Helsinki University of Technology Signal Processing Laboratory
Teknillinen korkeakoulu Signaalinkäsittelytekniikan laboratorio
Espoo 2006 Report 57

ADVANCED RECEIVERS FOR HIGH DATA RATE MOBILE COMMUNICATIONS

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Dissertation for the degree of Doctor of Science in Technology to be presented with due permission of the Department of Electrical and Communications Engineering for public examination and debate in Auditorium S4 at Helsinki University of Technology (Espoo, Finland) on the 8th of December, 2006, at 12 noon.

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ISBN-13 978-951-22-8486-3 (Printed)
ISBN-10 951-22-8486-3 (Printed)
ISBN-13 978-951-22-8487-0 (Electronic)
ISBN-10 951-22-8487-1 (Electronic)
ISSN 1458-6401

Otamedia Oy
Espoo 2006

Abstract

Improving the spectral efficiency is a key issue in the future wireless communication systems since the spectrum is a scarce resource. Both the number of users as well the demanded data rates are increasing all the time. Furthermore, in mobile communications the wireless link is required to be reliable even when the mobile is in a fast moving vehicle. Using Multiple-Input Multiple-Output (MIMO) antennas is a well known technique to provide higher spectral efficiency as well as better link reliability. Additionally, higher order modulation methods can be used to provide higher data rates. In order to benefit from these enhancements in practise, sophisticated signal processing methods as well as accurate estimates of time-varying wireless channel parameters are needed.

This thesis addresses the problem of designing multi-antenna receivers in high data rate systems. The case of multiple transmit antennas is also considered. System specific features of High Speed Downlink Packet Access (HSDPA) which is part of 3rd generation (3G) Wideband Code Division Multiple Access (WCDMA) evolution are exploited in channel estimation methods and in MIMO receiver design. Additionally, complexity reduction methods for Minimum Mean Square Error (MMSE) equalization are addressed.

Blind channel estimation methods are spectrally efficient, since no extra resources are needed for pilot signals. However, in mobile communications accurate estimates are needed also in fast fading channels. Consequently, semi-blind channel estimation methods where the receiver combines blind and pilot based channel estimation are an appealing alternative. In this thesis blind and semi-blind channel estimation methods based on knowledge of multiple spreading codes are derived. A novel semi-blind combining scheme for code multiplexed pilot signal and blind estimation is proposed.

Another important factor in receiver design criteria is the structure of interference in the received signals. Interference mitigation techniques in MIMO systems have been shown to be potential methods for providing improved performance. A chip level inter-antenna interference cancellation method has been developed in this thesis for HSDPA. Furthermore, this multi-stage ordered interference canceler is combined with the semi-blind channel estimation scheme to enhance the system performance further.

Abstrakti

Langattomassa tiedonsiirrossa radiospektrin tehokas käyttö on tulevaisuuden suuria haasteita. Taajuuksia on käytössä vain rajoitetusti, kun taas käyttäjien määrä sekä vaaditut siirtonopeudet kasvavat jatkuvasti. Lisäksi langattomien yhteyksien on toimittava luotettavasti myös nopeasti liikkuvissa kulkuneuvoissa. Moniantennijärjestelmät, joissa on useita antenneita sekä tukiasemissa että päätelaitteissa mahdollistavat radiospektrin tehokkaamman käytön sekä parantavat yhteyksien laatua. Tiedonsiirtonopeutta voidaan myös kasvattaa erilaisilla modulaatiotekniikoilla. Hyötyjen saavutamiseksi käytännössä tarvitaan sekä kehittyneitä vastaanotinrakenteita että tarkkoja estimaatteja aikamuuttuvasta radiokanavasta.

Tässä työssä on kehitetty vastaanotinrakenteita ja kanavan estimointimenetelmiä kolmannen sukupolven (3G) nopeiden datayhteyksien (HSPA) järjestelmissä. Työssä on johdettu menetelmiä, jotka hyödyntävät HSPA järjestelmien erikoispiirteitä tehokkaasti. Lisäksi on kehitetty laskennallisesti tehokkaita menetelmiä vastaanottimien signaalinkäsittelyyn.

Ns. sokeat menetelmät mahdollistavat taajuuskaistan tehokkaan käytön, koska ne eivät vaadi tunnettuja harjoitussignaaleja. Mobiileissa tietoliikennejärjestelmissä radiokanava saattaa kuitenkin muuttua hyvin nopeasti, jonka vuoksi kanavan estimoinnissa on tyypillisesti hyödynnetty tunnettua pilottisignaalia. Yhdistämällä pilottipohjainen ja sokea kanavaestimointimenetelmä, voidaan saavuttaa molempien menetelmien edut. Tässä työssä kehitettiin sokeita kanavaestimointimenetelmiä, jotka hyödyntävät useita tunnettuja hajoituskoodeja. Sokean ja koodijakoiseen pilottisignaaliin pohjautuvien kanavan estimaattien yhdistämiseksi kehitettiin uusi menetelmä.

Signaalin laatua ja siten vastaanottimen suorituskykyä voidaan langattomissa järjestelmissä parantaa vaimentamalla interferenssiä eli häiriöitä. Vastaanottimen toimintaa voidaan tehostaa oleellisesti, jos häiriösignaalin rakenne tunnetaan. Käytettäessä useampaa lähetysantennia HSPA järjestelmissä vastaanotetussa signaalissa olevia häiriötä voidaan kumota usealla eri tasolla. Tässä työssä on kehitetty chippitasolla häiriöitä kumoava vastaanotinrakenne, joka hyödyntää HSPA järjestelmän ominaisuuksia. Vastaanottimen suorituskykyä on edelleen parannettu yhdistämällä se aiemmin esitettyyn puolisokeaan kanavan estimointimenetelmään.

Preface

The work constituting this thesis was carried out in the Signal Processing Laboratory at Helsinki University of Technology during years 2001–2006. The Statistical Signal Processing research group, led by Prof. Visa Koivunen, is a member of SMARAD, Centre of Excellence of the Academy of Finland.

I wish to express my gratitude to my supervisor, Prof Visa Koivunen, for his continuous encouragement, guidance and support during the course of this work. It has been both a great honour and pleasure to work with such a dedicated scientist, who also values highly family.

I would like to thank the thesis pare-examiners Prof Saleem A. Kassam and Dr. Jukka Mannerkoski for their constructive and helpful comments that helped to improved the manuscript. I would also like to thank our laboratory secretaries Mirja Lemetyinen and Anne Jääskeläinen for their help with all the practical issues and arrangements.

I am grateful to all my colleagues and co-workers that I have had the pleasure to work with during the years. Especially, I would like to thank Pekka Jänis for the numerous memorable conversations and valuable comments. Additionally, I would like to thank Dr. Kimmo Kettunen for sharing his knowledge and encouraging me throughout this years. Also Prof. Risto Wichman, Dr. Andreas Richter, Dr. Timo Roman and Dr. Jan Eriksson are acknowledged for the many discussions both technical and non technical.

The research work has been funded by Academy of Finland, SMARAD CoE program and other research programs and Nokia Research Center. Also the financial support from Finnish Cultural Foundation and Nokia Foundation is gratefully acknowledged.

I wish to express my warmest gratitude to my parents Helena and Ari for their encouragement and support. Finally, I would like thank to my dear husband Lauri for his love and patience, and our daughter Veera, who has brought so much more love and happiness to my life.

Espoo, November 2006

Maarit Melvasalo

Contents

List of original publications

(I) M. Melvasalo and V. Koivunen. Blind Channel Estimation for Multi-code Transmission in Long Code CDMA Systems. In *Proc. of* 36^{th} *Conference on Information Sciences and Systems, CISS*, Princeton, New Jersey, USA, CD-ROM proceedings, March 2002.

(II) M. Melvasalo and V. Koivunen. Blind Channel Estimation in Multi-code CDMA System with Antenna Array. In *IEEE Workshop on Sensor Array and Multichannel (SAM) Signal Processing*, pp. 565-569, Virginia, USA, August 2002.

(III) M. Melvasalo and V. Koivunen. Blind Channel Estimation in Multi-code CDMA Using Multiantenna Receiver. In *Wireless Personal Communications*, Kluwer, vol. 28, pp. 163–178, 2004.

(IV) M. Melvasalo and V. Koivunen. Semi-blind channel estimation in HSDPA system. In *Conference Record of the* 39^{th} *Asilomar Conference on Signals, Systems and Computers, 2005*, pp. 1292-1296, California, USA, 2005.

(V) M. Melvasalo, P. Jänis and V. Koivunen. Low Complexity Space-Time MMSE Equalization in WCDMA Systems, In *Proc. of* 16^{th} *IEEE International Symposium on Personal, Indoor and Mobile Radio Communications, PIMRC*, pp. 306-310, Berlin, Germany, 2005.

(VI) M. Melvasalo, P. Jänis and V. Koivunen. MMSE equalizer and chip level inter-antenna interference canceler for HSDPA MIMO systems. In *Proc. of the IEEE* 63^{rd} *Vehicular Technology Conference, VTC Spring 2006*, vol. 4, pp. 2008 - 2012, Melbourne, Australia, 2006.

(VII) M. Melvasalo, P. Jänis and V. Koivunen. Low complexity semi-blind channel estimation in HSDPA systems. Submitted for publication. Material presented also in the technical report: Low complexity semi-blind channel estimation in HSDPA systems. *Signal Processing Laboratory Technical Report 55*, Helsinki University of Technology, 2006, ISBN on 951-22-8321-2, ISSN: 1458-6401.

Abbreviations and symbols

Abbreviations

3G	Third generation
3GPP	Third generation partnership project
4G	Fourth generation
AMC	Adaptive modulation and coding
AWGN	Additive white Gaussian noise
B3G	Beyond 3G generation
BER	Bit error rate
BS	Base station
CDMA	Code division multiple access
CIR	Channel impulse response
C-MOE	Constrained MOE
CPICH	Common pilot channel
CQI	Channel quality indicator
CSI	Channel state information
DD	Decision directed
FFT	Fast Fourier transform
GRAKE	Generalized RAKE
HARQ	Hybrid automatic repeat request
HSDPA	High speed downlink packet access
HS-DSCH	High speed downlink shared channel
HSPA	High speed packet access
HS-SCCH	Shared control channel for HS-DSCH
HSUPA	High speed uplink packet access
IAI	Inter antenna interference
IC	Interference cancellation
IFFT	Inverse fast Fourier transform
IMETRA	Intelligent multi-element transmit and receive antennas.
IPI	Inter path interference
ISI	Inter symbol interference
ITU	International telecommunications union
LMS	Least mean square
LS	Least squares

MAI	Multiple access interference
MAP	Maximum *a posteriori*
MIMO	Multiple-input multiple-output
MISO	Multiple-input single-output
ML	Maximum likelihood
MMSE	Minimum mean-square error
MOE	Minimum output energy
MOSIC	Multistage ordered serial interference cancellation
MPC	Multi-code principal component
MPI	Multipath interference
MRC	Maximum ratio combiner
MSE	Mean square error
MSWF	Multi-stage Wiener filter
MUD	Multiuser detector
MV	Minimum variance
MVDR	Minimum variance distortionless response
NLMS	Normalized least mean square
OSIC	Ordered Serial (successive) interference cancellation
PC	Principal component
PIC	Parallel interference cancellation
QAM	Quadrature amplitude modulation
RAKE	Rake
RLS	Recursive least squares
RRC	Root raised cosine
SB	Semi-blind
SB-MPC	Semi-blind multi-code principal component
SIC	Serial (successive) interference cancellation
SIMO	Single-input multiple-output
SINR	Signal to interference and noise ratio
SIR	Signal to interference ratio
SISO	Single-input single-output
SM	Spatial multiplexing
SMI	Sample matrix inversion
S-MMSE	Spatial minimum mean square error
SNR	Signal to noise ratio
ST	Space-time
ST-MMSE	Space-time minimum mean square error
SVD	Singular value decomposition
TTI	Transmission time interval
V-BLAST	Vertical Bell laboratories layered space-time
WCDMA	Wideband code division multiple access
VST	Virtual space-time
WSS	Wide sense stationary
ZF	Zero-forcing

Symbols

$\otimes$	Kronecker product
$\hat{\Theta}$	Estimate of Θ
$*$	Complex conjugate
$\star$	Convolution
$\dagger$	Pseudoinversion
$\forall$	For all
$\mathbf{a(1:N)}$	Elements from 1 to N of a vector $\mathbf{a}$
$\mathbf{a^T}$	Transpose of vector $\mathbf{a}$
$\mathbf{A^H}$	Hermitean transpose of matrix $\mathbf{A}$
$\mathbf{A^{-1}}$	Inverse of matrix $\mathbf{A}$
$\mathbf{A}(\tau,:)$	τth row of matrix $\mathbf{A}$
$\mathbf{0}$	Matrix or vector of all zeros
$\|\|\cdot\|\|$	Euclidean norm
a_l	Auto-correlation with lag l
$\mathbf{a}$	Auto-correlation vector
$\mathbf{a_m}$	Auto-correlation vector for antenna pair $\mathbf{m} = \{m_i, m_j\})$
$\tilde{\mathbf{a}}$	Zero padded auto-correlation vector
α	Semi-blind combining ratio
α_C	Combining weight for all C control codes in semi-blind estimation
α_p	Combining weight for one code p in semi-blind estimation
α_P	Combining weight for for all P HSDPA codes in semi-blind estimation
$\mathbf{b}$	Transmitted MIMO chip sequences for all transmit antennas
c	Control channel index
C	Number of control channels used in estimation (used also as index)
$\mathbf{c}_{nk}$	Code vector for nth symbol of kth user (size $G_k \times 1$)
$\mathbf{C}_{nk}$	Code convolution matrix for nth symbol of kth user (size $(G_k + L - 1) \times L$)
$\boldsymbol{\mathcal{C}}_{nk}$	Code convolution matrix $\mathbf{C}_{nk}$ for multiple receive antennas
$\boldsymbol{\mathcal{C}}_{nc}$	Control signal code convolution matrix for multiple receive antennas
$\boldsymbol{\mathcal{C}}_n$	Multi-code code convolution matrix $= [\boldsymbol{\mathcal{C}}_{n1}, \ldots, \boldsymbol{\mathcal{C}}_{nP}]$
$\tilde{\boldsymbol{\mathcal{C}}}_n$	ISI free part of the code convolution matrix $\boldsymbol{\mathcal{C}}_n$
δ	Dirac delta function, also referred to as the unit impulse
$\mathbf{d}$	Transmitted chip sequence from single antenna
$\mathbf{d}_n$	Transmitted chip sequence corresponding to nth symbol (size $G \times 1$)
$\boldsymbol{d}_n$	IPI extended $\mathbf{d}_n$ for nth symbol (size $(G + 2(L-1)) \times 1$)
$\mathbf{d}_n^{(q)}$	Transmitted chip sequence from q-th antenna for nth symbol
$\tilde{\mathbf{D}}_c$	Pilot chip matrix (size $N_c \times L$)
$\mathcal{D}(\cdot)$	Symbol estimator
$E[\cdot]$	Expectation
$\mathbf{f}$	Filter weight vector for all receive antenna (size $MF \times 1$)
$\mathbf{f}_m$	Filter weight vector for the mth receive antenna (size $F \times 1$)
$\mathbf{f}^{(q)}$	Filter weight vector for the qth transmit antenna (size $MF \times 1$)

F	Length of the equalizer per one receive antenna
$\mathbf{F}$	Filter convolution matrix for all M receive antennas
$\mathbf{F}_m$	Filter convolution matrix for mth antenna, $\mathbf{f}_m$
$\mathbf{g}^{(qr)}$	Impulse response from the r-th transmit antenna equalized using $\mathbf{f}^{(q)}$
g	Geometry factor
G	Smallest common spreading factor
G_k	Spreading factor of user k
γ	Quality measure of blind MPC method
$\mathbf{h}$	Channel impulse response (CIR) vector (size $LM \times 1$)
$\hat{\mathbf{h}}_b$	Blind estimate of CIR vector
$\hat{\mathbf{h}}_c$	Pilot based estimate of CIR vector
$\mathbf{h}_m$	CIR vector to the mth receive antenna (size $L \times 1$)
$\hat{\mathbf{h}}_{sb}$	Semi-blind estimate of CIR vector
$\mathbf{h}^{(q)}$	CIR vector from the q-th transmit antenna to all receive antennas
$\mathbf{H}$	Channel convolution matrix
$\mathbf{H}_m$	Channel convolution matrix for mth receive antenna
$\mathbf{H}^{(q)}$	Channel convolution matrix for qth transmit antenna
$\mathbf{H}_m^{(q)}$	Channel convolution matrix for qth transmit and mth receive antenna
$\bar{\mathbf{H}}$	Channel matrix for multiple codes ($\bar{\mathbf{H}} = \mathbf{I}_K \otimes \mathbf{h}$, size $MLK \times K$)
$\boldsymbol{\mathcal{H}}$	MIMO channel convolution matrix
$\dot{\boldsymbol{\mathcal{H}}}$	Conventional $M \times Q$ MIMO channel matrix (symbol level)
$\dot{\boldsymbol{\mathcal{H}}}_v$	Virtual MIMO channel matrix (size $ML \times Q$)
$\mathbf{I}_M$	$M \times M$ identity matrix
k	User index
K	Number of active codes
λ	Wavelength
L	Length of channel impulse response vector
m	Receive antenna index
$\mathbf{m}$	Pair of receive antennas m_i and m_j ($\mathbf{m} = \{m_i, m_j\}$)
M	Number of receive antennas
μ_C	Ratio between matrix traces for control signal matrices
μ_P	Ratio between matrix traces for HSDPA signal matrices
n	Symbol index
N	Number of symbols in one observation period
N_c	Number of chips in one observation period
$\boldsymbol{\varphi}_{\mathbf{m}}$	Fourier transform of correlation vector
$\Phi(\cdot)$	Channel estimator
p	(HSDPA) code index
P	Number of known HSDPA codes
q	Transmit antenna index
Q	Number of transmit antennas
$\mathbf{r}_{yd}$	Cross-correlation vector of the transmitted and desired signals
$\mathbf{R}$	Covariance matrix

$\mathbf{R_1}$	Spatial (SIMO) signal correlation matrix for 1st path (size $M \times M$)
$\mathbf{R_h}$	Channel auto-correlation matrix
$\mathbf{R}_{in}$	Interference and noise covariance matrix
$\mathbf{R}_{m_i,m_j}$	Cross-correlation matrix of signal received at antennas m_i and m_j
$\mathbf{R}_P$	Sum of post-despreading covariance matrix for P codes
$\mathbf{R_s}$	Desired signal covariance matrix
$\boldsymbol{\mathcal{R}}_{s_h}$	Spatial MIMO correlation matrix (size $Q \times Q$)
$\mathbf{R_{x_k}}$	Post despreading covariance matrix for kth code
$\mathbf{R_y}$	Sample covariance matrix
$\hat{\mathbf{R}}_\Delta$	Difference matrix between the post- and pre-despreading covariance matrices
$\hat{\mathbf{R}}_{\Delta_C}$	Difference matrix for control codes
$\hat{\mathbf{R}}_{\Delta_P}$	Difference matrix for P codes with same spreading factor
$\mathbf{R}_v$	Noise covariance matrix
σ_d	Chip power for the transmitted chip sequence $\mathbf{d}$
$\sigma^{(q)}$	Chip power for the qth transmit antenna
$\sigma^{(\tilde{q})}$	Chip power allocated to known codes at the qth transmit antenna
σ_v	Noise variance
ρ_k	Symbol power allocated to kth user (code)
ρ_c	Symbol power allocated to control (pilot) signal
s_{nk}	k-th users nth transmitted symbol
$s_{nk}^{(q)}$	k-th users nth transmitted symbol from qth antenna
$\mathbf{s}_n$	HSDPA transmitted symbol vector for nth symbol (size $PL \times 1$)
$\mathbf{s}$	Transmitted MIMO symbol vector (size $Q \times 1$)
$\mathbf{S}_N$	Matrix of transmitted HSDPA symbols (size $PL \times N$)
τ	Delay
$\mathcal{T}$	Toeplitz operator for constructing Toeplitz matrices
t	Time index for continuous time
T_c	Duration of one chip
$\mathbf{v}$	Noise vector
$\mathbf{x}_{nk}$	nth despread symbol vector for kth user (size $L \times 1$)
$\mathbf{x}_v$	Virtual despread symbol vector (size $ML \times 1$)
$\mathbf{X}_N$	Matrix of pre-processed HSDPA symbols (size $PL \times N$)
χ	Displacement rank
$\mathbf{y}_n$	Received signal vector for nth transmitted symbol
$\mathbf{y}$	Received signal vector for whole observation period
$\hat{\mathbf{y}}^{(q)}$	Interference free received signal vector qth antenna
$\check{\mathbf{y}}^{(q)}$	Estimate (with known codes) of the received signal vector from qth antenna
$\tilde{\mathbf{y}}_n$	ISI free part of the received signal vector
$\mathbf{z}^{(q)}$	Equalized chip vector for qth antenna

Chapter 1

Introduction

1.1 Motivation

In the past decades the cellular networks providing person-to-person wireless communications have been growing rapidly. Today both the number of users is increasing and new and better quality services are offered. Further increase in data rates, coverage and mobility are demanded. Since the spectrum is a scarce and expensive resource, higher spectral efficiency is needed to fulfil these requirements. The currently deployed third generation (3G) cellular systems are designed to enable multimedia communication. Recently first services proving data rates up to 1 Mbps have been introduced. These services are based on high speed downlink packet access (HSDPA) concept or so called 3.5G. The offered data rate is approximately three times compared to the previous stage 3G data rate which is 384 Kbps. However, these do not yet fulfil the original requirements. In order to provide reliable high quality image and video services in fast moving vehicles, further improvements in the system performance are required. Additionally, higher bit rates are also requested in the uplink. Consequently, evolution of 3G systems will continue even though a lot of research has already been carried out for future beyond third generation (B3G) and fourth generation (4G) wireless systems.

The wireless access in 3G systems is based on code division multiple access (CDMA) technique. It is a multiple access method developed from spread spectrum techniques originally developed for military to provide secure wireless links. The 3G standard deployed in Europe, as well in many other countries, is called Wideband CDMA (WCDMA). The first 3G services started in 2001 in Japan. Other countries followed slowly after this, for example first 3G service in Finland was opened in 2004. First HSDPA based services were opened in the fall of 2005, but the first mobile phones supporting HSDPA standard are only becoming available. Samsung recently announced to bring first HSDPA phone into Korean market in June 2006. Other vendors claim to follow later this year. Portable computer cards providing HSDPA connections have been available since fall 2005 from several suppliers. First laptops with included HSDPA support are introduced, even though are not yet available in stores.

The bit rates currently offered by the service providers are still far away from the 14.4 Mbps upper bound set by the different coding and modulation schemes, [HT04]. The current HSDPA terminals support only up to 1.8 Mbps connections due to technical challenges. The higher data rates require usage of higher order modulation, i.e. 16QAM, which needs more elaborate receivers. The main limitation for increasing the data rates beyond 14.4 Mbps is the available spectrum. As we all know, it is a scarce and expensive medium compared for example to copper. Consequently, efficient methods to exploit the available spectrum are needed. Besides higher order modulation, usage of multiple antennas both at the transmitter and receiver will be in future part of the high speed packet access (HSPA) systems. To benefit from these proposed enhancements, sophisticated signal processing methods as well accurate estimates of the time-varying channel parameters are needed.

The wireless channel is an unpredictable and difficult communications medium, [Gol05]. Furthermore, in mobile communications the transmitted signal experiences both temporal and spatial distortions due to the time-varying wireless channel. Additionally, in wideband system the channel is also frequency selective which introduces extra source of interference, called inter-symbol interference (ISI). On the other hand it also provides multipath diversity. When the channel impulse response (CIR) is known, it can be used at the receiver or transmitter to compensate the distortions due to wireless channel. Moreover, with advanced signal processing methods the channel impairments can be considered as different sources of diversity which provide robustness against fast fading. The quality of CIR estimate available at the receiver and transmitter has a strong impact on the system performance. In WCDMA downlink a code multiplexed pilot signal has been inserted to aid the estimation of channel parameters, [HT04]. However, the advanced receivers which are more complex than the conventional receiver, are also more sensitive to estimation errors. Furthermore, higher order modulation also increases the need for more accurate estimates.

Multiple input multiple output (MIMO) is a promising technology which can provide higher spectral efficiency and reliability, [PNG03]. Theoretically, the capacity can increase linearly proportional to $\min(Q, M)$, where Q and M are the number of transmit and receive antennas, [Tel99]. How well this is in practise achieved, depends on antenna placement and the wireless channel characteristics and the quality of available estimates. Transmitting independent data stream form multiple transmit antennas to increase the capacity is often called spatial multiplexing (SM). Alternatively, multiple transmit and receive antennas can be used to improve the quality of received signal, by adding diversity or by improving signal to noise ratio.

1.2 Scope of this thesis

The scope of this thesis is to develop advanced receivers structures and channel estimation methods for multi-code broadband (3G) CDMA systems with long spreading codes. The focus is on enhanced high speed downlink packet access (HSDPA) systems. Desired features of receivers designs are low complexity, fast convergence and robustness. Furthermore, channel estimation methods should be spectrally efficient and accurate. All these features can not be achieved simultaneously. There is always a trade-off between the different criteria when a new algorithm is designed.

The aim of this thesis is threefold. The first goal is to develop receiver algorithms which efficiently utilize the available information at the receiver. Both the channel estimation methods as well as the estimation of the filter coefficients should benefit from pilot signal and other system specific knowledge at the receiver. Consequently, faster adaptation to the changing wireless environment could be achieved. Alternatively, the amount of spectrum or power allocated to the pilot signal could be reduced.

The second goal is to reduce complexity of receiver structure for WCDMA downlink systems. This should be done without compromising the performance. Minimum mean square error (MMSE) equalization has been shown to be a promising technology to improve the system performance. However, finding the equalizing filter coefficients can be computationally demanding. Filter length is an other important feature influencing the receiver complexity. With efficient methods the performance of conventional receiver could be improved with only minor addition to the complexity.

The third goal of this thesis to develop receivers for multi-antenna transmission. Spatial multiplexing techniques can be used to provide higher data rates when multiple antennas are used both at the transmitter and receiver. When the channel is frequency selective, advanced signal processing techniques are needed to cope with the inter-antenna interference in HSDPA systems.

1.3 Contributions and organization of this thesis

Contributions of this thesis include introduction of novel blind and semi-blind channel estimation algorithms for multi-code CDMA systems. First, two preprocessing methods lowering the computational complexity are proposed for a blind subspace technique. These are shown to be more robust against code selection and interference than the original method. The technique is further extended to multi-antenna receiver case. Next, a principal component (PC) blind channel estimation method is extended to the multi-code scenario, and a quality measure of the channel estimate is proposed. The enhancement includes the case where the codes known at the receiver have different spreading factors. Furthermore, a semi-blind estimation method combining code multiplexed pilot signal based estimation with blind multi-code PC (MPC) channel estimation method is proposed. An extension of this semi-blind MPC (SB-MPC) channel estimation method to HSDPA MIMO systems is also introduced. Additionally, a low complexity method for finding the MPC channel estimate is introduced. Two novel semi-blind combining schemes are shown via simulations to outperform conventional combining of two estimates. An estimator for the combining weights is derived based on the traces of covariance matrices and other information already known at the receiver.

Next, low complexity frequency domain matrix inversion method is proposed for finding the filter coefficients for MMSE equalizer. This approximative method exploits the special circulant matrix structure of the signal covariance matrix. Furthermore, the performance of different estimation methods of the covariance matrix are compared in WCDMA downlink case. It is shown via simulations, that the channel auto-correlation based method outperforms the conventional sample covariance matrix based method. This holds especially for smaller sample supports. Furthermore, a novel method of finding sparse equalizer coefficients is derived. Proposed selection of tap placement is shown to provide an appealing trade-off between complexity and performance. Additionally, the influence of decoupled optimization of space-time (ST) MMSE equalizer coefficients for each antenna with different sample supports as well the influence of diagonal loading are studied.

Finally, a novel receiver structure for MIMO HSDPA systems is introduced. It combines ST-MMSE equalization and a chip level inter-antenna interference cancellation method. Furthermore, this method is enhanced by combining it with the proposed semi-blind channel estimation scheme.

This thesis consists of introductory part and seven original publications. The publications are listed at page vi, and appended at the end of this manuscript. The introductory part does not follow the chronological order of the original paper, because channel estimation has been studied in the first four publications, but also in the last publication. The organization of the introductory part is as follows. First, in chapter 2 the system model used through this thesis is given. Next, in chapter 3 receiver designs for single transmit antenna are reviewed and complexity reduction methods are proposed in section 3.5. In chapter 4 receivers for MIMO system are studied. A novel method for reducing the inter-antenna interference is introduced in section 4.4. Chapter 5 addresses the channel estimation in WCDMA systems. In section 5.2 blind channel estimation method for multi-code CDMA system with aperiodic spreading are introduced. Furthermore, in section 5.3 novel semi-blind channel estimation scheme, SB-MPC, for HSDPA systems is proposed. The extension of SB-MPC for MIMO systems is based in the MIMO receiver derived in section 4.4. Finally, section 6 summarizes the results and contributions of this thesis.

Table 1.1: Mapping between original publications and chapters of this thesis

Contributions introduced in this thesis	publication	chapter
Complexity reduction methods for equalization	V	3.5
A novel MIMO receiver design	VI and VII	4.4
Blind multi-code channel estimation methods	I-III	5.2
A novel semi-blind channel estimation method	IV and VII	5.3

In table 1.1 is given a mapping between the contributions of this thesis, the chapters of the introductory part and the original publications.

1.4 Summary of publications

The first three publications deal with blind channel estimation methods in multi-code CDMA systems with aperiodic spreading codes. The publications IV and VII address semi-blind channel estimation in HSDPA systems with multiple receive antennas. In publication V the issue of reducing complexity of finding the equalizer coefficients for MMSE equalizer is studied. Publication VI introduces a novel receiver algorithm for MIMO HSDPA systems, which is further enhanced with the semi-blind channel estimation method introduced in publication VII.

In Publication I blind channel estimation methods for multi-code CDMA systems with aperiodic spreading codes is addressed. A novel preprocessing method is introduced to lower the computational complexity and to improve the robustness of a subspace channel estimation method of [WF99]. Additionally, a single code principal component (PC) blind channel estimation method of [LZ97] is extend to multi-code systems.

In Publication II the blind subspace channel estimation method of Publication I is extended to multi-antenna receiver case. Simulation results are presented with different number of receive antennas.

In Publication III the blind multi-code channel estimation methods proposed in Publications I-II are studied in more detail. Additionally, an alternative preprocessing method is proposed to the blind subspace method. The influence of code selection, interference, number of receive antennas as well the antenna correlations are studied in extensive simulations in frequency selective channels. The novel modifications proposed to the subspace method are shown to more robust against code selection, interference and noise. However, all the blind subspace channel estimation methods are shown to be more sensitive to interference and noise than the multi-code PC (MPC).

In Publication IV the blind MPC method is further extended to the case when the known codes have unequal spreading factors. Additionally, a quality measure of the MPC method is developed. Furthermore, blind MPC method is combined with a code multiplexed pilot signal. As a consequence, a novel semi-blind (SB) channel estimation method, SB-MPC, is introduced with three different combining schemes. Estimates for the combining weights are derived from the traces of the different covariance matrices used in the MPC method.

In Publication V a low complexity matrix inversion method is proposed to reduce the complexity of finding space-time (ST) MMSE equalizer coefficients. A circulant approximation of the Toeplitz covariance matrix is used to enable a frequency domain matrix inversion. The computational gains are seen especially with long filter lengths. Additionally different covariance matrix estimates are compared using different sample supports. The performance of joint and decoupled spatial processing are also compared.

Publication VI introduces a novel chip level inter-antenna interference canceler for spatial multiplexed HSDPA MIMO systems. With spatial multiplexing the data rates of HSDPA systems can be further increased. However, the performance is degraded due to increased inter-antenna interference. The proposed interference cancellation scheme is shown via simulations to provide significant gains over space-time (ST) MMSE equalizer in HSDPA systems.

Publication VII is the main publication of this thesis. The semi-blind MPC (SB-MPC) channel estimation method proposed in Publication IV is further developed. First, the complexity of finding the channel estimate is lowered by using well known power method. The computational complexity is significantly reduced since the singular value decomposition (SVD) can be replaced by single matrix vector multiplication. Secondly, the MPC method is extended to MIMO systems. By using the MIMO receiver introduced in Publication VI the SB-MPC can be used in MIMO systems to improve the performance. Additionally, the performance of different covariance matrix estimates are evaluated with finite sample support. The minimum variance (MV) equalizer is compared to MMSE equalizer with estimated channel and covariance matrices.

The author of this thesis has derived all the algorithms proposed in this thesis, as well as planned and carried out all the simulations. The co-authors have provided guidance and constructive criticism. Additionally, major parts of the simulation software has been written by the author. The co-author Pekka Jänis has contributed a few functions to the simulation software used for publication V-VII. Most of the writing for the original papers of this thesis is done by the author. The co-authors have contributed to the final version of each paper.

Chapter 2

System model

In this chapter the system model used in this thesis is presented. The model is based on physical layer specifications given for Wideband CDMA 3GPP system, [gpp, HT04]. The model is not a full physical layer model, but includes the most important features for evaluating the performance of the receiver and channel estimation algorithms studied in this thesis. First, the model used for the wireless channel is introduced. Then, the used features of WCDMA system are presented. Mathematical models are separately given for single and multiple transmit antennas. Finally, key characteristics for HSDPA systems are presented.

2.1 Wireless channel

Designing of a cellular network where mobility of the users is a key issue, is a challenging task due the nature of wireless channel. The channel causes distortions such as fading, shadowing and self interference to the desired signal. On the other hand, it can also provide diversity due to multiple arriving copies of transmitted signal.

In wireless system signals propagate through a communication medium experiencing attenuation and reflections, diffractions and scattering from surrounding objects. Consequently, the received signal is a combination of multiple signals arriving with different phase shifts, delays and angles of arrivals. Moreover, in mobile systems temporal changes in channel characteristics are caused by movements of the transmitter, receiver or surrounding objects. The changes in the received signal power are typically categorised with mean propagation (path) loss, large and small scale fading. From the receiver design point of view the small scale fading is the most challenging. It refers to the rapid fluctuation of the received signal in time, frequency and space. Slower temporal fluctuation of signal components due to attenuation and shadowing caused by variations in both the terrain profile and the nature of surroundings is called slow fading or large scale fading.

In the following time, frequency and spatial selectivity are shortly described. Then the used mathematical notation is given. More detailed introduction of the wireless channel can be found for example in recent text books [PNG03, TV05] or more classical books [Rap96, Jak74, Pro95].

Temporal changes in signal powers are characterised by coherence time or by the Fourier transform of the time autocorrelation of the channel response, i.e. Doppler spectrum. Coherence time is generally defined as the time the channel coefficients have correlation larger than 0.7[1] [Rap96]. The rapid changes in the signal strength are due to multiple signal components arriving through different paths approximately at the same time instant. Since these multipath components are non resolvable, they can add up either constructively or destructively causing sudden changes in the signal strength. The channel is said to be in a fade if the signal power suddenly drops significantly. Consequently, fading can dramatically reduce the system performance. The mobile speed, the signal bandwidth and the scattering environment all influence the temporal fading. For example, with 85 km/h mobile speed the maximum Doppler spread at the WCDMA bandwidth is 150 Hz. This corresponds to coherence time of 0.67 ms which is duration of one WCDMA slot.

The signal is said to experience frequency selective fading if the delay spread between first and last arriving signal component is large compared to the inverse of the signal bandwidth. In this case the arriving multipath components are resolvable and cause signal distortion due to inter-path interference (IPI). On the other hand frequency selectivity provides diversity, since the separate multipath components are typically not in deep fade simultaneously. Consequently, by combining the different multipath components at the receiver the received signal power does not fluctuate as much as in flat channels, where no resolvable multipaths are available. In wideband systems, such as WCDMA, the channel is in general frequency selective. With the WCDMA chiprate of 3.84 Mcps, the chip duration is 0.26 μs, while the delay spread typically extends from 1 to 2 μs in urban and suburban areas.

Spatial selectivity needs to be taken into account when multiple antennas are used either at the transmitter or receiver. The coherence distance indicates how far apart two antenna elements need to be placed in order to gain from spatial diversity. It is inversely proportional to the angle spread, which depends on the angles of arriving signals at the receiver and on the departing angles at the transmitter. Due to varying scattering environments, the coherence distance can vary a lot. In [PNG03], it said to be between 0.25λ - 5λ[2] at the wireless terminals. For example, the often used rule of thumb to use $\lambda/2$ antenna spacing corresponds in WCDMA systems operating at roughly 2 GHz to a distance of 7-8 cm. Consequently, multiple antennas

[1] In some sources correlation of 0.5 is used.

[2] λ is the wavelength

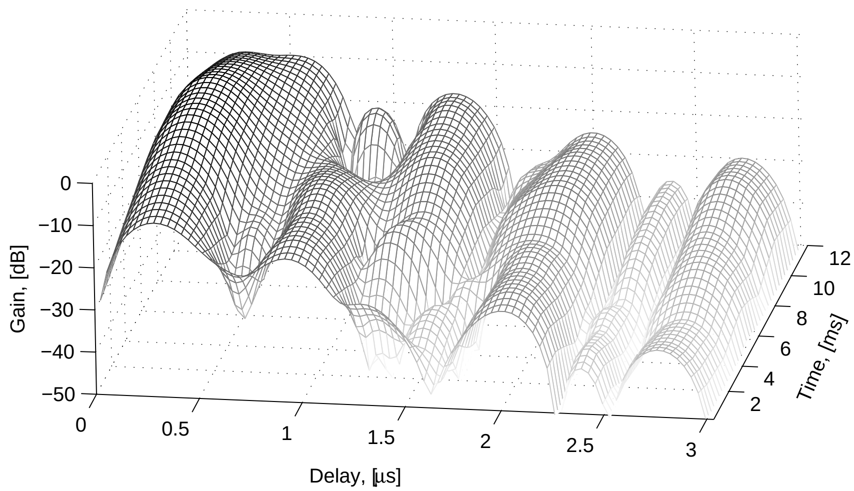

Figure 2.1: Example realization of time varying multipath channel, ITU Vehicular A with mobile speed 120 km/h

could be placed also in WCDMA terminals, if the cost to the users can be kept affordable.

2.1.1 Channel model

Modeling the unpredictable wireless communications medium is a difficult task and it has been intensively studied in the research community, see [Ric05, Cor06]. The channel parameters used in simulations for this thesis are generated with the IMETRA simulation software, [met]. It includes models for both receive and transmit antenna correlations which is the most difficult part of channel modeling. The used delay and power profiles are based on ITU channel profiles for Vehicular and Pedestrian cases, see [itu]. In Figure 2.1 is shown an example of time and frequency selective channel. Spatial channel model (SCM), [SCM03], is an alternative channel model developed with 3GPP for MIMO systems. Simulation softwares for SCM and it's extended version (SCME), [BJD+05], can be downloaded from [SDS+05].

The channel is typically modeled with impulse response function h which depends both on time t and delay τ. The continuous time low-pass channel impulse response is given as [Pro95]:

$$h(t;\tau) = \sum_{l} \alpha_l(t) e^{-j2\pi f_c \tau_l(t)} \delta(t - \tau_l(t)), \tag{2.1}$$

where $\alpha_l(t)$ is the attenuation factor for the lth multipath component at time t and the corresponding delay of the tap is $\tau_l(t)$. Carrier frequency is denoted with f_c and δ is the unit impulse function.

In this thesis, a discrete time vector model is used to combine the complex channel parameters for different delays to a vector. The impulse response is sampled once in a chip period and the length of the channel is denoted with L. The complex channel coefficients at the mth receive antenna are $\mathbf{h}_m = [h_m(1), \ \ldots \ , \ h_m(L)]^T$. In case of multi-antenna, receiver with M antennas the combined $LM \times 1$ channel vector can be written as:

$$\mathbf{h} = [\mathbf{h}_1^T, \ \ldots \ , \ \mathbf{h}_M^T]^T_. \tag{2.2}$$

The channel parameters are assumed to be constant during the observation period. This means that the sample support is less than the coherence time and the time index is dropped from this model.

The number of transmit antennas is denoted by Q. For each transmit antenna q, a channel vector $\mathbf{h}^{(q)}$ is defined as in equation (2.2). If $Q = 1$ and if there is only one receive antenna, the model is called single input, single output (SISO) model. With $M > 1$ antennas the channel is called single input multiple output (SIMO). Finally, if there are also multiple transmit antennas the channel is called multiple-input, multiple-output (MIMO). In this thesis multiple input, single output (MISO), i.e. $Q > 1$ and $M = 1$, channels have not been considered.

2.2 WCDMA system

The signal model considered in this thesis is based on the 3rd generation wideband CDMA downlink model, see [HT04, TS204] or documentation of 3rd generation partnership project (3GPP) [gpp]. The model is simplified containing only the basic features of WCDMA system. For example, channel coding, interleaving and rate matching are not considered. Consequently, the presented results throughout this thesis are for uncoded systems, i.e. so called raw bit error rates (BER) are considered. In Figure 2.2 is shown the system model used throughout this thesis. In the following, the basic WCDMA model with one transmit antenna and one or more receive antennas is given first. Then the usage of multiple transmit antennas is considered. Finally, high data rate specific features are reviewed.

In WCDMA downlink, the transmitted chip sequence $\mathbf{d}_n$ for one symbol period n can be presented as sum of K physical transport channels (referred as users or codes in this thesis) spread with different codes $\mathbf{c}_{nk}$ as:

$$\mathbf{d}_n = \sum_{k=1}^{K} \sqrt{\rho_k}\, \mathbf{c}_{nk} s_{nk}. \tag{2.3}$$

Here s_{nk} is the nth symbol for kth user and ρ_k is the power allocated to kth code. The transmitted symbols, s_{nk}, are assumed to be independent and identically distributed, such that $E\{s_{nk} s^*_{ap}\} = \delta(p-k)\delta(n-a)$, where $\delta(n)$

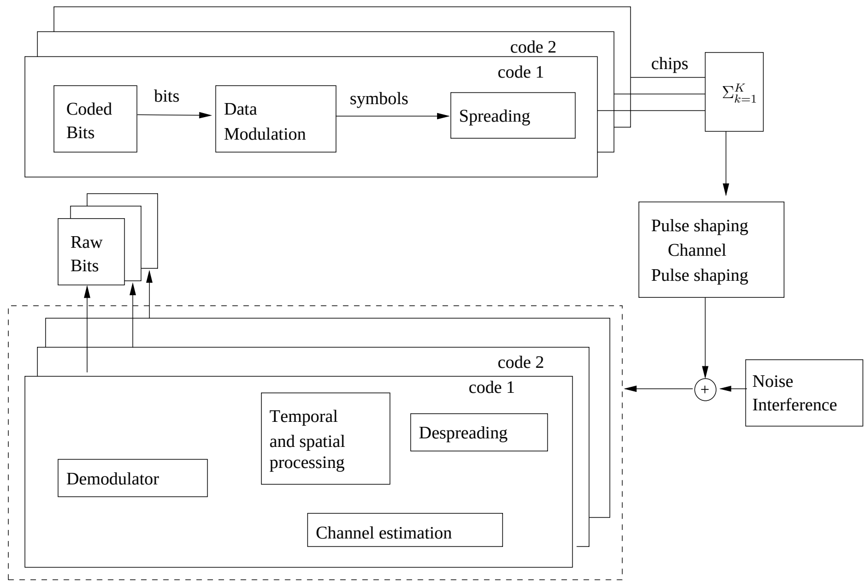

Figure 2.2: WCDMA downlink multi-code system model used in this thesis. This thesis addresses the signal processing techniques within the dashed box.

is the Dirac delta function. The codes $\mathbf{c}_{nk}$ are used to spread transmitted symbols into chips. They are composed of orthogonal spreading codes and complex valued long scrambling codes. The short spreading codes spread the symbols to common chiprate while the scrambling code is used to distinguish different base-stations (BS) and does not alter the chip rate. The same spreading code is used for all symbols of one user. The scrambling code is common to all users within one BS. Due to the long BS specific scrambling code, the used codes are aperiodic. This means that each symbol is spread with different code during one frame, which is the period for the long code. In WCDMA different spreading factors may be used for different users to alter the data rate. However, the model given in equation (2.3) can be used with the smallest G is used as a common spreading factor. The symbols spread with longer spreading codes can be considered as $N = G_p/G_k$ consecutive symbols with the common spreading factor. The kth code for nth symbol is denoted by vector $\mathbf{c}_{nk} = [c_{nk}(1), \ \ldots, \ c_{nk}(G_k)]^T$ where G_k is the spreading factor, i.e. the length of the code vector. Codes are scaled such that $\mathbf{c}_k^H \mathbf{c}_k = 1$. Additionally, due to orthogonality of the codes for $k \neq p$ and $G_p \leq G_k$, it holds that $\mathbf{c}_p^H \mathbf{c}_k((1 : G_p) + nG_p) = 0$, for all $n \in \{0, 1, \ldots G_k/G_p - 1\}$.

In ideal frequency flat channels there are no separable multipath signal components and the channel vector $\mathbf{h}$ diminishes into single complex value for each receive antenna. Consequently, inter-path interference (IPI) does not deteriorate the received signal. Additionally, due to the orthogonality of the codes there is no multiple access interference (MAI) due to own cell in the WCDMA downlink. However, the bandwidth for WCDMA is large enough to cause frequency selective fading and this will cause both IPI and MAI. In the following IPI is further divided into inter-symbol interference (ISI) and inter-chip interference, see figure 2.3.

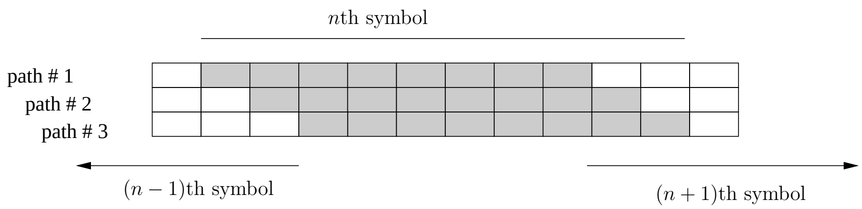

Figure 2.3: Inter-symbol and inter-chip interference in CDMA system in three tap channel with equally spaced tap delays.

The received signal is sampled at chip rate and stacked to a vector of length $M(G + L - 1)$. The received signal which is supercomposition of K multi-code signals transmitted from one base station may be written as:

$$\mathbf{y}_n = \mathbf{H}\boldsymbol{d}_n + \mathbf{v}_n \tag{2.4a}$$

$$= \sum_{k=1}^{K} \sqrt{\rho_k}\, \boldsymbol{\mathcal{C}}_{nk}\mathbf{h}s_{nk} + \mathrm{ISI} + \mathbf{v}_n \tag{2.4b}$$

where $\boldsymbol{d}_n = [\mathbf{d}_{n-1}((G - L + 2) : G)^T,\ \mathbf{d}_n^T,\ \mathbf{d}_{n+1}(1 : (L - 1))^T]^T$ is the transmitted chip sequence. It is influenced by the chip sequences of previous and following symbols, i.e. ISI is taken into account. The noise term is denoted with $\mathbf{v}_n$ and $\mathbf{H}$ is the channel convolution matrix. It is defined as $\mathbf{H} = [\mathbf{H}_1^T, \ \dots\ , \mathbf{H}_M^T]^T$. In this case the channel convolution matrix for one receive antenna is of dimension $G + L - 1 \times G + 2(L - 1)$ so that it may be written for the mth antenna as follows:

$$\mathbf{H}_m = \begin{bmatrix} h_m(L) & \dots & h_m(1) & & \dots\dots & 0 \\ \vdots & \ddots & \ddots & \dots & \ddots & \vdots \\ 0 & \dots & \dots & h_m(L) & \dots & h_m(1) \end{bmatrix}. \tag{2.5}$$

In the alternative formulation of the received signal given in equation (2.4b), the order of channel and code has been changed. The code convolution matrix for M receive antennas is $\boldsymbol{\mathcal{C}}_{nk} = \mathbf{I}_M \otimes \mathbf{C}_{nk}$, where $\otimes$ denotes Kronecker product and

$$\mathbf{C}_{nk} = \begin{bmatrix} c_{nk}(1) & 0 & \dots & 0 \\ c_{nk}(2) & c_{nk}(1) & \dots & 0 \\ \vdots & \vdots & \vdots & \vdots \\ c_{nk}(L) & c_{nk}(L-1) & \dots & c_{nk}(1) \\ \vdots & \vdots & \vdots & \vdots \\ c_{nk}(G) & c_{nk}(G-1) & \dots & c_{nk}(G-L+1) \\ 0 & c_{nk}(G) & \dots & c_{nk}(G-L+2) \\ \vdots & \vdots & \vdots & \vdots \\ 0 & 0 & \dots & c_{nk}(G) \end{bmatrix}, \tag{2.6}$$

is the code convolution matrix for one antenna. More compact definition for this matrix is $\mathbf{C}_{nk} = \mathcal{T}\left([\mathbf{c}_{nk}^T,\ \mathbf{0}_{1,L-1}]^T, [c_{nk}(1),\ \ \mathbf{0}_{1,L-1}]\right)$, where $\mathcal{T}$ is an operator for constructing of Toeplitz matrices[3]. This Toeplitz matrix has dimension $G+L-1 \times L$. The ISI term now includes the influence of symbols transmitted before and after the nth symbol:

$$\mathrm{ISI} = \sum_{k=1}^{K} \sqrt{\rho_k}\ [\boldsymbol{\mathcal{C}}^b_{(n-1)k}\mathbf{h}s_{(n-1)k} + \boldsymbol{\mathcal{C}}^f_{(n+1)k}\mathbf{h}s_{(n+1)k}], \tag{2.7}$$

where the partial kth code matrix for the $(n-1)$th symbol is:

$$\boldsymbol{\mathcal{C}}^b_{(n-1)k} = \mathbf{I}_M \otimes \begin{bmatrix} (G_k+1)\text{th row of} \quad \mathbf{C}_{(n-1)k} \\ \vdots \\ (G_k+L-1)\text{th row of} \quad \mathbf{C}_{(n-1)k} \\ \mathbf{0}_{G_k \times L} \end{bmatrix}. \tag{2.8}$$

Correspondingly, the contribution of the following symbol is obtained with a partial code matrix $\boldsymbol{\mathcal{C}}^f_{(n+1)k} = \mathbf{I}_M \otimes \mathbf{C}^f_{(n+1)k}$, where $\mathbf{C}^f_{(n+1)k}$ is comprised of G_k rows of zeros followed by $(L-1)$ first rows from the code matrix $\mathbf{C}_{(n+1)k}$, see equation (2.6).

Root raised cosine (RRC) filters with roll off 0.22 are used both at the transmitter and receiver for pulse shaping. These are included to the channel impulse response (CIR) prior sampling at chip rate. This compound channel can written as:

$$\tilde{h}(\tau) = f_p(\tau) \star h(\tau) \star f_p(-\tau), \tag{2.9}$$

where f_p is used to denote the pulse shaping filter and $\star$ is the convolution operator. This CIR $\tilde{h}(\tau)$ is further used in order to form the combined channel vector $\mathbf{h}$ given in equation (2.2).

[3]As defined in Matlab: toeplitz($[\mathbf{c}_{nk}^T,\ \mathbf{0}_{1,L-1}]^T, [c_{nk}(1),\ \ \mathbf{0}_{1,L-1}]$)

At the receiver, signal processing is applied either directly to the chip level received signal, $\mathbf{y}$, or to symbol level despread signal $\mathbf{x}$. It is defined for kth code and nth symbol as:

$$\mathbf{x}_{nk} = \boldsymbol{\mathcal{C}}_{nk}^{H}\mathbf{y}_{n}. \tag{2.10}$$

The correct delay of the received signal vector $\mathbf{y}_n$ and length of the CIR are assumed known.

In simulations, in order to facilitate channel estimation the WCDMA downlink common pilot channel (CPICH) is implemented with spreading factor $G_c = 256$ and spreading code of all ones and QPSK modulation, [HT04]. Other signalling channels are not implemented. Speech and data traffic are simulated with different number of codes and with spreading factors, signal powers and QPSK or 16QAM data modulation.

2.2.1 Multi-antenna transmission

Using multiple transmit antennas is an emerging technology which can be used to provide higher spectral efficiency and reliability. Traditional antenna array techniques are based on the assumption that there are multiple antennas only at one end of the transmission link, either at the transmitter or receiver. Using multiple antennas can offer several different type of gains. These gains are categorised as follows:

- Array gain - improves the signal to noise ratio (SNR)
- Multiplexing gain - provides increase in capacity
- Diversity gain - combats fast fading
- Interference rejection - mitigates the influence of interfering users

See e.g. [PNG03] for further details. Out of these multiplexing gain is the only one which requires usage of multiple antennas at both ends. Fundamental research on multiple input multiple output (MIMO) systems has been carried out at Bell Labs, see e.g. [Fos96]. In the past decades, MIMO systems have been very active research topic. Reviews on MIMO techniques can be found in [PNG03, GSS+03, TV05].

Multiple transmit antenna methods are typically categorised to diversity and spatial multiplexing (SM) techniques. Furthermore, the techniques can be divided depending on the availability of the channel state information (CSI) at the transmitter. In this thesis only the case where no knowledge of CSI at the transmitter is considered. Diversity methods are used to improve the quality of the received signal, whereas the aim of SM is to increase the link capacity. The capacity increase in MIMO systems was shown to grow linearly with $\min(M, Q)$ [Tel99], where Q is the number of transmit antennas and M is the number of receive antennas. Recently also trade-offs between SM and diversity transmission schemes have been studied in [ZT03, TV05]. Combining these techniques seems to be an viable alternative, since in practise, due to dimension and cost issues, base station could utilize more antennas than mobile terminals. Consequently, improvements both in rate and robustness of the link could be achieved.

MIMO techniques is one of the main enhancement considered for HSDPA systems, [HT04, TS805b]. MIMO can be used to increase either the bit rate or the robustness of the link. However, so far how to benefit from MIMO transmission has not been specified. In this thesis the focus is on spatial multiplexing (SM) techniques. This transmitter architecture is commonly called V-BLAST,[4] see e.g. [TV05, Ch. 8].

SM is used to increase the transmitted data rate by splitting the data stream across multiple transmit antennas. No increase in bandwidth or total transmission power are needed since multiple data pipes are created between the transmitter and the receiver. Under favourable channel conditions, using the CSI and advanced signal processing methods these pipes are separable. However, the optimal method is sensitive channel estimation errors and interference and has high computational complexity. In practise the drawback of SM is the increase in interference at the receiver due to inter-antenna-interference (IAI).

The mathematical model for SM in a WCDMA system is a straightforward extension of equation (2.4). The only addition is the summation over the transmit antennas:

$$\mathbf{y}_n = \sum_{q=1}^{Q} \mathbf{H}^{(q)} \boldsymbol{d}_n^{(q)} + \mathbf{v}_n = \boldsymbol{\mathcal{H}} \mathbf{b}_n + \mathbf{v}_n, \tag{2.11a}$$

$$= \sum_{q=1}^{Q} \sum_{k=1}^{K} \sqrt{\rho_k^{(q)}}\, \boldsymbol{\mathcal{C}}_{nk}^{(q)} \mathbf{h}^{(q)} s_{nk}^{(q)} + \mathrm{ISI} + \mathbf{v}_n \tag{2.11b}$$

where the $\mathbf{b}_n = [\boldsymbol{d}_n^{(1)T}, \ \ldots \ , \ \boldsymbol{d}_n^{(Q)T}]^T$ is the transmitted chip sequences for all transmit antennas. In order to empahasize IAI in these equations, the summations over Q transmit antennas could be divided to desired signal, e.g. $q = 1$, and due to IAI $q = 2, \ \ldots, \ Q$. The MIMO channel convolution matrix defined as:

$$\boldsymbol{\mathcal{H}} = \begin{bmatrix} \mathbf{H}_1^{(1)} & \ldots & \mathbf{H}_1^{(Q)} \\ \vdots & \ddots & \vdots \\ \mathbf{H}_M^{(1)} & \ldots & \mathbf{H}_M^{(Q)} \end{bmatrix}, \tag{2.12}$$

where each submatrix $\mathbf{H}_m^{(q)}$ is defined as in equation (2.5). Dimension of $\boldsymbol{\mathcal{H}}$ is now $(M * (G + L - 1)) \times (Q * (G + 2(L - 1)))$.

[4] Vertical Bell Labs Space Time Architecture

Typically the same set of orthogonal spreading codes are used across the antennas, i.e. the transmit antenna dependency can be neglected from the code convolution matrix, $\boldsymbol{\mathcal{C}}_{nk}^{(q)}$. For the second antenna, the control signal used for channel reference, i.e. CPICH, differs from the first antenna only by pilot pattern, [TS202].

Modeling inter-cell interference

Since the same frequency is used in all the BS, the model given for multi-antenna transmission in equation (2.11) can also be used to model asynchronous inter-cell interference from neighbouring cells. The channel parameters as well as the codes need to be independently designed for the interfering BS. A commonly used parameter to measure interference is the geometry factor, $g = I_d/I_i$, [HT04]. Here I_d is the power from the desired BS and I_i is the power from interfering BS.

2.2.2 HSDPA system

Higher data rates in WCDMA systems may be obtained with high speed downlink packet access (HSDPA) concept, [hsd01, HT04, HT06]. It is the key new feature included in the Release'5 of WCDMA specifications compared to Release'99. This development stage is often referred to as 3.5G. It is backward compatible with Release'99 WCDMA systems. The highest supported data rate in HSDPA system is 14 Mbps while 3G provides at most 2 Mbps.

The major differences in HSDPA compared to other packet access methods in WCDMA system are: fixing the spreading factor, disabling fast power control, introducing adaptive modulation and coding (AMC) and moving scheduling and retransmission decisions to base station. Since the decisions in HSDPA are made closer to mobile, the system can faster adapt to rapid term variations of the channel conditions. Consequently, most of the cell capacity can be allocated to one user for a very short time to obtain very high data rates when the channel conditions are favourable.

In Release'5, a new transport channel carrying the user data with HSDPA operation and two control channels have been added. The scheduling and retransmissions are based on measurements signalled from mobile to BS via the uplink high speed dedicated physical control channel (HS-DPCCH). It contains acknowledgement information of the previous packet as well a channel quality indicator (CQI). The new downlink control channel, high-speed shared control channel (HS-SCCH) carries information needed at the mobile to demodulate the data. Since in this thesis the focus is on link layer signal processing algorithms, only the features of the high-speed downlink shared channel (HS-DSCH) are considered in the following. It is the new transport channel introduced for data in Release'5. For further features of HSDPA, see for example [HT04, HT06].

The data channel, HS-DSCH, has a fixed spreading factor of 16, whereas other channels can have spreading factor varying from 4 to 512. Faster retransmission decisions are possible, since the transmission time interval (TTI) or interleaving period is set to three slots, i.e. 2ms. In 3G the interleaving is between 10-80ms and two separate interleavers (intra-frame and inter-frame) interleavers are used. Additional changes are allowing only turbo coding in channel coding and adding a new Hybrid ARQ (HARQ) functionality.

Higher data rates are obtained with multiple spreading codes allocated to one user or by reducing the encoding redundancy. Maximum number of parallel HS-DSCH, i.e. possible number of codes, is 5, 10 or 15 depending on the terminal capability. Also a higher order modulation scheme, 16QAM, can be used instead of QPSK to increase the instantaneous peak data rate. With 16QAM, estimation of amplitudes is needed to separate the constellation points. Additionally, since constellation points are closer to each other, sensitivity to estimation errors such as channel parameters is increased.

Further enhancements for HSDPA system include usage of multiple transmit antennas, [TS805b, LH05]. With HSDPA MIMO systems aim to support data transmission rates up to 20 Mbps. Both transmit diversity and spatial multiplexing methods are actively studied. However, in [HT04] it is stated that there is a conflict between fast scheduling and the benefits from transmit diversity in HSDPA. Also a high data rate system for the uplink is planned in the next phase of HSDPA, [TS306, TS805a, HT06]. High speed uplink packet access (HSUPA) aims to provide significant enhancements in terms of user experienced throughput and delay. The term high speed packet access (HSPA) is used in general case of both uplink and downlink.

Chapter 3

Advanced receiver structures

In order to mitigate channel dispersions and interference due to other users, sophisticated signal processing techniques, such as equalization, diversity combining, channel coding and interference cancellation can be used at the receiver. Since for different wireless systems the signal and interference structures can be quite different, the transceivers need to designed separately for each system. In this thesis HSDPA systems are studied. However, the results presented in this chapter are applicable to WCDMA downlink in general.

The focus is on equalization and diversity combining, but also interference cancellation and multiuser detection techniques are shortly reviewed. Channel coding and interleaving are beyond the scope of this thesis. The goal of equalization, in a broad sense, is to remove ISI. However, it simultaneously also provides diversity, [Rap96]. Diversity techniques are used to combat effects of fading channel. They exploit the random nature of the radio propagation by combining multiple copies of signal. If multiple receive (or transmit) antennas are assumed, so called space-time (ST) processing can be used to combine spatially and temporally separated signals. Alternatively the term smart antenna techniques [GSS+03] is also used.

In this chapter, receivers for single transmit antenna, i.e. both SISO and SIMO systems, are studied. First, linear receivers and related optimality criteria are reviewed. Then methods how to estimate the filter coefficients and their performance and computational complexity are discussed. Next, in section 3.5 novel methods for lowering the computational complexity of equalizers are proposed. This section is based on original Publication V and Publication VII. Finally, non-linear receiver structures are shortly reviewed.

3.1 Overview of SISO/SIMO receivers

There are alternative ways to classify different receivers. In Figure 3.1 a general classification of receiver for WCDMA SIMO case is shown. Estimation methods in general can also be classified based on what is known about the transmitted data at the receiver. This leads to three classes: pilot aided, blind, and semi-blind methods. In WCDMA downlink, a strong pilot signal is included to aid channel estimation or filter update. Consequently, only pilot aided methods are extensively studied in the literature. A few examples of semi-blind methods, where some additional information is used to update filter coefficients are available, see e.g. [LL99b, PLE^{+}01, Cho03].

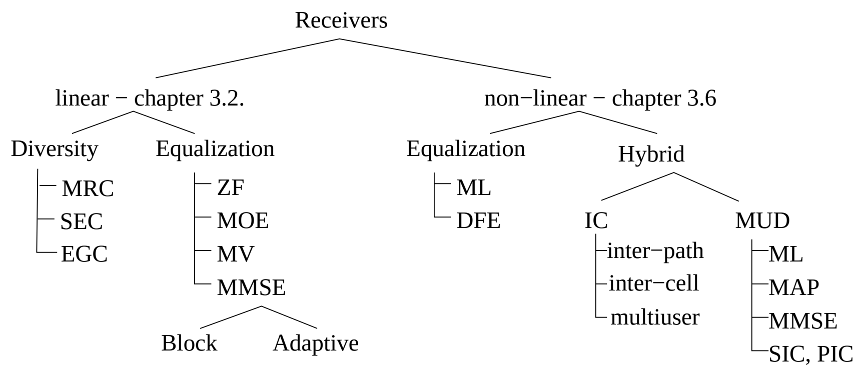

Figure 3.1: Categorisation of receivers for SISO and SIMO systems

Linear receivers, both diversity combiners and equalizers, are widely studied in the literature, see e.g. [Pro95, Rap96, Gol05]. The main part of this chapter deals with minimum mean square error (MMSE) equalizer, while the other linear receivers are only briefly discussed in section 3.2. In case the channel distortions or the interference is severe, non-linear receiver could be desirable. However, they are computationally more demanding than linear receiver. Non-linear receivers are here divided into equalization techniques and hybrid receiver. Hybrid receivers are combinations of two receivers, typically a linear receiver with either some interference cancellation (IC) technique or with a multiuser detection (MUD) method. A literature review of these techniques will be given at the end of this chapter. For details of the well known Maximum likelihood (ML) and decision feedback equalization (DFE) see any basic textbook on wireless communications, for example [Pro95, Rap96, Gol05].

A third type of non-linear receivers which are not shown in the Figure 3.1 is called turbo-receivers. These receivers combine channel decoding and for example equalization or multiuser detection in a iterative manner. Since channel coding is beyond the scope of this thesis, this receiver class is not further discussed. A recent review of turbo equalization is found in [KST04], or alternatively [Gol05]. For further references on turbo-multiuser detection see [WP98, YW02].

Analogy between array processing and spread spectrum systems has been used in various cases, see e.g. [GHST01, GSS^{+}03] and the reference there-after. Consequently, techniques extensively studied in the array processing community, such as minimum variance distortionless response (MVDR) beamformer, can be easily extended to CDMA context.

3.2 Linear receivers in WCDMA systems

In this chapter most used linear receivers, such as RAKE and MMSE equalizer, are briefly presented. Additionally some observations related to their implementation are given.

Linear filtering can be expressed as convolution

$$\mathbf{z} = \mathbf{F}^H \mathbf{y}, \tag{3.1}$$

where $\mathbf{F} = [\mathbf{F}_1; \ldots; \mathbf{F}_M]$ denotes filter convolution matrix which is defined for mth receive antennas as $\mathbf{F}_m = \mathcal{T}\left([\mathbf{f}_m^T,\ \mathbf{0}]^T, [\mathbf{f}_m(1),\ \ \mathbf{0}]\right)$. Here $\mathbf{f}_m$ is the filter coefficient vector for the mth antenna. The received signal vector $\mathbf{y}$ is defined in equation (2.4). Due to linearity of spreading operation linear receivers in WCDMA systems can be applied either at chip or symbol level, i.e. before or after despreading. The symbol level despread signal was defined in equation (2.10).

Optimization of linear filter coefficients, $\mathbf{f}$, are based on different assumptions on the received signal, interference and noise statistics. Unlike in other multiple access methods, in CDMA systems the filter coefficient optimization can be done either at chip or symbol level. Since the chip and symbol level signal statistics can be different, also the filters may differ.

In WCDMA systems the channel is typically frequency selective, i.e. even with single receive antenna multiple temporally separated copies of the signal are received. Furthermore, with multiple receive antennas there can also be spatially separated signal copies. Consequently, there are two dimension for the filter coefficients which can be optimized either separately or jointly.

3.2.1 Diversity receiver - decoupled temporal optimization

Typically spread spectrum systems are designed such that the codes have low correlation between successive chips. Consequently, after despreading the signals arriving via separable multipaths appear as uncorrelated noise at a spread spectrum receiver. This assumption simplifies the receiver design, since the different multipath diversity branches can be thought as independent and the filter coefficients can be separately optimized for each path. A conventional RAKE receiver, see [Pro95, Rap96], uses maximal ratio combining (MRC) to combine the different multipath components and to achieve maximal signal to noise ratio (SNR). Other combining schemes are equal gain combining (EGC) or selective combining (SEC), see e.g. [Pro95, Gol05].

The MRC filter coefficients for RAKE are related to channel coefficients $\mathbf{f}_{m_{\mathrm{RAKE}}} = [h_m^*(L) \ \ldots \ h_m^*(1)]$, i.e. it coherently combines the different signal components. The filter length is equal to the channel length, i.e. $F = L$.

The assumption of temporal whiteness (or negligible correlation) of the multiple replicas of received signal relies on auto-correlation properties of the spreading codes. The poor auto-correlation properties of used orthogonal spreading codes in WCDMA downlink are significantly improved by using a long scrambling code. However, due to usage of multiple spreading codes, also the cross-correlation properties will influence the interference due to frequency selectivity of the channel. If most of the available orthogonal spreading codes are used simultaneously, the intra-cell interference differs from conventional spread spectrum signal. This holds especially for HSDPA signals, which have low spreading factor [HL98].

It is well known that RAKE receiver does not work in other systems as well as in spread spectrum systems, see e.g. [Rap96]. There have been several proposals for improving the performance of RAKE receiver in WCDMA downlink. For example, in [Ket01] an enhanced MRC scheme which takes into account the different noise power in each finger has been proposed. For further improvements, joint temporal optimization of the RAKE finger combining coefficients has been proposed in [BOW00, LS01, TW02]. These schemes fall into to broad category of equalizers which will be discussed next.

3.2.2 Linear equalization - joint temporal optimization

In communication systems, linear filters which are designed to mitigate ISI by jointly optimizing the temporal filter coefficient are denoted as linear equalizer. These well known receiver have been extensively studied, see e.g. [Rap96, Pro95] and references therein. Compared to diversity receivers, such as RAKE, the difference is that the correlation of the received signal copies is taken into account by jointly optimizing the temporal filter coefficients.

Zero forcing (ZF) equalizer is one option to compensate the channel distortion. It inverts the channel and maximizes the signal to interference ratio (SIR). Consequently, in WCDMA downlink it restores the orthogonality of the codes used and removes both inter-path interference and intra-cell interference. However, it is well known that in practical cases it suffers from noise enhancement as well sensitivity to channel parameter errors.

Minimum mean square error (MMSE) equalizer minimizes the mean square error between the desired signal and the equalized signal, see [Pro95]. For the chip level received signal given in (2.4) the MMSE cost function is:

$$J_{\mathbf{f}} = E[|\mathbf{f}^H \mathbf{y} - d|^2], \tag{3.2}$$

where $\mathbf{f}$ is the equalizer, $\mathbf{y}$ is the received signal and d is the transmitted chip. The transmitted WCDMA downlink chip sequence $\mathbf{d}$ is close to i.i.d.

Consequently, if the channel is modeled wide sense stationary (WSS) the received signal is also WSS and the chip level equalizer does not depend on time. A causal finite impulse response equalizer that minimizes (3.2) is found by solving the *Wiener-Hopf* equation [Wie49]

$$\mathbf{R}\mathbf{f} = \mathbf{r}_{yd}, \tag{3.3}$$

where $\mathbf{R}$ is the covariance matrix of the received signal and $\mathbf{r}_{yd}$ is the cross-correlation vector of the transmitted and desired signals, i.e. the channel.

MMSE equalization balances between noise amplification and channel inversion. It can be seen as a good compromise between the SNR maximizing RAKE receiver and the SIR maximizing ZF receiver. The use of the MMSE equalizer in CDMA downlink was proposed in [Kle97, LZ97, GS98, FV98] and has been studied extensively, see e.g. [KHL00, KHZ02, FVM02, HJH$^+$02]. It has been shown to clearly outperform RAKE receiver in WCDMA systems.

In WCDMA system also a symbol level equalizer has been considered. In this case the optimization problem, given in equation (3.2), is time dependent, since after despreading the temporal structure of the interference + noise term is time-varying. However, when multiple spreading codes are used, the time dependency is reduced due to code averaging. This type of equalizer has been studied e.g. in [KHZ02, Hoo03].

An alternative symbol level equalizer for long code CDMA systems is based on minimizing the output energy (MOE) is proposed in [LL99b]. Assuming the unused spreading codes are known at the receiver, the signal energy spread with the unused codes is minimized. A constrain is set such that the equalized and despread desired signal is one. This blind equalization methods which is closely related to MVDR beamformer is not further considered here, for two reasons. First, in HSDPA system typically multiple spreading codes need to be despread. Secondly, the assumption of knowing which codes are not used, may not be realistic.

Equalization with 16QAM symbol modulation

In case constant modulus modulations e.g. QPSK are used, the symbol decisions can be made based on the phase of the soft symbol estimates, i.e. without knowledge (or estimate) of symbol amplitude. This is, however, not the case with 16QAM modulation where the amplitude information is needed in the decisions. Therefore, it is important that the cascade of channel and equalizer does not alter the amplitude of the symbols. This can be achieved by minimizing the MSE under constraint that the equalized impulse response is unity, i.e $\mathbf{f}^H\mathbf{h} = 1$, yielding:

$$\mathbf{f} = \frac{\mathbf{R}^{-1}\mathbf{h}}{\mathbf{h}^H\mathbf{R}^{-1}\mathbf{h}}, \tag{3.4}$$

see [GHST01, VT02]. Equation (3.4) is in thesis called minimum variance (MV) equalizer, and it is well know to obtain the same SINR as the MMSE equalizer. This receiver, denoted as Miminum Variance Distortionless response (MVDR) has been considered in [ZBM03] for WCDMA downlink. However, no clear comparisons to MMSE equalizer were given. Alternatively, the receiver could be called as linearly constrained MOE, as was done in [LS00], where the output energy was minimized at symbol level. A comparison of the raw BER performance obtained using MMSE and MV equalizers is given for a HSDPA system in Publication VII.

3.3 Finding the MMSE filter coefficients

The MMSE equalizer can be found by solving the Wiener-Hopf equation, (3.3):

$$\mathbf{f} = \hat{\mathbf{R}}^{-1}\sqrt{\sigma_d}\,\hat{\mathbf{h}}, \tag{3.5}$$

where $\hat{\mathbf{R}}$ denotes the covariance matrix estimate, $\sqrt{\sigma_d}\,\hat{\mathbf{h}}$ estimates the cross-correlation vector $\mathbf{r}_{yd}$ and σ_d is the transmitted signal power (chip level). The size of the matrix $\hat{\mathbf{R}}$ is $F \times F$, where F is the filter length. When $F > L$, as typically is, $F - L$ zeros are added to the channel vector $\hat{\mathbf{h}}$. Method for estimating the filter coefficients can be divided to two class those that involve sample matrix inversion (SMI) and those that use adaptive methods to avoid complex matrix inversion.

Adaptive methods have been extensively studied in case of WCDMA downlink. Less attention has been paid to SMI methods and how to different covariance estimation techniques influence the performance. In the following, different adaptive algorithms proposed for WCDMA downlink are briefly reviewed. Then different methods for estimating the sample covariance matrix are summarized and evaluated.

3.3.1 Adaptive implementations of MMSE

Adaptive implementation of MMSE equalizer can be obtained by either directly adapting the equalizer weights or by updating the weights based on estimates of the time varying channel coefficients. Methods performing adaptation at both chip level and symbol level have been proposed. Chip level adaptation suffers from low SNR and high computational complexity. Therefore, symbol rate adaptation is more appealing, see e.g. [Mai03, Hoo03]. A drawback in symbol level adaptation is that the despread signal covariance matrix has an increased eigenvalue spread [Hoo03], which deteriorates the performance of an adaptive equalizer [Hay96].

The well known normalized least mean squares (NLMS) and recursive least squares (RLS) based update rules have been considered at both chip and symbol level e.g. in [PME$^+$00, FVM02, PUZF02, Hoo03, Mai03]. The adaptation using LMS exhibits poor performance in time-varying and noisy channels due to excess MSE, [Pro95]. The improvement in the performance using RLS compared to NLMS is obtained with the cost of increased computational complexity. However, for example in [PUZF02] the sample matrix inversion based equalizer updated once in a slot has been shown to outperform RLS-type adaptive equalizer at 120 km/h mobile speeds.

An other method that utilizes LMS adaptation is the Griffiths' algorithm [HKL99, KHL00, PUZF02], which utilizes the channel estimate in the calculation of stochastic gradient vector. The complexity is lower than that of RLS, but performance is worse, especially in fast fading channels.

Several reduced-rank methods have been proposed for solving the equation (3.5) because of its high dimensionality. The multi-stage nested Wiener filter (MSWF), proposed by Goldstein and Reed in [GRS98], seeks the best reduced rank representation of the useful portion of the data by projecting it onto a reduced rank subspace based on the covariance matrix $\mathbf{R}$ and the cross-correlation $\mathbf{r}_{yd}$. MSWF has been proposed for equalization in CDMA downlink in [CZG00b, CZG00a, CZ02, MAMIL03a]. Conjugate gradient (CG) method, which is similar to MSWF, has been applied to CDMA downlink in [CZ01a, MAMIL03b, CZ01b] as a pilot symbol trained block based equalizer. Symbol level pilot trained MSWF and CG algorithms were both shown to have a relatively slow convergence but nevertheless substantially faster than a full-rank RLS.

Another adaptive method to find the filter coefficients is to adaptively update the inverse of the covariance matrix, similar to RLS algorithm, [WL99, WD03]. This method is based on matrix inversion lemma, see [Gv96, Hay96]. The performance of the this method is shown to be worse than that of the channel estimate based MMSE in [WD03] due to a bias caused by the correlation matrix inversion.

An adaptive algorithms based on constrained minimum output energy receiver have been studied in the context of blind multiuser detection in CDMA systems, [XT98, XT01] and [GHST01, Ch. 5]. The same idea is applied to adaptive equalization WCDMA downlink in [HJH$^+$02, Hoo03]. This leads to a structure closely related to generalized sideslobe canceler (GSC) that is well known from antenna array signal processing. The performance has been shown to be worse than SMI based implementation of MMSE equalizer.

Typically, adaptation using pure pilot signal based training are considered. However, semi-blind adaptation has been proposed in [PLE$^+$01] for code multiplexed pilots and for time multiplexed pilot signals in [PLD$^+$01]. The least squares (LS) minimization problem to find the optimum filter coefficients is reformulated using the known spreading codes and the pilot signal. This is done by projecting the estimate of the equalized chip sequence onto the orthogonal complement of the space spanned by the known code sequences. Alternatively, the unused spreading codes are used in [Cho03, SG00, LL99b] to improve the estimation of the MMSE equalizer.

In general, the adaptive methods have slow convergence and the sample matrix inversion methods provide better performance, see e.g. in [PUZF02, Hoo03].

3.3.2 Block approach - Sample matrix inversion (SMI)

In block based approaches for finding the MMSE filter coefficients, an estimate of the covariance matrix, $\hat{\mathbf{R}}$, is formed from a block a data. In this section, the estimation of $\hat{\mathbf{R}}$ is considered first. Different methods studied in the literature are presented here with a common notation. Then the performance of different SMI methods with finite sample support are compared via simulations.

Estimation of covariance matrix

Covariance matrix plays a key role in analysis and design of the discrete time filters, [Hay96]. Several different estimation methods have been proposed for WCDMA downlink systems, see e.g. [BOW00, FVM02, KHZ02, HJH$^+$02]. The performance obtained using different covariance matrix estimates depends on the sample support as well the system load and interference. The performances differ especially in fast fading channels. Furthermore, the computational complexity of the inversion of covariance matrix depends also on the matrix structure. Consequently, it is worthwhile to study how to estimate the signal covariance matrix used in equation (3.5) in WCDMA downlink systems.

In case of block based signal model, different ways to estimate the covariance matrix used for equalizer estimation include:

1. Traditional sample covariance matrix

$$\hat{\mathbf{R}}_{\mathbf{y}} = \frac{1}{N}\sum_{n=1}^{N} \mathbf{y}_n \mathbf{y}_n^H, \tag{3.6a}$$

   where N is the sample support.

2. Channel auto-correlation based estimation

$$\hat{\mathbf{R}}_{\mathbf{h}} = \sum_{q=1}^{Q} \sigma_d^{(q)} \hat{\mathbf{H}}^{(q)} (\hat{\mathbf{H}}^{(q)})^H + \hat{\mathbf{R}}_v, \tag{3.6b}$$

   where q is summation over different transmit antennas and interfering base stations. $\hat{\mathbf{R}}_v$ is the noise covariance matrix which is typically estimated as $\hat{\mathbf{R}}_v \approx \sigma_v \, \mathbf{I}$, where σ_v is the noise power.

3. Post despreading covariance matrix, which corresponds to symbol level equalization

$$\hat{\mathbf{R}}_{\mathbf{x}_k} = \frac{1}{N}\sum_{n=1}^{N} \mathbf{x}_{nk}\mathbf{x}_{nk}^H, \tag{3.6c}$$

where $\mathbf{x}_{nk}$ is despread symbol vector with multiple delays for nth symbol of kth user.

4. Interference + noise covariance matrix, corresponding to

$$\hat{\mathbf{R}}_{in} = \hat{\mathbf{R}}_{\mathbf{x}} - \rho\hat{\mathbf{h}}\hat{\mathbf{h}}^H. \tag{3.6d}$$

The first two estimates are widely used. They both may be used to form a chip level MMSE equalizer. Equation (3.6b) can be obtained straightforwardly from the signal model $\mathbf{y} = \mathbf{Hd}$ given in equation (2.4a) and with assumption that the chip level signal $\mathbf{d}$ is i.i.d. and the noise is uncorrelated with the signal. Similarly to $\hat{\mathbf{R}}_{\mathbf{h}}$ in equation (3.6b), the traditional sample covariance matrix $\hat{\mathbf{R}}_{\mathbf{y}}$ has nearly Toeplitz structure[1]. The Toeplitz structure is a direct consequence of the assumption that received signal is well modeled as WSS signal, [Hay96]. The channel length is assumed to be known and the auto-correlation length is limited to channel length.

The third approach corresponds to a symbol level MMSE equalizer and the fourth to the Generalized RAKE (GRAKE) [BOW00]. All the estimators give asymptotically the same equalizers up to a positive multiplicative constant. This is because the their expectations differ only by a rank one matrix $\mathbf{hh}^H$, e.g. $\hat{\mathbf{R}}_{\mathbf{x}} = \hat{\mathbf{R}}_{\mathbf{y}} + \gamma\mathbf{hh}^H$. This rank one modification only scales the equalizer, as shown in [FVM02, Hoo03], based on the matrix inversion lemma [Gv96].

The structure of the third covariance matrix estimate is not Toeplitz. Therefore, finding an equalizer based on it is more tedious. Additionally, due to an increased eigenvalue spread of the post-despreading covariance matrix, $\mathbf{R}_{\mathbf{x}}$, an equalizer based on it is less robust against channel estimation errors. Even tough $\mathbf{R}_{\mathbf{x}}$ is actually code and time dependent, if the number of orthogonal spreading codes used is large, i.e. cell load is high, the code dependency diminishes, as the γ parameter gets smaller. Value of γ depends on signal powers and spreading factors and the number of used codes. In Publication IV and Publication VII values of γ are studied in the context of semi-blind channel estimation.

To emphasize the similarities of the linear receivers, also the ZF equalizer and RAKE receiver can be defined using covariance matrices and equation

[1] When $\hat{\mathbf{R}}_{\mathbf{y}}$ is estimated with different auto-correlation lags of received signal, the structure is exactly Toeplitz.

(3.5). For ZF and RAKE the covariance matrix estimates are:

$$\mathbf{R}_{\mathbf{z}} = \sigma_d \mathbf{H}^{(q)} (\mathbf{H}^{(q)})^H, \tag{3.6e}$$

$$\mathbf{R}_{\mathbf{r}} = \sigma_d \mathbf{I}. \tag{3.6f}$$

These equations reveal nicely how the MMSE equalizer balances between ZF and RAKE receiver and what means that RAKE treats the temporal signal copies as uncorrelated.

Performance comparison

The performance of a linear receiver is often analyzed using signal to interference and noise ratio (SINR) at the receiver as a criterion. This can be in general case written as:

$$\text{SINR} = \frac{\mathbf{f}^H \mathbf{R}_s \mathbf{f}}{\mathbf{f}^H \mathbf{R}_{in} \mathbf{f}}, \tag{3.7}$$

see [GHST01, Ch. 5]. Here $\mathbf{R}_s$ denotes the desired signal covariance matrix, which for example is $\rho_k \mathbf{h}\mathbf{h}^H$ for symbol level signal. In WCDMA downlink, the problem is the time varying nature of the interference + noise covariance matrix $\mathbf{R}_{in}$. It depends on the systems load (number of codes versus spreading factor) and signal powers. Furthermore, in fast fading channels, the quality of available channel and covariance matrix estimates will also have an impact on the performance. Consequently, in the following as well in the original publications of this thesis, only simulation are used to validate the performance.

Simulations indicate that the channel estimate based covariance matrix estimator of equation (3.6b) yields better results than the sample covariance $\mathbf{R}_{\mathbf{y}}$ based estimator of equation (3.6a), see Publication V and Publication VII. This is due to the strong pilot signal. In WCDMA downlink 10% of the total transmit power is allocated to the common pilot channel, CPICH. Channel estimates are obtained by conventional correlator:

$$\hat{\mathbf{h}}_c = \frac{1}{N} \sum_{n=1}^{N} \boldsymbol{\mathcal{C}}_{nc} \mathbf{y}_n s_{nc}^* / \rho_c, \tag{3.8}$$

where the index c denotes pilot channel. For a fair comparison, the signal power and noise covariance matrix need to be estimated when the approach given in (3.6b) is used. In Publication VII a simple algorithm to estimate jointly the signal and noise powers is proposed. Even with this algorithm the channel auto-correlation based approach outperforms the sample matrix approach, (3.6a), when the geometry factor is high or moderate, i.e. the mobile is not close to cell edge. With lower geometry factor, more accurate signal and noise power estimates are needed.

The channel auto-correlation based approach seems to be especially well suited for HSDPA systems. The signal structure can change after every TTI if HSDPA codes on consecutive TTI's are allocated to different users, that may have different power and modulation schemes. In this case the estimate of $\mathbf{R_y}$ is formed by averaged over one TTI, i.e. three slots. Simulations indicate clearly that the shorter the sample support is, the greater is the performance difference between $\mathbf{R_y}$ and $\mathbf{R_h}$. With long enough observation period both estimators seem to perform equally well. It can be observed that time variations in channel parameters and signal and noise powers can be faster tracked than changes in covariance matrix. This is due to the strong pilot signal.

Diagonal loading is a well known technique to robustify covariance matrix prior to inversion, see e.g. [VT02]. It improves the conditioning of the covariance matrix. In the context of WCDMA downlink equalization it has been considered e.g. [Hoo03, Mai05] and Publication VII. It is not straightforward to define how much diagonal loading is needed. However, by observing equation (3.6f), it can be seen that the larger the load parameter the closer the receiver filter gets to RAKE receiver. Since the need for diagonal loading is only seen in simulations with very high SNR values, it might not be needed in practise.

3.4 Computational complexity

In this section the complexities of linear RAKE and MMSE receivers are studied. The results are only to compare different receivers, not to give exact numbers how complex receivers are in terms of complex additions and multiplications. To simplify the comparison further the tasks common to all receivers are not counted. These include for example channel estimation. The complexities of the receiver differ due to method for solving the MMSE weights and due to different filter lengths. For MMSE weight estimation only the SMI method, which requires solving of Wiener Hopf equation, (3.3), is considered. The complexities are estimated based on number of complex multiplications.

3.4.1 Solving SMI with Toeplitz covariance matrix

The Toeplitz structure of the covariance matrix can be used to lower the computational complexity of finding the MMSE equalizer coefficients. In case of single receive antenna, the $\mathbf{R}$ is a finite order Toeplitz matrix:

$$
\begin{bmatrix}
a_0 & a_1 & \dots & a_{L-1} & 0 & \dots & & & \\
a_{-1} & a_0 & \ddots & \ddots & \ddots & & & & \\
\vdots & \ddots & \ddots & & & & & & \\
a_{-L+1} & a_{-L+2} & & & & & & & \\
0 & a_{-L+1} & & & & & & & \\
 & \ddots & \ddots & & & & & & \\
\vdots & 0 & a_{-L+1} & \dots & a_0 & \dots & a_{L-1} & 0 & \dots \\
 & & \ddots & \ddots & \ddots & & & & \\
0 & & \ddots & 0 & \ddots & \ddots & \ddots & a_0 & a_1 \\
0 & & \dots & & 0 & a_{-L+1} & \dots & a_{-1} & a_0
\end{bmatrix}. \tag{3.9}
$$

The elements a_l denote auto-correlation with lag l. The matrix has finite order since $a_l = 0,\ \forall l > L-1$ and the size of $\mathbf{R}$ is $F \times F$, where F is the filter length. The auto-correlations can be estimated either from the channel estimates $\hat{\mathbf{h}}$ or from the received signal. The resulting matrices are defined in equations (3.6b) and (3.6a), respectively.

There are several well know techniques to solve a linear system with Toeplitz structure. For example, the well known Levinson-algorithm solves the problem with computational complexity of order $\mathcal{O}(F^2)$, where F is the filter length, [Gv96, KS99]. This has been used in WCDMA downlink equalization in e.g. [HJH$^+$02] for SISO systems.

In case of multiple receive antennas, i.e. SIMO case, the space-time covariance matrix is a Toeplitz-block or block-Toeplitz matrix depending on the system model. The model used in this thesis leads in case of M receive antennas to following Toeplitz-block structure :

$$
\mathbf{R} = \begin{bmatrix} \mathbf{R}_{1,1} & \dots & \mathbf{R}_{1,M} \\ \vdots & \ddots & \vdots \\ \mathbf{R}_{M,1} & \dots & \mathbf{R}_{M,M} \end{bmatrix}_{FM \times FM} . \tag{3.10}
$$

Here each block $\mathbf{R}_{m_i,m_j}$ has the Toeplitz structure shown (3.9). When $m_i \neq m_j$ the auto-correlation are replaced by cross-correlation between two antennas m_i and m_j. Equivalent system model could be written such that the covariance matrix structure is block-Toeplitz. It is a matrix consisting of blocks that are arranged in Toeplitz-fashion, i.e.

$$
\mathbf{R} = \begin{bmatrix} \mathbf{R}_1 & \dots & \mathbf{R}_F \\ \vdots & \ddots & \vdots \\ \mathbf{R}_F^H & \dots & \mathbf{R}_1 \end{bmatrix}_{FM \times FM} . \tag{3.11}
$$

Table 3.1: Computational complexities in terms of complex multiplications of Cholesky decomposition and Block-Levinson algorithms.

	$\times$	$F = 20,\ M = 2$
Cholesky	$(MF)^3/3 + 2(MF)^2$	≈ 24500
Block-Lev.	$4M(MF)^2$	≈ 12800

In this case each block $\mathbf{R}_f$ is size of $M \times M$ and is Hermitian, but not Toeplitz. This type covariance matrix can be obtained by stacking first the received signal for one delay and all the M antennas. Then these $M \times 1$ vectors for all delays are stacked into one long vector [2]. The filter coefficients found by solving the block-Toeplitz or Toeplitz-block systems are the same.

There are several methods which have been designed for solving block-Toeplitz systems. For example, an extension of the Levinson algorithm to solve block-Toeplitz can be found in [KS99]. Alternatively, the $\mathcal{O}(F^2)$ Trench algorithm for inverting Toeplitz matrices (see e.g. [Gv96]) can be adapted to block-Toeplitz case as is done in [Aka73]. Another alternative to lower the computational complexity is to utilize the Lanczos algorithm as has been proposed in [JMK06]. It does not assume any structure for the covariance matrix. Hence, it is well suited for use in multi-antenna receivers. In Table 3.1 are given the computational complexities of solving a general positive-definite Hermitian system of equations using Cholesky decomposition and a Levinson-like algorithm with receive antenna diversity. The estimate is based on the concept of displacement rank [KS99]. The complexity of the Levinson-algorithm is $4F^2$. Since a Toeplitz matrix has displacement rank $\chi = 2$, the complexity is approximately $2\chi F^2$. With multiple antennas the dimensionality is MF and the displacement rank is $\chi = 2M$ which in turn gives approximately the complexity of $4M(MF)^2$.

3.4.2 HSDPA receiver complexity

A lot of attention has been paid to filter coefficients update method. However, in HSDPA processing the computationally most demanding task is the filtering itself. This is due to low spreading factor ($G_p = 16$) and multiple codes to be despread. This is shown with a simple example given in Table 3.2. Naturally, chip and symbol rate processing will yield different complexities, but the difference is not as significant since the spreading factor is low and multiple codes need to be processed. The ratio between the different complexities is $1/16 \leq P/G_p \leq 15/16$. Consequently, chip rate processing might be useful, since it reduces the number of needed despreading units as

[2] System model leading to block Toeplitz covariance structure has been used e.g. in Publication III.

Table 3.2: Computational complexities in terms of complex multiplications for chip and symbol rate processing. Assuming $P = 10$ HSDPA codes, $L = 5$ and duration of one slot, i.e. 2560 chips. Two filter lengths are assumed $F = L$ and $F = 2L$. Number of receive antennas is $M = 1$ or $M = 2$

	$\times$	$F = L$ $M = 1$	$F = 2L$ $M = 1$	$F = 2L$ $M = 2$
Chip rate filtering	$F\ 2560$	12800	25600	51200
Symbol rate filtering	$\frac{PF}{G_p}\ 2560$	8000	16000	32000
Finding $\mathbf{R_h}$	$\frac{L(L+1)(M^2+M)}{2*2}$	15	15	45
Finding $\mathbf{R_y}$	$\frac{(L-1)(M^2+M)}{2}\ 2560$	10240	10240	30820
Solving $\mathbf{f}$ (Block-Lev.)	$4M(MF)^2$	200	400	3200

well memory. The complexity of estimating the covariance matrix is given for both $\mathbf{R_h}$ and $\mathbf{R_y}$.

Complexity of finding the filter coefficients depends also on the channel coherence time, i.e. how often the equalizer coefficients need to updated. For example, if the update is done once in a slot, the complexity of finding the filter coefficients is less critical. This is shown in the example given in Table 3.2. Furthermore, as proposed in [WZB04] linear interpolation can be used to update the filter coefficients more regularly.

3.5 Complexity reduction methods

In this section, methods for lowering the computational complexity are presented. First, a frequency domain matrix inversion is introduced. It was originally proposed in Publication V. Then two simple alternatives which offer trade-offs between complexity and performance are presented. Decoupled spatial optimization was evaluated in Publication V, while the idea of a simplified sparse equalizer rose during writing this thesis.

3.5.1 Matrix inversion in frequency domain

In order to lower the complexity of matrix inversion, an efficient frequency domain methods can be used. This approximative approach is based on knowledge that circular convolution in time domain and multiplication in frequency domain are duals. Fast Fourier transform (FFT) based matrix inversion methods for equalization in CDMA systems have been proposed in

Publication V and independently in [ZBM03, GZMC04, GZMC05]. Alternatively, the equalizer coefficients can be found in the frequency domain, see [BRH+03, BRFF03, Mai04]. In this methods FFT is performed to the estimated channel impulse response. For a tutorial on the Fourier and discrete Fourier transforms see e.g. [OS99].

The frequency domain matrix inversion methods are based on the property that a large finite order Toeplitz matrix can be approximated with a circulant matrix, see [Gra02]. With one receive antenna the sample covariance matrix in WCDMA system is a large Toeplitz matrix with its energy concentrated near the main-diagonal. Consequently, it is well approximated by a circulant matrix. Furthermore, circulant matrices can be diagonalized by Fourier transform. Than in the frequency domain the matrix inversion is simple element-wise inversion due to diagonality of the matrix.

A finite order Toeplitz matrix is asymptotically equivalent to a circulant matrix, where the upper right and lower left corners are filled with appropriate entries to make the matrix exactly circulant, see [Gra02]. The circulant approximation of the correlation matrix given in equation (3.9) is:

$$\mathbf{A} = \begin{bmatrix} a_0 & a_1 & \dots & 0 & \dots & a_{-2} & a_{-1} \\ a_{-1} & a_0 & \ddots & \ddots & & a_{-3} & a_{-2} \\ \vdots & \ddots & \ddots & & & & \\ a_{-L+1} & a_{-L+2} & & & & & \\ & \ddots & \ddots & & & & \\ 0 & \dots & a_{-L+1} & \dots & a_{L-1} & \dots & 0 \\ & \ddots & \ddots & \ddots & & & \\ a_2 & & & \ddots & \ddots & a_0 & a_1 \\ a_1 & a_2 & \dots & 0 & \dots & a_{-1} & a_0 \end{bmatrix}. \tag{3.12}$$

The inverse of a circulant matrix can be derived using any row of the matrix. This row is denoted as $\tilde{\mathbf{a}}$, in Table 3.3, where the frequency domain matrix inversion algorithm is given. It is a block-Toeplitz extension of the algorithm given in Publication V. Similar approach has been proposed in [ZBM03].

The influence of the FFT length and equalizer length are studied with decoupled processing in Publication V. The performance of FFT based matrix inversion method is shown to be close to exact inversion. Only differences are seen at high SNR region. Also the number of zeros needed to pad $\mathbf{a}_m$ for adequate approximation has been tested via simulations. For computationally efficient FFT implementation, the number of zeros should be selected such that the length of $\tilde{\mathbf{a}}_m$ is a power of two.

Table 3.3: FFT matrix inversion for M receive antennas

For each antenna pair	**for** $\forall\ \mathbf{m} = \{m_1, m_2\},\ m_i \in \{1, 2, \dots M\}$
cross correlation vector	$\mathbf{a_m} = \mathbf{R}_{m_1 m_2}(\tau, :),\ \tau = \max(F/2, L)$
Zero pad	$\tilde{\mathbf{a}}_\mathbf{m} = [0, \dots, 0, \mathbf{a}_m, 0, \dots, 0]$
Move to frequency domain	$\boldsymbol{\varphi}_\mathbf{m} = \mathrm{FFT}(\tilde{\mathbf{a}}_\mathbf{m})$
	end
For each frequency bin	**for** $i = 1 : \mathrm{length}(\tilde{\mathbf{a}}_\mathbf{m})$:
$M \times M$ matrix inversion	$\Psi^{(i)} = \begin{bmatrix} \boldsymbol{\varphi}_{1,1}(i) & & \boldsymbol{\varphi}_{1,M}(i) \\ & \ddots & \\ \boldsymbol{\varphi}_{M,1}(i) & & \boldsymbol{\varphi}_{M,M}(i) \end{bmatrix}^{-1}$
	end
For each antenna pair	**for** $\forall\ \mathbf{m}$
	$\psi_\mathbf{m} = [\Psi^{(1)}(m_1, m_2), \dots, \Psi^{(i)}(m_1, m_2), \dots]$
Return to time domain	$\boldsymbol{r}_\mathbf{m} = \mathrm{IFFT}(\psi_\mathbf{m})$
Inverse matrix	$\hat{\mathbf{R}}_\mathbf{m} = \mathrm{Toeplitz}(\boldsymbol{r}_\mathbf{m}(\tau : \Gamma))$ (*
	end
Formulate	$\hat{\mathbf{R}}^{-1} \approx \begin{bmatrix} \hat{\mathbf{R}}_{1,1} & & \hat{\mathbf{R}}_{1,M} \\ & \ddots & \\ \hat{\mathbf{R}}_{M,1} & & \hat{\mathbf{R}}_{M,M} \end{bmatrix}$

(* Choose Γ such that dimension of $\hat{\mathbf{R}}_\mathbf{m}$ is equal to filter length F.

Complexity

The computational complexity of the frequency domain matrix inversion method is dominated by the FFTs and inverse FFTs. One FFT of length F takes $F \log_2 F$ complex multiplications and some additional overhead from the computation of sines and cosines [Gv96]. The method given in Table 3.3 for block-Toeplitz matrices needs $(M^2 + M)/2$ FFTs to transform the covariance matrix to the frequency domain and the same amount of inverse FFTs (IFFT) to transform the elements of the inverse matrix back to time domain. In addition, inverting an $M \times M$ Hermitian positive definite matrix is approximated to take $M^3/3 + 2M^2$ complex multiplications [Gv96], which is a bit over-kill for a small number of antennas, especially for $M = 2$. Further, while Levinson recursion or Cholesky decomposition with back substitution solve the equations, the proposed FFT method only inverts the channel. Additional FL multiplications are needed to find the filter coefficients. Consequently, the total complexity of the FFT-inversion method is of order: $(M^2 + M)F \log_2 F + F(M^3/3 + 2M^2) + FL$. The complexity with FFT for the example given in Table 3.1 is ≈ 700. Here it's assumed that $M = 2$, $L = 10$ and $F = 32$ instead of 20 used in Table 3.1 in order to facilitate efficient FFT implementation. Consequently, in this case the FFT matrix inversion method reduces the complexity approximately by decade compared to Levinson method.

3.5.2 Decoupled processing

When there are multiple antennas at the receiver, joint space-time equalization is commonly employed. This means that the linear filter coefficients are jointly optimized for spatial and temporal filter taps. However, the gains obtained via joint spatial optimization are rarely considered even though it causes an increase in complexity. In case of the example used in Table 3.1, the complexity reduction from block-Levinson algorithm (size FM) to M Levinson (size F) is order of magnitude. Furthermore, the gains of joint processing are not always evident due to errors in parameter estimation. This was shown via simulations in Publication V using estimates obtained with finite sample support and different covariance matrix estimators.

3.5.3 Filter length - Sparse equalizers

Filter length is a dominant factor to complexity of a linear receiver. In literature equalizers are assumed to be quite long, for example $F \geq 2L$ and in general, longer equalizer improve the performance. With RAKE the number of fingers is in practise limited to $4-5$ fingers [HT04]. The fingers are placed on strongest multipaths, which may lead to a sparse filter structure since the delay spread in chips can expand the number of fingers. In equalization the term sparse is conventionally used to identify channels where long delays may exist between significant channel taps. In this case the equalizer coefficients can estimated using reduced dimension covariance matrix, see e.g. [CZG01, RB03, CM99].

In case of WCDMA downlink receivers which enjoy the benefits of equalizer but with RAKE structure have been proposed e.g. in [BOW00, LMS00, LS01, KC05]. Joint temporal optimization for finger coefficients are derived using reduced size covariance matrix. The generalized RAKE, [BOW00, Bot03], uses interference and noise covariance matrix, $\mathbf{R}_{in}$, whereas in [KC05] uses channel auto-correlation based covariance matrix is used. An algorithm for finding the positions for a sparse receiver taps is given in [KC05] and the receiver is called sparse chip equalizer. In [BOW00, LMS00, LS01, Bot03] no explicit algorithms are given how to select number of fingers or their placement.

In practise, due to pulse shaping and rich scattering environment the WCDMA channel is typically not sparse. Consequently, neither the covariance matrix nor the equalizer are sparse. In the following an alternative sparse filter, which approximates MMSE equalizer is derived. First, estimate the equalizer coefficients for entire delay spread using full size covariance matrix. After that select the strongest equalizer taps, depending the number of available RAKE fingers. Consequently, the complexity of filtering is equal to RAKE, whereas the estimation of the filter weights is similar MMSE equalizer. This is viable alternative, since for example using the low complexity method proposed in this thesis, the finding the equalizer coefficients is not overwhelmingly complex.

In Figure 3.2 the bit error rate (BER) performance is plotted for different ST-MMSE equalizer lengths compared to ST-RAKE receiver and sparse ST-MMSE equalizer. This example shows that even in less dispersive channels, such as ITU Ped A, the ST-MMSE equalizer can provide gains over ST-RAKE. Furthermore, sparse ST-MMSE equalizer provides a nice trade-off between RAKE and ST-MMSE equalizer. Depending on the affordable computational complexity the covariance matrix used for sparse equalizer coefficient estimation should have dimension of at least $2L$ while the actual number of filter taps used F can be much less.

The sparse equalizer considered here has several advantages. Firstly, it provides a nice trade-off between complexity and performance, since complexity is close to that of RAKE and performance close to that of ST-MMSE. Secondly, with sparse equalizer the number of multipaths that can be taken into account is increased. This might be useful since all the weak channel taps can be used without fixing their number beforehand. For example, in this thesis the channel impulse response is estimated for the entire delay spread without thresholding any of the channel taps. In some channel realizations [3] there are multipath components close to zero, but the performance compared to optimal ST-MMSE is not drastically degraded due to this overestimation of the channel length. Furthermore, with sparse equalization no complex algorithm is needed for estimating the finger placements, since simply the strongest coefficients of equalizer vector are selected.

3.6 Non-linear receiver structures

Due to multiple sources of interference in WCDMA, non-linear receivers can be used to provide improved performance. In this section a brief literature review is given to receivers proposed for WCDMA downlink or similar systems. These receivers include different kind interference cancellation methods combined with RAKE receiver or MMSE equalizers. Receivers such as

[3] All the simulations are averaged over 200 (or more) channel realizations.

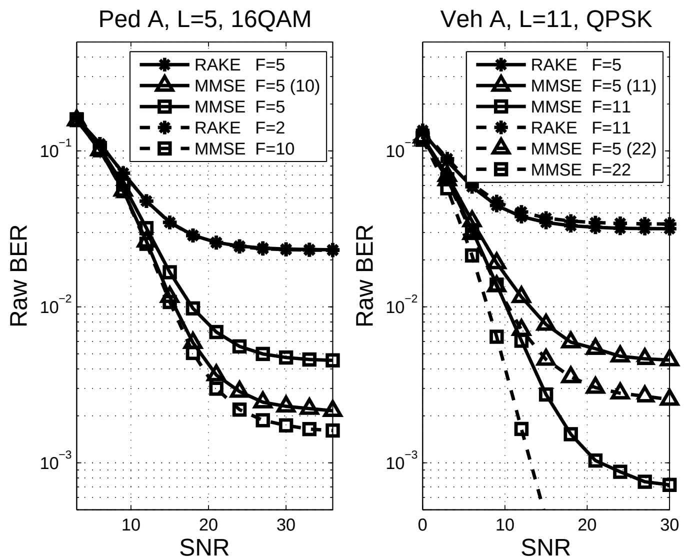

Figure 3.2: *The raw BER performance comparison of RAKE and ST-MMSE equalizer and sparse ST-MMSE equalizer.* $P = 5$ *HSDPA signals and 40 speech users as intra-cell interferers. One interfering BS with geometry factor* $g = 6$ *dB. Two receive antennas* $M = 2$*. Sparse equalizers are denoted with '*Δ*' and for example* $F = 5(11)$ *denotes that* 5 *strongest taps are used while the dimension of covariance matrix* $\mathbf{R_h}$ *(for single antenna) used in equalizer derivation is* 11*. Covariance matrix estimate is based on the channel estimates from desired and interfering BS with sample support of one TTI. Conventional ST-MMSE equalizers with different lengths are drawn with '* $\Box$*' and RAKE receiver with'* $\star$*'*

non-linear equalizers are not discussed due to their high computationally complexity.

In HSDPA systems the received signal consists of the desired signal and many types of interference. Therefore, a natural choice to improve the detection is to use interference cancellation (IC) type detectors. Alternatively, when multiple spreading codes are known (such as in HSDPA and in WCDMA uplink) also other type of multiuser detection algorithms can be applied.

A multipath interference cancellation method for WCDMA downlink, has been proposed and further developed in [KHKS02, HFS02, YCKH05]. In this type of multistage receiver, estimates of the multipath interference (MPI) are generated separately for each path based on symbol estimates. In the next stage, the interfering $L - 1$ MPI replicas are canceled from the received signal and separate estimates for each path are formed. New symbol estimate is obtained by combining these path-wise estimates with MRC.

The most studied IC detector for CDMA systems is parallel interference cancellation (PIC) used to reduce the influence of other users signals. These methods have been widely studied for different CDMA systems over the past years. Basic types of interference cancellation techniques are successive (SIC) and parallel (PIC) interference cancellation, see [Ver95, Mos96] for extensive reviews. Due to common scrambling code and knowledge of multiple codes, IC algorithm are applied in HSDPA systems, see e.g. [CWS02, MGW02, HLS04, BS05]. A modified partial PIC receiver, is used for HSDPA in [WHL02], where it is applied after conventional RAKE receiver.

Even though MMSE equalization aims to restore the orthogonality of the spreading codes, typically there is some residual interference due to other codes. Consequently combination of MMSE equalizer and multiuser detection methods have been reported to provide some performance gains in long code CDMA systems in [CZG00a, PVB03, WCGB04]. In [BS05] multicode multistage IC is used after ZF- equalizer, which is approximated with a polynomial expansion (PE) technique. The actual benefit of these type receivers depends on the accuracy of MMSE equalizer. Since complexity is one key issue determining the receiver, it might not be worth of combining complex multiuser detection algorithms with equalization.

Pilot symbol interference cancellation is considered e.g. in [PVB03] as a part of more complex receiver structure combining MMSE equalization and multi-user detection. Pilot symbol interference can be subtracted as part of the multi-code interference or separately, [HLS04].

Typically interference due to neighbouring cells has been considered as Gaussian noise, and no separate cancellation is considered. Alternatively, it's well known that equalization with multiple receive antennas can provide interference suppression. In [TFD00] interference canceler matched filter is combined with maximum SINR receiver in order to reduce neighbouring cell interference. A joint HSDPA interference and inter-cell interference cancellation is considered also in [BS05].

3.7 Discussion

In this chapter advanced receivers for WCDMA downlink were studied. The main focus was on the MMSE equalizer, which in optimal case without noise is able to restore the orthogonality of the spreading codes. Consequently, it cancels out the intra-cell interference which is the main source of interference in the downlink. Furthermore, with multiple receive antennas ST-MMSE equalizer provides suppression of inter-cell interference. In practical case, due to estimation errors and system model imperfections, the performance of ST-MMSE depends on method used for estimation of the equalizer weights. The choice of this estimation method influences also computational complexity of the receiver.

Estimation of the signal correlation matrix based on channel autocorrelation approach was shown to provide both computational saving as well as improved BER performance over the conventional sample covariance matrix estimation approach. Furthermore, the computational complexity was reduced by introducing a frequency domain matrix inversion method and sparse selection of filter taps. The proposed sparse equalizer was shown to provide a viable alternative for RAKE receiver in WCDMA downlink. Even though the same number of fingers was used, it had clearly improved performance over conventional RAKE, since the correlation structure of interference and desired signal was taken into account. The performance was closer to that of MMSE equalizer than RAKE, depending how many filter taps were used. The more computational power the receiver had, the closer the performances of the sparse and the MMSE equalizer were.

In this chapter, a conventional correlator based channel estimator was used. It was shown that the performance of equalizer depends clearly on the quality of the channel estimate. Consequently, for further improvements in performance the channel estimation as well as signal to noise ratio estimation methods should be studied in more detail.

Chapter 4

Advanced receivers for MIMO systems

Multiple transmit and receive antennas can be used to increase system capacity or improve the quality of received signal. In this chapter, advanced receivers for spatial multiplexing (SM) MIMO systems are studied. SM is a technique which aims to increase the data rate by transmitting different data streams from different transmit antennas. Unfortunately, SM also introduces a new dimension of interference to the received signal because the signals transmitted from different antennas share the same spectrum. This so called inter-antenna interference can be especially harmful in CDMA systems where the same spreading codes are reused across the antennas. With code reuse the receiver can distinguish the symbols transmitted from different antennas based only on the different channel coefficients. In HSDPA systems code reuse and SM based MIMO are included in the future revision for very high data rate transmission, [TS805b, HT06]. Consequently, new sophisticated received design is needed.

In this chapter an overview of existing MIMO receiver structures is given first. After that, a brief review of receivers for frequency flat MIMO channels is presented. Next, space-time (ST) receivers for frequency selective channels are reviewed. Both linear and hybrid receiver structures combining a linear receiver with a non-linear receiver are considered. Finally, a new joint ST-MMSE equalizer and inter-antenna interference canceler for HSDPA MIMO systems is introduced. It was originally proposed in Publication VI.

4.1 Overview of MIMO Receivers

A general classification of receivers designed for MIMO systems with spatial multiplexing is given in Figure 4.1. The classification for frequency flat channels is based on well known receivers described e.g. in [PNG03, TV05]. Receivers for the frequency selective channels, which are more relevant to

WCDMA systems, are less studied in the literature. The extensions of linear receivers to frequency selective MIMO systems are rather straightforward. In this thesis, following the terminology of [PNG03], S (space only) and ST (space-time) are used to distinguish the different receivers for frequency flat and frequency selective cases. Furthermore, VST is used to denote virtual space-time receiver, where the multipath components are considered as virtual antennas. Iterative MIMO receivers combining decoding and detection, see e.g. [PNG03, HTR03], are beyond the scope of this thesis. The organization of this chapter follows the structure given in Figure, 4.1. The new receiver structure introduced in section 4.4 is called ST-MMSE-MOSIC.

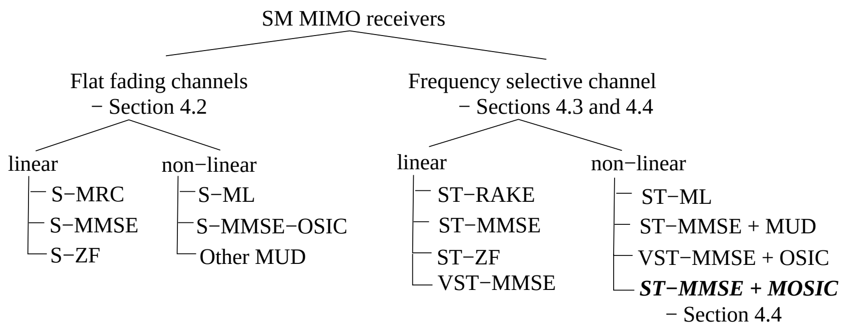

Figure 4.1: Categorisation of receivers for SM MIMO systems.

In receiver design it is important to understand the signal and interference structure. There are five signal dimensions which can provide diversity in MIMO CDMA systems. These are two spatial (transmit and receive), frequency, time and code dimensions. Furthermore, diversity due to antenna polarization could be utilized, but this is beyond the scope of this thesis, see e.g. [PNG03]. Typically the interference is divided into intra-cell and inter-cell interference. In case of SM MIMO, the intra-cell interference can be further divided into inter-antenna, inter-symbol, and inter-code interference. The inter-code interference denotes the interference from other users (codes) transmitting from the same antenna.

In WCDMA downlink, where typically the number of used codes is relatively high, both the chip and symbol level sequences transmitted from different antennas can be modeled as uncorrelated. Furthermore, due to the orthogonality of the spreading codes, the inter-symbol and inter-code (i.e. intra-cell) interference are present only in frequency selective channels. An other important feature in HSDPA MIMO with SM is the code reuse across transmit antennas. It will increase the inter-antenna interference after despreading.

Due to linearity and code orthogonality, performance differences of chip and symbol rate processing are in theory only seen in frequency selective channels with non-linear receivers. However, similar to the SIMO case, the optimization of linear filter coefficients differs for symbol and chip rate cases.

4.2 Receivers for flat fading channels

In this section, well known receivers for flat fading MIMO are revised. They are important in order to understand the more complex receiver structures needed for the frequency selective case. Recall the general received signal model for MIMO system from equation (2.11a):

$$\mathbf{y}_n = \boldsymbol{\mathcal{H}}\mathbf{b}_n + \mathbf{v}_n, \tag{4.1}$$

where $\mathbf{b}$ is the transmitted chip sequence vector and $\boldsymbol{\mathcal{H}}$ is the MIMO channel matrix and $\mathbf{v}$ is noise. This notation is commonly used in the MIMO literature. Majority of the MIMO studies are carried out assuming the channel is frequency non-selective, which reduces the channel matrix, defined in equation (2.12), to a $MG \times QG$ matrix:

$$\boldsymbol{\mathcal{H}} = \begin{bmatrix} h_1^{(1)} & h_1^{(2)} & \dots & h_1^{(Q)} \\ h_2^{(1)} & h_2^{(2)} & \dots & h_2^{(Q)} \\ \vdots & \vdots & & \vdots \\ h_M^{(1)} & h_M^{(2)} & \dots & h_M^{(Q)} \end{bmatrix}_{M\times Q} \otimes \mathbf{I}_G = \dot{\boldsymbol{\mathcal{H}}} \otimes \mathbf{I}_G, \tag{4.2}$$

where $\dot{\boldsymbol{\mathcal{H}}}$ is the $M \times Q$ symbol level MIMO channel matrix. Here Q denotes the number of transmit antennas and M the number of receive antennas and G is the spreading factor. With this notation the symbol rate presentation of received (and despread) signal can be written as:

$$\mathbf{x}_n = \dot{\boldsymbol{\mathcal{H}}}\mathbf{s}_n + \dot{\mathbf{v}}_n, \tag{4.3}$$

where $\mathbf{s}_n$ is the nth transmitted symbol vector. In flat fading channels it is size of $Q \times 1$. Symbol level noise vector is denoted with $\dot{\mathbf{v}}_n$.

4.2.1 Linear receivers

The simplest MIMO receiver is a diversity receiver, called space MRC (S-MRC) receiver, see [PNG03]. It ignores the inter-antenna interference and combines the different receive diversity branches for each data stream at the time. For the qth stream this receiver is simply:

$$\mathbf{f}_{SMRC}^{(q)} = \mathbf{h}^{(q)*} = \dot{\boldsymbol{\mathcal{H}}}(:,q)^*, \tag{4.4}$$

where $\mathbf{h}^{(q)}$ is the $M \times 1$ channel vector from transmit antenna q to all M receive antennas.

More advanced receivers take into account the inter-antenna interference. The same techniques used to cancel ISI in previous chapter can now be used to cancel out the inter-antenna interference. By jointly optimizing linear filters for all the Q signals transmitted from different antennas, the well known MMSE and Zero Forcing (ZF) [1] MIMO receivers, [PNG03, TV05], can be derived. Following the terminology of [PNG03] the MMSE MIMO receiver is called spatial MMSE (S-MMSE). The straightforward derivation for this receiver gives the filters for all Q antennas:

$$\mathbf{F}^H = (\sigma^{(q)} \dot{\boldsymbol{\mathcal{H}}}^H \dot{\boldsymbol{\mathcal{H}}} + \dot{\mathbf{R}}_v)^{-1} \sqrt{\sigma^{(q)}}\, \dot{\boldsymbol{\mathcal{H}}}^H \tag{4.5}$$

The noise, $\dot{\mathbf{v}}$, is typically assumed white, i.e. $\dot{\mathbf{R}}_v \approx \sigma_{\mathbf{v}}\, \mathbf{I}$ and $\sigma^{(q)}$ is the power used to transmit the sequence from qth antenna. For further use the term spatial correlation matrix of size $Q \times Q$ is defined as:

$$\boldsymbol{\mathcal{R}}_{s_h} = \sigma^{(q)} \dot{\boldsymbol{\mathcal{H}}}^H \dot{\boldsymbol{\mathcal{H}}} + \dot{\mathbf{R}}_v. \tag{4.6}$$

In this chapter, only MIMO system where the number of transmit and receive antennas are equal, i.e. $M = Q$, are considered. This is a natural consequence of the well know result that the maximum capacity increase due to SM is proportional to $\min(M, Q)$. If $M > Q$, the extra antennas can be used to provide array gain, i.e. improve the SINR. In the downlink, the more probable case of $M < Q$ allows usage of different transmit diversity schemes, see for example [PNG03, TV05, HTR03].

4.2.2 Non-Linear receivers

There are two important non-linear MIMO receivers, [PNG03, TV05], the spatial maximum likelihood (S-ML) receiver and successive interference cancellation (SIC) receiver. Both the optimal ML receiver and SIC suffer from high computational complexity but on the other hand they have been shown to provide clear performance gains.

In flat fading channels the consecutive symbols are assumed independent and the ML estimator can be formulated as:

$$\hat{\mathbf{s}}_n = \arg\min_{\mathbf{s}} ||\mathbf{x}_n - \sqrt{\sigma^{(q)}}\, \dot{\boldsymbol{\mathcal{H}}} \mathbf{s}||^2. \tag{4.7}$$

Here $\mathbf{s}$ is the symbol level sequence with Q symbols. This receiver is equivalent to ML multiuser detection, see [TV05]. The complexity of optimal S-ML detection method can be lowered with sphere decoder, [PNG03]. Alternative suboptimal methods have been extensively studied in CDMA multiuser context, see [Ver95].

An ordered successive inter-antenna interference cancellation (OSIC) [2] method for V-BLAST systems is proposed in [GFVW99]. It has been further studied and analysed e.g. in [TV05, Ch. 8]. The basic idea of this

[1] ZF receiver is also called decorrelator or (inter-antenna) interference nuller in the literature.

[2] Also term ordered successive cancellation (OSUC) is used.

receiver is straightforward ordered serial interference cancellation applied to the received signals in descending order of signal powers. First, S-MMSE (or other linear receiver) is applied to the strongest signal. Then with the estimated symbols, interference from this antenna is canceled from the received signal $\mathbf{y}$. Consequently, for the next transmit antenna, the interference due to the strongest signal is canceled. Then, the spatial covariance matrix is re-estimated for the next antenna and the symbols are estimated using this new S-MMSE receiver. After that, the influence of the second antenna is canceled. The procedure is repeated for all antennas. This receiver is denoted here as S-MMSE-OSIC.

4.3 Receivers for frequency selective channels

In WCDMA systems the channel is typically frequency selective, which causes both inter-symbol and inter-code interference. On the other hand it also provides multipath diversity. The signal model was presented in equation (2.11) and the corresponding MIMO channel matrix in equation (2.12). In this section, linear receivers for frequency selective MIMO systems are discussed first. Then, non-linear receiver structures proposed for WCDMA downlink are reviewed.

4.3.1 Linear receiver

Space-time RAKE

A straightforward extension of the RAKE receiver for the multipath MIMO case is to use the S-MRC receiver and assume that the multipath components are independent after despreading and are treated as noise. Consequently, the MRC diversity receiver given in equation (4.4) is directly applicable also to the multipath case.

Alternatively, a slightly more sophisticated space-time RAKE receiver, [PNG03], takes the inter-antenna correlation into account. The multipath components are still assumed to be independent, but prior to finger combining a S-MMSE receiver is applied to separate the different transmit antenna sub-streams. The S-MMSE given in equation (4.5) is derived separately for each multipath, but the spatial covariance matrix, $\boldsymbol{\mathcal{R}}_{s_h}$, see equation (4.6), is assumed to be same for all multipath delays.

Space-time MMSE equalizer

The derivation of space-time (ST) MMSE equalizer for MIMO systems is based on the observation that the chip sequences transmitted form different antennas are uncorrelated. Considering the interfering transmit antenna as noise the ST-MMSE equalizer can be found as a solution to the Wiener-Hopf equation, given in equation (3.3). The filter coefficients for the qth transmit antenna are:

$$\mathbf{f}^{(q)} = \sqrt{\sigma^{(q)}}\,\mathbf{R}^{-1}\mathbf{h}^{(q)}, \tag{4.8}$$

where $\sigma^{(q)}$ is the transmitted signal power from the qth antenna. The ST covariance matrix, $\mathbf{R}$, can be estimated in many different ways, see equations (3.6)(a-c). For example, the channel auto-correlation based expression given in equation (3.6b), is in the MIMO case:

$$\mathbf{R}_h = \sum_q \sigma^{(q)}\mathbf{H}^{(q)}\mathbf{H}^{(q)H} + \mathbf{R}_v, \tag{4.9}$$

where $\mathbf{R}_v$ is the noise covariance matrix.

To gain insight on the behaviour of the equalizer with inter-antenna interference, the equalized chip sequence $\mathbf{z}^{(q)}$ is written as

$$\mathbf{z}^{(q)} = \mathbf{F}^{(q)H}\mathbf{y} = \mathbf{F}^{(q)H}\mathbf{H}^{(q)}\mathbf{d}^{(q)} + \sum_{r,r\neq q}\mathbf{F}^{(q)H}\mathbf{H}^{(r)}\mathbf{d}^{(r)} + \mathbf{F}^{(q)H}\mathbf{v}(n), \tag{4.10}$$

where $\mathbf{F}^{(q)}$ is an equalizer convolution matrix $\mathbf{F}^{(q)} = [\mathbf{F}_1^{(q)};\ldots;\mathbf{F}_M^{(q)}]$, where the sub-matrix for one receive antenna is $\mathbf{F}_m^{(q)} = \mathcal{T}\left([\mathbf{f}_m^{(q)T},\ \mathbf{0}]^T, [\mathbf{f}_m^{(q)}(1),\ \mathbf{0}]\right)$ and $\mathcal{T}$ is the Toeplitz operator. The three terms on the right-hand side correspond respectively to the equalized desired signal, the residual inter-antenna interference, and the noise. The filter coefficients should be selected taking into account all these terms. Consequently, the optimization task is three-fold. One needs to equalize the desired signal, mitigate the inter-antenna interference and suppress the noise. To get a better idea of these three tasks, the equalized impulse responses $\mathbf{g}^{(qr)}$ is defined using a convolution sum as

$$\mathbf{g}^{(qr)} = \sum_m \tilde{\mathbf{f}}_m^{(q)} \star \mathbf{h}_m^{(r)}, \tag{4.11}$$

where $\tilde{\mathbf{f}}_m^{(q)} = [\mathbf{f}_m^{(q)*}(F)\ldots\mathbf{f}_m^{(q)*}(1)]$ for the mth receive antenna. Now, equalization of the desired signal can be achieved by requiring $\mathbf{g}^{(qq)}$ to be close to a delta function for $q = 1,\ldots,Q$. The second goal is to suppress inter-antenna interference, i.e. obtain $\mathbf{g}^{(qr)}$ close to zero for $q \neq r$. The third task is the suppression of the noise, i.e. keeping the noise term after equalization $\mathbf{F}^{(q)H}\mathbf{v}(n)$ as small as possible. In an MMSE equalizer, all this is achieved by obtaining MMSE estimates of the transmitted chips. In Figure 4.2, an example of the residual interferences (both inter-symbol and inter-antenna) is presented. The equalized impulse responses $\mathbf{g}^{(qr)}$ are plotted for one realization of 2×2 MIMO in the ITU vehicular A channel.

As discussed in the context of the MMSE equalizer in subsection 3.3.2, the different methods for estimating the covariance matrix yield different performances with finite sample support. In Figure 4.3 the raw BER performance with estimated channels for a 2×2 MIMO system is shown. Interference from one neighbouring cell with a single transmit antenna was

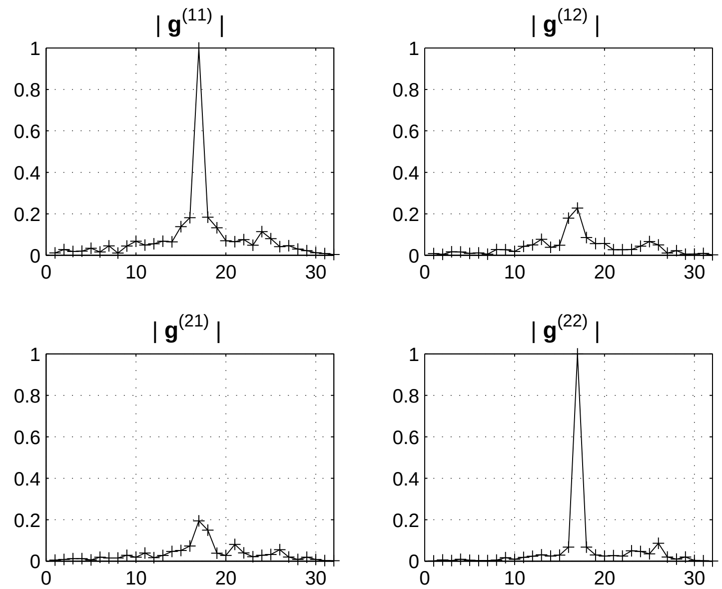

Figure 4.2: *Equalized channel responses, with ITU veh A channel,* $L = 11$*,* $F = 2L = 22$ *and SNR* $= 14dB$ *for* 2×2 *MIMO system.*

assumed. Even though the channel estimator, given in equation (3.8), is a simple correlator using only the known pilot signal, the channel estimate based method, equation (3.6b), for obtaining the covariance matrix gives the best performance. This method is denoted by HHH in the plot. Curves for the post-despreading covariance matrix, see equation (3.6c), method, denoted by RX H are plotted with $\mathbf{R}$ averaged over all known codes. This improves the performance in half-loaded systems over the conventional pre-despreading covariance matrix method (RY H given in equation (3.6a)), as reported in [YHKJ04]. However, with increased system loads the performances of RX H and RY H are similar For comparison, the ST-MMSE equalizer with known channels is also shown.

ST-MMSE equalization in long code CDMA systems with MIMO channel model has been studied [BRFF03, PBDM02, WMV02, Mai05]. Normalized least squares (NLMS) adaptation of ST-MMSE equalization has been proposed in [WMV02, Mai03, Mai05]. In [YHKJ04, YHKJ05] derivation of equalizer with post despreading correlation matrix has also been considered.

Frequency domain MMSE equalization has been studied in [BRFF03, BRH$^+$03]. The equalizer coefficients are calculated in the frequency domain from zero padded channel impulse responses. In this approach, equalizers are derived separately for each transmit-receive antenna pair in frequency domain. After equalization, the signals are combined to obtain estimates

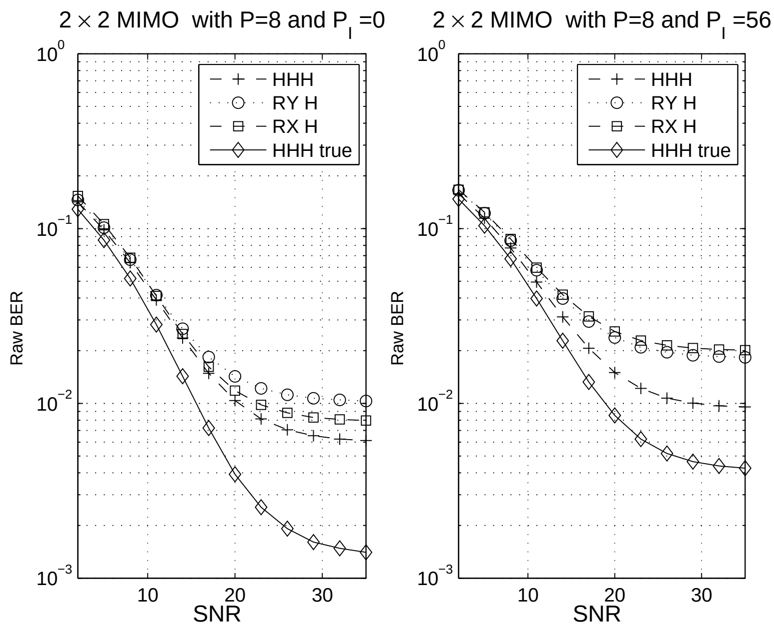

Figure 4.3: *Different covariance matrix estimates in* 2×2 *MIMO system. ITU Veh A channel model* ($L = 11$) *averaged over 256 random realizations.* $P = 8$ *HSDPA codes and* $P_I = 0$ *or* $P_I = 56$ *speech users. Interfering BS with one transmit antenna and* $g = 8dB$. *Channels are estimated with* 20 *pilot symbols based on a simple correlator with pilot signal.*

for all Q transmitted signals. Finally, the equalized signal is transformed to time domain for despreading. A frequency domain approach with NLMS adaptive update is derived in [Mai05].

Virtual space time (VST) MMSE receiver

An alternative formulation for a MIMO receiver is based on a virtual antenna concept, see e.g. [ABB$^+$03, SWL01]. After despreading, the multipath received signals can be considered as independent diversity sources, called virtual antennas. The virtual channel matrix is defined as:

$$\dot{\boldsymbol{\mathcal{H}}}_v = \begin{bmatrix} \mathbf{h}_1^{(1)} & \mathbf{h}_1^{(2)} & \dots & \mathbf{h}_1^{(Q)} \\ \mathbf{h}_2^{(1)} & \mathbf{h}_2^{(2)} & \dots & \mathbf{h}_2^{(Q)} \\ \vdots & \vdots & & \vdots \\ \mathbf{h}_M^{(1)} & \mathbf{h}_M^{(2)} & \dots & \mathbf{h}_M^{(Q)} \end{bmatrix}_{ML\times Q,} \tag{4.12}$$

where $\mathbf{h}_m^{(q)}$ is the $L\times 1$ multipath channel vector from qth transmit antenna to mth receive antenna. Similar to the RAKE, the received signal is despread with different delays and stacked to a $ML \times 1$ vector $\mathbf{x}_v$. Due to code reuse this can be written in form of the common MIMO system model as: $\mathbf{x}_v = \dot{\boldsymbol{\mathcal{H}}}_v \mathbf{s}_v + \dot{\mathbf{v}}_v$. With the above virtual matrix notation the ST-RAKE can be expressed as $\mathbf{f}^{(q)}_{STMRC} = \dot{\boldsymbol{\mathcal{H}}}_v(:, q)^*$.

A virtual space-time MMSE (VST-MMSE) receiver can be defined with equation (4.5) and using the virtual channel matrix $\dot{\boldsymbol{\mathcal{H}}}_v$ given in previous equation. This symbol level receiver treats the inter-symbol interference after despreading as AWGN, similarly to the RAKE receiver. VST-MMSE receiver was proposed as part of more complex receiver in [ABB$^+$03, SWL01], and it has not been studied in detail. However, VST-MMSE and the corresponding VST-ZF have been used as a front-end for interference cancellation receivers for example in [ABB$^+$03, SWL01, WWC$^+$02b, WWC$^+$02a].

4.3.2 Non-Linear receiver

In this section non-linear receivers with hybrid structure are reviewed. ST-ML which is a optimum spatial receiver for frequency selective channels is straightforward extension of the S-ML receiver, [PNG03].

Linear receivers are not capable of removing all the interferences in frequency selective channels in WCDMA downlink systems with multiple transmit antennas. Some residual interference remains, even if the channels would be perfectly known, see Figure 4.2. Furthermore, due to code reuse across transmit antennas, despreading will increase the residual inter-antenna interference. Consequently, with SM transmission, combining a linear receiver with some multiuser detector is a viable solution for improvement of the performance.

For WCDMA downlink joint detection of the Q symbols transmitted from the different transmit antennas has been considered. In HSDPA systems with multi-code[3] transmission all PQ HSDPA signals could also be jointly detected. However, this adds significantly to the computationally complexity. Furthermore, ST-equalization can be used to suppress both temporal and inter-code interference due to the used orthogonal spreading codes.

Usage of Maximum A Posteriori (MAP) multiuser detector after ST-MMSE equalizer for WCDMA downlink MIMO systems has been studied in [CKM03, KC03, YHKJ04, YHKJ05]. Similar idea, but with a slightly different approach is presented in [GMB03], where MAP is combined with generalized RAKE (GRAKE). Complexity reduction methods for MAP detection are presented e.g. in [CKM03, KHYJ05]

In order to reduce the complexity of spatial processing, other group-wise multiuser detection algorithms such as PIC could be applied. In [PBDM02, CKM03] a ST-MMSE equalizer is followed by multistage PIC. A hybrid

[3] P is the number of known spreading codes transmitted from each Q antennas.

receiver combining GRAKE and SIC has been considered in [GCK+04].

Another spatial interference cancellation method was proposed in [ABB+03, SWL01]. It extends the S-MMSE-OSIC receiver to frequency selective WCDMA systems, i.e. it is a symbol level inter-antenna interference canceler. The extension is based on the virtual antenna concept, described in section 4.3.1. Consequently, the receiver is denoted here by virtual space time (VST) MMSE-OSIC. At initial stage a VST-MMSE (or VST-ZF) front-end is used to separate the symbol level signals from different antennas. For the next antenna, the interference is canceled and the VST-MMSE receiver is updated similarly to original S-MMSE-OSIC receiver.

In [WWC+02b, WWC+02a] VST-ZF and ST-MRC are used as initial stages prior to the partial multistage parallel IC (PIC). The partial PIC, [DSR98], uses an attenuation factor in the cancellation. Attenuation is beneficial, since especially in the early stages the symbol estimates are not necessarily reliable.

4.4 Chip level ST-MMSE MOSIC receiver

Inspired by the inter-antenna interference cancellation scheme of [ABB+03, SWL01] a scheme suitable for HSDPA systems is proposed in Publication VI and further improved in Publication VII. The receiver is denoted as ST-MMSE-OSIC and it combines a ST-MMSE equalizer with a chip level inter-antenna interference cancellation. Furthermore, the receiver with multiple ordered successive cancellation stages (MOSIC) is denoted ST-MMSE-MOSIC.

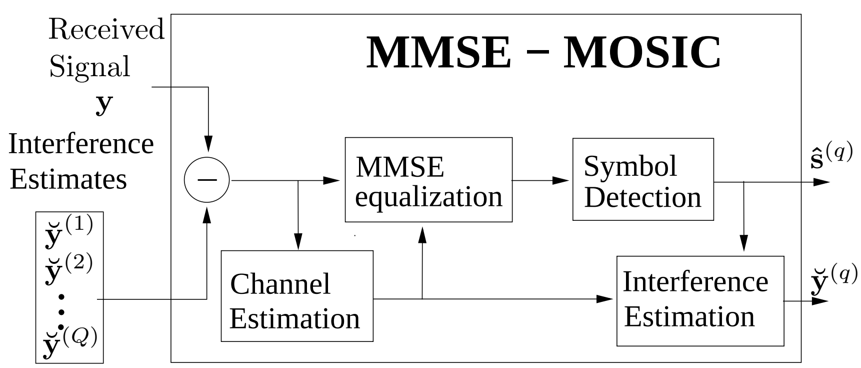

Figure 4.4: Generic block diagram of the MMSE-MOSIC algorithm.

The basic idea of ST-MMSE-MOSIC is straightforward multistage ordered interference cancellation, see Figure 4.4. At the initial stage, a conventional channel estimation, ST-MMSE equalization and symbol detection for the strongest layer are performed. Then, the interference from the strongest layer is canceled prior to estimation and detection of the next layer. The

Table 4.1: Single stage ST-MMSE-OSIC for Q transmit antennas

Initialization:	$\hat{\mathbf{R}}^{(1)} = \hat{\mathbf{R}}, \hat{\mathbf{y}}^{(1)} = \mathbf{y}$, $\hat{\mathbf{h}}^{(1)} = \hat{\mathbf{H}}^{(1)}(:, \tau)$
For all antennas	**for** $q = 1, \ldots, Q$
MV equalizer	$\mathbf{f}^{(q)} = \hat{\mathbf{R}}^{(q)\,-1}\hat{\mathbf{h}}^{(q)}/(\hat{\mathbf{h}}^{(q)H}\hat{\mathbf{R}}^{(q)-1}\hat{\mathbf{h}}^{(q)})$,
Equalization	$\hat{\mathbf{z}}^{(q)} = \mathbf{F}^{(q)H}\hat{\mathbf{y}}^{(q)}$, see equation (4.10)
For all symbols	**for** $n = 1, \ldots, N$
Symbol estimation	$\hat{s}_{np}^{(q)} = \mathcal{D}[\mathbf{c}_{np}^{(q)H}\hat{\mathbf{z}}_n^{(q)}], \ \forall p$
Interference estimation	$\check{\mathbf{y}}_n^{(q)} = \sum_p \mathcal{C}_{(np)}\sqrt{\rho_p^{(q)}}\ \hat{\mathbf{h}}^{(q)}\hat{s}_{np}^{(q)}$
	end
Interference cancellation	$\hat{\mathbf{y}}^{(q+1)} = \hat{\mathbf{y}}^{(q)} - \check{\mathbf{y}}^{(q)}$
Update channel estimate	$\hat{\mathbf{h}}^{(q+1)} = \Phi(\hat{\mathbf{y}}^{(q+1)})$
and covariance matrix	$\hat{\mathbf{R}}^{(q+1)}$ (use eq. (3.6a) or (3.6b) or (4.13))
	end

chip level interference is estimated using the known codes, and the obtained symbol and channel estimates. Consequently, the signal used to detect the symbols for the second antenna, i.e. $\hat{\mathbf{y}}^{(2)} = \mathbf{y} - \check{\mathbf{y}}^{(1)}$, is free (or partially free) of inter-antenna interference. The estimation and interference cancellation steps can be repeated in a successive manner for all Q layers.

After all the layers are estimated once, the procedure can be repeated for all layers. The use of multiple stages is beneficial since now the inter-antenna interference can be canceled simultaneously from all the interfering antennas. The algorithm for the single stage ST-MMSE-OSIC is described in Table 4.1 and for multiple stages in Table 4.2. In Publication VII, a simplified algorithm for 2×2 MIMO system is given. The notation used in the ST-MMSE-MOSIC algorithm follows the notation used in this thesis. Additionally, τ is the delay used for equalizer, $\mathcal{D}$ is used to denote the symbol decision operator and $\Phi(x)$ denotes channel estimator. Similar idea of combining MMSE equalization and one stage interference cancellation was independently depicted in [HMF$^+$03]. However no derivations are given.

Just as in section 3.3.2, the signal covariance matrix can be formed at the initial stage in two alternative ways: using the sample covariance matrix or the estimated channel, see equations (3.6a) and (3.6b) (or (4.9)) respectively. These two methods can also be used to update the covariance matrix, using updated signal or updated channel and power estimates. Alternatively, the

Table 4.2: The multistage ST-MMSE-MOSIC for Q transmit antennas

First stage	Apply ST-MMSE-OSIC
For all stages	**for** $w = 1, \ldots, W$
For all TX antennas	**for** $q = 1, \ldots, Q$
Interference cancellation	$\hat{\mathbf{y}}^{(q)} = \mathbf{y} - \sum_{r \neq q} \breve{\mathbf{y}}^{(r)}$
Update channel estimate	$\mathbf{h}^{(q)} = \Phi(\hat{\mathbf{y}}^{(q)})$
Update covariance matrix	$\hat{\mathbf{R}}^{(q)}$, (use eq. (3.6a) or (3.6b) or (4.13))
Update ST-equalizer	$\mathbf{f}^{(q)} = \hat{\mathbf{R}}^{(q)\,-1}\hat{\mathbf{h}}^{(q)}/(\hat{\mathbf{h}}^{(q)H}\hat{\mathbf{R}}^{(q)-1}\hat{\mathbf{h}}^{(q)})$,
Equalization	$\mathbf{z}^{(q)} = \mathbf{F}^{(q)H}\mathbf{y}^{(q)}$, see equation (4.10)
For all symbols	**for** $n = 1, \ldots, N$
Symbol estimation	$\hat{s}_{np}^{(q)} = \mathcal{D}[\mathbf{c}_{np}^{(q)H}\hat{\mathbf{z}}_n^{(q)}], \ \forall p$
Interference estimation	$\breve{\mathbf{y}}_n^{(q)} = \sum_p \mathcal{C}_{(np)}\sqrt{\rho_p^{(q)}}\ \hat{\mathbf{h}}^{(q)}\hat{s}_{np}^{(q)}$
	end
	end
	end

following update rule can be used:

$$\hat{\mathbf{R}}_{q+1} = \hat{\mathbf{R}}_q - \sum_{p=1}^{P} \rho_p^{(q)}\hat{\mathbf{H}}^{(q)}\hat{\mathbf{H}}^{(q)H}, \tag{4.13}$$

where $\rho_p^{(q)}$ is the power allocated to the pth code at the qth antenna. For the auto-correlation based method and equation (4.13), very accurate power estimates are needed. To give a more realistic view of the ST-MMSE-MOSIC performance, the conventional sample matrix estimate given in equation (3.6a) is used in Publications VI-VII. The improvement in uncoded BER performance over the ST-MMSE equalizer are clearly seen in both vehicular and pedestrian channels. In the shown examples, $55 - 65$ % of the total transmit power was allocated to known HSDPA codes. The drawback of using ST-MMSE-MOSIC is the increase in the computational complexity.

The performance of the ST-MMSE-OSIC depends on the power allocated to known codes versus the total transmitted power. To illustrate this, a 2×2 MIMO system is assumed in the following. The ST-covariance matrix for the second antenna after the interference from the first antenna has been canceled is:

$$\mathbf{R}_{\breve{y}} = (\sigma^{(1)} - \sigma^{(\tilde{1})})\mathbf{H}^{(1)}\mathbf{H}^{(1)H} + \sigma^{(2)}\mathbf{H}^{(2)}\mathbf{H}^{(2)H} + \mathbf{R}_v, \tag{4.14}$$

where $\sigma^{(1)}$ and $\sigma^{(\tilde{1})}$ are the total transmitted power from transmit antenna 1 and the power used to transmit known codes respectively. With multiple known HSDPA codes, the interference from the interfering antenna can be estimated to the level that the ST-MMSE-OSIC will show some improvements in the BER performance. This means that $(\sigma^{(1)} - \sigma^{(\tilde{1})}) << \sigma^{(1)}$.

Additionally, similarly to other interference cancellation schemes, the performance is also sensitive to errors in the interference estimate used in the cancellation. For example, the estimation error of transmitted symbols is illustrated in Figures 2-4 of Publication VI, where the results are shown for both know HSDPA symbols and symbols estimated after despreading with hard decision (uncoded). These reference curves show that the quality of available symbol estimates has a strong impact on the MMSE-MOSIC performance. In HSDPA systems, further improvement can be obtained by more accurate symbol estimates achieved with channel decoding see [HT04]. Also, the quality of the available channel estimates has great influence on the system performance. Therefore, in Publication VII an enhanced channel estimation method for ST-MMSE-MOSIC has been proposed. This method is introduced also in the next chapter.

4.5 Discussion

In this chapter advanced receiver structures for HSDPA MIMO systems were reviewed. A novel multistage chip level inter-antenna interference canceler was proposed. The new receiver combines a space-time (ST) MMSE equalizer and multistage ordered successive interference canceler. It is denoted ST-MMSE-MOSIC. It provides a viable alternative for HSDPA transmission, when the power allocated to known codes is large enough. Furthermore, when the inter-antenna interference is canceled at chip level, ST-MMSE-MOSIC also improves the ability to suppress inter-cell interference. This is achieved by using multiple receive antennas.

There are several possible future research topics related to the proposed ST-MMSE-MOSIC receiver. First, the trade-offs between the computational complexity and performance with sparse equalization, introduced in chapter 3 of this thesis should be studied. Secondly, since the performance of the ST-MMSE-MOSIC is reduced due to estimation errors, combining of turbo-equalization with ST-MMSE-MOSIC should be considered. Finally, further performance and complexity comparisons with ST-MMSE-MOSIC and hybrid ST-MMSE and multiuser detectors should also be carried out.

Chapter 5

Channel estimation in HSDPA systems

Knowledge of the channel impulse response (CIR) at the receiver plays important role in the receiver design in wireless communication systems. CIR can be used to compensate for the effects of the radio channel and consequently to improve the system performance. In mobile wireless systems, a radio signal may fluctuate randomly while propagating from transmitter to receiver. Due to the physical environment and mobility, i.e. time-varying nature of the channel CIR is difficult to measure. With higher order modulation, such as 16QAM, the closeness of the signal constellation points further increases the sensitivity to errors in CIR estimate. Also advanced receivers required in future systems to meet the capacity and robustness demands are more sensitive to errors in CIR measurements than conventional receivers. Consequently, high quality channel estimates are needed for adequate quality reception in many wireless systems.

In this chapter, novel blind and semi-blind channel estimation methods for HSDPA systems are introduced. These methods are derived and further studied in Publications I-IV and Publication VII. This chapter is organized as follows. First, a literature review of channel estimation methods in WCDMA downlink is given. Then blind channel estimation methods stemming from subspace method of [WF99] are presented. Next blind and semi-blind multi-code principal component (MPC) methods are introduced. Finally, the semi-blind MPC method is extended to MIMO case.

5.1 Overview of channel estimation methods

Channel estimation methods are typically in the literature classified as pilot-based, blind and semi-blind methods, see Figure 5.1. Methods may process the received data in batch mode or in an adaptive manner. However, each communication systems has it's own special design which needs to taken into account in derivation of the channel estimator. In WCDMA systems the aperiodic spreading codes introduce an extra challenge to the channel estimation. Furthermore, the downlink and uplink signals have different structures leading to different interference scenarios. Also HSDPA systems have their own special characteristics which differ from other WCDMA downlink signals.

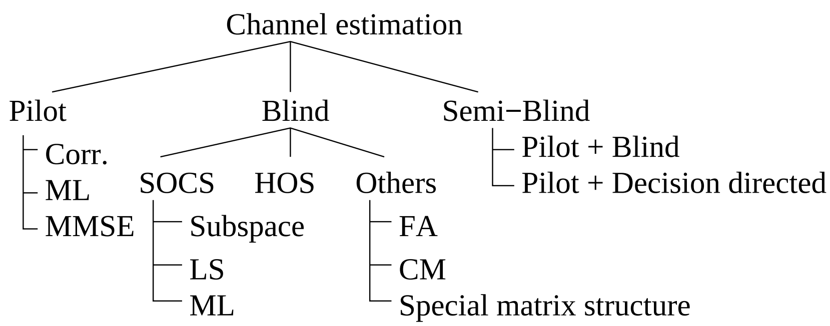

Figure 5.1: Categorisation of channel estimation methods

Channel estimation in most communication standards is based on a known training signal, called pilot signal in this thesis. A pilot signal in CDMA systems is either time multiplexed or code multiplexed with the other transmitted signals. Furthermore, in MIMO systems spatially multiplexed pilots can be used. Majority of studies on channel estimation methods are based on time multiplexed pilot signal. This holds for both non-CDMA and CDMA systems, see e.g. recent a review article of pilot-aided channel estimation [TSD04]. Three classical pilot aided channel estimation approaches are MMSE, ML and correlator, see e.g. [DU05] for further details.

In general, the use of a pilot signal reduces the resources available for user data signal, decreasing the throughput of the system. A trade-off between quality of available channel estimate and amount of resources allocated to pilot signal is an important design criteria in wireless systems. The ultimate goal would be to have high quality channel estimate, without allocating any extra resources to pilot signal and with low computational complexity. In practise, the system designer needs to find a good balance with these three requirements (quality, complexity and used resources).

Due to limited amount of available resources in wireless communications, so called blind channel estimation methods have been extensively studied in literature since Sato's paper [Sat75], see overviews [KLB02, TP98] and the references therein. Blind channel estimation methods are based on some statistical or structural properties of the transmitted signal. These properties include the second order cyclostationarity (SOCS), higher order statistics (HOS), finite alphabet (FA), constant modulus (CM), and special matrix structures arising from the system model. Originally, only the temporal properties were considered but nowadays also the spatial properties are often exploited. Some prior information, such as known symbols, are needed in order to resolve the inherent ambiguity of blind estimates.

Recently there has been growing interest in so called semi-blind channel estimation methods which combine pilot based channel estimation with blind estimation. Semi-blind methods have drawn much attention because they could improve the effective data rates by reducing the number of pilot bits. Alternatively, the enhancement in channel estimation obtained through a semi-blind method could allow usage of e.g. higher order modulation and hence provide increased data rates. The basic idea in semi-blind estimation is to utilize all the information known at the receiver. Consequently, the estimators have larger sampler support, since both pilot and data symbols are used. Most wireless standards include some kind of pilot signal. Therefore semi-blind channel estimation methods seems to be an viable alternative for improving the performance. Decision directed (DD) estimation methods are iterative estimators where estimates of the unknown symbols are used to aid the channel estimation. Typically DD methods are combined with time multiplexed pilot symbols to track fast fading channel coefficients. Consequently, DD methods are classified in the semi-blind estimation methods.

Semi-blind channel estimation has mainly been considered with time multiplexed pilot. The blind methods are used to track the varying channel parameters in fast fading channels during the time period when no pilot is available. A tutorial of semi-blind methods with time multiplexed pilots is given in [GHST01, Ch. 7]. Alternatively, in CDMA systems semi-blind methods can be combined with code multiplexed pilots. However, only little attention have been paid to this problem in the literature so far. Furthermore, proposed methods are mainly intended for uplink, see [AM04].

Next, a literature review of channel estimation in WCDMA systems is given. Then in section 5.2 novel blind methods based on SOCS and special matrix structures applicable to multi-code CDMA systems with long spreading codes are presented. Furthermore, a novel semi-blind combining scheme of a blind and a code multiplexed pilot signal based channel estimation method is introduced in section 5.3. The new method have been originally proposed in publications I-IV and VII.

5.1.1 Channel estimation in WCDMA

Channel estimation in WCDMA systems differs from conventional short code CDMA channel estimation. In WCDMA systems, unlike in short code CDMA systems, the received signal is not bit-interval cyclostationary, [BP00, XT00, GS98]. Therefore, many of the advanced signal processing algorithms proposed for CDMA systems, are not directly applicable or are ineffective to WCDMA. However, with multiple receive antennas (or time-domain oversampling) the chip level cyclostationarity properties may be obtained, see [Gar91]. Second order cyclostationary means that the signal auto-correlation is periodic.

Uplink and downlink WCDMA have different information signal and pilot structures, see [HT04]. Furthermore, the uplink signals are asynchronous, while downlink is synchronous and the fading channel for all users is the same. Consequently, the uplink channel estimation methods are not directly applicable to downlink. Unfortunately, majority of the proposed long-code estimation methods have been designed for the uplink, see e.g. [TEX97, XT00, BP01, EMS01, BA02, AM04, ST04, BP04, FM05] and the references therein.

A dominant feature of WCDMA downlink systems is a strong code-multiplexed pilot signal transmitted to enable channel estimation. The common pilot channel, CPICH, is allocated 10% of the total transmission power. Despite this, a conventional pilot-based channel estimator might not provide accurate enough estimates. In [Yli02] a combined channel estimation scheme with primary and secondary pilots is proposed. Combined channel estimation with CPICH and time multiplexed dedicated channel pilot symbols are proposed, e.g. in [MS03, BMS05]. However, the control channel in HSDPA systems, i.e. high-speed shared control channel (HS-SCCH) does not include any time multiplexed pilot symbols, which could be used to aid channel estimation [HT04]. Therefore, these methods are not applicable to HSDPA systems.

Pilot signal based estimation

Conventional pilot signal based estimation is based on correlation and averaging. In WCDMA systems the despread symbol estimates are correlated with known pilot symbol, s_{nc}, and averaged over N symbols :

$$\hat{\mathbf{h}}_c = \frac{1}{N}\sum_{n=1}^{N} \boldsymbol{\mathcal{C}}_{nc}^H \mathbf{y}_n s_{nc}^* / \rho_c, \tag{5.1}$$

where ρ_c is the symbol power of pilot signal and $\boldsymbol{\mathcal{C}}_{nc}$ is the code convolution matrix.

The interference at the weaker multipath components is more significant than at the strong taps due to inter-path-interference (IPI). Consequently, due to lower SINR, the channel estimates of the weaker taps are less reliable than the strong taps. This holds especially when the taps are separately estimated, as is the case with conventional correlation based estimation.

Maximum likelihood (ML) pilot based channel estimator has been considered for WCDMA downlink in [BP02, WZB04]. Assuming Gaussian noise, ML estimation problem can be formulated as:

$$\hat{\mathbf{h}}_{ML} = \arg\min_{\mathbf{h}} ||\mathbf{y} - \tilde{\mathbf{D}}_c^H \mathbf{h}||^2, \tag{5.2}$$

see e.g. [Kay93]. Here $\tilde{\mathbf{D}}_{\mathbf{c}}$ is the pilot chip matrix of size $N_c \times L$, where N_c is the number of chips in the observation period. $\tilde{\mathbf{D}}_{\mathbf{c}}$ has a block Toeplitz structure, see e.g. [WZB04, DU05]. Now the ML solution, which is also called zero forcing or least squares is [DU05]:

$$\hat{\mathbf{h}}_{ML} = (\tilde{\mathbf{D}}_c^H \mathbf{R}_{in}^{-1} \tilde{\mathbf{D}}_c)^{-1} \tilde{\mathbf{D}}_c^H \mathbf{R}_{in}^{-1} \mathbf{y}. \tag{5.3}$$

where $\mathbf{R}_{in}$ denotes the interference + noise covariance matrix. In [BP02, WZB04] approximation $\mathbf{R}_{in} = \rho_i \mathbf{I}$ has been used for WCDMA downlink. The solution can be further simplified due to the long scrambling code and pilot signal specifications. The spreading code used for pilot is all ones. Consequently, the chip level pilot signal has same auto-correlation properties as the long code which is complex gold code. This means that $\tilde{\mathbf{D}}_c^H \tilde{\mathbf{D}}_c \approx \rho_c \mathbf{I}$, where ρ_c is pilot power. With these assumptions, the ML approach and conventional correlation approaches are the same. In [BP02] a decision directed mode is further used to enhance the ML estimation after initial symbol estimation. Asymptotic performance of ML channel estimation is studied in [MF02] with assumption that observation period is a multiple of the scrambling code length.

Wiener (MMSE) channel estimator using long term channel statistics is not directly applicable to long code WCDMA systems, due to time varying nature of the covariance matrix. Therefore, a modified Wiener filter channel estimation is proposed in [CLC01] with time multiplexed pilot symbols. It is based on assumption that the channel vector is random Gaussian vector with time invariant second order statistics. The proposed scheme is related to blind channel estimation method proposed in [NP94, LZ97]. The modified Wiener solution may be applied to code multiplexed pilot directly. However, it may not be practical in HSDPA systems, since the number of slots needed to average the long term second order statistics is much larger than TTI.

Blind channel estimation

Due to strong pilot signal little attention has been paid to blind channel estimation methods suitable for WCDMA downlink. However, blind methods could be combined with pilot based methods especially in fast fading channels to provide faster convergence. Alternatively, the amount of resources allocated to pilot signal could be reduced.

For WCDMA downlink both symbol level and chip level subspace channel estimation methods have been proposed, see [WF99, XLZ04] and publications I-III. Blind subspace detectors exploit the orthogonality of the signal and noise subspaces. These methods, such as the one proposed in [MDCM95], need only relatively small number of observations. One problem with subspace methods is the complexity of calculating the eigenvalue decomposition of the covariance matrix or singular value decomposition (SVD) of the data matrix. Furthermore, not all channels are identifiable due to an ill-conditioned channel matrix [LXTK96].

In [WF99] a symbol level subspace channel estimation was proposed for synchronous downlink system with assumption that the receiver knows all spreading codes used at the transmitter. A noise subspace channel estimator is obtained from received signal which is first preprocessed with the pseudoinverse of the code matrix. Only the ISI free part of the signal is used. This symbol level channel estimator is designed for single receive antenna. Also pilot aided and decision directed (DD) iterative methods are presented. The DD method is based on assumption that the symbols are ± 1, which is not valid for HSDPA systems.

A chip level subspace channel estimation method for WCDMA downlink has been proposed in [XLZ04]. This method relies on oversampling and the observation that the transmitted chip level signal from desired BS is uncorrelated. The columns of the block Toeplitz channel convolution matrix span the signal subspace of covariance matrix $\mathbf{R}$. With multiple receive antennas (or temporal oversampling) subspace techniques can be used to estimate the channel parameters. This chip level subspace channel estimator is shown to outperform the symbol level subspace method of [WF99].

The special matrix structure of the received signal is exploited in the cross relation method (CRM). This blind channel estimator was originally presented in [XLTK95] and has been extended for long code CDMA systems in [LZ99a] and in [LZ99b] for semi-blind channel estimation. The idea is based on the observation that the signal received and despread in one antenna convolved with the channel impulse of the other antenna is equal to the signal received in the other antenna convolved with channel impulse response of the first antenna. This leads to a linear equation which can be solved e.g. with least squares (LS) method as in [XLTK95] or with recursive LS (RLS) as considered in [LL99a]. The CRM method can also be applied to estimate the channel when transmit diversity is used instead of receiver diversity, [LZ00].

In [LZ97] the principal component channel estimation method of [NP94] was used for long code channel estimation in frequency selective channels. The idea was first used to estimate the array response in frequency flat channels in [SNXP93], where optimum beamformer was constructed based on this array response estimate. The method is based on second order statistics of both the pre- and post-correlation data. After correlation the processing gain strengthens the desired path but the interfering paths yield same covariance after and before [LZ97] despreading. Therefore, the channel can be estimated using the principal eigenvector of the difference matrix between the post- and pre-despreading covariance matrices, see equation (3.6c) and (3.6a) respectively. The obtained method is denoted as in this thesis as principal component (PC) channel estimate.

5.2 Blind channel estimation in multi-code systems

In this section, blind channel estimation methods designed for long code CDMA systems, where multiple spreading codes are known at the receiver, are introduced. They are applicable for both downlink and uplink, i.e. HSDPA and HSUPA. These three methods have been originally proposed in Publications I-III. First, two methods reduce the complexity of the multi-code subspace method proposed in [WF99]. They also provide robustness to code selection and to interfering users. They differ in the preprocessing approach. Also, extension of the method given in [WF99] to multiple receive antenna case is presented, see Publications II and III for further details. After the preprocessing, the channel estimation is obtained using the commonly used noise subspace method [MDCM95]. The third new blind method introduced in this section is a multi-code extension of the blind PC method proposed in [LZ97]. The extension also includes the case where the spreading factors are unequal.

The notation used in Publications I - III is slightly different than the one used in this thesis. Notation was altered to simplify the presentation for the receiver structures introduced in chapters 3 and 4. There are three differences, which will not influence the results. First, the order of stacking the signals received by multiple antennas and multiple paths is altered[1]. Secondly, the summation over codes used in equation (2.4b) is replaced by matrix operations. The multi-code channel matrix is denoted as $\bar{\mathbf{H}} = \mathbf{I}_K \otimes \mathbf{h}$ and a joint code convolution matrix for all K users is denoted as: $\boldsymbol{\mathcal{C}}_n = [\boldsymbol{\mathcal{C}}_{n1}, \ldots, \boldsymbol{\mathcal{C}}_{nK}]$ Consequently the received signal can be written as:

$$\mathbf{y}_n = \boldsymbol{\mathcal{C}}_n \bar{\mathbf{H}} \mathbf{s}_n + \mathrm{ISI} + \mathbf{v}_n, \tag{5.4}$$

where $\mathbf{s}_n$ is the symbol vector for all users (size $K \times 1$). The third difference is that the model in Publications I - III is uplink model, where the interfering users have unique long scrambling codes and channels. Consequently the interference structure is different.

Throughout this thesis a block fading channel model is used. This means that the channel is assumed to be constant during the observation block, i.e. the block duration is less than the coherence time, T_C, of the channel. Each block is assumed to be independent from the previous and following blocks. If number of symbols N needed for the channel estimation is small, also faster fading channels can be tracked. Therefore, a small value of N is desirable from channel estimation point of view.

[1] The model used in Publications I - III leads to block Toeplitz signal covariance matrix.

5.2.1 Preprocessing methods

For symbol level subspace channel estimation the received signal needs to be preprocessed. In this case preprocessing means the algorithm which is used to convert chip level signal to symbol level signal. Three different methods are considered in the following. The first method is a pseudoinverse preprocessing, introduced in [WF99]. This preprocessing method is called decorrelating matched filter, see [TvDS03]. The two alternative methods proposed in Publication III are partial despreading and normal despreading.

In [WF99] only the ISI free part of the signal vector is used. The partial received signal vector, denoted with $\tilde{\mathbf{y}}$ is not deteriorated by previous or following symbols and the disturbance to the desired signal is due to inter path interference (IPI) and noise $\tilde{\mathbf{v}}$. Alternatively, the preprocessing operation may be applied to the whole received signal vector $\mathbf{y}$.

The three different preprocessing methods can be written as follows:

$$\bar{\mathbf{x}}_n = \tilde{\boldsymbol{\mathcal{C}}}_n^{\dagger}\tilde{\mathbf{y}}_n \approx \bar{\mathbf{H}}\mathbf{s}_n + \tilde{\boldsymbol{\mathcal{C}}}_n^{\dagger}\tilde{\mathbf{v}}_n \tag{5.5a}$$

$$\tilde{\mathbf{x}}_n = \tilde{\boldsymbol{\mathcal{C}}}_n^{H}\tilde{\mathbf{y}}_n \approx \bar{\mathbf{H}}\mathbf{s}_n + \tilde{\boldsymbol{\mathcal{C}}}_n^{H}\tilde{\mathbf{v}}_n \tag{5.5b}$$

$$\mathbf{x}_n = \boldsymbol{\mathcal{C}}_n^{H}\mathbf{y}_n \approx \bar{\mathbf{H}}\mathbf{s}_n + \boldsymbol{\mathcal{C}}_n^{H}(\mathrm{ISI} + \mathbf{v}_n), \tag{5.5c}$$

where $\tilde{\boldsymbol{\mathcal{C}}}_n$ is the ISI free part of the joint code convolution matrix of size $M(G-L+1) \times MLK$ and $\dagger$ denotes pseudoinversion. The pseudoinverse method sets limitations to the number of codes that can be used, since the code matrix $\tilde{\boldsymbol{\mathcal{C}}}_n$ needs to be tall and full rank. It also adds computational complexity, since the code matrix in symbol dependent. Consequently, even though the approximation of equation (5.5a) is more accurate than the two other, the pseudoinverse method is less appealing. Furthermore, in HSDPA system, where the spreading factor is only 16, using only the ISI free part can also cause problems, if the channel delay is long.

5.2.2 Multi-code Subspace method

After preprocessing stage, the channel estimate may be obtained using widely used noise subspace method [MDCM95]. In the absence of noise, $\mathbf{x}(n)$'s lie in a subspace spanned by the columns of $\bar{\mathbf{H}}$ (i.e., the latter term may be neglected in equations (5.5)). The preprocessed symbol level signals $\mathbf{x}$ are collected to a $MLK \times N$ size matrix:

$$\mathbf{X}_N = [\mathbf{x}_1, \ \dots \ , \mathbf{x}_N] \approx \bar{\mathbf{H}}\mathbf{S}_N, \tag{5.6}$$

where $\mathbf{S}_N = [\mathbf{s}_1, \ \dots \ , \mathbf{s}_N]$ is the $K \times N$ matrix of N transmitted data symbols and $\mathbf{x}$ is defined as in equations (5.5)(a-c). The signal and noise subspaces can be found by performing the SVD of the preprocessed data matrix. For more detailed description of the subspace method, see Publication III.

The influence of the neglected noise terms on the performance has been studied via simulations in Publications I-III. They show clearly that the pseudoinverse method of [WF99] is more sensitive to both selection of used spreading codes and to interference, than the despreading methods. The pseudoinversion is theoretically optimal, but in practical case it enhances the interference and noise term more. Furthermore, due to aperiodic scrambling code, the pseudoinversion preprocessing is computationally more complex.

5.2.3 Multi-code Principal Component (MPC) method

The PC method proposed in [LZ97] finds a blind channel estimate as the largest eigenvalue of the the difference matrix between the post- and pre-despreading covariance matrices, see equation (3.6c) and (3.6a), respectively. For one code p, the difference matrix is:

$$\hat{\mathbf{R}}_{\Delta} = \hat{\mathbf{R}}_{x_p} - \hat{\mathbf{R}}_y \approx \gamma \mathbf{h}\mathbf{h}^H, \tag{5.7}$$

where γ is a scalar. PC has the same problem as the symbol level equalizer, i.e. the interference and noise term is time varying due to the aperiodic spreading code. Consequently, as stated in [CZR$^+$00] the basic assumptions given for PC method in [SNXP93] is violated. This is easily seen by looking at the despread received symbol expressed as follows:

$$\boldsymbol{\mathcal{C}}_{np}\mathbf{y} = \boldsymbol{\mathcal{C}}_{np}^H \boldsymbol{\mathcal{C}}_{np}\mathbf{h}\ \sqrt{\rho_p} s_{np} + \boldsymbol{\mathcal{C}}_{np}^H(\mathrm{ISI} + \sum_{r \neq p}^{P} \mathcal{C}_{nr}^H \mathbf{h}\ \ \sqrt{\rho_r} s_{nr} + \mathbf{v}_n).$$

However, since in WCDMA systems the system load is typically high, the time dependency is less significant. For single code systems the long code assumption would be critical. With short orthogonal spreading codes, if all the codes and equal symbol powers are used, the time varying term is negligible. Furthermore, if multiple spreading codes are known at the receiver, the time dependency can be reduced and the quality of PC can be improved. The multi-code PC (MPC) channel estimate can be found as the principal eigenvalue of the following difference matrix:

$$\hat{\mathbf{R}}_{\Delta} = \hat{\mathbf{R}}_P - P\hat{\mathbf{R}}_y \approx \gamma \mathbf{h}\ \mathbf{h}^H, \tag{5.8}$$

where $\hat{\mathbf{R}}_P = \sum_p \hat{\mathbf{R}}_{x_p}$, i.e. is the sum of of post-despreading covariance matrices. The MPC is proposed and further studied in Publications I-IV and VII.

Th scalar γ in equations (5.7) and (5.8) can be interpreted as a quality indicator of the blind MPC channel estimate. After some straightforward math it can be approximated as:

$$\gamma \approx P(\rho_p - \sigma_d), \tag{5.9}$$

where ρ_p is the symbol power allocated to each of the P known codes and σ_d is the total chip power used for transmission. Consequently, γ is influenced by the total number of codes and their powers and spreading factors. In Publication VII γ is plotted for different number of known codes, versus number of interfering users.

In WCDMA downlink the used spreading codes can have different spreading factors. For example, the control signals have different spreading factors than the HSDPA signals. To benefit from all the known codes regardless of their spreading factor, an extension of MPC method is proposed in Publication VII and Publication IV.

In Publication III it was shown via simulations that the MPC method is more robust against interference and noise than the other subspace methods. Consequently, the further studies in Publications IV and VII were carried out only with MPC.

Finding the largest eigenvalue

MPC channel estimate up to complex scaling factor can be found as the largest eigenvector of the difference matrix in equation (5.8). Finding the largest eigenvector of a matrix via SVD is computationally very demanding. Alternatively, it can be found more efficiently using the power method, [Gv96], which works especially well when there is only one distinct large eigenvalue. With MPC this holds in case γ is clearly positive. The use of the power method with MPC channel estimation has been proposed and studied in Publication VII. It is shown that with an initial estimate based on pilot signal, one iteration step is sufficient enough to converge to the solution. Alternatively, fast subspace tracking algorithms could be used to update the channel estimates, see e.g. [XHZGK94, KS99].

5.3 Semi-blind channel estimation

In this section, a semi-blind channel estimator combining blind MPC with pilot-based channel estimation is introduced. This semi-blind (SB) estimator is denoted as SB-MPC. In the following $\hat{\mathbf{h}}_c$ and $\hat{\mathbf{h}}_b$ and $\hat{\mathbf{h}}_{sb}$ denote pilot based, blind and semi-blind channel estimate respectively.

Three different semi-blind combining methods are proposed and tested in simulations in Publications IV and VI. These approaches are :

1. Conventional combining of two estimates

$$\hat{\mathbf{h}}_{sb} = \alpha_c \hat{\mathbf{h}}_c + \alpha_b \hat{\mathbf{h}}_b, \tag{5.10a}$$

2. Blind difference matrix combination

$$\hat{\mathbf{R}}_\Delta = \alpha_P \hat{\mathbf{R}}_{\Delta_P} + \alpha_C \hat{\mathbf{R}}_{\Delta_C} \approx \gamma \mathbf{h}\, \mathbf{h}^H. \tag{5.10b}$$

3. Semi-blind difference matrix combination

$$\hat{\mathbf{R}}_{\Delta} = \alpha_P \hat{\mathbf{R}}_{\Delta_P} + \alpha_c \rho_c \hat{\mathbf{h}}_c \hat{\mathbf{h}}_p^H \approx \gamma \mathbf{h}\,\mathbf{h}^H. \tag{5.10c}$$

Here $\hat{\mathbf{R}}_{\Delta_P}$ and $\hat{\mathbf{R}}_{\Delta_C}$ denote difference matrix estimates with two different spreading factors, estimated with equation (5.8). The combining weights, α_i, where $i \in \{b, c\}$ correspond to weights defined for blind and control signal based (single code) estimates respectively. The indexes $i \in \{P, C\}$ correspond to multi-code estimates with P and C codes. Determining α_i's is not straightforward. However, with the difference matrix combining methods, equations (5.10)(b and c), a logical way to define the weights α_C and α_P is based on the chip powers of the known codes, ρ_p and ρ_c. In practise these powers are not necessarily known, but they can be estimated. An estimator for the relative powers is:

$$\alpha = \frac{\alpha_P}{\alpha_C} = \frac{\rho_p P/G_p}{\rho_C C/G_c} \approx \frac{ML\mu_P - PML + P}{ML\mu_C - CML + C}\,\frac{G_c}{G_p}, \tag{5.11}$$

where P and C are the number of known codes with two different spreading factors G_p and G_c. The number of receive antennas is M and L is channel length. Furthermore, μ_P is the ration between post- and pre-despreading covariance matrices for P codes:

$$\mu_P = \frac{\mathrm{trace}(\hat{\mathbf{R}}_P)}{\mathrm{trace}(\hat{\mathbf{R}}_y)}. \tag{5.12}$$

For codes with another spreading factor μ_C is defined correspondingly. See Publications IV and VI, for further details.

The improvement of SB-MPC over pilot-based channel estimation depends on the systems setting. In general possible gains obtained with semi-blind estimation are reduction of resources allocated to pilot signal, faster convergence or improved quality of the estimates. The simulation results shown in Publications IV and VII illustrate these different type of gains. Furthermore, as discussed earlier, higher order data modulation is more sensitive to channel estimation errors. This is emphasized, for example, in Figure 4 of Publication VII, where the SB-MPC is shown to provide clear gains with 16QAM HSDPA systems with sample support of one slot.

In Publication IV the three semi-blind combining schemes are compared in HSDPA systems with QPSK data modulation. Simulations show clearly, that with the conventional combining approach, see equation (5.10a) a threshold parameter is needed to limit the influence of the blind estimate in cases when γ, see equation (5.9), is small and the quality of the blind estimate is not so good. Consequently, the two other alternatives are preferred. They scale naturally with the value of γ. There are two alternative scenarios when γ is small. First, when the system is fully loaded and secondly when the power allocated to known codes is small.

5.3.1 MPC in MIMO systems

In this section an extension of the semi-blind channel estimation method (SB-MPC) with multiple transmit antennas is presented. It is presented in detail in Publication VII. The method combines SB-MPC with the chip level inter-antenna interference canceler, ST-MMSE-MOSIC, introduced in Publication VI for HSDPA MIMO systems.

The SB-MPC method is based on observation that the difference matrix $\mathbf{R}_\Delta$ in (5.8) is a rank one matrix. With multiple transmit antennas and code reuse this does not hold anymore. Both the post- and pre-despreading matrices contain a summation over the transmit antennas. Consequently, the difference matrix is not rank one, but a sum of Q rank one matrices:

$$\hat{\mathbf{R}}_\Delta = \hat{\mathbf{R}}_P - P\hat{\mathbf{R}}_y \approx \gamma \sum_{q=1}^{Q} \sigma^{(\tilde{q})}\mathbf{h}^{(q)}\, \mathbf{h}^{(q)H}, \tag{5.13}$$

where $\sigma^{(\tilde{q})}$ is the power allocated at qth transmit antenna to codes known at the receiver. In order to restore the rank one property, the hybrid ST-MMSE-MOSIC receiver can be used. In this receiver, after the initial cancellation stage, the signal transmitted from the qth antenna is estimated with $\hat{\mathbf{y}}^{(q)}$, which is free from inter-antenna interference, see Table 4.2. If all codes are not known at the receiver, some residual interference remains. However, with multiple known HSDPA codes the amount of inter-antenna interference is clearly less than in the original signal $\mathbf{y}$. Consequently, the two largest eigenvalues of the difference matrix given in (5.13) are distinct, and the SB-MPC can be used to enhance channel estimation. This means that estimator Φ in Table 4.2 can be evaluated with equation (5.10c) instead of conventional correlator given in equation (5.1). With the single IC stage ST-MMSE-OSIC, the SB-MPC can only be used for the last transmit antenna Q.

In Publication VII the performance improvement of SB-MPC in MIMO is demonstrated in Figure 7 with a simulation in HSDPA system. Due to inter-antenna interference and multiple sources of error in the ST-MMSE-MOSIC receiver, larger values of γ are needed than in SIMO case.

5.4 Discussion

In this chapter channel estimation in WCDMA systems has been addressed. In order to improve the effective data rates of HSDPA systems both blind and semi-blind channel estimation methods have been developed. A novel semi-blind SB-MPC method which benefits from all the known codes regardless of the spreading factor is shown to provide a viable alternative for HSDPA systems, where the receiver knows multiple spreading codes. With

SB-MPC all the data and control signal codes can be used to aid the channel estimation. Due to similarities between the HSDPA and HSUPA, the SB-MPC is also applicable in the uplink.

Due to small amount of research on semi-blind methods in case of code multiplexed pilot signal, there are number of unsolved questions in this area. For example, how to combine other blind methods, such as [XLZ04], with code multiplexed pilot. Furthermore, analytic performance analysis of SB-MPC would be interesting to perform.

Chapter 6

Conclusion

The demand for high data rate mobile systems, such as HSDPA, has recently emerged. When more and more people start to use high data rate services, such as video streaming, the spectral efficiency as well as the quality and coverage of the service need to be improved. For further enhancements of HSDPA, higher order modulation and multiple transmit and receive antennas will be used. These two methods can offer better spectral efficiency and even higher data rates. However, the receivers used in current WCDMA system are not able to the provide the promised increase in bit rates and capacity. Both better quality channel estimates and advanced signal processing techniques are needed in order to benefit from the proposed enhancements. In this thesis advanced receiver structures for high data rate wireless mobile systems have been developed. Three important aspects considered in this thesis are the quality of the available estimates, the complexity of the receiver, and the inter-antenna interference rejection capability in HSDPA MIMO systems with spatial multiplexing.

Joint optimization of space-time (ST) filter coefficients, e.g. ST-MMSE equalization, is known to outperform conventional ST-RAKE receiver with increased computation complexity. In this thesis complexity reduction methods for ST-MMSE equalization are developed. First, a frequency domain matrix inversion method is introduced to lower the computation burden of finding the equalizer coefficients. This approximate solution is shown via simulations to have performance close to the exact time domain inversion method and computational savings are seen especially with long filter lengths. For HSDPA system with low spreading factors, updating the filter coefficients once or twice in a slot is less complex than filtering itself. Since matrix inversion can be efficiently implemented even for long filters, a viable trade-off of receiver complexity and performance is achieved with sparse equalization. As comparison to earlier proposals, the selection for sparse filter taps in this thesis is done only after equalizer coefficients are calculated. Experimental result shown, that this type sparse equalizer provides improved performance over both RAKE and a equalizer derived with sparse channel estimates. These improvements can be achieved with only small increase in computational complexity.

The method used to estimate the signal covariance matrix needed for MMSE equalization has also impact on both the receiver performance as well as complexity. In WCDMA downlink, due to strong pilot signal, channel auto-correlation based covariance matrix estimation is shown to provide improved bit error rates compared to conventional sample covariance matrix estimation based methods. This holds especially with 16QAM modulation, which is more sensitive to estimation errors. Furthermore, the channel auto-correlation based method is clearly computationally less complex than sample covariance matrix estimation. In order to have a fair comparison between the two methods, a simple algorithm is introduced for signal to noise ratio estimation. Close to the cell borders or in case of MIMO systems, more accurate signal to noise ratio estimates are needed than the one used in this thesis.

Spatial multiplexing (SM) MIMO techniques are used to increase the transmission data rate in the future evolution of HSPA systems. When the channel impulse response is not known at the transmitter, the receiver needs to cope with the interference from the other transmit antennas as well from the inter-path interference caused by frequency selective channels. Canceling both inter-antenna and inter-path interference is a challenging task, especially in HSDPA systems where the same spreading codes are reused across the transmit antennas. In this thesis, a novel receiver combining ST-MMSE equalizer and chip level inter-antenna interference cancellation method is introduced. It differs from the earlier methods presented in the literature by performing the interference cancellation (IC) at chip level instead of symbol level. Additionally, the benefits of using multiple IC steps in an ordered manner is evaluated via simulation. The proposed hybrid ST-MMSE-MOSIC receiver is shown to provide performance improvements in HSDPA system, where the receiver knows multiple codes.

The final part of this thesis focuses on blind and semi-blind channel estimation in multi-code CDMA systems. First, enhancements to a blind multi-code estimation method based on subspace techniques are introduced. These approaches are shown to be more robust against interference and to code selection than the original method. Next, a blind principal component (PC) technique is extended to multi-code systems with variable spreading factors. The blind multi-code extension of PC (MPC) method is further combined to a conventional pilot-based channel estimator in a novel way. The proposed semi-blind (SB) MPC method can exploit all the known codes, regardless of different spreading factors and signal powers. The traces of the covariance matrices are used in estimating the combining weights for each code. Furthermore, a low complexity method for finding the SB-MPC channel estimate is introduced, and an extension of the semi-blind method to multi-antenna transmission (MIMO) scenario is proposed. This extension combines the SB-MPC channel estimation method with the ST-MMSE-MOSIC receiver introduced also in this thesis. The performance of SB-MPC has been tested using ST-MMSE equalizer in HSDPA systems with different number of known codes and with different interference scenarios. The BER performance is improved in majority of the test cases. The amount of improvement varies depending on the power allocated to known and unknown codes, the cell load, the sample support and the method used to estimate the signal covariance matrix. With higher order modulation, the performance improvements are more clearly seen. When a cell is fully loaded, or only very small portion of power is allocated to known HSDPA codes, the SB-MPC method is not able to improve the performance.

In this thesis the receiver design for HSDPA systems has been studied from several different viewpoints. As a conclusion, following observations are made. First, using of an sparse equalizer instead of RAKE receiver is highly recommended, since the complexity of equalization can be reduced close to that of RAKE receiver. The equalizer coefficients should be estimated using channel auto-correlation based approach and with low complexity matrix inversion method. The linear filter taps could be sparsely selected corresponding to the strongest equalizer weights and available computational resources. Furthermore, all available knowledge of the transmitted signal should be efficiently utilized at the receiver. For example, the proposed SB-MPC channel estimation method uses HSDPA system specific knowledge of multiple spreading codes in order to improve the channel estimates. Finally, in case of spatial multiplexed HSDPA signals the introduced ST-MMSE-MOSIC receiver could be used to mitigate interference caused due to multi-antenna transmission.

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