessary assumption for calculating interference resistant receivers according to both time delay acquisition and data demodulation.
In this chapter the concept of time delay acquisition for CDMA signals, using short spreading codes, is studied. Firstly, the adaptive MMSE receiver is investigated. It is discussed that this type of receiver can be used to implement the near-far resistant interference suppression algorithm for joint acquisition and demodulation of data. In [Xie90] the application of the MMSE criterion to centralized multiuser detection was suggested. Due to the fact that for short spreading-code class of CDMA signals, the received signal is statistically cyclostationary, decentralised implementation of the algorithm is possible. This type of receiver, which does not use any explicit knowledge of the interference parameters was suggested in [Abdu90], [Madh94] and [Rapa94]. All of these schemes are based on assumption that a perfect knowledge about the multipath delays of the desired user is available. In reference [Madh98], an adaptive LMMSE receiver, which actually does not need to know the delay information, was proposed for the AWGN channel. In [Smit94] the idea was extended to blind reception and again for the AWGN channel.

In this chapter we extend the method introduced in [Madh98] to the mobile fading channel as a reference approach, and in particular we study the effect of training
sequence structure on the acquisition time. This concept is also extended to MMSE blind receivers in a mobile multipath fading channel, where we propose a practical method of mismatch handling that obtains a good acquisition performance, even in the presence of strong MAI and near-far effect.

In a blind acquisition scheme, mismatch is caused because of unavoidable uncertainty surrounding the received signature vector of the desired user. This concept is also defined and investigated in this chapter. Finally, we study the effect of multi-path diversity on the acquisition performance, complexity of the algorithm and acquisition time analysis.

### 4.1 Chip and bit levels time uncertainty

In equation \{3.24\} we assumed a chip level synchronised system to provide a simpler notation. However as far as the synchronisation process is concerned, there is always some ambiguity about the start time of spreading chips. In other words, it is always possible for the receiver to be asynchronous with the received signal at both chip and symbol level. Although it is straightforward to account for this effect in equations \{3.24\} and \{3.25\}, it is better to look at this problem and its effect on the received signal from a distinct perspective. Let us consider the \(n\) observation vector \(\bar{r}_n\) as

\[
\bar{r}_n = \sum_{k=1}^{K} \bar{r}_n^{(k)} + n_n(n) \in C^{2x(N+L)} \tag{4.1}
\]

![Figure 4.1: Structure of the received signal in a chip and bit level asynchronous system](image)

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\]


![Figure 4.1: Structure of the received signal in a chip and bit level asynchronous system](image)
where, $\tilde{r}_n^{(k)}$ denotes the $n_{th}$ vector of $k_{th}$ user's received signal and $n_{th}$ sampled vector of a white Gaussian noise process. "L" in equation (4.1) denotes the channel delay spread in chips and $N$ is the number of chips in one period of spreading code. The sampling rate "p" is also considered as chip rate and period of spreading code is the same as symbol duration, $M=1$ (see section 3.6 for the definition of "M" and "p").

Figure 4.1 shows the structure of $\tilde{r}_n$, where there are some ambiguities about both bit and chip levels. If the sampling rate is at the chip rate and there are $N$ chips in every symbol interval, then the size of the vector $\tilde{r}_n$ is $2 \times N + L$, where $L$ is assumed small in comparison to the spreading factor $N$. Therefore the observation vector $\tilde{r}_n$ covers two complete symbols of the received signal. However, because the start time of a symbol is unknown, generally, three symbols of every user contribute to one observation vector $\tilde{r}_n$. In Figure 4.1, $b_{k,n}$ denotes the $n_{th}$ bit of $k_{th}$ user. (For simplicity of illustration, effect of multi-path has been neglected in this figure). Vertical arrows in Figure 4.1 show the start times of the integration process. It is also seen that because of chip level uncertainty, two adjacent chips of each user in each path contribute to the output of integration process. This introduces an unavoidable inter chip interference (ICI).

The delay $\tau_{k,j}$ in Figure 4.1, is presented as $\tau_{k,j} = (n_{k,j} + \delta_{k,j})T_c$, where $n_{k,j}$ is an integer value between 0 and $N-1$ and $\delta_{k,j} \in [0,1)$. Without loss of generality, we assume that there is no over sampling. To be able to show the effect of chip level and bit level uncertainties, let $\bar{a}_k$ denote a vector of length $2N+L$, consisting of $N$ elements of spreading code of $k_{th}$ user followed by $N+L$ zeros; $\bar{a}_k = [a_k[0], ..., a_k[N-1], 0, ..., 0]^T_{2 \times N + L}$. Let $T_{L}^{n}(\bar{x})$ and $T_{R}^{n}(\bar{x})$ denote left hand and right hand shifts operator, which shifts vector $\bar{x}$, "n" times in the left or right direction, respectively. Considering the effect of a multi-path fading channel, the following equations can be derived directly from equation (4.1), Figure 4.1 and based on the previous discussion.
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\[ F_n^{(k)} = d_{k,n-1}u_k^{-1} + d_{k,n}u_n^0 + d_{k,n+1}u_k^1 \] \hspace{1cm} (4.2)

\[ u_k^{-1} = \sum_{l=1}^{L} c_{k,l}[(1 - \delta_{k,l}) T_L^{n_l} + \delta_{k,l} T_L^{n_l - N}] \] \hspace{1cm} (4.3)

\[ u_k^0 = \sum_{l=1}^{L} c_{k,l}[(1 - \delta_{k,l}) T_R^{n_l} + \delta_{k,l} T_R^{n_l + N}] \] \hspace{1cm} (4.4)

\[ u_k^1 = \sum_{l=1}^{L} c_{k,l}[(1 - \delta_{k,l}) T_R^{n_l + N} + \delta_{k,l} T_R^{n_l - N}] \] \hspace{1cm} (4.5)

The definitions of parameters \( d_{k,n}, c_{k,l}, \delta_{k,l}, \) and \( \eta_{k,l} \) are the same as were given in section 3.6. Clearly, \( \delta_{k,l} \) shows the effect of chip asynchronicity, while \( \eta_{k,l} \) represents the symbol asynchronous effect. Now, it is easy to show the received signal (4.1) as a synchronous model:

\[ r_n = d_0[n]u_0 + \sum_{l=1}^{L} d_l[n]u_l + n_0(n) = b_0[n]u_0^0 + \sum_{l=1}^{L} b_l[n]u_l^1 + n_0(n) \] \hspace{1cm} (4.6)

where \( b_0[n] = b_{k,n} \) is the desired user's symbol in \( n \)th observation window and \( u_0 \) is its signature vector, which is representing the effect of multi-path and chip level and symbol level asynchronous effects (see equation (4.4)). \( b_l[n] \) are interference symbols due to inter symbol interference and multiple access interference and \( u_l \) are their relating signature vectors. It is easy to see that \( u_0 = u_0^0 \) in equation (4.4), and vector \( u_l \) is one of three possible interference signatures \( \{u_0^0, u_1^1, u_0^1\} \). The summation at the right hand side of equation (4.6) is performed over all inter symbol interference and multiple access interference, where \( L = 3K - 1 \) and \( K \) is the number of active users.

### 4.1.1 Effect of chip level uncertainty on power of signal

For chip asynchronous systems, using chip rate sampling, equation (4.4) denotes the desired signature vector of \( k \)th user, where \( k = 1 \) represents the desired user. It is seen that the delay \( \delta_{k,l} \) results in a loss of desired signal energy, coming through the \( l \)th
Chapter 4: Interference resistant time-delay acquisition

path. Assume that the adjacent shifts of the spreading sequences are approximately orthogonal and the multi-path fading coefficients are independent random variables. Therefore, the average power of desired signal in chip asynchronous and chip synchronous systems is calculated in equations 4.7 and 4.8 respectively.

\[ p_m = \sum_i E[c_i^2] (1 - \delta_{it})^2 + \delta_{it}^2 \]  
(4.7)

\[ p_s = \sum_i E[c_i^2] = 1 \]  
(4.8)

The loss of signal energy, due to the relative delay less than a chip, is defined as
\[ L_g = 10 \log_{10} \left( \frac{p_m}{p_s} \right) \text{ dB}. \]
From equations 4.7 and 4.8, it is clear that \( L_g \) depends on the channel power delay profile, as well as on chip level time delay uncertainty. For example, when there is only one dominant path, or there are multi-paths but the delay of every path, relative to the sampling time of the receiver, is modelled as an integer multiple of chip duration, \( \delta_{it} \) in equation 4.7 is a constant value (i.e. \( \delta_{it} = \delta_t \)) and the loss is shown as: \( L_g = 10 \log_{10} ((1 - \delta_t)^2 + \delta_t^2) \).

For \( \delta_t = 0.5 \), which is the worst case when there is only one sample per chip, a loss of 3 dB in comparison to the chip synchronous system is obtained.

One possible way to overcome this loss of energy is to over sample the received signal and apply the acquisition algorithm over the extended received vector (extension of order “p”, where “p” is the over sampling rate). Another simpler way in terms of computational complexity is to run the acquisition algorithm for “p” different sub-streams in parallel, where the sampling rate of each process is the same as chip rate. Based on the outcomes of these parallel runs, the final decision is made.

Note that the standard feedback-based delay tracking [Vite95] does not work for chip timing recovery in this type of receiver, since even a small adjustment of the chip sample times results in a completely different interference structure and correlation matrix. Therefore, an interference suppression receiver that works well for a set of relative delays typically does not work if the sampling times are adjusted (e.g. by a feedback loop) [Madh98]. This problem does not happen for the conventional
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4.2 Training approach for JAD

In chapter 3, the LMMSE receivers were presented. It was explained that some type of known sequences could be used to adapt the receiver's FIR filter. The same approach can be used to perform joint acquisition and demodulation (JAD) of the received signal, where the output of acquisition mode is also an interference suppression receiver suitable for demodulation. In a linearly modulated CDMA system, assuming short spreading codes, the sampled blocks \( \mathbf{r}_n \) of the received signal are cyclo-stationary. This means that the statistical characteristics of the received signal vector \( \mathbf{r}_n \) over the observation window \( n \in [0..n_{\text{max}}] \), remain unchanged during a sufficient period of time. Therefore an adaptive algorithm can exploit the cyclo-stationary characteristic of the received signal and adapt itself to the received signal.

In acquisition mode the receiver is still asynchronous with the network, and in contrast with a synchronous system, the receiver neither knows the time/sample delay of the \( l \)th path of the desired symbol (shown as \( \tau_{x,l} \) in Figure 4.1 where \( k=1 \)) nor the current symbol of the periodic training sequence that falls within the observation interval.

The idea, which was originally proposed in [Madh95] for an AWGN channel, is based on running the adaptation algorithm during the acquisition mode for different phases of the training sequence. This will give a set of different FIR receiver filters. The decision process would be based on a defined criterion that the final receiver should satisfy; the MMSE (minimum mean square error) is a well-known criterion, which is suitable for adaptive implementation.

A possible candidate for the training sequence is an all-one sequence [Smit94], but in this case only one user can be in acquisition mode at each time. This is unacceptable for scenarios such as dense traffic network, or where the users send their information as data packets and may need to repeat the acquisition process for every packet session. Another scenario is an ad-hoc environment where the structure of network
may change very fast and acquisition should be repeated frequently.

Another possibility for selecting the training sequence can be based on application of a specific pseudo noise (PN), codes with proper correlation properties. The cross-correlation property is used to remove the effect of multiple access interference and auto-correlation is used to give the correct time delay (code phase). For an efficient use of resources (here the number of codes), users that are in acquisition mode can use the same training symbols as their spreading codes.

If $t(n)$ shows the $n_{th}$ bit of the training sequence and if there is $i$ symbols delay between the transmitted training sequence and its locally generated replica at the receiver side, then $b_0[n] = t[n+i]$. Because $i$ is unknown during the acquisition mode, the adaptation algorithm is running in serial or parallel for $N$ different hypotheses. The $i_{th}$ hypothesis $H_i$ is defined as follows:

$$H_i: b_0[n] = t[n+i] \quad i = 0, ..., N-1$$

Assuming that the training sequence is uncorrelated with the bits of interfering users, the best MMSE solution for $i$ hypothesis is the Wiener filter $f^i = R_n^{-1} r^i_n$ solution.

Referring to equation (4.2) and (4.6) the cross-correlation $r^i_n$ is calculated as:

$$r^i_n = E[t[n+i] t^*_n] = R_n (i-i') a_0^i b_0^0 + R_n (i-1-i') a_0^i b_0^{-1} + R_n (i+1-i') a_0^i b_0^1$$

and

$$R_n(k-k') = E[t[n+k] t^*[n+i]]$$

$R_n(i-i')$ is the auto-correlation function of sequence $t[n]$. If we assume a perfect auto correlation characteristic for the training sequence, there is a zero correlation result under all hypothesis except the $i_{th}$ hypothesis, and thus $r^i_n$ is zero, which leads to $MSE = 1$. Therefore the criterion for selection of the best hypothesis is:

$$i_{min} = \arg \min_i \{MMSE_i\} \quad \text{and} \quad [MMSE_{i_{min}} = 1 - (r^{i_{min}}_n)^H f^{i_{min}}]$$
In practice, $\bar{R}_n$ and $\bar{r}_n$ are replaced by their empirical averages as follows:

$$\hat{R}_n = (1/X)^n \sum_{n=1}^{N-1} \bar{r}_n \bar{r}_n^H \quad \hat{r}_n = (1/X)^n \sum_{n=1}^{N-1} \bar{r}_n$$  \hspace{1cm} (4.11)$$

In equation (4.11), $X$ denotes the total number of received blocks, which contribute to the delay estimation process and is a key factor for the acquisition time.

It is seen from equations (4.2)-(4.5) that the effect of multi-path fading and chip level uncertainty are included in the structure of $\bar{u}_0$. Here we do not need to have an assumption about this combined vector at the receiver side, because the receiver filter converges to this structure automatically during the adaptation period. This suggests that the training algorithm can benefits from the multi-path characteristic of the channel without suffering from mismatch problem.

Mismatch is defined as uncertainty about the structure (not just delay of the first path) of the desired signature waveform $\bar{u}_0$ (see equation (4.3), (4.4) and (4.5)), which is caused by time delay uncertainty, multi-path characteristic and fading effect. Mismatch is a source of performance degradation in blind algorithms, where a training sequence is not available. Therefore it is very important for blind receivers to handle the mismatch properly.

### 4.2.1 Training sequence structure

The structure of the training sequence is an important issue. The length of the spreading codes has a direct influence on the acquisition time, as well as number of users that can be supported in the acquisition process simultaneously and without collision. High spreading factor is needed to support a large number of users at the same time. However, using the same training sequence as the dedicated spreading code for each user can unnecessarily increase the acquisition time. This effect can be explained in the following manner:

While there are usually many active users in the network, a few percent of these users are in acquisition mode and most of them are in the communication process. Random bits modulate the spreading codes of the users in communication mode and are
uncorrelated with the periodic training sequences. Therefore, as far as the acquisition process is concerned, it makes sense that a user in acquisition mode uses a periodic code sequences from the same family of its spreading code, but with shorter period. Figure 4.2 shows the structure of CDMA signal in this case, while its acquisition performance is investigated through the simulation results in the next section. Actually we will see that although shorter spreading codes have poorer cross correlation characteristics, give a better acquisition performance if the acquisition time, which is an important parameter, is also considered.

![Diagram](image)

**Figure 4.2:** A possible structure of modulating symbols for signals in delay acquisition (training algorithm) and communication modes.
4.3 Simulation results and discussion

The following simulations assess the performance of the training algorithm. We assume a downlink scenario, where a family of Gold 31 are used as spreading codes for all users in the network. For the users in acquisition mode, Gold-31 and Gold-7 are used as training sequences in two different scenarios. The periods of spreading codes are selected to be the same as the symbol duration, $M = 1$, although other integer values of $M$ can be used at a cost of longer acquisition time. A mobile fading channel using carrier frequency, $f_c = 1$ GHz, mobile speed $v_m = 110$km/h and chip rate $1/T_c = 3.84$ MHz is considered. Complex fading channel coefficients are generated using a deterministic model proposed in [Patz94] where the fading channel is a Rayleigh channel with Jakes power spectral density.

Table 4.1 shows the statistical parameters of the multi-path channel model.

<table>
<thead>
<tr>
<th>Tap</th>
<th>Rel. Delay (Chips)</th>
<th>Rel. Av. Power (dB)</th>
<th>Doppler Spectrum</th>
</tr>
</thead>
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<td>0.0</td>
<td>CLASSIC</td>
</tr>
<tr>
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<td>CLASSIC</td>
</tr>
<tr>
<td>3</td>
<td>7</td>
<td>-3.0</td>
<td>CLASSIC</td>
</tr>
</tbody>
</table>

Table 4.1: Multi-path fading channel parameters, used for simulation runs.

The outcome of this acquisition algorithm is actually an interference suppression demodulator; therefore the receiver performance is tested according to the probability of acquisition error, as well as the demodulator BER. The BER performance is evaluated and averaged over successful acquisition steps.

For the evaluation of acquisition performance, the algorithm is run 100000 times and for each run the channel is initialised to different conditions (with a uniform distribution). In this way we are able to see the effect of time varying multi-path
channel on the acquisition performance, which is a very short-term process. Delay of the first path of the desired user, $\tau_{i,t}$, start phase of the training sequence and information bits of all other users are selected randomly at the start of each “Monte Carlo” simulation run.

Figure 4.3 illustrates the probability of acquisition error, where received samples are collected over 100, 200 and 300 observation intervals. To see the effect of SNR and collection length, one path flat fading channel has been considered for this case, where there are 10 multiple access users with the same power in communication mode and only one user in training mode.

Figure 4.4 shows the same scenario as Figure 4.3, except that Gold 7 was used as training sequence of the desired user. It is seen that with the same load percentage, the shorter Gold sequence results in a same performance for fewer symbols. As a result, faster acquisition process is obtained with shorter training sequence. The reason can be explained as follows.

For a good empirical estimation \{(4.11)\}, a reasonable number of observation windows “$X$” is necessary. A proper value of “$X$” (as it was understood through many simulations for different scenarios) depends on the number of periods of training sequence that contribute to our estimation. Therefore, by using a shorter spreading code as training sequence, the same complete periods of training sequence can be seen for fewer observation symbols “$X$”. As a result, a faster acquisition is obtained at a cost of being able to support smaller number of users in acquisition mode at the same time.

However, decreasing the period of the training sequence degrades the cross-correlation characteristic of the codes. This characteristic is necessary for removing the effect of other users that are in acquisition mode at the same time and use training sequences as modulating symbols. For a specific channel, these parameters should be optimised to give a good overall acquisition performance, as well as acquisition capacity. Therefore it makes sense to use the proposed structure in Figure 4.2 because there are usually more users in communication in comparison with
acquisition mode. It is also worth mentioning that during the training process, the interference part in equation \(4.6\) is suppressed because of the randomness of their modulating bits, and not because of the cross correlation characteristics between their spreading codes and the training sequence. This fact again suggests that training bits and spreading codes can have different periods without loss of performance.

The effect of number of users, which are in the training mode instantaneously, is also illustrated in Figure 4.4. By increasing the number of users in acquisition mode, the cross correlation property of the training codes (in this case Gold 7) is taken into account. These periodic cross correlations are not zero and increasing the averaging time \(X\) in equation \(4.11\) does not cancel out them. This is because of periodic characteristic of training sequences. This effect is important especially in high Eb/No and high acquisition load, where the effect of interference caused by the other users' signals, which are in acquisition mode, becomes the dominant issue.

Near-far resistant and interference suppression performance of acquisition process, using the three-path fading channel introduced in Table 4.1, is illustrated in Figure 4.5. “Pmai” in Figure 4.5 represents the offset power of MAI users in comparison to the desired user in dB.

Figure 4.6 shows the near-far resistant characteristic of the calculated receiver in the best acquisition hypothesis, according to the demodulation performance. Both figures suggest a good performance even in highly loaded scenarios and in the presence of near-far effect. The only disadvantage of this algorithm is its training sequence that increases the redundancy of the system especially in packet radio networks where a new acquisition for every packet session may be necessary. The large overhead introduced by the training sequences motivates the study of blind receivers, where the only available information about the desired user is its spreading code.

Figure 4.7 illustrates the performance of the acquisition process in multi-path frequency selective fading channel (Table 4.1), as well as for flat fading channel. Significant improvement of the acquisition performance is obtaining for the frequency selective channel compared to the flat fading case. This gain is obtained
because of multi-path diversity.

As explained before, in a training algorithm it is not necessary to have a good assumption about the structure of received spreading waveform $\bar{u}_q$. Actually during the training mode, copies of the transmitted signal, which are coming through the multi-paths, are combined automatically. As a result, $\bar{u}_q$ usually has reasonable energy to obtain a good estimation of demodulator filter $\tilde{f}$.

On the other hand it is shown that according to the blind algorithms, one important problem is how to make a good assumption about signature vector $\bar{u}_0$ and how to handle the error on construction of $\bar{u}_0$. 
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Figure 4.3: Probability of acquisition error vs. Eb/No (dB).
Gold 31 as training sequence

Figure 4.4: Probability of acquisition error vs. Eb/No (dB).
Gold 7 as training sequence
Figure 4.5: Near-far resistance of the acquisition process in multi-path fading scenario. Time delay acquisition error less than half a chip in one path, has been assumed as a correct delay estimation.

Figure 4.6: Near-far resistance of the demodulation process in multi-path fading scenario (Table 4.1)
Figure 4.7: Comparison of the acquisition performance in Rayleigh flat fading and frequency selective fading (Table 4.1)
4.4 Blind reception of CDMA signals

Blind reception of a desired user signal enables the receiver to asynchronously tune into the transmission of interest at any time, which is desirable in a mobile wireless network supporting broadcast or multicast services. Also in packet-based systems, using a training sequence for time acquisition of each transmitted packet would require significant overhead. In these cases, blind reception is an interesting concept.

Moreover, any adaptive receiver architecture needs continuous adaptation to follow the channel variations; decision-directed mechanism is an interesting method. In this algorithm, demodulated bits act like a training sequence and adaptively refine the receiver to be able to follow the channel transitions. Decision directed adaptation works well when the demodulator is working with BER less than 10^{-2}. However, it is still sensitive to fast variations of channel such as appearance or disappearance of a strong multi-path or interfering user. In these cases, the blind adaptive mechanism can be an interesting option. Blind algorithms can be used to recover the connection to the network without sending a new training sequence. Recovery a connection means obtaining a new time delay acquisition and setting a new interference suppression receiver filter. Again, this motivates the idea of blind receivers for joint acquisition and demodulation of data.

4.4.1 Classification of blind receivers

Blind algorithms are classified according to the available knowledge at the receiver side into the three categories [Madh97]. These are as follows:

- The receiver knows the propagation channel and spreading sequence of desired user/users.
- The receiver only knows the spreading sequence of desired user.
- The only information about the desired user, which is available by the receiver, is the fact that it is digitally modulated at a given symbol rate.

According to the time delay acquisition, the first category is out of our interest and
the third one simply doesn't use the available information in a mobile communication network at least for a serial reception case. Works published in [Madh97] cover the third category and it was concluded that by using the additional available information such as spreading waveform of the desired user, a simpler implementation as well as better performance is achievable. Through the following section we extend the constraint minimum output energy, C-MOE, version of the MMSE algorithm to the time delay acquisition over multi-path channels. We also look at different important concepts, which are relating to this approach. A practical method of handling the available uncertainty, referred to as “mismatch”, is introduced in section 4.4.5. Mismatch handling is an important issue according to the blind acquisitions.

4.4.2 Blind C-MOE algorithm for time delay acquisition

It is seen from equation (4.6) that what is finally calculated through the cross-correlation (4.10) in a MMSE receiver, is the signature waveform \( \bar{u}_g \). Therefore if at the receiver side, some estimate of this signature waveform was available, the training sequence, which was needed for empirical calculation of the cross correlation in equation (4.10), becomes unnecessary. Also it is worth noting that the auto-correlation matrix, introduced in equation (4.9), does not need any training sequence.

Of course there is always some uncertainty in terms of time delay, at both chip and symbol level at the receiver side. We have to add to the above problem the instantaneous multi-path characteristic of the mobile channel, which is an unknown quantity during the acquisition mode. Therefore due to these facts, a correct assumption about the desired signature waveform, \( \bar{u}_g \), at the receiver side is impossible.

Therefore, we have to make a hypotheses set over discrete area of uncertainty (Fig. 2.1). By running the acquisition algorithm over these different hypotheses and testing the results against a predefined figure of merit, it is possible to achieve a sense of the best hypothesis and accordingly the best estimation of \( \bar{u}_g \). On the other hand, this estimation is used to make an interference suppression receiver vector \( \bar{f} \). In MMSE
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criterion the receiver filter $\hat{f}$ is calculated as:

$$\hat{f} = \bar{R}_m^{-1}\hat{\alpha}_0 \in \mathbb{C}^{2^{N_L} \times L}.$$  \hspace{1cm} (4.12)

where $N$ and $L$ are the spreading factor and the delay spread of the channel in number of chips. Now the problem is how to select the best hypothesis among all hypotheses.

A constrained minimum output energy (C-MOE) receiver tries to minimise the average output energy, but freezes the contribution of the desired user's vector to the output. Clearly this receiver is calculated to suppress the sum of the noise and interference energies at the output, and only extract the energy of the desired user.

This optimisation problem is shown by equation (4.13).

$$\min E(\langle \hat{f}_{\text{MOE}}^*\vec{r}_n \rangle^2) \quad \text{Subject to} \quad \langle \hat{f}_{\text{MOE}}^*\hat{\alpha}_i \rangle = 1$$  \hspace{1cm} (4.13)

In (4.13), $E[\cdot]$ denotes the statistical expectation and $\hat{\alpha}_i$ is the estimation of the desired signature vector at the receiver side. $\hat{f}_{\text{MOE}}$ is the receiver filter that is calculated based on MOE criterion and operator $\langle \cdot, \cdot \rangle$ denotes the inner product of two vectors (i.e. $\langle x, y \rangle = x^H y$).

Therefore, if $\hat{\alpha}_i$ is a correct estimate of the signature vector of interest and if the signature of the desired user (see chapter 3 and equations (4.2)-(4.6) for the notation of CDMA signal) is nearly orthogonal to the interference signal space, then the effect of the undesired signals (MAI and noise) would be minimised. At the same time a constant energy relating to the desired user is guaranteed at the output of the receiver. In other words, the receiver vector $\hat{f}_{\text{MOE}}$ is constructed from two orthogonal vectors. One is in the direction of $\hat{\alpha}_0$, which is fixed, and the other one is orthogonal to the $\hat{\alpha}_0$. The latter component is actually calculated to minimise the energy of undesired signals at the output of the receiver filter.

However, in a mobile environment, where different users arrive with different delays and see different channel conditions (in uplink), or just because of the multi-path effect (in downlink), the desired signature vector is not actually orthogonal to the
interference signal space. In this case, the varying part of $\tilde{f}_{\text{nov}}$, should become large to be almost orthogonal to the interference space and reject the interference properly (see Figure 4.8). As a result, the contribution of noise to the output energy would be gained. Then the constraint optimisation \{4.13\} provides a trade-off between augmenting the noise and interference suppression.

If $\hat{u}_i$ is not the correct hypothesis, the behaviour of the algorithm is the same as before, except that in this case the calculated filter $\tilde{f}_{\text{nov}}$ suppresses both the desired and interference signals. This is because now, $\tilde{f}_{\text{nov}}$ freezes the contribution of a wrong vector to the output, while its relating signature vector actually does not exist in the received signal and its energy can’t be extracted by the receiver. Therefore by comparing the output energies relating to the different hypothesis, the correct hypothesis can be selected.

Figure 4.8 illustrates the geometry of the algorithm. In this figure $\hat{u}_i$ shows a nominal signature vector, where it is assumed that $\| \hat{u}_i \| = 1$. If the noise can be neglected, $\tilde{f}_{\text{nov},1}$ is the ideal constructed vector based on the MOE criterion. In the presence of the noise, less interference rejection is obtained to avoid gaining the energy of the noise at the output. In this case the solution changes to a vector such as $\tilde{f}_{\text{nov},2}$ (with smaller norm value in comparison to the $\tilde{f}_{\text{nov},1}$) to provide a trade-off between the noise boosting and interference suppression.

Using a Lagrange multiplier technique, the constraint problem \{4.13\} is changed to an unconstrained one as follows.

$$E\{<\tilde{f}_{\text{nov}}, \vec{r}_n>^2\} + \lambda_{\text{nov}} (<\tilde{f}_{\text{nov}}, \hat{u}_i> - 1)$$  \hspace{1cm} \text{\{4.14\}}

where $\lambda_{\text{nov}}$ is a real valued scalar and denoted the Lagrange multiplier. $\tilde{f}_{\text{nov}}$ is the desired receiver vector that minimises equation \{4.14\}. Desired filter $\tilde{f}_{\text{nov}}$ is calculated by setting the gradient of equation \{4.14\} to zero and is derived as follows:

$$\frac{\partial}{\partial f_{\text{nov}}}[E\{<\tilde{f}_{\text{nov}}, \vec{r}_n>^2\} + \lambda_{\text{nov}} (<\tilde{f}_{\text{nov}}, \hat{u}_i> - 1)] = 0$$ \hspace{1cm} \text{\{4.15\}}
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\[
\frac{\partial}{\partial f_{\text{moe}}} \left\{ \tilde{f}_{\text{moe}}^H \tilde{R}_{rr} \tilde{f}_{\text{moe}} + \lambda_{\text{moe}} (\tilde{f}_{\text{moe}}^H \tilde{u}_i) - \lambda_{\text{moe}} \right\} = 0
\]
\[
2 \tilde{f}_{\text{moe}}^H \tilde{R}_{rr} + \lambda_{\text{moe}} \tilde{u}_i^H = 0
\]
\[
\tilde{f}_{\text{moe}}^H = -\frac{\lambda_{\text{moe}}}{2} \tilde{u}_i^H \tilde{R}_{rr}^{-1}
\]  \[4.16\]

The constant multiplier $\lambda_{\text{moe}}$ is calculated by applying the constraint $\langle \tilde{f}_{\text{moe}}, \tilde{u}_i \rangle = 1$ on equation \[4.16\] as:

\[
\lambda_{\text{moe}} = -\frac{2}{\tilde{u}_i^H \tilde{R}_{rr}^{-1} \tilde{u}_i}
\]  \[4.17\]

By substituting the equation \[4.17\] into the \[4.16\] it is straightforward to find out that:

\[
\tilde{f}_{\text{moe}} = \frac{\tilde{R}_{rr}^{-1} \tilde{u}_i}{\tilde{u}_i^H \tilde{R}_{rr}^{-1} \tilde{u}_i}
\]  \[4.18\]

Therefore, the minimum output energy (using $\tilde{u}_i$ as the signature vector) is calculated as in equation \[4.19\].

\[
MOE(\tilde{u}_i) = E[\langle \tilde{f}_{\text{moe}}, \tilde{r}_n \rangle^2] = \frac{\tilde{u}_i^H \tilde{R}_{rr}^{-1} \tilde{u}_i}{\tilde{u}_i^H \tilde{R}_{rr}^{-1} \tilde{u}_i} \frac{\tilde{R}_{rr}^{-1} \tilde{u}_i}{\tilde{u}_i^H \tilde{R}_{rr}^{-1} \tilde{u}_i} = \frac{1}{\tilde{u}_i^H \tilde{R}_{rr}^{-1} \tilde{u}_i}
\]  \[4.19\]

Figure 4.8: The geometry of the MOE algorithm
4.4.3 The array signal processing application

In array signal processing context, one interesting problem is the estimation of the direction of arrival (DOA) of a desired signal $s(t)$, impinging on a receiving array. This is performed by using a finite data set $\{e(t)\}$ observed over duration $t = t_0, ..., t_{N-1}$. Assume that $s(t)$ denotes the base-band part of the received signal at time "t" with direction of arrival $\theta_a$. Also assume that all array sensors see the same base-band signal at time "t".

In the array processing literature, this is referred to as the narrowband assumption [Kar1996]. For an antenna array of size $L_a$ and with arbitrary geometry, the base-band array's output vector at time "t" is obtained as:

$$\vec{r}(t) = \vec{a}(\theta_a)s(t) \in C^{L_a \times 1}$$  \hspace{1cm} (4.20)

where $\vec{a}(\theta_a) = [a_1(\theta_a) ... a_{L_a}(\theta_a)]^T$ is referred to as steering vector and depends on the pattern of array sensors as well as array geometry. $\theta_a$ is the direction of arrival of $s(t)$. If $K$ different signals impinging on the array from distinct DOAs $\theta_{a,1}, ..., \theta_{a,K}$, the output vector at time "t" is written in the form

$$\vec{r}(t) = \sum_{n=1}^{K} \vec{a}(\theta_{a,n})s_{n}(t)$$  \hspace{1cm} (4.21)

Also assume that $\vec{r}(n) \in C^{L_a \times 1}$ represents the $n_{th}$ sample vector of the received vector $\vec{r}(t)$ at $n_{th}$ observation interval. In beam-forming techniques, which belong to the class of spectral-based methods in array signal processing [Kar1996], the array response is steered by forming a linear combination of the array outputs. Therefore we can write

$$y(n) = \vec{f}_b^H\vec{r}(n)$$  \hspace{1cm} (4.22)

where $\vec{f}_b \in C^{L_a \times 1}$ denotes the beam-forming vector. The average output power is also given as:

$$p(\vec{f}_b) = \vec{f}_b^H\overline{R}_r\vec{f}_b$$

where $\overline{R}_r$ is the correlation matrix of the received vector $\vec{r}(n)$, defined in equation (4.11).

In comparison to the concept of time delay acquisition and the notations used in the previous section, one can see from (4.21) and (4.22) that the steering vector $\vec{a}(\theta_{a,n})$. 

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acts like a signature vector for the \(m\)th user and the beam-former \(\tilde{f}_b\) is actually the receiver filter. Now the interested problem can be defined as follows.

We are interested to form the array antenna beam so as to receive the signal of interest (coming from the direction \(\theta_{a,o}\)) and to suppress the effect of other signals coming from other direction at the same time. A well-known method of computing the beam-former \(\tilde{f}_b\), which was proposed by Capon [Capo69], (also known as the Minimum Variance Distortion-less Response filter (MVDR), in the acoustics literature) is as follows:

\[
\tilde{f}_b = \arg\min_{f_b} p(f_b) \text{ subject to } \tilde{f}_b^H \tilde{a}(\theta_{a,o}) = 1 \tag{4.23}
\]

According to the demodulation concept in equation (4.18), where the signature vector \(\tilde{u}_0\) or similarly \(\tilde{a}(\theta_{a,o})\) in equation (4.23) is known, (4.23) and (4.13) are exactly the same. The optimum solution to (4.23), \(\tilde{f}_b\) and its relating minimum output energy \(p(\tilde{f}_b)\) are easily calculated by substituting the signature vector \(\tilde{u}_0\) with steering vector \(\tilde{a}(\theta_{a,o})\) in equations (4.18) and (4.19), respectively.

A situation similar to the time delay acquisition occurs when there is some uncertainty about the direction of arrival of the desired user, \(\tilde{a}(\theta_{a,o})\). However, in this case, one cannot construct a set of hypothesis just based on different states of the desired signature vector as \(\tilde{a}(\theta_{a,o})\). This is because some of the signature vectors in this hypothesis set would be the same as steering vectors relating to the other users (the same direction of arrival). Therefore the hypotheses set cannot be constructed just by spatial signature vector \(\tilde{a}(\theta_{a,o})\); one needs to introduce and exploit some other types of orthogonality between the users to construct a proper hypotheses set at the receiver for the desired signal. Ideally, all of these hypotheses should be orthogonal (or nearly orthogonal) to the signature vectors of the other users. If this orthogonality is provided by time domain spreading, then the receiver should be also synchronised with the desired user to exploit this characteristic. Otherwise it needs to run the acquisition process as well. Due to this similarity between these two concepts, it is also interesting to study the C-MOE algorithm for joint time delay acquisition and direction of arrival estimation in a WCDMA system.
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using array antenna. However, this is out of scope of this thesis and is considered as possible future work.

4.4.4 The mismatch problem in blind approaches

In CDMA networks, spreading codes are used to ensure that the signature vectors of different users are nearly orthogonal. However, in a real environment, mainly due to the asynchronous structure of the network in the uplink (where different users experience different channel conditions and delays) and multi-path characteristics of the channel in downlink, there is always an amount of inter-symbol interference (ISI) and inter-chip interference (ICI). These effects reduce the orthogonality of the signature vectors. Consequently, the received signature vector of the desired user cannot be assumed orthogonal to the space of interfering users (see Figure 4.8). This issue and its effect on the receiver vector $\bar{f}_{\text{mix}}$ were explained already in section 4.4.2.

On the other hand, in acquisition mode there is always some uncertainty about the correct signature vector $\bar{\alpha}_0$. As explained before, this uncertainty arises because of the unknown time delays and channel conditions. Although the receiver runs the acquisition algorithm over different hypothesis, none of these hypothesis matches the correct signature vector $\bar{\alpha}_0$ perfectly. Consequently, there is always a distance between the best hypothesis vector and desired signature vector $\bar{\alpha}_0$. Referring to the Figure 4.8, one can see that in this case, desired signal is actually located in the interference space, where our best hypothesis is relatively close to the desired signature vector.

As a result the receiver vector $\bar{f}_{\text{mix}}$ try to suppress both the interference signal and the desired user's signal. It was explained during the section 4.4.2 that in the presence of noise, the MOE receiver is calculated to provide a trade-off between the energy of interference and noise at the receiver output. Therefore in low interference conditions and in the presence of mismatch, the trade-off is obtained by effectively rejecting the desired signal and unnecessary noise gain. The same effect is seen under low noise condition, where there is no effective controlling parameter to control the norm of the
received vector \( \tilde{f}_{\text{moe}} \) and to prevent the algorithm to reject the desired signal.

In all of these cases, and because our best hypothesis is relatively very close to the desired signature, the norm of \( \tilde{f}_{\text{moe}} \) should become too much (in comparison with the case that there is no mismatch) to reject the desired signal.

What was explained in above paragraphs suggest that the mismatch between the nominal signal vector \( \tilde{u}_0 \) and the correct vector \( \tilde{u}_0 \) is a very important issue according to the acquisition process.

The constraint minimum output energy receiver (C-MOE), was proposed in [Madh94][Honi94] and [Honi95] for a known propagation channel scenario. In [Madh97] the known delay assumption was removed for an AWGN channel, and noise enhancement problem was explained as well. The work in [Madh97] is based on setting a constraint on the length of the interference suppressor vector \( \tilde{f}_{\text{moe}} \). But no practical method for automatically handling the mismatch was proposed. This constraint is to control the effect of mismatch by controlling the length of \( \tilde{f}_{\text{moe}} \).

Therefore, the unconstraint optimisation problem \{4.14\} can be modified by adding a penalty factor on the norm of \( \tilde{f}_{\text{moe}} \). This gives equation \{4.24\}. The desired vector \( \tilde{f}_{\text{moe}} \) should minimise equation \{4.24\} and at the same time keeps the contribution of the desired signature vector to the output energy, constant.

\[
E[<\tilde{f}_{\text{moe}}, \tilde{R}_n >^2] + \lambda_{\text{moe}} (<\tilde{f}_{\text{moe}}, \tilde{u}_i > -1) + \nu \| \tilde{f}_{\text{moe}} \|^2
\]

\{4.24\}

The solution steps are the same as what explained for \{4.14\}

\[
\frac{\partial}{\partial \tilde{f}_{\text{moe}}} \left[ E[<\tilde{f}_{\text{moe}}, \tilde{R}_n >^2] + \lambda_{\text{moe}} (<\tilde{f}_{\text{moe}}, \tilde{u}_i > -1) + \nu \| \tilde{f}_{\text{moe}} \|^2 \right] = 0
\]

\{4.25\}

\[
2\tilde{f}_{\text{moe}}^H \tilde{R}_r + \lambda_{\text{moe}} \tilde{u}_i^H + 2\nu \tilde{f}_{\text{moe}}^H = \overline{0}
\]

\[
\tilde{f}_{\text{moe}}^H = -\frac{\lambda_{\text{moe}} \tilde{u}_i^H (\tilde{R}_r + \nu \overline{I})^{-1}}{2}
\]

\{4.26\}

By applying the condition \(<\tilde{f}_{\text{moe}}, \tilde{u}_i > = 1\), the desired receiver vector is calculated as
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\[ \tilde{f}_{\text{noe}} = \left( \tilde{R}_r + \nu \tilde{I} \right)^{-1} \tilde{u}_i \tilde{u}_i^H \tilde{R}_r + \nu \tilde{I} \right)^{-1} \tilde{u}_i \in \mathbb{C}^{(2N+L)x1} \]  \( (4.27) \)

\( \tilde{R}_r \in \mathbb{C}^{2N+L} \) is the auto correlation matrix of the received sample vector \( \tilde{r}_i \). \( \tilde{I} \) is the identity matrix. Also, it is easy to see that \( \nu \) in equation (4.27) acts the role of fictitious noise power, which is added to the diagonal elements of matrix \( \tilde{R}_r \).

It was explained already that proper selection of \( \nu \), depends on the condition of MAI and noise in the channel and has an important effect to prevent the suppression of the desired user.

A large values of \( \nu \) cause less signal loss, because it sets a bigger penalty factor on the length of \( \tilde{f}_{\text{noe}} \). This results in a better tolerance against the mismatch at the expense of lower interference suppression. On the other hand, a small value of \( \nu \) creates more interference suppression, and is more suitable for large MAI scenarios. However in low noise conditions, this can also cause the suppression of the desired signal as well as increasing the power of noise at the output of the demodulator.

Therefore, proper selection of \( \nu \) is an important issue. The simulation results in section 4.4.7 show the effect of \( \nu \) on the acquisition performance of the C-MOE algorithm in the AWGN channel.

A simple and practical algorithm for automatic selection of \( \nu \) is proposed in section 4.4.6 and its performance is studied in section 4.4.10.

Let us first assume a single path channel condition. For this simplified case, list 4.1 represents the acquisition algorithm. For general fading channel the algorithm is the same but the structure of hypothesis can be different. This is addressed in section 4.4.8.

**List 4.1**

**Hypothesis construction:** for \( i = 0, 1, \ldots, N-1 \), \( N \) hypotheses are constructed, which correspond to \( N \) different chip delays. \( N \) represents the spreading factor. For hypothesis \( H_i \) corresponding to the delay \( t_i^0 = iT_c \), the nominal vector \( \tilde{u}_i \) is
constructed as $\tilde{a}_0^{\prime} = T_{g}^{\prime}(\tilde{a}_0) / \|T_{g}^{\prime}(\tilde{a}_0)\|$, where the shift operator $T_{g}^{\prime}(\tilde{a}_0)$ and signature vector $\tilde{a}_0$ were defined in section 4.1 (page 61). By proper combining of the adjacent hypotheses, a more precise set of hypotheses can be achieved. Delays less than half a chip can be captured by increasing the sampling rate. However, this increases the dimension of collected vectors and results in a more processing time and complexity.

The best hypothesis selection: In the $i_{\text{th}}$ hypothesis, $\tilde{f}_{\text{noe}}^\prime$ is calculated as (4.27), where $\tilde{u}^\prime$ is replaced by $\tilde{u}_0^{\prime}$. The relating output energy is calculated as $MOE^i = E[\tilde{f}_{\text{noe}}^\prime, \tilde{f}_{\text{noe}}^\prime] = (\tilde{f}_{\text{noe}}^\prime)^{\prime} \hat{R}_{rr} \tilde{f}_{\text{noe}}^\prime$.

The best hypothesis is the one that maximises its related $MOE$. We use a normalised version of the $MOE^i$ to compare the hypotheses. This is to compare the hypotheses according to the amount of received signal energy that contribute to the receiver output. Therefore the best hypothesis is selected as follows

$$i_{\text{max}} = \max_i \frac{MOE^i}{f_{\text{noe}}^i f_{\text{noe}}}.$$  \hspace{1cm} (4.28)

Hypotheses combining: In the case that a better delay estimation, (less than half a chip) is needed, two adjacent hypotheses, which relatively maximise the equation (4.28) can be combined. For example, a linear combination can be performed as $\tilde{u}_0(\chi) = \chi \tilde{u}_0^{\prime \prime} + (1 - \chi) \tilde{u}_0^{\prime \prime}$, where the parameter $\chi$ is calculated to properly combine the two hypotheses and give a better estimation of the signature waveform.

In [Madh97] a normalised output energy function was introduced in AWGN channel for the refinement step, which is a function of the parameter $\chi$ (the desired signature vector). The desired parameter $\chi$ is the argument that maximises the corresponding function. In this reference, $\chi$ is calculated analytically by solving a quadratic equation.

However, for a coarse acquisition block an analytically calculation of $\chi$, introduces unnecessary computational complexity. In the next section a linear approach for
refinement of the delay is proposed that can obtain an approximate estimation of the delays less than a chip, which its accuracy is enough according to a coarse acquisition block.

After calculating the parameter $\chi$, the refined version of the receiver and its relating delays can be obtained from equations (4.29) and (4.30).

Although the calculated values of $n$ and $\delta$ are not used explicitly here, they are useful information that may be passed to the tracking algorithms to start the tracking of channel transitions after successful acquisition.

$$\tilde{f}_{\text{new}} = \chi \tilde{f}_{\text{new}}^{\text{max}} + (1 - \chi) \tilde{f}_{\text{new}}^{\text{max}}$$  \hspace{1cm} (4.29)

$$\tau_i = \chi \tau_i^{\text{max}} + (1 - \chi) \tau_i^{\text{max}}$$  \hspace{1cm} (4.30)

### 4.4.5 Delay estimation refinement

In digital implementation of a receiver, the input chip matched filter samples the received signal by a specified sampling rate. This means that the receiver has a limited time resolution and it is not necessary to evaluate the delay, $\tau_i$, analytically.

A simple and practical method, according to the result and complexity is as follows.

After coarse timing acquisition, the receiver makes a set of new hypotheses. This is performed by linear combination of the selected adjacent hypotheses $i_{\text{max}}$ and $i_{\text{next}}$. The number of these new hypotheses depends on the actual resolution that the receiver (e.g. the tracking block) expects. For example if the resolution of $\%25$ of chip duration is enough, the receiver needs to make just a set of 4 new hypotheses. Equation (4.31) shows a general member of the new set, where signature-vectors $\hat{u}_0^{\text{max}}$ and $\hat{u}_0^{\text{max}}$ are related to the selected adjacent hypotheses $i_{\text{max}}$ and $i_{\text{next}}$, respectively.

$$\hat{u}_0(\chi) = \chi \hat{u}_0^{\text{max}} + (1 - \chi) \hat{u}_0^{\text{max}} \text{ where } 0 < \chi < 1.$$  \hspace{1cm} (4.31)

The C-MOE algorithm, described by list 4.1, is implemented again using this new set.
Because this second step actually needs a very few hypotheses, the added complexity is negligible.

It is worth noting that generally, the false alarm rejection is performed (e.g. by testing the received information bit sequence) after the acquisition step, where there is always a possibility that the initial selected hypotheses $\hat{h}^{(1)}$ and $\hat{h}^{(2)}$ are actually wrong (false alarm). Therefore the coarse acquisition should be fast and does not need to be very accurate. Tracking loop, which is closed after successful coarse acquisition, is responsible for carrying out fine alignment and is out of scope of this thesis.

Furthermore, some simulation studies show that the simple refinement approach proposed in this section generally works properly, while the mentioned analytical approach can go wrong especially when the correct delay is exactly equal to one of the first step hypotheses. For example Table 4.2 compares the percentage of successful refinements between [Madh97] and the simple approach proposed in this section for an AWGN channel, where the coarse acquisition has been already performed successfully. In our discrete simulator, the delays less than a chip were set as $\delta_i \in [0,0.25,0.5,0.75]$, where $\delta_i$ denotes a fraction of one chip duration delay. Also for the analytical selection of $\chi$, the corresponding $\delta_i$ was rounded to the nearest expected value of $\delta_i$ ($\delta_i \in [0,0.25,0.5,0.75]$). The results were averaged over 100 tries. 10 users with the same power and Eb_No=10 dB were considered in this comparison.

<table>
<thead>
<tr>
<th>Successful selections</th>
<th>[Madh97]</th>
<th>Second set of hypotheses</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>%65</td>
<td>%73</td>
</tr>
</tbody>
</table>

Table 4.2: Percentage of successful selections in two compared approaches

Therefore the desired value of $\chi$ can be selected using maximisation problem (4.32).

Equation (4.32) is the same as (4.28) but uses the new set of hypotheses.
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\[
\chi_{\text{max}} = \arg(\max_x \left( \frac{(\tilde{f}_{\text{max}}(\chi))^H \hat{R}_e \tilde{f}_{\text{max}}(\chi)}{(\tilde{f}_{\text{max}}(\chi))^H \tilde{f}_{\text{max}}(\chi)} \right))
\]

where

\[
\tilde{f}_{\text{max}}^t = \chi_l \tilde{f}_{\text{max}}^l + (1 - \chi_l) \tilde{f}_{\text{max}}^m \quad \text{and} \quad 0 \leq \chi_l \leq 1, \quad i=1...h
\]

In equation (4.33), "h" is the number of new hypotheses, which depends on the desired resolution.

4.4.6 Automatic selection of \( v \)

It was already explained during section 4.4.4 that the selection of \( v \) in equation (4.24) has a significant effect on the trade-off between interference suppression, noise enhancement and desired user's signal rejection. It was also concluded that the proper selection of \( v \) depends on the level of noise and interference. Therefore, dynamical and practical selection of \( v \) is an important task for the receiver. This fact, which is also studied through the simulation results in section 4.4.7, insists that inserted fictitious noise should be adjusted for different conditions of the channel, MAI and noise power to obtain a desirable acquisition-error performance. In this section we propose a practical method for automatic selection of \( v \) that is assumed as a pre-acquisition step.

Let us assume a zero value for fictitious noise factor \( v \) in equation (4.24). Following this assumption and in a wrong hypothesis \( \hat{u}_0 = \bar{u}_{\text{wrong}} \), the first and second terms of equation (4.24) become independent (especially when the wrong hypothesis vector \( \hat{u}_0 = \bar{u}_{\text{wrong}} \) is orthogonal to all of the signature vectors \( \bar{u}_j \) that have contributed to the received signal). As a result, the algorithm tries to minimise both the desired signal and interference signal terms. Thus the receiver vector \( \tilde{f}_{\text{max}} \) is calculated to be orthogonal or nearly orthogonal to the signal space, which is near to the direction of \( \bar{u}_{\text{wrong}} \). Noise is the most effective term that can control the length of \( \tilde{f}_{\text{max}} \) effectively and its contribution to the output energy is proportional to the \( \| \tilde{f}_{\text{max}} \| \). Increasing the power of the noise, decreases the \( \| \tilde{f}_{\text{max}} \| \) and vice-versa because this norm is directly
related to the receiver output energy.

On the other hand, as the wrong hypothesis is not exactly orthogonal to the space of the received signal, MAI always has a minor contribution to the output energy or equivalently has a minor effect on the $\| \vec{f}_{\text{mae}} \|$.

The behaviour becomes very important in low noise conditions where during interference suppression a significant amount of desired user signal is also rejected. This happens because the mismatch is present and enough power of noise does not exist to control the length of receiver vector $\vec{f}_{\text{mae}}$. What was mentioned in the last paragraphs, initiates this idea that the norm of $\vec{f}_{\text{mae}}$ in a known wrong hypothesis can provide a valid measure for selection of $\nu$.

Now, consider a unique spreading code that is not used for communication by any of the users in the network. This code and its different shifts can provide a known set of wrong hypotheses $\{\vec{\mu}_i\}$. All active receivers in the network use this known code before the coarse acquisition step, where $\nu = 0$, and calculate the $\vec{f}_i = R_{\nu}^{-1}\vec{\mu}_i$. Then the mean value of the normalised norm of $\vec{f}_i$ (i.e. $D = \text{mean}(\| \vec{f}_i \|)$) is used as a measure for the instant power of noise and MAI in the system. Based on the calculated value of “$D$” and an available lookup table (optimised for current traffic zone and other cell specifications) the receiver can select the value of $\nu$ from the available options. The simulation results in section 4.4.6.1 illustrate the appropriate characteristics of “$D$” for different conditions of noise and MAI.

4.4.6.1 Numerical example

Figure 4.9 presents the mean and standard deviation (shown by sigma) of “$D$” obtained from 400 simulation runs for near-far ratios of $-20$ dB and $+20$ dB. The 3 path channel model, presented in table 4.1, is used. The first path delays and spreading codes of users are selected randomly. There was no attempt to optimise vector $\vec{\mu}$ over the available resource of spreading codes in the network. All spreading codes are selected randomly and remain constant during the simulation runs for each
point. PMAI(offset) in Figure 4.9 denotes the offset power of the interfering users over the desired one in dB.

It is seen from Figure 4.9 that "D" increases as the Eb/No is increased, which results in a larger value of \( \nu \). This is necessary to limit the signal suppression, particularly in low noise conditions, where in the absence of noise, both the interference and desired signal can be rejected because of mismatch. On the other hand, in low noise condition, the effect of MAI is more significant and "D" decreases as the MAI increases. This results in a smaller value of \( \nu \), which is again desirable and helps more interference suppression in strong MAI conditions. Generally, the characteristic of "D" in all situations is desirable and can be used for dynamic selection of \( \nu \).

![Graph showing mean value of D vs. Eb/No](image)

**Figure 4.9**: Mean value of \( D = \text{mean}(\|n_j^i\|) \) in a wrong hypothesis vs. Eb/No.
4.4.7 Simulation results and discussion

To investigate the sensitivity of the interference suppression algorithm to the selection of $v$, the probability of acquisition error is calculated in different conditions of noise and MAI.

One thousand least square (LS) simulation runs give the mean and variance of the acquisition error in each hypothesis. Using the Gaussian approximation, suggested in [Madh98] and [Madh97] the probability of acquisition error is calculated semi analytically as equation (4.34).

$$P_{\text{seq.-error}} \leq \sum_{i=0}^{\text{max.}} Q\left(\frac{\mu_i}{\sigma_i}\right)$$  \hspace{1cm} \text{(4.34)}

where $\mu_i$ and $\sigma_i$ denote the mean and standard deviation of the random variable $Z = (N_{\text{MOE}}) - (N_{\text{MOE}}^i)$ where $i$ is the correct hypothesis and $N_{\text{MOE}}^i = \frac{\text{MOE}^i}{\text{trace}(\hat{R}_{\text{ref}})}$. To evaluate the equation (4.34), the summation is applied over all hypotheses except the selected ones. In all simulation runs, 10 interfering users with Gold spreading codes of length 31 in the AWGN channel are considered. Delays were selected randomly and $v = \alpha x \text{trace}(\hat{R}_r)$ where $\hat{R}_r$ denotes the estimation of the received signal covariance matrix and was defined by equation (4.11). trace($\hat{R}_r$) gives the summation of the main diagonal elements of the correlation matrix $\hat{R}_r$ as a measure for power of noise and interference. Parameter “$\alpha$” is used to map the value of trace($\hat{R}_r$) to a proper value of $v$. This parameter is changed in following figures to show how the selection of $v$ could affect the acquisition performance in different conditions.

In Figure 4.10 a low interference power and good Eb/No condition are considered. However $\delta \neq 0$ results in a chip level asynchronous system and introduces the mismatch between the signature waveform of the desired user and the best hypothesis. The parameter $\delta$ denotes a fraction of one chip delay relating to the
desired user signal and in the general case was defined in section 4.1. As a result, the algorithm suppresses the desired signal if the penalty factor $\nu$ is not selected properly. In this case the probability of acquisition error will be high if the value of “$\alpha$” is under-estimated. Increasing “$\alpha$” limits the length of $f_{\nu e}$ and limits the suppression of desired signal effectively. For example we see that $\alpha = 1.0E-1$ makes a very low probability of acquisition error. In Figure 4.11 there is no mismatch between the correct hypothesis and actual received signature waveform. This is because $\delta = 0$ and system is chip level (but not bit level) synchronous.

The plots in Figure 4.11 are relatively more compressed and a better acquisition performance for fewer bits is obtained. This results in a faster time delay acquisition process.

Figure 4.12 shows a scenario where there is a high level of MAI in the system. All 10 interfering users send their information bits with the same power, which is 20 dB stronger than the desired user’s signal. As was expected, in this case the best performance is achieved by using smaller value of “$\alpha$” (i.e. $\alpha = 1.0E-4$) in comparison to the last two figures. Also Figure 4.13 illustrates the results for the same conditions as Figure 4.12, except that $E_b/N_0 = 7$. In this case the only effective way to obtain a low probability of error is increasing the number of information bits, “X” used for the acquisition process.

These figures generally illustrate the degree of sensitivity of the acquisition performance to the selection of parameter $\nu$. It is worth noting that in the presence of fading channel, the mismatch problem becomes worse and dynamic selection of $\nu$ becomes increasingly important.
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![Diagram 1](image1.png)

Figure 4.10: Semi-analytic approximation of the probability of acquisition error vs. "X" (See equation (4.11)). \( \delta \neq 0 \).

![Diagram 2](image2.png)

Figure 4.11: Semi-analytic approximation of the probability of acquisition error vs. "X" (See equation (4.11)). \( \delta = 0 \).
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Figure 4.12: Semi-analytic approximation of the probability of acquisition error vs. "X", in presence of strong MAI power. Eb/No=10

Figure 4.13: Semi-analytic approximation of the probability of acquisition error vs. "X", in presence of strong MAI power. Eb/No=7
Table 4.3 illustrates the simulation results of the probability of acquisition error and also investigates the effectiveness of the proposed method of refinement, proposed in section 4.4.5. In total, 200,000 simulation runs in each case were performed for an AWGN channel. The parameter $\nu$ is selected dynamically from a pre-defined lookup table, based on the measuring method proposed in 4.4.6. In each run all delays are selected randomly and any error more than half a chip between the estimated and correct delay is considered as an acquisition error. The number of received blocks $X=50$ (see equation (4.11)), $Eb/No=15$ and other simulation parameters remain the same as mentioned previously. The mean and standard deviation of error are also presented in Table 4.3. These values show a good performance according to the fine acquisition ability of the algorithm. This suggests that the information provided by the acquisition mode can also be reliable for the tracking process. Table 4.4 presents the same information as Table 4.3 for $Eb/No=7$.

<table>
<thead>
<tr>
<th>$\Delta$ (PMAI)</th>
<th>-20dB</th>
<th>0dB</th>
<th>20dB</th>
</tr>
</thead>
<tbody>
<tr>
<td>$P_{acq _error}$</td>
<td>0.00</td>
<td>$3 \times 10^{-5}$</td>
<td>$5 \times 10^{-4}$</td>
</tr>
<tr>
<td>$\bar{\Delta}_r$</td>
<td>-</td>
<td>0.0017</td>
<td>$10^{-3}$</td>
</tr>
<tr>
<td>$\sigma_{\Delta_r}$</td>
<td>-</td>
<td>0.094</td>
<td>0.019</td>
</tr>
</tbody>
</table>

Table 4.3: Probability of acquisition error, mean and standard deviation of error for different near-far conditions. $Eb/No=15$.

<table>
<thead>
<tr>
<th>$\Delta$ (PMAI)</th>
<th>-20dB</th>
<th>0dB</th>
<th>20dB</th>
</tr>
</thead>
<tbody>
<tr>
<td>$P_{acq _error}$</td>
<td>0.00</td>
<td>$10^{-3}$</td>
<td>$2 \times 10^{-2}$</td>
</tr>
<tr>
<td>$\bar{\Delta}_r$</td>
<td>-</td>
<td>0.006</td>
<td>0.003</td>
</tr>
<tr>
<td>$\sigma_{\Delta_r}$</td>
<td>-</td>
<td>0.25</td>
<td>0.05</td>
</tr>
</tbody>
</table>

Table 4.4: Probability of acquisition error, mean and standard deviation of error for different near-far conditions. $Eb/No=7$. 

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4.4.8 Acquisition performance in multi-path fading channel

The results presented in the previous section, show the effect of MAI and mismatch on the acquisition performance. The only source of mismatch, which was considered for the simulation results of section 4.4.7, was the chip level uncertainty according to the time delay of the received signal.

The multi-path characteristic of channel is another important source of mismatch. Because the acquisition is performed before any channel estimation, there is no pre-known information about the instantaneous condition of the fading channel. This means that the mismatch cannot be overcome effectively just by increasing the number of hypotheses or equivalently increasing the resolution of the delay estimator.

Equations (4.2) to (4.6) show that in the presence of multi-path fading channel, we are faced with a more complicated signature waveform. As a result, it is very difficult to perform an appropriate set of hypotheses at the receiver. Furthermore, because of mismatch, the receiver filter suppresses the desired signal and the BER performance of JAD receiver is degraded, even if the acquisition was obtained correctly.

The next chapter addresses the BER improvement of the final JAD receiver. In this section we are just interested in the acquisition performance.

Different methods of constructing an appropriate set of hypotheses, \( \{H_i\} \), are possible and are as follows.

*Statistically based hypotheses:

Usually in a communication networks, some pre-defined statistical model of the channel is available. This model generally specifies the statistical characteristics of the channel in terms of relative average power of paths, relative delays and so on. Statistical model can be used to provide some information about average behaviour of the channel, especially when a long-term process is considered. However we should note that, because time delay acquisition is a short-term process, the instantaneous behaviour of channel could be too far from its statistical characteristics.
For example, the main path may be completely under fade for the whole duration of the acquisition process, while a weak transmission path dominates the power of the received signal effectively. Therefore using statistical information of the channel can cause a large amount of mismatch and significant performance degradation.

**Single path hypotheses:**

Another way of performing an appropriate set of hypotheses is to consider different hypotheses for different copies of the desired user's signature waveform coming from different paths (assuming that the relative delays are known). In this way the dimension of \( \{ H_i \} \) increases from \( N \) to \( L \times N \), where \( L \) is the maximum number of paths and is defined by the channel model. Mismatch is still a problem. In comparison to the statistical based method of combining (that is not realistic in a real system), complexity increases because of increasing the dimension of the hypotheses set.

**Equal gain combining hypotheses:**

\( \{ H_i \} \) can also be constructed by combining the important paths (in terms of their relative powers) presented in the channel model with equal gain coefficients. In this way, a general member of the hypotheses set is denoted as:

\[
\hat{\Phi} = \sum_{i \in [L]} \left[ (1 - \delta_j) T_{R}^{n_{k,i}} (\bar{a}_o) + \delta_j T_{R}^{n_{k,i+1}} (\bar{a}_o) \right]
\]

where vector \( \bar{a}_o \) is presented as \( \bar{a}_o = [a_o[0], ..., a_o[N-1], 0, ..., 0]^T_{2 \times N+L} \) and was defined in section 4.1. The summation is performed over selected paths of the channel model. Equation (4.35) is obtained from equation (4.4) by just considering the selected paths and neglecting the channel fading coefficients that are unknown during the acquisition mode. In this way we only need \( N \) hypotheses, assuming the channel model is reliable (in terms of relative delays of multi-paths). In comparison to the previous one, this method uses a smaller hypotheses set. Naturally, in this way the acquisition algorithm would benefit from time diversity characteristic of the channel, because in each hypothesis the effect of the all-important paths are considered.
resulting in better acquisition performance.

4.4.9 UMTS RACH structure & process

In the current standard of the UMTS network, the reverse link (uplink) synchronisation is performed using the random access channel (RACH). The random access channel is typically used for signalling purposes. It is used for registration of the terminal after power on, or to update the location because of mobility or to initiate a call.

In UTRA the RACH procedure has the following phases [ETSI98] [Holm02][3GPP03]

- The terminal decodes the BCH (Broadcast Channel) to find out the available RACH sub-channels and their scrambling code and signature.
- The terminal selects randomly one of these sub-channels. The signature is also selected randomly from among the available signatures.
- Downlink power is measured and initial RACH power level is set accordingly.
- A 1 ms. RACH preamble is sent with the selected signature.
- The terminal decodes AICH (Acquisition Indicator channel) to find out if the base station has detected the preamble.
- If there is no acquisition acknowledgement, the terminal will increase its transmission power and send a new preamble signal.
- When an acquisition was acknowledged by AICH, the terminal transmits the 10 ms message part of the RACH.

Figure 4.14 shows the RACH procedure where the terminal transmits the preamble and listens to the AICH for acknowledge indication.

The mobile terminal can start the transmission on the random access channel at a number of well-defined time offsets, relative to the frame boundary of the received broadcasting control channel of its cell.
These time offsets are known as access slots. The method is based on a slotted ALOHA approach as is illustrated in Figure 5.14 [3GPP03].

There are 16 signature bits in every preamble, which modulate the spreading codes of period 256. This means that there are in total 4096 chips in each preamble.

The essential method of uplink synchronisation is based on matched filter correlator, as shown in Figure 2.10. In the next section the acquisition performance of this conventional receiver is compared with the interference resistant algorithm, C-MOE, using the same number of chips. Therefore, the evaluation time is the same, but the structure of data is different in these two approaches. It is seen that for the same evaluation time, significant improvement is achieved in terms of acquisition error probability, by using the interference resistant approach instead of the conventional receiver.
4.4.10 Simulation results and discussion

To investigate the performance of the C-MOE combined with the dynamic selection of $\nu$, some simulations were performed and the results are reported in this section. The channel is a Rayleigh fading channel and in frequency selective scenarios, the channel model, presented in Table 4.1, is used. Spreading factor is always 31. The complex fading channel is modelled using the deterministic approach (Equal Area) introduced by [Patz94]. The carrier frequency, velocity of mobile and chip rate are set to 1GHz, 70 Km/h and 3.84 Mc/s respectively. Acquisition error is counted when none of the propagation paths was captured correctly, and the results were averaged over 10000-simulation runs. All interference users have the same "PMAI(offset)" in each case, where "PMAI(offset) dB" is defined as $10\log(P_{\text{ul}}/P_{\text{detected,ul}})$. In all cases relating to the blind C-MOE approach, equal gain combination of multi-paths is used to construct the hypotheses set \{H_i\}.

Figures 4.16 and 4.17 show the probability of acquisition error vs. near-far ration. In both cases, equation (4.18) is used to calculate the receiver filter and there is no mismatch handling. Figure 4.16 illustrates the results in a frequency selective UL channel (Table 4.1), while a flat fading channel is used for Figure 4.17. Eb/No=10 dB for both cases. The main source of mismatch in Figure 4.16 is the multi-path effect, while in the flat fading case the uncertainty about the start of chips is the only source of mismatch. It is seen from these figures that the acquisition algorithm is an interference resistant algorithm. However it is interesting to note that the results in the frequency selective channel are worse than the flat fading results. This is because the level of mismatch in frequency selective channel is greater than flat fading and therefore the desired signal suppression is more probable.

Figure 4.18 shows the near-far resistant characteristic of the C-MOE algorithm as well as its interference suppression performance, where equation (4.27) is used to calculate the receiver filter. Eb/No=10 dB and the channel is a frequency selective fading channel as shown by Table 4.1. Delay error less than half a chip (for at least one path) is considered as correct acquisition. From Figure 4.18, it is seen that in terms of near-far resistant characteristic and for a specific number of multiple access
interferer users, the algorithm behaves very well. However, increasing the number of interfering users does not have the same effect as increasing the power. This can be explained by referring to Figure 4.8. Increasing the number of interfering users increases the dimension of MAI space. Therefore it becomes more possible that the desired user signature vector has a significant power in the direction of MAI space. As a result, this increases the risk of desired signal rejection and degradation of the acquisition performance.

Comparison between Figure 4.18 and 4.16 shows that a considerable gain according to the acquisition performance is obtained by handling the mismatch as explained in sections 4.4.4 and 4.4.6.

Figure 4.19 illustrates the probability of acquisition error for the same conditions, as Figure 4.18, except that the channel is a flat fading. By comparing this figure with Figure 4.18, we see that in a frequency selective multi path fading, better performance is achievable. However in a blind approach there is always a trade-off between diversity and mismatch. Increasing the number of paths increases the diversity, but also increases the mismatch. Therefore, the improvement is not too significant. We already saw (Figure 4.7) that the diversity has a considerable effect on the acquisition performance of the training algorithm. This is because in a training approach there is no mismatch problem. Therefore as far as the maximum delay spread of the channel is not comparable with symbol time duration, an appropriate combination of multi-path is directly obtained during the training mode. This desirable characteristic of the training algorithm is achieved at the expense of training preamble and greater acquisition time.

Also, comparing Figures 4.17 and 4.19 shows that even in a flat fading conditions, handling the mismatch improve the acquisition performance. However this improvement is not as significant as multi-path fading scenario.

Effect of Eb/No is shown in Figure 4.20. Although it is known that the performance of blind reception algorithms degrades in low Eb/No, still the figure shows that the dynamical selection of $\nu$ can still limit the noise power amplification.
Figures 4.21 and 4.22 illustrate the performance of a conventional acquisition system using the preamble structure of Figure 4.15. In comparison to the previous figures, the conventional receiver uses the same number of chips as interference suppression receivers, which is approximately the same as the standard size of the preamble part of RACH for UMTS. The number of preamble symbols in Figure 4.21 is 127 and the spreading factor is 31, while in figure 4.20, 31 preamble symbols are used and the spreading factor is 127. Other simulation conditions are the same as in Figure 4.18.

Both figures show how much the conventional acquisition receivers suffer from MAI and near-far effect, and confirm that in the presence of these hostile effects, interference resistant acquisition algorithms are promising. In these simulations, a fixed power of preamble section is used.

Also it is worth noting that using longer spreading codes and smaller signature sequences, while the total number of chips is fixed, increases the acquisition performance in low MAI conditions. However, it still provides poor performance when MAI and near-far effect are increased.
Figure 4.16: Probability of acquisition error in an UL scenario for the MOE algorithm using equation (4.18) (No mismatch handling). Three paths Rayleigh fading channel (Table 4.1).

Number of symbols in acquisition mode $X=100$

Figure 4.17: Probability of acquisition error in an UL scenario for the MOE algorithm using equation (4.18) (No mismatch handling). One path Rayleigh fading channel

Number of symbols in acquisition mode $X=100$
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Figure 4.18: Probability of acquisition error in an UL scenario for the MOE algorithm using equation (4.27). Three Paths Rayleigh fading channel (Table 4.1)

Number of symbols in acquisition mode \( X = 10.0 \)

Figure 4.19: Probability of acquisition error in an UL scenario for the MOE algorithm using equation (4.27). One path Rayleigh fading channel, Number of symbols in acquisition mode \( X = 100.0 \)
Figure 4.20: Probability of acquisition error vs. near-far ratio in an UL scenario. The MOE algorithm using equation (4.27), Three Paths Rayleigh fading channel (Table 4.1), $X=100$.

Figure 4.21: Probability of acquisition error vs. near-far ratio in an UL scenario. Three Paths Rayleigh fading channel (Table 4.1), Conventional receiver, 127 symbols & spreading factor 31.
Figure 4.22: Probability of acquisition error vs. near-far ratio in an UL scenario. Three Paths
Rayleigh fading channel (Table 4.1), Conventional receiver,
31 symbols & spreading factor 127
4.5 Acquisition time analysis

The algorithms presented in this chapter can be considered as parallel acquisition algorithms according to the classification introduced in chapter 2. The flow diagram of this kind of algorithm is shown in Figure 4.23.

![State diagram of the acquisition algorithm](image)

Referring to section 2.7 and the access slot illustrated in Figure 2.11, it is straightforward to find out that the number of cells, C, for these parallel methods is 1, and we have:

$$H_D(z) = p_d z^{XNT_e}$$  \[4.36\]

$$H_m(z) = p_m z^{XNT_r + T_r} + p_f z^{XNT_r + T_r}$$  \[4.37\]

$$H_0(z) = 1$$  \[4.38\]

where X, N and T represent the number of symbols used for acquisition algorithm, spreading factor and reset time, respectively. The latter is the time that receiver has to wait until the next access slot is arrived (Figure 2.11). In these cases T is equivalent to the penalty time, which is the necessary time for rejecting the false alarm and returning to the acquisition mode. The false alarm recovery time depends on the structure of receiver. However for the time slotted mode, this recovery time should be shorter than time interval between two consecutive slots. Also, for these parallel structures we have: $p_m + p_f = (1 - p_d)$.

Substituting \[4.36\], \[4.37\] and \[4.38\] in \[2.35\] and using \[2.36\] the average
acquisition time $T_{acq_{-av}}$ is obtained as follows:

$$T_{acq_{-av}} = \left[ \frac{d}{dz} H(z) \right]_{z=1} = XNT_c / p_d + (1 - p_d)T_r / p_d$$  \hspace{1cm} (4.39)$$

where $P_d$ is the probability of correct acquisition, $T_c$ denotes the chip time interval and $p_m + p_f = (1 - p_d)$ for these parallel algorithms.

Figure 4.23 and Figure 4.24 use the derived formula in equation (4.39) to illustrate the average acquisition time of the training and C-MOE algorithms, respectively. The reset time $T_r$, which in this case is the same as penalty time, is selected 1000 symbols, which is less than 10 ms. (For example in UMTS scenario, the frame length of RACH is 10 ms). Of course the actual acquisition time performance depends on the actual values of penalty time, reset time, evaluation time and so on. Therefore the Figures 4.24 and 4.25 are valid just for the scenario assumed in this section.

![Graph](figure4_24.png)

**Figure 4.24**: Average acquisition time of the training algorithm in a single path and multi-path fading channel (Table 4.1) vs. number of symbols “X”.

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4.6 Summary

In this chapter adaptive algorithms for time delay acquisition were studied. When there is the possibility of transmitting a training sequence, training algorithm can be used for joint acquisition and demodulation of a CDMA signal. Acquisition time can be adjusted by proper selection of training sequence structure, using different period of Gold codes for training and spreading purposes.

It was also shown that a training algorithm does not suffer from the mismatch and exploit the time diversity much better than a blind approach.

However, the application of training sequences for packet-based networks will not be feasible. The usage of training sequences, and reacquisition for every packet service will result in unacceptable overhead. Therefore this has motivated interest into blind techniques.
Blind adaptive receivers for joint acquisition and demodulation were studied here; in particular we focused on the constraint minimum output energy (C-MOE) scheme. The mismatch problem was addressed and its effect on the acquisition performance was investigated. A novel method for adaptive selection of the optimisation constraint, which results in adaptive handling of the mismatch, was proposed.

Performance evaluation of the algorithm for the uplink and downlink of a mobile channel shows a good near far resistant, as well as interference suppression characteristic of the acquisition process.

Simulation results show that the method of handling the mismatch improves the acquisition performance in both frequency-selective and flat fading channels in comparison with a MOE receiver without handling the mismatch. However the performance improvement for a flat fading channel is not as much as a frequency selective condition.

Also the acquisition performance of a conventional acquisition system in the presence of MAI and near-far effect was evaluated and finally the acquisition time analysis was addressed.
In chapter 4, the acquisition characteristic of a practical blind adaptive CDMA receiver, which is based on constraint minimum output energy (C-MOE), was studied. According to the time delay acquisition, receiver performance in a multi-path fading mobile channel was investigated and the problem of mismatch handling was discussed. In this section we discuss that for any practical blind receiver based on the idea of joint acquisition and demodulation of data (JAD), a single step constraint is not sufficient especially if a multi-path channel is considered. In this case, multiple constraints are necessary. In other words, an acceptable acquisition performance of this type of receivers, does not guarantee to give a good multi-path-combining characteristic for the final demodulator.

This is mainly because the acquisition process is a short-term process, and any further information about instantaneous conditions of the channel is not available during this stage. Note that the traditional channel estimation algorithms exploit the timing information, and could be implemented only after a successful synchronisation step.

It is worth noting that regarding a JAD receiver, it is very important to have an
acceptable initial BER. In this way, the receiver is able to perform the demodulation of data immediately after the acquisition, or initiate some adaptive algorithms, such decision-directed techniques, to refine itself during time.

The idea of integrating an interference suppression channel estimation step, which uses the same constraint minimum output energy structure and the same collected data as acquisition mode, is the main motivation behind this chapter. To achieve this goal, a new optimisation problem is defined, which uses the information provided by the acquisition mode. This gives enhanced knowledge surrounding the channel condition and results in a better receiver filter.

However, it is clear that after acquisition mode, mismatch still exists due to unknown channel coefficients and imperfect multi-path combining by the acquisition step. It is shown in this chapter that by extending the method of mismatch handling used for the acquisition process to this refinement step, a considerable improvement in terms of starting BER performance is obtained.

### 5.1 Near-far resistant characteristic of JAD receiver

It was explained in the previous chapter that the output of acquisition mode in a JAD algorithm is a near-far resistant demodulator. Here we show this characteristic using simulation results. However, in the next section it is explained that because of lack of time diversity due to poor combining, this characteristic does not necessarily result in an acceptable BER performance in a multi-path-fading channel.

Figure 5.1 illustrates the near-far resistant characteristic of the receiver FIR filter.

The SIR (Signal to Interference-Noise Ratio) of the resulting demodulator, which is obtained through the successful acquisition process, is calculated analytically using equation (5.1). The results are averaged over 100 simulation runs. In this calculation it is assumed that the receiver rejects the false alarm (FA), immediately and thus the averaging is performed over 100 successful acquisition steps relating to 100 different channel conditions.
In equation (5.1), \( \tilde{f}_{\text{out}} \) denotes the general receive filter and is equivalent to \( \tilde{f}_{\text{meas}} \) for the problem in hand. \( \bar{u}_0 \) and \( \bar{u}_j \) are the signature vectors of the desired user and interfering signals respectively. \( \sigma^2 \) is the variance of noise and \( J \) shows the number of interfering signals that for this case is 10. It is seen from Figure 5.1 that increasing the power of the interfering users by 20 dB over the desired signal, results in a maximum 3dB loss of SIR, where 50 symbols have been used for simulations. However, increasing the number of symbols that are contributing to the acquisition process (see "X" in equation (4.11)) can decrease the loss of SIR, effectively. Although this is a desirable characteristic, this does not mean that the calculated receiver provides good multi-path combining and gives a good BER performance.

\[
SIR = \frac{\left| \langle \tilde{f}_{\text{out}}, \bar{u}_0 \rangle \right|^2}{\sum_{j=1}^{J} \left| \langle \tilde{f}_{\text{out}}, \bar{u}_j \rangle \right|^2 + \sigma^2 \left\| \tilde{f}_{\text{out}} \right\|^2} \tag{5.1}
\]

Figure 5.1: SIR of the blind JAD algorithm vs. number of symbols used for acquisition process. 10 MAI users in a 3-path UL fading channel (Table 4.1) were considered.
5.2 Conventional receiver’s exploitation of time diversity

In transmission of wideband CDMA signals, the chip rate is much larger than the coherency bandwidth of the channel. The channel acts as a frequency selective environment, where its delay spread can provide time diversity. In this case, we assume uncorrelated (in an ideal case) versions of the transmitted signal, the coherent combination of which can provide improved detection of data. In addition, an important characteristic of wideband CDMA is exploiting this diversity to increase the system capacity in channel fading conditions.

A conventional WCDMA receiver uses a Rake structure to benefit from this phenomenon. Figure 5.2 represents a Rake structure. There are different possibilities to select the combining coefficients, $\beta_i$, presented by equation (5.2).

$$ Z = \sum_i \beta_i x_i $$

For example, the maximum ratio combining (MRC) is a well-known method of combining, which is optimal when the effect of multiple access interference can be neglected or assumed as AWGN noise.

Now, remember that a blind interference resistant acquisition algorithm generally starts by constructing a set of hypotheses about the signature waveform of the desired signal. Therefore the final interference suppression receiver, calculated during the acquisition mode, completely depends on this construction.

Referring to section 4.4.8, it is seen that each type of hypothesis construction results in a different combination of multi-paths in the final receiver vector.

For example, assume that the selected hypothesis according to the acquisition mode is a member of single path hypotheses set as described in section 4.4.8. Therefore, the receiver vector $\tilde{f}_{\text{near}}$, actually does not exploit the available time diversity, while the acquisition algorithm may benefit from this diversity. As a result, the BER performance of the JAD receiver may be degraded effectively.
However, even by using equal gain combining hypotheses set for the acquisition process, an appropriate exploitation of multi-path effect cannot be achieved. This is mainly because the dimension of ambiguity during the coarse acquisition mode is too high (uncertainty about chip level and symbol level timing, as well as the unknown channel coefficients and even un-detected paths). Therefore, even the best hypothesis suffers from considerable amount of mismatch with the instantaneous desired signature waveform.

It is possible to start from an available channel profile, but this just reflects the average characteristic of the channel not the instantaneous situation of the multi-path.

The idea of handling the mismatch by inserting the fictitious noise can prevent the complete signal rejection, and can provide correct time delay acquisition on most occasions. However, the receive filter generally becomes matched, as closely as possible, to the weakest path to decrease the output energy. This also results in a degraded BER performance.

Furthermore, it is well known that a conventional receiver, as illustrated in Figure 5.2, suffers from MAI floor and is highly near-far and MAI limited.

The facts mentioned in the previous paragraphs, initiates a study of the integration of interference suppression channel estimation with the acquisition step.
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It is worth noting that we are interested to do this refinement, without any new collection of data or increasing overall complexity excessively. It is well known that in this type of receiver, re-adjustment of the sampling time (compensating the first path delay) after acquisition process is not applicable because any shift in sampling time changes the structure of the multiple access interference and hence the receiver does not remain an interference suppression detector any more. This means that the receiver needs to collect a new block of data and calculate a new correlation matrix (to construct a new receiver filter) if it adjusts its sampling time according to the acquisition outcomes [Smith99].

Therefore, the interference resistant algorithms for demodulation of data and channel estimation, which generally assume a known multi-path delay and synchronised receiver, need to collect a new set of symbols to estimate some statistical characteristics of the channel such as its covariance matrix $R_m$. Some types of these methods were studied in references [Tsat97] [Veen00].

### 5.3 JAD receiver's exploitation of time diversity

In chapter 4, it was shown that in a multi-path fading channel the received signal could be represented as:

$$
\bar{r}_n = d_0[n]u_0 + \sum_{i=1}^{J} d_i[n]u_i + n_0(n) = b_0[n]u_0 + \sum_{i=1}^{J} b_i[n]u_i + n_0(n)
$$

where $b_0[n]$ is the desired symbol and $u_0$ denotes its signature vector, assuming the effect of a multi-path channel and chip and symbol level asynchronous conditions.

It is clear that in a linear solution, we are going to find a receiver vector $\tilde{f}$ that:

$$
\tilde{f}^T \tilde{r}_n = \delta_0[n] = b_0[n]
$$

If the acquisition mode is completed successfully, then some ambiguity surrounding the signature vector $\tilde{u}_0$ can be removed using the information provided by the acquisition step. This is observed by rewriting the signature vector $\tilde{u}_0$ as, $\tilde{u}_0 = \tilde{A}_0 \times \tilde{c}_0$. 

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Vector $\overrightarrow{e}_0 \in C^{1\times 1}$ represents the multi-path channel coefficient vector, and still remains unknown after acquisition mode. Matrix $\overrightarrow{A}_0$ shows the code signature waveform of the desired user, where each column represents the related shifted version of its spreading code. $\overrightarrow{e}_c(n)$ denotes nth chip of the desired user's spreading code sequence. Now the vector $\overrightarrow{e}_0$ is the new unknown vector to be found.

In equation (5.4), the number of zeros in the first column and before $\overrightarrow{e}_0(0)$ equals the sample delay of the first path; this is the case when generally the strongest path, which is captured during the acquisition mode, is the first one.

For a general case $\overrightarrow{A}_0$ is shown by equation (5.6). In this equation, the arrow points to the column that represents the strongest path and its time delay estimation has the maximum certainty.

$$\overrightarrow{A}_0 = \begin{bmatrix}
0 & \ldots & 0 & \ldots & 0 \\
\vdots & \ddots & \vdots & & \vdots \\
\overrightarrow{e}_c(0) & \ldots & \overrightarrow{e}_c(0) & \ldots & \overrightarrow{e}_c(0) \\
\vdots & \ddots & \vdots & \ddots & \vdots \\
0 & \ldots & \overrightarrow{e}_c(N-1) & \ldots & 0 \\
\vdots & \ddots & \vdots & \ddots & \vdots \\
0 & \ldots & 0 & \ldots & 0
\end{bmatrix}_{(2\times N + L \times 2L)} \quad (5.6)$$

In this case, the number of columns of $\overrightarrow{A}_0$ is increased to capture the highest possible number of paths (according to the delay spread) on either side of the strongest path.
Now, the ambiguity in the MOE optimisation problem is reduced to the unknown channel coefficients. As a result the optimisation criterion $\tilde{f}_{\text{moe}}^H \tilde{R}_r = 1$ in equation \begin{equation} \label{5.7} \end{equation} 

$$\min_{\tilde{f}_{\text{moe}}} \tilde{R}_r, \tilde{f}_{\text{moe}} \quad \text{Subject to} \quad \tilde{f}_{\text{moe}}^H \tilde{R}_r = 1,$$

which is the same as equation \begin{equation} \label{4.13} \end{equation}, can be changed to $\tilde{R}_r^H \tilde{f}_{\text{moe}} = \tilde{e}_o$.

Therefore, another minimum output energy optimisation can be applied, searching for the channel estimation $\tilde{e}_o$ where $\tilde{R}_r$ is known. The new optimisation problem and its relating solution are presented as follows.

$$\min_{\tilde{f}_{\text{op}}} \tilde{R}_r, \tilde{f}_{\text{op}} \quad \text{Subject to} \quad \tilde{R}_r^H \tilde{f}_{\text{op}} = \tilde{e}_o$$

\begin{equation} \label{5.8} \end{equation}

$$\tilde{f}_{\text{op}} = \tilde{R}_r^{-1} \tilde{A}_0 (\tilde{A}_0^H \tilde{R}_r^{-1} \tilde{A}_0)^{-1} \tilde{e}_o$$

\begin{equation} \label{5.9} \end{equation}

It is worth noting that in this case, we do not use the constraint on the norm of the receiver filter or the penalty factor in the unconstraint form, as was considered in section 4.4.4. This is because we assume the delays, which were estimated during the acquisition mode, are correct and also we like to use the well-known analytic solution to the optimisation problem \begin{equation} \label{5.8} \end{equation}. Therefore a finite number of hypotheses are not used in this case.

The best solution for $\tilde{e}_o$, according to the minimum output energy criterion, is such a complex vector $\tilde{e}_{\text{op}}$ that its corresponding receiver filter \begin{equation} \label{5.9} \end{equation} (i.e. the answer to the optimisation problem \begin{equation} \label{5.8} \end{equation}) maximises its output energy. Therefore $\tilde{e}_{\text{op}}$ is the solution to the maximisation problem presented by equation \begin{equation} \label{5.10} \end{equation}.

$$\tilde{e}_{\text{op}} = \max_{\tilde{e}_o} \text{moer}(\tilde{e}_o) = \tilde{e}_o^H (\tilde{A}_0^H \tilde{R}_r^{-1} \tilde{A}_0)^{-1} \tilde{e}_o / \| \tilde{e}_o \|^2$$

\begin{equation} \label{5.10} \end{equation}

The analytic solution to this problem is simply the eigen-vector of matrix $\tilde{E} = (\tilde{A}_0^H \tilde{R}_r^{-1} \tilde{A}_0)$, corresponding to its minimum eigen-value [Hayk85]. It is worth noting that $\tilde{R}_r^{-1}$ and $\tilde{A}_0$ are known after the acquisition mode. Therefore, this step uses the same collected information bits, correlation matrix and its inverse as the
acquisition mode, and there is no need for more collection of data or inverse calculation of the autocorrelation matrix. The dimensions of matrix $\mathbf{E} \in \mathbb{C}^{L \times L}$ are defined by number of paths, which are much smaller than the dimensions of the autocorrelation matrix $\mathbf{R}_r$.

5.4 Mismatch handling and receiver refinement

Although the $\hat{c}_{ep}$ calculated in the previous section can be used to give the receiver vector, using equation (5.9), still mismatch between the estimated signature vector of the desired user and its actual value exists. This is due to error in the multi-path delay estimation (i.e. error in $\hat{A}_p$), as well as error in channel estimation. In [Tsat97], a receiver with full knowledge of all the multi path delays uses this approach to obtain an estimate of the channel coefficients, where mismatch handling is not considered.

We propose to use the same idea of mismatch handling as acquisition mode, using the technique of fictitious noise presented in equation (4.27). The idea is explained as follows.

The channel estimation step uses the same received signal structure as acquisition mode, to obtain a better estimation of the signature vector. This new version (refined version) is applied to equation (4.27) that benefits from the mismatch handling technique to prevent the desired signal rejection. Therefore, the final receiver filter is calculated as equation (5.11). This equation is obtained by replacing $\hat{u}_c$ in equation (4.27) with the new version $\hat{u}_q = \mathbf{A}_p \times \hat{c}_0$. In following simulation results, the receiver presented by equation (5.11) is referred to as enhanced receiver for joint acquisition and demodulation of data, E-JAD, while the receiver proposed in the previous chapter (Equation (4.27)) is referred to as JAD receiver.

$$
\tilde{f}_{opp} = \frac{(\mathbf{R}_r + \nu \mathbf{I})^{-1} \mathbf{A}_q \hat{c}_{ep}}{\hat{c}_{ep}^H \mathbf{A}_q^H (\mathbf{R}_r + \nu \mathbf{I})^{-1} \mathbf{A}_q \hat{c}_{ep}} \quad (5.11)
$$

The parameter $\nu$ has the same value as was calculated during the first step of the
delay acquisition process.

Simulation results in the next section show that using the proposed equation (5.11) instead of standard form (presented by equation (5.9)) improves the BER performance, even in comparison with a receiver with full knowledge of the multi-path delays, but without considering the effect of mismatch [Tsat97]. The receiver in equation (5.9), which assumes that the multi-path time delays are known, and was compensated after acquisition mode, is referred to as basic minimum output energy, B-MOE receiver. By adjusting the estimated delays the structure of interference is completely changed [Smit99] and a new estimation of the autocorrelation matrix for the demodulation step becomes necessary.

Equation (5.11) also outperforms the receiver calculated by equation (4.27), which tries to handle the mismatch but doesn’t exploit the multi-path diversity.

In summary, the proposed modified algorithm for joint acquisition and demodulation of CDMA signal consists of the following steps:

List: 5.1

1- Compute the fictitious noise factor $\nu$ as explained in section 4.4.6.

2- Estimate the delay of the first arrival path or the strongest path.

3- Calculate the vector $\hat{\delta}_o$, which obtains a more realistic model of the channel, using (5.10) and its associated eigen-value problem.

4- Compute the receiver filter $\hat{f}_{app}$, as proposed in equation (5.11)

Figure 5.3 illustrates the block diagram of the final B-JAD receiver. It is worth noting that the refinement process uses the available information after acquisition mode such as the auto-correlation matrix and its inverse.
5.4.1 Pilot channel aided receiver

The eigen-value decomposition of matrix $E$, gives $\hat{c}_{op}$ with a multiplicative complex coefficient. This effect can rotate the detected complex bits and resulting in incorrect detection, especially for higher order modulations.

In the down link channel, a reasonable assumption that can be used for many different purposes is pilot channel assumption.

One interesting characteristic of blind algorithms for joint acquisition and detection of data is that it supports data modulation during the synchronisation phase. Hence, we may have a modulated pilot channel, while the proposed receiver can use a very short duration of this pilot signal for phase correction. In the downlink, this pilot comes through the same channel as the desired user and experiences the same delay and fading.

Furthermore, because the algorithm is near-far resistant and effectively suppress the MAI, it can recover the pilot channel even in presence of strong interfering signals. The acquisition process, as well as the channel refinement is performed using this
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pilot channel. This pilot is common between all users in downlink, and therefore the receiver actually does not need to know even the spreading code of the desired user for the acquisition purpose. Finally the calculated receiver vector in equation (5-11) is passed to the reception block for demodulation of data.

The same receiver can be used in the uplink, where the uplink pilot channel is considered, or equivalently a few preamble symbols at the start of each packet session are used to correct the phase. This preamble section is very shorter than what is necessary for acquisition, according to a conventional acquisition approach. Simulation results in the next section show that generally four or five symbols are enough to average out the effect of noise and correct the phase.

5.5 Simulation results and discussions

To investigate the BER performance of the enhanced blind CMOE receiver for joint acquisition and demodulation of data (illustrated in Figure 5.3), E-JAD, a Mont Carlo simulation was set for a mobile fading channel. A three-path WSSUS channel is considered, where Doppler frequency is 120 Hz and relative powers and delays of multi-paths are fixed as [0, -3, -3] dB and [0, 3, 10] chips, respectively.

After every successful delay acquisitions, the BER of the receiver in equation (5.11) is calculated (over 20000 information bits) and results were averaged over 500 different fading channel conditions (500 successful acquisition with different quality of calculated receiver). Mobile velocity was set to 80km/h and Eb/No=15dB.

The results are also compared with the basic MOE receiver of equation (5.9) (B-MOE) proposed in [Tsat97], where it was assumed that perfect knowledge of all path-delays is available (time delay acquisition was performed) and delays are already adjusted. Therefore a new collection of data after acquisition is necessary. Also the mismatch handling has not been considered in their approach.

To see how much gain is obtained through the refinement, the BER performance of the basic CMOE for joint acquisition and demodulation of data (B-JAD) is also presented (based on equation (4.27)). For the downlink channel, these results are also
compared with a Rake receiver, where it is assumed that the delays and channel coefficients are completely known and thus the simulation is biased in favour of the Rake receiver. In all simulations the spreading factor is 31 and delays of users as well as channel conditions are selected randomly at the start of every simulation run. It is worth noting that the presented BER in following figures is the initial BER just after the acquisition mode.
Figure 5.4: BER vs. PMAI (offset) dB. MAI = 15 users.
Acquisition window size = 100 symbols. DL scenario.

Figure 5.5: BER vs. PMAI (offset) dB. MAI = 15 user.
Acquisition window size = 200 symbols. DL scenario.
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Figure 5.6: BER vs. PMAI(offset) dB. MAI = 5 users.
Acquisition window size = 100 symbols. UL scenario.

Figure 5.7: BER vs. PMAI(offset) dB. MAI = 5 users.
Acquisition window size = 200 symbols. UL scenario.
Figure 5.8: BER vs. PMAI (offset) dB. MAI = 10 users.
Acquisition window size = 100 symbols. UL scenario.

Figure 5.9: BER vs. PMAI (offset) dB. MAI = 10 users.
Acquisition window size = 200 symbols. UL scenario.
Figure 5.10: BER vs. PMAI(offset) dB. MAI = 15 users.
Acquisition window size = 100 symbols. UL scenario.

Figure 5.11: BER vs. PMAI(offset) dB. MAI = 15 users.
Acquisition window size = 200 symbols. UL scenario.
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Figure 5.12: BER vs. acquisition window size. MAI=10 users.

UL scenario.

Figure 5.13: Mean and variance of channel phase detection error vs. PMAI(offset) dB. UL scenario.
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Figure 5.14: The Histogram of error energy in detection of the channel amplitudes.

UL scenario.

The BER performance of the Rake receiver, basic MOE (B-MOE), basic joint acquisition and demodulation (B-JAD) and enhanced receiver for joint acquisition and demodulation (E-JAD) were compared in Figure 5.4 for a DL channel scenario where there are 15 active users in the network. E-JAD clearly outperforms the other algorithms, even in the presence of strong MAI. Although it is not significant, B-JAD still works better than B-MOE. This shows the effectiveness of the joint acquisition and demodulation approach.

Figure 5.5 illustrates the same results as Figure 5.4, except that 200 symbols are used for the acquisition process. Increasing the number of symbols clearly improves the BER performance, but at a cost of longer acquisition time. It is seen from this figure that by increasing the power of interference, the BER of B-MOE and E-JAD degrade relatively faster in comparison to Figure 5.4. This happens because the channel estimation performance degrades. Although, using 200 symbols improves the acquisition
performance, the starting point BER suffers more from channel variations during the collection of data. In this case, the autocorrelation function actually presents the average channel conditions, and therefore the degradation is a result of channel refinement error. It is also seen that the B-JAD that does not use the channel refinement step, has a relatively flat performance.

Figure 5.6 shows that in the UL scenario, the performance improvement by B-JAD and E-JAD is more significant than DL, where both outperform the B-MOE algorithm.

Figure 5.7 illustrates the same results as Figure 5.6, except that 200 symbols are used for acquisition. Increasing the acquisition time improves the BER performance. We see that the performance of B-JAD and B-MOE converge together, and it shows that the mismatch handling of B-JAD and the channel refinement of B-MOE are balanced in this case.

Figures 5-8 to 5-11 compare the BER performances in an UL scenario, where the system load is increased. It is seen that by increasing the system load and in the presence of strong near-far effect, BER of E-JAD and B-MOE are increased relatively faster than BER of B-JAD. Again, this is due to poor channel estimation performance for this highly loaded case.

Figure 5.12 illustrates the BER performance of the receiver, presented in Figure 5.3 vs. the number of symbols used for the acquisition process. Number of MAI users is 10 and PMAI was set to 10 dB. It is seen that until we reach a given point (round 250 symbols in this case), BER decreases as the acquisition window size or, equivalently, the autocorrelation estimation length is increased. For B-JAD and E-JAD, this is obtained at the expense of longer acquisition time, and for B_MOE this means collection of more received signal vectors. On the other hand after some point, the BER increases by increasing the number of received symbols. This is due to the time varying characteristic of mobile channel that can't be assumed constant during the acquisition mode; also it clarifies the necessity of a tracking algorithm after the short interval acquisition process. Tracking algorithms aren't considered in this thesis.

Mean and variance of phase estimation error for different paths of UL channel is
illustrated in Figures 5.13.

Also, the histogram in Figure 5.14 shows the energy of error in the detection of multi-path amplitudes.

5.6 Complexity discussion

One important issue about blind adaptive algorithms is their implementation complexity. The most part of the computational complexity is due to the need to invert a large autocorrelation matrix $\mathbf{R}$. Different techniques of approximate low complexity matrix inversion are proposed in literature, two of these are presented here.

Polynomial expansions (PE) are one group of approximately matrix inversion methods.

One characteristic of these methods, according to our problem in hand, is that we need to know the autocorrelation matrix $\mathbf{R}$, or at least a reliable estimate of this matrix. When this knowledge is available, some expansions such as the Taylor series, presented in following paragraphs, can be used to obtain the inverse matrix. This does not need any new collection of data.

For more information on expansions with different convergence rates and complexity see [Helm96] and [Lei98].

Taylor series:

Using the Taylor series, the expansion of $\mathbf{R}^{-1}$ is expressed as:

$$\alpha_\tau^j\mathbf{R}^{-1} = \sum_{i=0}^{\infty} (1 - \alpha_\tau \mathbf{R})^i$$  \hspace{1cm} (5.12)

where $|1 - \lambda_{i,j}(\alpha_\tau \mathbf{R})| < 1$ is the convergence condition of equation (5.12). $\lambda_{i,j}(\alpha_\tau \mathbf{R})$ denotes the $i^{th}$ eigen-value of matrix $\alpha_\tau \mathbf{R}$. For positive semi-definite matrix $\mathbf{R}$ the above condition can be expressed as: $0 < \alpha_\tau < \frac{2}{\lambda_{e,max}^2(\mathbf{R})}$. 

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In a simpler way, $\alpha_r$ can be selected as $\alpha_r = \frac{2}{\text{trace}(R)}$. Also this is a practical constraint, when the calculations of eigen-values of $R$ are not desirable in terms of time and complexity.

Some discussion about rate of convergence, and some modifications can be found in [Moza00]. Here we just address the fact that complexity reduction is possible by using the approximate polynomial expansions.

The other different approach of calculating the inverse of autocorrelation matrix is based on a lemma presented by equation 5.13.

In this approach the inverse matrix is calculated during the collection of data. The simulation results, which are presented at the end of this section, show that this method is faster and more practical than the polynomial expansion. On the other hand, before applying the acquisition process, the receiver needs to wait until it receives enough information blocks. Therefore it is more convenient for the receiver to adaptively estimate the inverse of the autocorrelation matrix during this waiting time. The following section addresses the matrix inversion lemma [Scha91].

Matrix inversion lemma:

One approach to avoid explicit matrix inversion is based on the matrix inversion lemma in equation (5.13) [Scha91].

\[(\overline{A} + \overline{BCD})^{-1} = \overline{A}^{-1} - \overline{A}^{-1}\overline{B}(\overline{DA}^{-1}\overline{B} + \overline{C}^{-1})^{-1}\overline{DA}^{-1}\]  

By replacing the matrixes $\overline{A}$, $\overline{B}$, $\overline{C}$ and $\overline{D}$ in equation (5.13) with autocorrelation matrix $\overline{R}$, received vector $\overline{r}_i$ in $i_{th}$ iteration, unitary matrix $\overline{I}$ and $\overline{r}_n$, respectively, a recursive solution for inverse matrix $\overline{R}^{-1}$ is obtained. This is shown by equation (5.14). Index $i$ in this equation represents the number of iteration steps, which is the same as the number of collected symbols during the acquisition process.

\[\overline{[R^{-1}]}_i = [\overline{R}_{i-1} + \overline{r}_i \overline{r}_i^H]^{-1} = \overline{[R^{-1}]}_{i-1} - \frac{[\overline{R}^{-1}]_{i-1} \overline{r}_i \overline{r}_i^H [\overline{R}^{-1}]_{i-1}}{1 + \overline{r}_i^H [\overline{R}^{-1}]_{i-1} \overline{r}_i}, \quad i = 0, 1, \ldots\]  

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If there is no initial assumption, it is sufficient to start from identity matrix (i.e. $R_0^{-1} = \bar{I}$).

### 5.6.1 Numerical examples and discussions

To show the effectiveness of the inverse algorithms explained in the previous section, some simulations were run for an uplink communication link. We consider 10 users in the network where the near-far ratio is 5dB; the other simulation parameters are the same as section 5.5. A measure of convergence is the Frobenius norm\(^2\), $F-N$, of the error between matrixes $R^{-1} \times \bar{R}$ and identity matrix $\bar{I}$. This norm is also normalised to the $F-N$ of identity matrix $\bar{I}$.

For the Taylor series the autocorrelation matrix is first estimated based on 100 received symbols and then the inversion is applied for a different number of polynomial terms in equation 5.12.

Moreover, for the lemma presented in equation (5.13), the inversion is carried out adaptively during the collection of information blocks. It is seen that by increasing the number of symbols, the error norm decreases rapidly. For example, it is observed that 100 symbols (necessary for reliable acquisition performance) can provide a good approximation of the inverse matrix.

Figures 5.15 and 5.16 illustrate the simulation results for the error norm of the two different approaches that have been mentioned here. It is seen that the lemma of equation (5.13) provide a good approximation of the inverse of autocorrelation matrix and, does not need further delay for additional calculation after collection of symbols. This is also desirable for a faster acquisition process.

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\(^2\)Frobenius-norm($X$) $= \sqrt{\text{Trace}(X^* \times X)}$
Figure 5.15: F-N of error in calculating the inverse matrix (using (5.12))

vs. the number of polynomial terms. “alfa” denotes the convergence parameter $\alpha$ in equation (5.12).

Figure 5.16: F-N of error in calculating the inverse matrix (using (5.14))

vs. the number of iterations in equation (5.14)
5.7 Summary

The BER performance of the JAD algorithm was studied in this section. It was shown that although the JAD algorithm can give a good acquisition performance and the calculated FIR filter coefficients from the acquisition mode is a near-far resistant receiver, it does not exploit the channel diversity appropriately. Therefore this class of receiver does not provide acceptable BER performance.

By combining an appropriate method of channel estimation, which is compatible with the structure of acquisition mode, and does not need any further collection of data, a novel receiver structure was proposed in Figure 5.3.

The idea of mismatch handling proposed in chapter 4, was also combined with the channel refinement procedure. As a result a novel formula for calculation of final receiver vector was proposed and presented by equation (5.11).

Simulation results for different channel conditions were used to investigate the effectiveness of the proposed algorithm, as well as the behaviour of the receiver by changing the acquisition time length, level of MAI and the near-far effect. These results show that the E-JAD approach outperforms the B-MOE receiver presented by equation (5.9). This is mainly because this receiver does not consider the effect of mismatch. Also the E-JAD algorithm clearly outperforms the B-JAD and the Rake receiver in terms of BER performance.

Finally, an approximate matrix inversion scheme and its effectiveness in a mobile channel environment were studied. It was discussed that the acquisition time length is enough for adaptively calculating the inverse of the autocorrelation matrix using the lemma presented by equation (5.13).

This lemma can also be used for tracking the autocorrelation matrix or, in other words, tracking the mobile channel.
Chapter 6

6 Conclusions and future work

6.1 Conclusions

In this dissertation the concept of interference resistant algorithms for joint acquisition and demodulation of CDMA signal was studied.

The limitations of conventional receivers, which are based on the matched filter structure, for acquisition of time delays in a CDMA communication network were studied. It was concluded that the acquisition-based capacity of the mobile communication network could be a bottleneck for the CDMA system design. This motivated the study for interference resistant algorithms for time delay acquisition in an attempt to maximise system throughput.

Furthermore, the hostile problems such as multiple access interference, near-far effect, multi-path channel characteristics and the mobile environment are common issues for both the acquisition and demodulation processes. Therefore the idea of adaptive interference suppression algorithms for joint acquisition and demodulation of CDMA signal (JAD) is another interesting concept. In this regard, the training algorithm and the C-MOE algorithm for performing the JAD approach were studied.

It was shown that the training algorithm could provide good acquisition performance as
well as perform an interference resistant receiver. It does not suffer from mismatch and
therefore can exploit the multi-path diversity to improve the acquisition performance,
effectively. Also by proper selection of the training sequence, shorter acquisition time is
possible, at the expense of smaller number of possible transmissions in simultaneous
acquisition mode.

In addition, new systems such as packet-based radio access networks, ad-hoc networks
and high mobility environments are strong initiative behind the study of blind adaptive
algorithms for time delay acquisition. In these scenarios, it is very important to remove
the preamble part from every packet session. This results in a considerable reduction of
redundancy in comparison to the conventional approaches that exploit preamble symbols
for acquisition purpose.

In this thesis a linear blind interference resistant receiver, based on the C-MOE algorithm
was studied, and a novel structure of such a receiver was proposed that is capable to cope
with the variable characteristics of the mobile channel. The proposed structure was also
evaluated in terms of both code acquisition and demodulation performance.

The JAD receiver shortcoming was due to inability to exploit the multipath diversity
properly. This drawback can degrade the demodulation performance. Therefore this
motivated the concept of joint application of a low additive complexity channel
estimation algorithm with the acquisition process. This was investigated comprehensively.

By applying the idea of mismatch handling to the process of channel estimation,
considerable improvement is achieved in the BER performance of the final JAD receiver.
Therefore it is concluded that the blind C-MOE JAD receiver is an interesting and
practical type of receiver that can improve the acquisition and demodulation performance,
effectively.

Based on the work accomplished in this thesis the following publications have been
provided.

- N. Neda, R.Tafazolli, "Simplified near-far resistant technique for joint time
  acquisition and demodulation of CDMA signals", IEE, Electronics Letters
Chapter 6: Conclusions and future work


6.2 Future work

Mobile wireless communication systems are facing the new demands for higher bit rate services, large-scale mobility and more efficient use of network resources (e.g. the spectrum). These already initiated considerable research amongst the communication community. Two different approaches for satisfying these demands are; designing more sophisticated CDMA receivers and using new multiple access techniques.

On the receiver sub-system, as was shown in this thesis the acquisition step could act as a potential bottleneck to the system capacity if not designed properly.

Further research is needed to extend the work which has been carried out in this thesis to the interested multiple access technologies as well as more sophisticated receivers.

More sophisticated receivers:

a) In section 4.4.3, it was mentioned that the MOE structure has also been considered in array signal processing area as a beam-forming algorithm, where the direction of arrival is the parameter to be detected. The similarity between the fundamental approaches means that it would be interesting to combine these two concepts in a CDMA system, which uses array antenna facility, for joint blind estimation of time delays and angle of arrivals.

b) Delays and channel tracking processes were not considered in this thesis. Based on the algorithms investigated in this research, tracking process can be viewed as tracking the correlation matrix (or its inverse) and the corresponding eigen-values (e.g. see [Cham94]). Studying the practical algorithms related to these concepts and combining them with the structure of receiver for joint acquisition and
demodulation (Figure 5.3) is another interesting issue to be considered.

Systems beyond the 3G:

Multi-carrier CDMA (MC-CDMA) already has attracted lots of interest among the worldwide communications community and is considered as a promising candidate for the systems beyond the 3G (Third Generation of Mobile Systems) or 4G. Extending the blind MOE acquisition algorithm to this multiple access scheme would be an interesting area of research.

Moreover, MC-CDMA systems use many carrier frequencies, known as sub-carriers, to transmit the information symbols. Therefore, the frequency offset is an important issue in these systems. Current methods of carrier synchronisation proposed for the orthogonal frequency division multiple access or OFDM (OFDM is a base technology for the MC-CDMA air-interface), usually use the preamble symbols [Heis02]. Extending the C-MOE approach and other blind estimation schemes to perform joint time delay and frequency offset estimation for MC-CDMA systems, is another interesting work to be considered.
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