Transceiver Design
Over
Space-Time Fading Channels

Tingting Zhang

A thesis submitted for the degree of
Doctor of Philosophy (PhD)
in the Faculty of Engineering of the University of London
and the Diploma of Imperial College (DIC)

April 2008

Communications and Array Processing
Department of Electrical and Electronic Engineering
Imperial College London
University of London
Abstract

The objective of the study is to develop novel space-time transceivers for DS-CDMA (Direct-Sequence Code Division Multiple Access) arrayed MIMO (Multiple-Input Multiple-Output) systems operating in multiple-access multipath fading channels. The ever increasing demand for performance and capacity in cellular wireless systems has prompted the development of DS-CDMA MIMO systems. Unlike other reported research on MIMO systems which often assumes the use of multiple antennas, in this thesis, arrayed MIMO system frameworks are proposed, which, by harnessing the geometrical information of the antenna arrays and received signals, provide a richer description of the channel characteristics and additional degrees of freedom in designing communication systems. Firstly, the framework of MIMO arrayed system in a DS-CDMA system is developed. Subspace type blind channel estimator and receiver based on Doppler Spatial-Temporal Array (Doppler-STAR) are developed for time varying frequency selective fading channels, providing the system with more accurate channel estimation, higher resolution and superior interference cancellation capabilities. Secondly, joint transmitter and receiver optimisation in multipath and multiple access downlink system is studied. The two dimensional Spatio-Temporal Array manifold matrix (STAR matrix) is introduced. An iterative and a closed-form solution are proposed to solve the optimisation problem of minimising the MSE over the entire network. The proposed schemes offer a performance advantage over transmitter only (Tx-beamforming) and receiver only (Rx-beamforming) beamforming techniques under the system framework. Finally, the framework of an OFDM-CDMA arrayed system is developed. With the aid of Circular Spatial-Temporal ARray (C-STAR) manifold vector, two types of receivers (i.e. pre-FFT and post-FFT) are proposed to eliminate the effect of multiple access interference (MAI) and inter-symbol interference (ISI) in a completely blind way.
## Contents

Abstract 2

Contents 3

List of Figures 6

List of Tables 9

Acknowledgements 10

Notation 11

List of Symbols 13

Abbreviations and Acronyms 15

### 1 Introduction

1.1 Multiple-Access Wireless Communications 17

1.2 Antenna Array System and Space Processing 20

1.3 Space-Time Propagation CDMA Channels 22

  1.3.1 Channel Dispersiveness 22

  1.3.2 Parametric Channel Modelling 25

1.4 Space-Time Processing 25

  1.4.1 Space-Time CDMA Receivers 26

  1.4.2 Space-Time CDMA Transmitter 31

1.5 Research Objective and Thesis Organisation 33

### 2 Space-Time Channel Modelling and System Design 36

2.1 Array Manifold Vector 37

2.2 Space-Time Channel Modelling 39

  2.2.1 Scalar-Input Scalar-Output (SISO) Channel 40

  2.2.2 Scalar-Input Vector-Output (SIVO) Channel 41

  2.2.3 Vector-Input Scalar-Output (VISO) Channel 43

  2.2.4 Vector-Input Vector-Output (VIVO) Channel 44

2.3 Space-Time System Framework 47

2.4 Conclusion 51
3 Arrayed MIMO System

3.1 Introduction .................................................. 54
3.2 Arrayed MIMO System Model ................................. 55
  3.2.1 Spread Spectrum Transmission .......................... 55
  3.2.2 Spatial Temporal ARray Manifold Vector (STAR) .... 57
  3.2.3 Doppler-STAR Manifold Vector ......................... 59
  3.2.4 Discrete-Time Signal Model ............................ 60
3.3 Blind Channel Estimation ...................................... 62
  3.3.1 Angle, Delay and Doppler Estimation ................. 62
  3.3.2 Path Power Estimation ................................. 64
3.4 Reception .................................................. 65
  3.4.1 Space-Time Multiuser Receiver ......................... 65
  3.4.2 The Subspace Based Single-User Rx for Interference Can-
cellation ..................................................... 66
  3.4.3 Correlation Analysis Assignment ....................... 69
3.5 Simulation Studies .............................................. 72
3.6 Conclusions ................................................ 82
3.7 Appendix .................................................. 83
  3.7.1 Proof of Eqn. 3.19-3.21: .................................. 83
  3.7.2 Complexity Analysis .................................... 85

4 Joint Transmitter-Receiver Beamforming ....................... 87

4.1 Introduction ................................................ 88
4.2 Downlink Signal Modelling ................................... 89
4.3 Joint Tx-Rx Beamforming .................................... 92
  4.3.1 An Iterative Method Based On Lagrange Multipliers .... 93
  4.3.2 A Closed-Form Solution Based On Channel Eigendecom-
position ..................................................... 94
4.4 Numerical Studies ........................................... 95
4.5 Conclusions ................................................ 101
4.6 Appendix .................................................. 101
  4.6.1 Derivation of the Iterative Solution: .................. 101
  4.6.2 Derivation of the Closed-Form Solution: ............... 103

5 Arrayed OFDM-CDMA System ................................... 105

5.1 Introduction ................................................ 106
5.2 OFDM-CDMA Signal Modelling ............................... 108
5.3 Angle, Delay and Path Fading Coefficient Estimation ..... 114
  5.3.1 Angle and Delay Estimation ......................... 114
  5.3.2 Estimation of Complex Fading Coefficients .......... 115
5.4 Blind Reception for Interference Suppression .............. 117
  5.4.1 Post-FFT Signal Reception ......................... 117
  5.4.2 Pre-FFT Signal Reception ......................... 120
5.5 Simulation Studies ........................................ 122
5.6 Conclusion ................................................ 126
6 Conclusions and Further Work
  6.1 Thesis Summary .............................................. 129
  6.2 List of Contributions ........................................ 131
  6.3 Suggestions for Future Work ................................. 133

References ......................................................... 137
## List of Figures

<table>
<thead>
<tr>
<th>Figure</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.1</td>
<td>CDMA classification</td>
</tr>
<tr>
<td>1.2</td>
<td>A beamformer structure</td>
</tr>
<tr>
<td>1.3</td>
<td>Multipath propagation environment</td>
</tr>
<tr>
<td>1.4</td>
<td>Types of fadings experienced by a CDMA signal due to (a) delay spread and (b) Doppler spread.</td>
</tr>
<tr>
<td>1.5</td>
<td>Classification of space-time CDMA receivers</td>
</tr>
<tr>
<td>1.6</td>
<td>Decoupled space-time processing</td>
</tr>
<tr>
<td>1.7</td>
<td>Joint space-time processing</td>
</tr>
<tr>
<td>2.1</td>
<td>Planewave propagation model</td>
</tr>
<tr>
<td>2.2</td>
<td>Channel Classification based on scalar input/output and vector inputs/outputs of the channel</td>
</tr>
<tr>
<td>2.3</td>
<td>Scalar-Input Scalar-Output (SISO) channel model</td>
</tr>
<tr>
<td>2.4</td>
<td>Scalar-Input Vector-Output (SIVO) channel model</td>
</tr>
<tr>
<td>2.5</td>
<td>Scalar-Input Vector-Output (SIVO) channel model with Doppler frequency shift</td>
</tr>
<tr>
<td>2.6</td>
<td>Vector-Input Scalar-Output (VISO) channel model</td>
</tr>
<tr>
<td>2.7</td>
<td>Vector-Input Vector-Output (VIVO) channel model with CSI at the transmitter</td>
</tr>
<tr>
<td>2.8</td>
<td>Vector-Input Vector-Output (VIVO) Channel Model without CSI at the transmitter</td>
</tr>
<tr>
<td>2.9</td>
<td>Block diagram of the space-time system model as used in Chapter 3. Also note $\mathcal{NK}_1 = \mathcal{K}_1$ and the notations will be defined in Chapter 3.</td>
</tr>
<tr>
<td>2.10</td>
<td>Block diagram of the space-time system model as used in Chapter 4</td>
</tr>
<tr>
<td>2.11</td>
<td>Block diagram of the space-time system model as used in Chapter 5</td>
</tr>
<tr>
<td>3.1</td>
<td>The tapped-delay lines (TDL) structure</td>
</tr>
<tr>
<td>3.2</td>
<td>Data flow diagram of the proposed space-time single user detector</td>
</tr>
<tr>
<td>3.3</td>
<td>An illustration of the Correlation Analysis Assignment. Paths belonging to different antennas are represented using different colours.</td>
</tr>
</tbody>
</table>
3.4 (a) 3D surface plot and (b) 2D contour plot of the joint angle (DOA) and delay (TOA) estimation of all the multipath components of the desired user based on Eqn. 3.20. Taking position $k = 3$ as an example, it represents the estimates of the desired user’s 3rd path of the 1st antenna that arrives from direction 70° with delay $25T_c$ as can be found in Table 3.1.  

3.5 Doppler frequency shift estimation of all the multipath components due to the desired user based on Eqn. 3.22. For example, peak $k = 10$ at $f = -120$Hz is the estimated Doppler frequency of the desired user’s 10th path.  

3.6 The performance comparison of (i) the conventional MUSIC and (ii) the STAR-MUSIC algorithms in resolving closely spaced sources (70° and 70.1°) with identical delays ($5T_c$) and without Doppler frequency shift.  

3.7 Fading coefficient estimation of (Antenna 1, Path 3) of the desired user based on Eqn. 3.25. The deep drop corresponds to $|\beta_{13}|^2$ as listed in Table 3.1. Searching step size is set to be 0.0001.  

3.8 The standard deviation of the channel estimates to the true channel vector in the presence of noise as the searching resolution improves. The composite channel estimation method proposed in [60] is also plotted for comparison.  

3.9 The comparison of the effectiveness of the proposed receiver in combating Doppler frequency shift as compared with i) Receiver proposed in [41]; ii) Doppler-STAR decorrelating receiver; iii) Doppler-STAR RAKE receiver; iv) STAR decorrelating receiver; iv) STAR RAKE receiver.  

3.10 The behaviour of the proposed receiver is investigated as the total number of multipaths increases. PN codes of length $N_c = 31$ and a linear array of $N = 5$ are used in the simulation which make the dimension of the observation space $2N_cN = 310$. ISR is kept constant at -20dB for a fair comparison. The performance of the proposed receiver degrades sharply when the number of multipaths exceeds the dimension of the overall space.  

3.11 The SNIR performance of the proposed receiver in the presence of near-far problem as compared with i) Receiver proposed in [41]; ii) Doppler-STAR decorrelating receiver and iii) Doppler-STAR RAKE receiver.  

4.1 The Vector-Input Vector-Output (VIVO) slow fading channel.  

4.2 Convergence study for the iterative method in terms of total MSE with input $E_b/N_0 = 0, 5, 10$dB. It is bounded by the closed-form solution.  

4.3 Convergence study for the iterative method in terms of output SNIR with input $E_b/N_0 = 0, 5, 10$dB. It is bounded by the closed-form solution.
4.4 Overall MSE performance of proposed methods in systems with varied number of users \((M = 5, 15, 31)\).  

4.5 Output SNIR performance of proposed methods in systems with varied number of users \((M = 5, 15, 31)\).  

4.6 BER performance of joint Tx-Rx beamforming, Tx-beamforming and Rx-beamforming as different number of transmitting antennas are employed \((\bar{N} = 1, 4, 8)\).  

4.7 Output SNIR performance of joint Tx-Rx beamforming, Tx-beamforming and Rx-beamforming as different number of transmitting antennas are employed \((\bar{N} = 1, 4, 8)\).  

4.8 BER performance of joint Tx-Rx beamforming and Rx-beamforming as the number of receiver elements are chosen as \(N = 2, 4, 6\).  

4.9 Output SNIR performance of joint Tx-Rx beamforming and Rx-beamforming as the number of receiver elements are chosen as \(N = 2, 4, 6\).  

5.1 Block diagram of the OFDM-CDMA transmitter.  

5.2 The circulant structure of received symbols and the intersymbol interferences (after CP removal).  

5.3 Block diagram of the post-DFT equalizer.  

5.4 Block diagram of the subspace based pre-DFT channel equalizer.  

5.5 (a) Surface plot and (b) Contour plot of the space-time estimation of all the desired user’s multipath.  

5.6 Output SNIR performance versus input SNR with the employment of (i) post-FFT receiver, (ii) pre-FFT receiver and (iii) STAR MMSE receiver and (iv) STAR RAKE receiver respectively.  

5.7 Output SNIR performance versus input SNR as the number of subcarriers increases.  

5.8 BER performance versus input SNR as the number of subcarriers increases.  

5.9 Output SNIR performance versus input SNR as the number of antenna elements increases.  

5.10 BER performance versus input SNR as the number of antenna elements increases.
List of Tables

3.1 User’s parameters ........................................ 73
3.2 Estimated set and correlation analysis .................. 78

5.1 User’s parameters ........................................ 124
5.2 Complex fading coefficients estimation of the desired user .... 124
Acknowledgements

This thesis is the result of three and half years of research work at Imperial College London whereby I have been accompanied and supported by many people. It is a pleasure that I have now the opportunity to express my gratitude for all of them who gave me the possibility to complete this thesis.

First of all, I owe my most sincere gratitude to my supervisor Prof. A. Manikas who introduced me to the field of array digital communications and gave me the most important guidance throughout my PhD studies. His wide knowledge and insightful thinking have been of great value for me. I would have never reached new heights without his stimulating suggestions and great encouragement in all the time of the research and writing of this thesis.

Secondly, I wish to express my warm thanks to my colleagues in the communications and signal processing group whom I have worked with for the last few years. Special thanks go to Dr. Jason W. P. Ng for his enormous guidance during my first steps into array processing studies. I also owe my thanks to my lab mates Xiaolei Wang, George Elissaios, George Efstatopoulos, Azibananye Mengot, Farrukh Rashid and Victor Li for enormous technical discussions and support that I have received and also for providing me such a pleasant and enjoyable working environment to conduct my research. I am also thankful to Dr. Andy Khong, Dr. Xiao Huang, Dr. Shujun Chen and Dr. Zhiguo Ding for sharing with me their perceptiveness and knowledge on diverse areas of wireless communications.

Thirdly, I am indeed grateful to all my friends for being the surrogate family during the years I stayed in London and for always standing by me in good and bad times. Their moral support and encouragement has made my experience more enriching and valuable than I expected it to be.

Lastly, and most importantly, I wish to thank my parents. They bore me, raised me, supported me, taught me, and loved me. To them I dedicate this thesis.
### Notation

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>$a, A$</td>
<td>Scalar</td>
</tr>
<tr>
<td>$\mathbf{A}$</td>
<td>Column vector</td>
</tr>
<tr>
<td>$\mathbf{A}$</td>
<td>Matrix</td>
</tr>
<tr>
<td>$\mathbf{A}^b$</td>
<td>Element by element power</td>
</tr>
<tr>
<td>$\mathbf{0}_N$</td>
<td>Zero vector of $N$ elements</td>
</tr>
<tr>
<td>$\mathbf{0}_N$</td>
<td>Zero matrix of $N \times N$ dimension</td>
</tr>
<tr>
<td>$\mathbf{1}_N$</td>
<td>$N \times 1$ column vector of ones</td>
</tr>
<tr>
<td>$\mathbf{I}_N$</td>
<td>$N \times N$ identity matrix</td>
</tr>
<tr>
<td>$\mathbf{P}_A$</td>
<td>Projection matrix onto the subspace of $\mathbf{A}$</td>
</tr>
<tr>
<td>$\mathbf{P}_A^\perp$</td>
<td>Orthogonal projection matrix onto the subspace of $\mathbf{A}$</td>
</tr>
<tr>
<td>$\text{diag}(\mathbf{A})$</td>
<td>Diagonalisation of $\mathbf{A}$</td>
</tr>
<tr>
<td>$\text{diag}(\mathbf{A})$</td>
<td>A column vector formed from the main diagonal of $\mathbf{A}$</td>
</tr>
<tr>
<td>$\exp(\mathbf{A})$</td>
<td>Element by element exponential of $\mathbf{A}$</td>
</tr>
<tr>
<td>$\text{trace}(\mathbf{A})$</td>
<td>Trace of $\mathbf{A}$</td>
</tr>
<tr>
<td>$\text{vec}(\mathbf{A})$</td>
<td>Column-wise vectorisation of $\mathbf{A}$</td>
</tr>
<tr>
<td>$\text{sign}(a)$</td>
<td>Signum function</td>
</tr>
<tr>
<td>$\text{eig}_i(\mathbf{A})$</td>
<td>The $i^{th}$ eigenvalue of $\mathbf{A}$</td>
</tr>
<tr>
<td>$(\cdot)^T$</td>
<td>Transpose</td>
</tr>
<tr>
<td>$(\cdot)^H$</td>
<td>Conjugate transpose (Hermitian)</td>
</tr>
<tr>
<td>$(\cdot)^*$</td>
<td>Complex conjugate</td>
</tr>
<tr>
<td>$(\cdot)\dagger$</td>
<td>Penrose Moore pseudo-inverse</td>
</tr>
<tr>
<td>$[\cdot]$</td>
<td>Round up to integer</td>
</tr>
<tr>
<td>$</td>
<td>\cdot</td>
</tr>
<tr>
<td>$|\cdot|_F$</td>
<td>Frobenius norm</td>
</tr>
<tr>
<td>$\odot$</td>
<td>Hadamard (schur) product</td>
</tr>
<tr>
<td>$\odot$</td>
<td>Hadamard (elementwise) division</td>
</tr>
<tr>
<td>$\otimes$</td>
<td>Kronecker product</td>
</tr>
<tr>
<td>$\mathcal{L}[\mathbf{A}]$</td>
<td>Subspace spanned by columns of $\mathbf{A}$</td>
</tr>
<tr>
<td>Notation</td>
<td>Description</td>
</tr>
<tr>
<td>----------</td>
<td>-------------</td>
</tr>
<tr>
<td>$\mathcal{E}{\cdot}$</td>
<td>Expectation operator</td>
</tr>
<tr>
<td>$\mathbb{C}$</td>
<td>Set of complex numbers</td>
</tr>
<tr>
<td>$\mathbb{N}$</td>
<td>Set of natural numbers</td>
</tr>
<tr>
<td>$\mathbb{R}$</td>
<td>Set of real numbers</td>
</tr>
<tr>
<td>$\mathbb{Z}$</td>
<td>Set of integers</td>
</tr>
<tr>
<td>$\Omega$</td>
<td>Parameter space</td>
</tr>
<tr>
<td>$\mathcal{M}$</td>
<td>Array manifold</td>
</tr>
</tbody>
</table>
## List of Symbols

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>$a[n]$</td>
<td>Data symbol associated with the $n^{th}$ interval</td>
</tr>
<tr>
<td>$c$</td>
<td>Speed of light</td>
</tr>
<tr>
<td>$c_{PN,i}(t)$</td>
<td>Pseudo-noise spreading waveform of the $i^{th}$ user</td>
</tr>
<tr>
<td>$\mathbb{D}_s$</td>
<td>Diagonal matrix of signal eigenvalues</td>
</tr>
<tr>
<td>$\mathbb{E}_n$</td>
<td>Matrix with columns containing the noise eigenvectors</td>
</tr>
<tr>
<td>$\mathbb{E}_{\text{unwanted}}$</td>
<td>Matrix with columns eigenvectors of the unwanted signal</td>
</tr>
<tr>
<td>$\mathbb{E}_s$</td>
<td>Matrix containing columns of the signal eigenvectors</td>
</tr>
<tr>
<td>$f_{ik}$</td>
<td>Doppler frequency shift of the $k^{th}$ path of the $i^{th}$ user</td>
</tr>
<tr>
<td>$F_c$</td>
<td>Carrier frequency</td>
</tr>
<tr>
<td>$\mathbb{F}$</td>
<td>Inverse DFT matrix</td>
</tr>
<tr>
<td>$\mathbb{F}^{-1}$</td>
<td>DFT matrix</td>
</tr>
<tr>
<td>$\mathbf{h}_{ik}$</td>
<td>Doppler-STAR manifold vector of the $k^{th}$ path of the $i^{th}$ user</td>
</tr>
<tr>
<td>$\mathbb{H}_i[n]$</td>
<td>Channel matrix of the $i^{th}$ user</td>
</tr>
<tr>
<td>$\mathbb{H}_i^{\text{prev}}[n]$</td>
<td>Channel matrix associated with the previous symbols</td>
</tr>
<tr>
<td>$\mathbb{H}_i^{\text{next}}[n]$</td>
<td>Channel matrix associated with the next symbols</td>
</tr>
<tr>
<td>$\mathbb{J}_c$</td>
<td>Circular shifting operator</td>
</tr>
<tr>
<td>$\mathbb{J}$</td>
<td>Shifting operator</td>
</tr>
<tr>
<td>$\mathbf{k}_{ik}$</td>
<td>Wavenumber vector of the $k^{th}$ path of the $i^{th}$ user</td>
</tr>
<tr>
<td>$K$</td>
<td>Number of multipaths</td>
</tr>
<tr>
<td>$l_{ik}$</td>
<td>Discrete path delay of the $k^{th}$ path of the $i^{th}$ user</td>
</tr>
<tr>
<td>$L$</td>
<td>Number of snapshots (observation interval)</td>
</tr>
<tr>
<td>$L_{cp}$</td>
<td>Length of cyclic prefix</td>
</tr>
<tr>
<td>$m(t)$</td>
<td>Baseband modulated signal</td>
</tr>
<tr>
<td>$M$</td>
<td>Number of users</td>
</tr>
<tr>
<td>$\mathbf{n}(t)$</td>
<td>Complex noise vector</td>
</tr>
<tr>
<td>$\mathbf{n}[n]$</td>
<td>Discrete complex noise vector associated with the $n^{th}$ interval</td>
</tr>
<tr>
<td>$N$</td>
<td>Number of receive antenna elements</td>
</tr>
<tr>
<td>$\mathbb{N}$</td>
<td>Number of transmit antenna elements</td>
</tr>
</tbody>
</table>
List of Symbols

$N_{sc}$ Number of subcarriers
$N_c$ Processing gain
$P_i$ Transmission power of the $i^{th}$ user
$\mathbb{P}_{\text{unwanted}}$ Projection operator on the orthogonal complement subspace of unwanted signals
$[r_x, r_y, r_z]$ Cartesian coordinates of antenna elements
$\mathbb{R}_{xx}$ Covariance matrix of received signals
$\mathbb{R}_{nn}$ Noise covariance matrix
$S_{ik}$ Array manifold vector of the $k^{th}$ path of the $i^{th}$ user
$T_{cp}$ Duration of cyclic prefix
$T_{cs}$ Channel symbol duration
$T_s$ Sampling period
$\mathbb{W}$ Receiver beamforming matrix
$\mathbb{W}$ Transmit beamforming matrix
$x[n]$ Discrete baseband signal
$\alpha_i$ A vector of $\pm 1$s representing one period of the $i^{th}$ user’s spreading code
$\beta_{ik}$ Complex fading coefficient of the $k^{th}$ path of the $i^{th}$ user
$(\theta, \varphi)$ (Azimuth, elevation) angle
$\lambda_i$ Lagrange multiplier
$\xi$ Cost function
$\Xi_i$ Estimates set of the $i^{th}$ user
$\sigma_n^2$ Noise power
$\tau_{ik}$ Time delay of the $k^{th}$ path of the $i^{th}$ user
<table>
<thead>
<tr>
<th>Abbreviation</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>1D/2D/3D</td>
<td>One/ Two/ Three-Dimensional</td>
</tr>
<tr>
<td>1G/2G/3G/4G</td>
<td>First/ Second/ Third/ Fourth Generation</td>
</tr>
<tr>
<td>AWGN</td>
<td>Additive White Gaussian Noise</td>
</tr>
<tr>
<td>BER</td>
<td>Bit Error Rate</td>
</tr>
<tr>
<td>BPSK</td>
<td>Binary Phase Shift Keying</td>
</tr>
<tr>
<td>BS</td>
<td>Base Station</td>
</tr>
<tr>
<td>CCI</td>
<td>Co-Channel Interference</td>
</tr>
<tr>
<td>CDMA</td>
<td>Code Division Multiple Access</td>
</tr>
<tr>
<td>CP</td>
<td>Cyclic Prefix</td>
</tr>
<tr>
<td>CSI</td>
<td>Channel State Information</td>
</tr>
<tr>
<td>DBPSK</td>
<td>Differential Binary Phase Shift Keying</td>
</tr>
<tr>
<td>DQPSK</td>
<td>Differential Quadrature Phase Shift Keying</td>
</tr>
<tr>
<td>DFT</td>
<td>Discrete Fourier Transform</td>
</tr>
<tr>
<td>DOA</td>
<td>Direction of Arrival</td>
</tr>
<tr>
<td>DOD</td>
<td>Direction of Departure</td>
</tr>
<tr>
<td>Doppler-STAR</td>
<td>Doppler-Spatial-Temporal ARray</td>
</tr>
<tr>
<td>DS-CDMA</td>
<td>Direct-Sequence Code Division Multiple Access</td>
</tr>
<tr>
<td>EGC</td>
<td>Equal Gain Combining</td>
</tr>
<tr>
<td>FDD</td>
<td>Frequency Division Duplex</td>
</tr>
<tr>
<td>FDMA</td>
<td>Frequency Division Multiple Access</td>
</tr>
<tr>
<td>FFT</td>
<td>Fast Fourier Transform</td>
</tr>
<tr>
<td>GPRS</td>
<td>General Packet Radio Service</td>
</tr>
<tr>
<td>GSM</td>
<td>Group Speciale Mobile</td>
</tr>
<tr>
<td>IDFT</td>
<td>Discrete Fourier Transform</td>
</tr>
<tr>
<td>IFFT</td>
<td>Inverse Fast Fourier Transform</td>
</tr>
<tr>
<td>IS-95</td>
<td>Interim Standard for U.S. Code Division Multiple Access</td>
</tr>
<tr>
<td>ISI</td>
<td>Intersymbol Interference</td>
</tr>
<tr>
<td>JADE</td>
<td>Joint Angle and Delay Estimation</td>
</tr>
<tr>
<td>Abbreviation</td>
<td>Full Form</td>
</tr>
<tr>
<td>--------------</td>
<td>-----------</td>
</tr>
<tr>
<td>MAI</td>
<td>Multiple Access Interference</td>
</tr>
<tr>
<td>MC-CDMA</td>
<td>Multi-Carrier Code Division Multiple Access</td>
</tr>
<tr>
<td>MC-DS-CDMA</td>
<td>Multi-Carrier Direct Sequence Code Division Multiple Access</td>
</tr>
<tr>
<td>MIMO</td>
<td>Multiple Input Multiple Output</td>
</tr>
<tr>
<td>ML</td>
<td>Maximum Likelihood</td>
</tr>
<tr>
<td>MLSE</td>
<td>Maximum Likelihood Sequence Estimator</td>
</tr>
<tr>
<td>MMSE</td>
<td>Minimum Mean Square Error</td>
</tr>
<tr>
<td>MRC</td>
<td>Maximal Ratio Combining</td>
</tr>
<tr>
<td>MS</td>
<td>Mobile Station</td>
</tr>
<tr>
<td>MSE</td>
<td>Mean Squared Error</td>
</tr>
<tr>
<td>MMSE</td>
<td>Minimum Mean Squared Error</td>
</tr>
<tr>
<td>MUSIC</td>
<td>MUltiple SIgnal Classi…</td>
</tr>
<tr>
<td>NFR</td>
<td>Near-Far Ratio</td>
</tr>
<tr>
<td>OFDM</td>
<td>Orthogonal Frequency Division Multiplexing</td>
</tr>
<tr>
<td>PN</td>
<td>Pseudo Noise</td>
</tr>
<tr>
<td>RMSE</td>
<td>Root Mean Squared Error</td>
</tr>
<tr>
<td>Rx</td>
<td>Receiver</td>
</tr>
<tr>
<td>SIR</td>
<td>Signal-to-Interference Ratio</td>
</tr>
<tr>
<td>SIVO</td>
<td>Scalar Input Vector Output</td>
</tr>
<tr>
<td>SNIR</td>
<td>Signal-to-Noise-plus-Interference Ratio</td>
</tr>
<tr>
<td>SNR</td>
<td>Signal-to-Noise Ratio</td>
</tr>
<tr>
<td>STAR</td>
<td>Spatial-Temporal ARray</td>
</tr>
<tr>
<td>STBC</td>
<td>Space-Time Block Code</td>
</tr>
<tr>
<td>TDD</td>
<td>Time Division Duplex</td>
</tr>
<tr>
<td>TDL</td>
<td>Tapped Delay Line</td>
</tr>
<tr>
<td>TOA</td>
<td>Time of Arrival</td>
</tr>
<tr>
<td>Tx</td>
<td>Transmitter</td>
</tr>
<tr>
<td>ULA</td>
<td>Uniform Linear Array</td>
</tr>
<tr>
<td>UMTS</td>
<td>Universal Mobile Telecommunications System</td>
</tr>
<tr>
<td>VIVO</td>
<td>Vector Input Vector Output</td>
</tr>
<tr>
<td>WCDMA</td>
<td>Wideband Code Division Multiple Access</td>
</tr>
</tbody>
</table>
Chapter 1

Introduction

1.1 Multiple-Access Wireless Communications

The recent and anticipated growth of wireless communication systems has fueled research efforts investigating methods to increase system capacity. Doubtlessly, the exploding demand for efficient high-quality services of digital wireless communications, on the one hand, and the rapid advancement in modern digital signal processing, on the other hand, play a dramatic role in this trend. One of the main issues involved in the development of the next generation system is the choice of multiple access (MA) technology to efficiently share the available scarce bandwidth among a large number of users. Multiple-access communication is key for a wireless system to be commercially viable. As a consequence of the conflict between the increasing demand for wireless services and the extremely scarce spectrum, spectrally efficient modulation and multiple-access techniques are highly demanded for the wireless communication systems designs. Among many multiple-access techniques, the three most important techniques that encompass the evolution of almost all wireless mobile systems around the world are

- **Frequency division multiple access (FDMA):** The total frequency spectrum is divided into a number of frequency channels where each channel occupies a portion of the total available bandwidth and is given to a single user. Multiple users using separate frequency channels could access the system simultaneously without significant interference from other users. It is the simplest way of having MA in a multi-user system.

- **Time division multiple access (TDMA):** The time axis is divided into a
number of time slots, each occupied by a single user to transmit data information. Data from a single user is assigned to the same time slot of a data frame and all the data from that slot are collected and aggregated at the receiver to form the original information data.

- **Code division multiple access (CDMA):** An information-bearing signal is transformed into a transmission signal with a much larger bandwidth by encoding the information signal with a unique pseudo random (PN) code sequence that has a much larger spectral width than the data signal. The receiver, with the knowledge of the modulating code signal, then correlates the received signal with a synchronously generated replica of the code signal to recover the original information-bearing signal.

The first generation (1G) mobile systems, introduced in 1980s, is characterised by using the FDMA technique to support multiuser communications. The 1G systems are primarily analog systems for voice services, such as the advanced mobile phone service (AMPS) in the United States. The advancement of real time digital signal processing led to the development of the digital mobile communication system based on TDMA in the late 1980s, known as the second generation (2G) communication systems [1]. To remain compatible with the spectrum allocation already in existence in the 1G system, most of the 2G systems, such as Groupe Speciale Mobile (GSM), use FDMA in conjunction with TDMA to improve system capacity. Compared to 1G systems, 2G systems not only offer speech but also low bit rate data service with higher spectrum efficiency.

A major step improvement came later with the development of 3G and beyond 3G (B3G) systems which provide higher bit rates for data transmission and more flexible multimedia services satisfying different quality requirements, for example voice, video, data or a combination of different multimedia types [2]. 3G aims to provide packet data connections of 144kbps for high mobility, 384kbps with restricted mobility, and 2Mbps in an indoor office environment. In B3G mobile systems, higher data rate transmission on the order of 10Mbps and more is expected [3]. CDMA as a means of multiple access communications has formed the basis for the third generation (3G) systems and is essentially capable of becoming the MA technique for B3G. Several 3G system standards have been proposed such as WCDMA (Universal Mobile Telecommunications System (UMTS)) [4], CDMA2000 [5] and time duplex synchronous CDMA (TD-SCDMA) [6].

As a spread-spectrum based technique, CDMA can be classified into four categories shown in Fig. 1.1: direct sequence (DS), frequency hopping (FH),
time hopping (TH) protocols and hybrid CDMA. DS-CDMA systems average interference across the entire signalling bandwidth, whereas TH-CDMA and FH-CDMA attempt to avoid interference. Hybrid CDMA techniques also emerged as combinations of the former three types. The work presented in this thesis is concentrated with DS-CDMA which has been selected for application within 3G standards around the world.

CDMA schemes have some impressive advantages over TDMA and FDMA schemes. The orthogonality of the set of spread spectrum code signatures gives the CDMA systems their multiple access capability [7]. All users in a CDMA network share the entire available bandwidth resource and each user is distinguished from the others by its unique pseudo noise (PN) sequence. Multipath diversity of the channel can be exploited because paths that arrive with delay differences greater than one chip interval can be separately resolved, due to the autocorrelation and crosscorrelation properties of the PN codes. The RAKE receiver can be used to exploit the multipath diversity and improve reception by collecting delayed copies of the required signals. However, a major problem in CDMA is the interference from other users, known as Multiple Access Interference (MAI). The quasi-orthogonality between different users’ spreading sequences causes unwanted signals from other users to remain in the desired user’s signal and the near-far problem occurs if the undesired user has a high detected power as compared to the desired user. Thus, the successful implementation of the 3G multiuser wireless communication systems requires the development of novel, practical, and high-performance signal processing algorithms for channel estimation, interference cancellation, and multiuser detection. That is the motivation of this study.

![CDMA Classification Diagram](image-url)

Figure 1.1: CDMA classification
1.2 Antenna Array System and Space Processing

Two considerable impairments that limit the capacity and performance of mobile communication systems are multiple access interference (MAI) and co-channel interference (ICI) (interference from users in the adjacent cell sharing the same channel, e.g. frequency, time, code) [8]. Antenna array system is one of the most promising technologies that provides a higher capacity in wireless networks by effectively reducing MAI and ICI [9][10]. The early smart antenna systems were designed for use in military applications to suppress interfering signals [11][12]. Since interference suppression is a feature in these systems, this technology is borrowed to apply to personal wireless communications where interference is limiting the number of users that a network could handle. It is a major challenge to apply smart antenna technology to personal wireless communications due to the denser network traffic and the limited time for complex computations. However, the advent of powerful, low-cost, digital processing components and the development of software-based techniques has made smart antenna systems a practical reality for cellular communication systems.

An antenna array system is a number of antenna elements that form an array system of a given geometry with measurements taken with respect to the array reference point. Antenna arrays come in various geometrical configurations, for instance, circular, grid and linear arrays. The beam pattern of the array depends on the geometry, the amplitude and phase excitation of the elements, and also the radiation pattern of individual elements. Without loss of generality, isotropic antennas will be assumed throughout the thesis.

Fig. 1.2 depicts a typical beamformer that exploits the space diversity at the base station receiver. It is a narrowband beamformer which assumes that the bandwidth of the signal is narrow enough that the baseband signal waveform stays almost constant at each array element. The beamformer operates by adjusting the phase and amplitude of the signals induced on each element of the antenna array using the weights \( \{w_1, w_2, \ldots, w_N\} \) and combining to form the beamformer output, thus a beam has been formed towards the direction where maximum gain is required. The collection of different weights creates different beamformers. The simplest weights are formed for diversity combining, which is used to overcome the problem of fading in radio channels and utilizes the fact that the signals arriving at different locations fade at different rates. The combining techniques that are
1. Introduction

Currently used in various wireless and cellular systems are selection diversity, equal gain combining (EGC) and maximal ratio combining (MRC).

The MRC diversity combiner is also known as the Wiener-Hopf beamformer which maximises the signal-to-noise-plus-interference ratio (SNIR) at the array output, thus reducing CCI and providing spatial diversity at the same time [13][14][15]. However, a known problem with the Wiener-Hopf beamformer is that it suffers from diminished resolution as the SNR is reduced, implying that closely located interference may not be cancelled. Several super-resolution beamformers have been proposed to solve this problem. The high resolution subspace-based approach, due to the introduction of the well-known MUltiple Signal Classification (MUSIC) algorithm [16], hence draws tremendous interest. The MUSIC algorithm is derived by introducing the concept of array manifold as the set of all possible array responses as a function of the parameters to be estimated. The intersection between this manifold and the signal subspace determined from the array measurements is then sought. MUSIC yields the results which are known to be asymptotically unbiased and efficient. Different channel estimation approaches based on MUSIC are proposed in [17][18] for a DS-CDMA system over frequency selective fading channels. Other notable subspace-based algorithms include Estimation of Signal Parameters via Rotational Invariance Techniques (ESPRIT) which was proposed by Roy and Kailath [19] and was later adapted to MC-CDMA system in [20], and Subspace Fitting (SF) Approach which is computationally attractive for the ubiquitous case of a uniform linear array [21].

Figure 1.2: A beamformer structure
1.3 Space-Time Propagation CDMA Channels

In general, mobile radio channels are multipath and time-varying caused by the mechanisms of reflection, diffraction and scattering. An illustrative example is shown in Fig. 1.3. A good understanding of the wireless channel and its key physical parameters lays the foundation for the rest of the thesis. The variation of the received signal strength broadly exhibits two types of characteristics,

- **Large-scale fading**: Due to path loss of signal as a function of distance and shadowing by large objects such as buildings and hills. It represents the average signal power attenuation or path loss resulting from the motion over large areas (cell size). The statistics of large-scale fading provide a way of computing an estimate of path loss as a function of distance. This is described in terms of a mean-path loss and a log-normally distributed variation about the mean.

- **Small-scale fading**: Due to the constructive and destructive contributions of signals coming from different paths between the transmitter and receiver. Since the transmitted signals typically arrive at the receiver with independent phase, time and amplitude, the sum of the received signals varies rapidly over very short time durations (on the order of seconds) or over very short travel distances (a few wavelengths).

The understanding of the small-scale fading is important in designing a reliable and efficient space-time communication system - the focus of this thesis. Therefore, small-scale fading will be studied next by examining the two mechanisms, namely, the delay spread (or time dispersion) and the Doppler spread (time-variant behavior) of the channel.

1.3.1 Channel Dispersiveness

It is often desirable to characterise the channel by several statistical parameters of concern. The propagation channel is considered as an operator that introduces dispersion to the transmitted signal characterised in terms of delay spread and Doppler spread [22][23]. It has significant effects on the transmitted signals and also profoundly influences the receiver architecture design.

- **Delay spread** is caused by multipath arrival with different time delays. The transmitted signals typically arrive at the receiver with independent phase,
time and amplitude, resulting in a boost or a deep fade in the envelope of the received signal. The multipath intensity profile shown in Fig. 1.4(a) plots the average received signal power as a function of the time delay $\tau$. The delay spread $T_{\text{spread}}$ is the maximum excess delay between the first and last arrived path components, during which the signal power falls to some threshold level below that of the strongest received power. The spaced-frequency correlation function is the Fourier transform of the multipath intensity profile where the coherence bandwidth $B_{\text{coh}}$ is defined as the frequency range over which the signals have a strong potential for amplitude correlation. Delay spread ($T_{\text{spread}}$) and coherence bandwidth ($B_{\text{coh}}$) are inversely proportional to one another, i.e.

$$B_{\text{coh}} \approx \frac{1}{T_{\text{spread}}}$$

In a CDMA system, if the chip period $T_c$ is much greater than delay spread, the effects of time dispersion are negligible at the receiver and multipath fading is frequency flat. Therefore, the channel is known as the flat fading channel if

$$T_c \gg T_{\text{spread}}$$

As the signal bandwidth $B_s$ is inversely proportional to $T_c$, Fig. 1.4(a) also
Figure 1.4: Types of fadings experienced by a CDMA signal due to (a) delay spread and (b) Doppler spread.

implies

\[ B_s \ll B_{\text{coh}} \]

Otherwise Inter-symbol-interference (ISI) arises and multipath fading is frequency selective, resulting in the frequency selective channel.

- **Doppler Spread** Doppler effect causes frequency dispersion which leads to signal distortion. The relative phase shifts between multipath components of the received signal change with the spatial location of the transmitter resulting in the received envelope varying rapidly. Fig. 1.4(b) shows the Doppler power spectrum as a function of the Doppler frequency shift \( f \) which yields the knowledge about the spectral broadening of the channel caused by the motion of the transmit or receive antennas. The width of the Doppler power spectrum is referred to as the Doppler spread \( B_{\text{Dop}} \). The spaced-time correlation function is the Fourier transform of the Doppler power spectrum with \( T_{\text{coh}} \) defined as the channel coherence time over which the channel’s response to a transmitted signal has a correlation greater than some threshold. Doppler spread \( B_{\text{Dop}} \) and coherence time \( T_{\text{coh}} \) are parameters that describe the time varying nature of the channel in a small scale region. Analogous to the delay spread, the approximate relationship between the two parameters is given as

\[ T_{\text{coh}} \approx \frac{1}{B_{\text{Dop}}} \]
As shown in Fig. 1.4(b), if the baseband signal bandwidth $B_s$ is much greater than the Doppler spread, the effects of Doppler spread are negligible at the receiver and the fading channel can be categorized as the slow fading (time flat) channel. Therefore, a signal undergoes slow fading if

$$B_s \gg B_{Dop}$$

and

$$T_c \ll T_{coh}$$

Otherwise the channel changes during one symbol period and it can be classified as the fast fading (time selective) channel.

1.3.2 Parametric Channel Modelling

To exploit antenna arrays in wireless communications, it is necessary to obtain an accurate, yet tractable, modelling of the channel. In general, two main approaches can be found in the existing models. On the one hand are the parametric physical models which are more accurate descriptions of the actual propagation environment whose parameters represent physically meaningful quantities, such as path delays, directions, Doppler shifts and complex path gains. Parametric modelling is especially important in developing channel parameter estimation and detection techniques that can remove channel distortions [24][25][26]. In this thesis, parametric channel models based on the array manifold concept will be developed. Note that the concept of antenna array manifolds is a natural representation and thus a perfect abstraction of the channel propagation characteristics.

On the other hand is the less stringent but simple to implement approach, the non-parametric models. It models the overall stochastic channel as a random matrix conformed to certain distributions without characterising individually the probability density function (pdf) of the random channel variables involved in the modelling. Such models have been heavily used in capacity calculations for the development of given communication techniques, such as MIMO systems [27][28] and space-time coding techniques [29][30].

1.4 Space-Time Processing

The space-only processing of Fig. 1.2 is for narrow-band signals. The performance of the beamformer using this structure deteriorates as the signal bandwidth in-
creases. For processing CDMA signal experiencing multipath fading, a tapped delay line (TDL) structure is normally used.

### 1.4.1 Space-Time CDMA Receivers

CDMA receivers can be grouped into two categories, single user (SU) receiver and multiuser (MU) receiver. The single user receiver operates with the knowledge of only the desired user’s signature waveform, path delays and path coefficients. The advantage of the single user receiver is its simplicity and robustness. However, it is unable to suppress MAI, also known as the near-far problem, thus it needs accurate power control. The multi-user receivers are designed to eliminate MAI completely (in the absence of additive noise) despite the lack of orthogonality between the user’s signature waveforms. The main concept of multiuser receiver is to take into account the PN sequences of all users in forming the receiver weight and jointly detect all active users so as to outperform the single user receivers that treat MAI as noise. A hierarchy of various single and multiuser receivers is given in Fig. 1.5. According to Fig. 1.5, both the single and multiuser receivers
can be further categorised into two families: the *decoupled space-time receiver* and the *joint/integrated space-time receiver*. The decoupled receiver performs interference cancellation in two stages while the integrated space-time processing exploits jointly the spatial and temporal signatures of the received signal.

The *decoupled space-time receiver* is shown in Fig. 1.6. It involves a beamformer at the front-end followed by a temporal processor. The beamformer weights \([w_1, w_2, \ldots, w_N]\) are used to steer the array in a given looking direction. Following the steering array, the tapped delay line (TDL) (of length \(L\)) samples the signal and passes it to an FIR filter with weight \(v_i^H \forall i = 1, \ldots, N\). The filter weight is designed to exploit the ISI structure of the desired signal and removes the associated interference in the looking direction. The cleaned signal is then passed through the decision device to regenerate the information data symbols.

On the other hand is the *integrated space-time receiver* shown in Fig. 1.7. The received signals are first sampled and passed through a bank of TDLs. Signals at the output of the TDLs are concatenated to form a \(NL \times 1\) vector-signal which contains the spatial temporal signature of the channel. It is then processed by applying a \(NL \times 1\) weight vector which is a despreader and a beamformer that removes both the MAI and ISI components from the signals. It is worthy of note that the number of users/multipaths that can be resolved by the integrated receiver is \(NL - 1\) which is much greater than its decoupled counterparts. The decoupled receiver soon reaches the limits of their performance once the number of users exceeds the number of antennas. The integrated subspace type space-time receiver will be focused in this study. The unique spatial temporal signature of each multipath based on the Spatial Temporal ARray Manifold (STAR) is developed and exploited, which is powerful in the suppression of both the ISI and MAI effects of the channel due to the greater degrees of freedom it offers.

### 1.4.1.1 Single User CDMA Receiver

A standard single user CDMA receiver that operates in the presence of multipath is the RAKE receiver originally proposed by Price and Green [31]. It was later developed into the 2D RAKE receiver which identifies both the temporal and spatial structure of the individual paths arriving at the receiver. The 2D RAKE receiver has the decoupled structure shown in Fig. 1.6 where the front-end is a beamformer which helps to isolate the wanted signal component from a particular direction prior to despreading. Each temporal finger is a TDL line with a weight, the delayed copy of the desired user’s PN sequence. The beamformer output are
1. Introduction

Figure 1.6: Decoupled space-time processing
Figure 1.7: Joint space-time processing
correlated with the weight and diversity combined before passing through the decision device. Such a receiver can outperform the conventional RAKE receiver due to the presence of the added spatial dimension [32]. However, in this study, the integrated RAKE receiver (Block 8 in Fig. 1.5) which has a structure shown in Fig. 1.7 will be derived and used for performance analysis. The beamforming weights will be formed based on the spatial-temporal array manifold (STAR) which filters the received signal in joint space-time domain and despreads the CDMA signals to form the receiver output.

1.4.1.2 Multi-user Receiver

It has been well established that multiuser detection techniques can substantially enhance the receiver performance and increase the capacity of CDMA communication systems. An optimum multiuser receiver was proposed in [33] in an AWGN channel and an optimum space-time multiuser receiver for DS-CDMA system was proposed by Kohno et al. in [34]. It is demonstrated that the optimum receiver can be near-far resistant at the expense of computational complexity which increases exponentially with the number of users and the number of channel parameters. So the practical use of the optimal receiver is prohibitive and a class of linear and nonlinear (suboptimal) receiver have been proposed to trade-off the computational burden of the optimal receiver. Such linear receivers are MMSE detector, decorrelator, parallel/successive interference cancellers and etc. A comprehensive treatment of space-time multiuser detection in multipath CDMA channels with receiver antenna arrays have been provided in [35] and several space-time multiuser detection structures were derived, including the optimum MLSE detector, low complexity linear space-time multiuser detectors based on iterative interference cancellation, and blind adaptive space-time multiuser detectors.

Blind space-time receivers have also been proposed. By shortening or eliminating training sequences, the transmission efficiency is increased, particularly for those rapidly varying channels whose training signals must be transmitted periodically for conventional equalization methods. One of the most effective blind methods is the subspace technique which relies on the second-order statistics of the received signal. By applying the eigenvalue decomposition on the data covariance matrix, the signal subspace and noise subspace can be identified. Based on orthogonality property of subspaces, channel parameters are then estimated by minimising projections of signature waveforms of input symbols onto the noise subspace or maximising their projections onto the signal subspace. The
method has been successfully applied in DS-CDMA systems for flat fading channels [36], frequency-selective fading channels [37][38][39], single-input multiple-output (SIMO) systems [40][35] and multiple-input multiple-output (MIMO) systems [41].

In this thesis, blind subspace based single-user receiver (Block 9 in Fig. 1.5) will be developed. The receiver is composed of a subspace estimator and an interference canceller. The performance of the proposed receiver will be compared with other existing single and multiuser space-time receivers. As the proposed receiver is an integrated type and is based on the STAR manifold vector, the standard 2D RAKE receiver, the decorrelating receiver and MMSE receiver will be extended to the STAR RAKE, decorrelator and MMSE receiver (Block 8, 12 and 13 in the classification described in Fig. 1.5), which use conjointly the spatial temporal signature attributed to each multipath. We will show that the proposed single-user receiver has a performance much greater than the STAR RAKE receiver and also has a performance comparative to the multiuser receivers.

### 1.4.2 Space-Time CDMA Transmitter

The space-time processing can also be carried out at the transmitter prior to the transmission. This is very different from receiver processing which is carried out after the channel has affected the signal. The transmitters use a variety of techniques to maximise diversity, minimise generated CCI and also in some situations pre-equalise the channel for ISI [42]. Most transmit algorithms can be separated into two categories as follows:

- **Close loop** transmission: The receiver sends back the channel state as side information using a feedback link. However, this technique requires the channel to be sufficiently slowly varying and has a loss in spectral efficiency due to the utilization of part of the bandwidth to transmit the channel state.

- **Open loop** transmission: A feedback link is not required between transmitter and receiver. Space-time block coding based schemes are open loop transmission techniques. In these schemes, the signals are transmitted in a balanced way via multiple antennas which provides maximal path diversity at the receiver [43]. Open loop transmit (Tx-) beamforming techniques have also been developed. Channel is estimated at the transmitter based on the channel reciprocity which assumes that the angles of arrival of uplink
signals are almost the same as the angles of departure of downlink transmitting signals to the mobile station. There are also algorithms proposed to mitigate the channel response mismatch between uplink and downlink [44][45].

Furthermore, based on the degree of channel state information (CSI) available at the transmitter, most of the transmit algorithms can be separated into two categories as follows:

- **No CSI at transmitter**: Transmission processing can be done without the knowledge of the channel. As discussed above, Space-time block coding (STBC) is such a technique that is designed to maximise the spatial and temporal diversity of the channel [46].

- **Known/partially known CSI at transmitter**: Although techniques that require no channel information are desirable for their simplicity, they cannot attain the performance achievable if channel knowledge is available, at least partially, at the transmitter. In this case, the transmitter can employ Tx-beamforming or linear precoding/decoding algorithms to equalize the channel distortion in joint space-time domain. In the case when the transmitter has perfect knowledge of the channel, the Tx-beamforming weight can be designed to satisfy different criterion such as maximum ratio transmission considered in [44][47] and maximum SNIR [47][48]. The latter method provides desired levels of isolation among mobiles by directing the signal power to the desired mobile station (MS) while minimising the interferences created at other MSs. [49][50] study the case when partial information is available at the transmitter, that is the transmitter has only the knowledge of either the mean or the covariance of the channel coefficients.

The advantage of transmit diversity can also be demonstrated by the fact that various kinds of transmit diversity schemes have been implemented in practical wireless communication standards. A significant effort has been devoted in 3GPP to develop efficient transmit diversity solutions to enhance downlink capacity through the use of antenna arrays [51]. WCDMA supports three transmit diversity concepts. The open loop scheme is the $2 \times 2$ space-time block code proposed by [52], known as Space Time Transmit Diversity (STTD) in WCDMA. The close loop schemes, known as Mode 1 and 2, apply a 2- or 4- quantization for the feedback weight, respectively, to parameterize the downlink beamforming
weight matrix. As more powerful MIMO techniques emerge, they will certainly be considered as enabling techniques for future high-speed wireless systems (i.e. 4G and beyond).

1.5 Research Objective and Thesis Organisation

This chapter has provided a brief introduction of topics covered by the thesis. Recent research interests in the field of wireless personal communications have been moving to the 3G and beyond cellular systems for higher quality and variable speed of transmission for multimedia information. The CDMA modulation scheme supports the service requirement in third generation mobile radio systems due to its capabilities to provide higher capacity over conventional TDMA and FDMA schemes. However, the multiple access interference (MAI) and the co-channel interference (CCI) due to multiple access, as well as the intersymbol interference (ISI) due to the multipath fading channels are three major limiting factors facing wireless systems today.

An antenna array can be used to reduce the interference, improving the Signal-to-Noise-plus-Interference Ratio (SNIR) and thus increasing the system capacity. Therefore, space-time processing based on antenna array technology becomes a breakthrough technique for the third generation of wireless personal communications.

It can be seen that the antenna arrays can be applied at the base station, the mobile unit or at both locations. The differences in propagation environment, physical limitations and the cost constraints result in different choices of type and number of antennas at the base station and mobile unit. Similarly, space-time processing techniques can be used in receive alone, in transmit alone or in both links. It can be blind or nonblind, open loop or close loop. It is directly affected by the physical layer infrastructure and channel propagation properties. As a result, the primary objective of this thesis is the design of array transmitters and receivers (i.e. transceivers) through space-time exploitation. Several signal processing functions within the wireless communication systems will be looked into, including modulation/demodulation, transmit diversity/precoding, channel estimation and reception. The rest of the thesis is structured as follows:

- In Chapter 2, the parametric model based on the array manifold concept will be derived and an overview of the space-time architectural system will also be outlined. The space-time channel models are categorised based on
the number of inputs and outputs. Various types of space-time channels are covered, from the simplest Scalar-Input Scalar-Output (SISO) channel to the more advanced Vector-Input Vector-Output (VIVO) channel. Both the diagrammatical and mathematical representations of these models are provided. This modelling forms the basic mathematical framework for later formulation in the subsequent chapters.

- In Chapter 3 an arrayed MIMO system employing antenna arrays at both transmitter and receiver is investigated. Many reported research works on MIMO, often assume that the channel is flat fading and known. However, the MIMO array receiver proposed in this study is developed for asynchronous time-varying multipath fading channels and can be operated in a completely blind way. Subspace type blind algorithms have been proposed for joint DOA, TOA, Doppler frequency and path power estimation. With the channel estimator employed as the front-end, a subspace based space-time single-user receiver is proposed which requires no information of the interfering users while achieving a Signal-to-Noise-and-Interference-Ratio (SNIR) enhancement comparative to a multiuser decorrelator. Simulation results have shown that the proposed single-user receiver achieves, asymptotically and blindly, the performance comparative to the existing multiuser receivers. The proposed receiver is also robust to channel estimation errors in the event of any unidentified (incomplete) or erroneous (incorrect) channel parameter.

- Chapter 4 focuses on the problem of joint transmitter and receiver (Tx-Rx) beamforming optimisation in a DS-CDMA system over multipath fading channels. Unlike the majority of works which often ignores the array geometry, the proposed approach is built on the array manifold concept and thus, the space-time properties of the channel can be fully exploited. Thus in this chapter, the beamforming weights are designed to minimise the mean-squared-errors (MSE) over the entire network. An iterative solution to the optimisation problem is firstly proposed under the system framework. A closed-form solution based on channel eigendecomposition is then proposed. The convergence of the iterative method to the closed-form solution and the equivalence of the two methods are verified through numerical simulations. The empirical simulation results of both joint Tx-Rx beamforming schemes, illustrate a clear advantage over the transmitter (Tx-) beamforming and linear multiuser receiver in terms of both SNIR and bit
error probability.

- Chapter 5 is concerned with an arrayed OFDM-CDMA system operating in the presence of $M$ co-channel single antenna transmitters. OFDM modulation technique is combined with CDMA scheme where the entire channel is divided into many narrow parallel subchannels, thereby increasing the symbol duration and reducing or eliminating the intersymbol interference (ISI) caused by the multipath environments. The signals can be easily transmitted and received using the Fast Fourier transform (FFT) device without increasing the transmitter and receiver complexities. The receiver is decoupled into channel estimation process and reception process. A blind channel estimator is proposed which explores the space-time properties of the channel and jointly estimates the DOA and TOA of each path. Two space-time receivers have been proposed. One is a post-FFT type which takes the advantage of fast FFT/IFFT device at both transmitter and receiver which greatly simplifies the transceiver design. A pre-FFT type subspace based signal detection algorithm is then proposed, which is subspace type and has superior performance over the post-FFT approach especially at low SNR levels.

- In Chapter 6 the thesis is concluded and a list of the original contributions is presented together with potential directions for future research.
Chapter 2

Space-Time Channel Modelling and System Design

The dispersive effects introduced by the propagation channel cause the transmitted signal to be distorted at the receiver front-end of the communication link. In this chapter, parametric channel models based on the concept of array manifold will be derived forming the basis for the subsequent channel estimation and reception algorithms proposed in Chapters 3-5 in the thesis. Furthermore, general frameworks of designing the space-time CDMA communication systems are presented with the main building blocks properly outlined. Note that, a detailed mathematical modelling of each of these building blocks will be described in later chapters.
2.1 Array Manifold Vector

To exploit the space-time property of the channel, the antenna array is equipped at the front-end of the receiver. The mathematical way of modelling the spatial information provided by the antenna array is using the array manifold vector which is a function of a number of channel parameters, including the direction of arrival (DOA) of the multipath, the array geometry and the carrier frequency. The array manifold is an important concept in capturing the spatial characteristics of the channel and it forms a preliminary part of the modelling of the space-time channel, alongside with other intrinsic channel parameters.

Fig. 2.1 shows the propagation of the planewave signals in a 3D real space. The narrowband model will be assumed in the thesis\(^1\). The direction of arrivals (DOAs) can be well captured using the array manifold vector. Assume that the \(k\)th path of the \(i\)th user arrives at the \(m\)th element of the antenna array from (azimuth, elevation) direction \((\theta_{ik}, \varphi_{ik})\), \(e_{ik} = [\cos \theta_{ik} \cos \varphi_{ik}, \sin \theta_{ik} \cos \varphi_{ik}, \sin \varphi_{ik}]^T\) which is the real unit-vector pointing towards the direction \((\theta_{ik}, \varphi_{ik})\). It is known that the relative phase variation at the \(m\)th antenna element with respect to the reference point can be expressed as

\[
\exp(-j2\pi F_c \frac{E_{ikm}}{c})
\]

where \(E_{ikm}\) is a 3 \times 1 vector denoting the Cartesian coordinates in metres of the location of the \(m\)th antenna element. Let’s define \(k_{ik} = \frac{2\pi F_c}{c} e_{ik}\) as the wavenumber vector with \(F_c\) the carrier frequency and \(c\) the speed of light, the array manifold vector \(S_{ik}\) associated with the \(k\)th path of the \(i\)th user’s DOA \((\theta_{ik}, \varphi_{ik})\) is given as

\[
S_{ik} = \exp(-j2\pi F_c \frac{E_{ikm}}{c})\]

where \(E_m = [E_1, E_2, \ldots, E_N] = [x, y, z]^T\) is a \(3 \times N\) matrix with its \(m\)th column the location \(E_m\) of its \(m\)th antenna element. The set of array response vectors, \(\{S(\theta, \varphi) \mid \theta, \varphi \in [0, 2\pi]\}\) forms the array manifold. In many cases, the signals are assumed to be on the \(x-y\) plane (i.e. \(\varphi_{ik} = 0^\circ\)). Therefore, the array manifold vector is simplified to

\[
S_{ik} = \exp(-j\pi(r_x \cos \theta_{ik} + r_y \sin \theta_{ik})) \in C_{N \times 1}
\]

\(^1\)This implies the ratio of the signal bandwidth over the carrier is small (even if the signal is wideband).
Figure 2.1: Planewave propagation model
with sensor location measured in units $\lambda_c/2$. A popular class of array is that of linear arrays, i.e. $r_y = r_z = 0_N$ and in this case, Eqn. 2.1 is simplified to

$$S_{ik} = \exp(-j\pi r_x \cos \theta_{ik}) \in \mathbb{C}^{N 	imes 1}$$

### 2.2 Space-Time Channel Modelling

Space-time wireless channel can be classified, based on its number of inputs and outputs, into four main categories using the terms

- **Scalar input/output** to indicate a single input/output and
- **Vector input/output** to indicate a set of more than one inputs/outputs

In a similar fashion the terms scalar-signal and vector-signal indicate a single signal or a set of more than one signals respectively. Using the above terms and with reference to Fig. 2.2, the communication channels can be classified into four basic types as follows:

- **Scalar-Input Scalar-Output (SISO)** channel: e.g. In the single link transmission, the single antenna elements are employed at both ends of the transmission link;

- **Scalar-Input Vector-Output (SIVO)** channel: e.g. In the single link transmission, the system employs a single antenna transmitter and an antenna array receiver;

- **Vector-Input Scalar-Output (VISO)** channel: e.g. According to Fig. 2.2, in the case of a single link transmission, VISO channel is formed when an antenna array is employed at the transmitter and a single antenna is used at the receiver; In a multiple access system, VISO channel can be further categorised into: (i) **VISO MA-1** with multiple users employing a single antenna at the transmitters and a single antenna is used at the receiver; (ii) **VIVO MA-2** with the employment of antenna arrays at the transmitters and a single antenna at the receiver.

- **Vector-Input Vector-Output (VIVO)** channel: e.g. Similar to the classification of VISO channel, there are different types of VIVO channels depending on how vector-signals are formed at the transmitter and receiver. In a single link system, VIVO channel is formed with the employment of antenna
2.2.1 Scalar-Input Scalar-Output (SISO) Channel

SISO channel is the simplest wireless model which does not include the spatial information. This is best illustrated in the Scalar-Input Scalar-Output (SISO) channel model as shown in Fig. 2.3 representing the transmission channel from a single antenna transmitter to a single antenna receiver. Consider a multipath slow fading channel where the scalar-signal $m_i(t)$ is transmitted to the receiver via $K_i$ multipaths. In this case, the channel impulse response between the transmitter and the receiver can be expressed as the summation of the multipath components:

$$
\text{SISO channel impulse response} = \sum_{k=1}^{K_i} \beta_{ik} \delta(t - \tau_{ik}) \quad (2.2)
$$
The channel is characterised by its own delay (TOA) \( \tau_{ik} \) which is uniformly distributed within the range \([0, T_{cs})\) and the fading coefficient \( \beta_{ik} \) described by the complex Gaussian distribution.

According to Fig. 2.3, the signal at the output of the channel \( x_i(t) \) is the convolution of the transmitted signal \( m_i(t) \) with the channel impulse response and is given as

\[
x_i(t) = \sum_{k=1}^{K_i} \beta_{ik} m_i(t - \tau_{ik})
\]

### 2.2.2 Scalar-Input Vector-Output (SIVO) Channel

If the antenna array of multiple elements is employed at the receiver, the Scalar-Input Vector-Output channel is formed. An extra dimension (i.e. the spatial dimension) is added to the model due to the employment of the antenna array. This is exploited using the array manifold vector derived in Section 2.1 which provides information on the spatial structural property of the transmission channel.

The multipath propagation environment can be modelled as a composition of a number of multipath components of the channel as shown in Fig. 2.4, each characterised with its own DOA, delay and path fading. The channel impulse response between the transmitter and the receiver containing the array manifold vector can be expressed as

\[
\text{SIVO channel impulse response} = \sum_{k=1}^{K_i} \beta_{ik} s_{ik} \delta(t - \tau_{ik})
\] (2.3)
The vector-signal at the output of the channel can thus be modelled as

\[ x_i(t) = \sum_{k=1}^{K_i} \beta_{ik} S_{ik} m_i(t - \tau_{ik}) \]

To describe the time varying nature of the channel, the Doppler frequency shift can be further incorporated into the model to obtain the time varying SIVO channel model as shown in Fig. 2.5. The symbol \( F_{ik}(t) \) is used to denote the frequency shift component caused by the Doppler frequency for the \( k \)th path of the \( i \)th user and is defined as

\[ F_{ik}(t) = \exp(j2\pi f_{ik}t) \] (2.4)

where \( f_{ik} \) explicitly models the Doppler shift due to motion of the transmitter and the frequency offset caused by the local oscillator frequency offset. The channel impulse response can thus be expressed as

\[ \text{SIVO channel impulse response} = \sum_{k=1}^{K_i} F_{ik}(t) \beta_{ik} S_{ik} \delta(t - \tau_{ik}) \] (2.5)

The output signal \( x_i(t) \) including Doppler effect can be expressed as

\[ x_i(t) = \sum_{k=1}^{K_i} F_{ik}(t) \beta_{ik} S_{ik} m_i(t - \tau_{ik}) \] (2.6)

Note that if Doppler effect can be ignored, that is \( f_{ik} = 0 \), the phase shift
Figure 2.5: Scalar-Input Vector-Output (SIVO) channel model with Doppler frequency shift

\[ F_{ik}(t) = 1 \] thus Fig. 2.5 reduces to Fig. 2.4. A final note concerning the channel parameters \( \theta_{ik}, \beta_{ik}, \) and \( f_{ik} \) is that they are time-varying in the general situation. However, the rate of variation is very low compared to the symbol rate, so these parameters can be considered to be constant over the observation interval (a block of symbols). The precise number of symbols in the block primarily depends on the user’s speed of motion.

### 2.2.3 Vector-Input Scalar-Output (VISO) Channel

From Fig. 2.2, it is clear that there are different types of VISO channels

- VISO in a single link system
- VISO MA-1 and VISO MA-2 in a multiple access system

Fig. 2.6 represents the VISO channel model in a single link system. The transmit array manifold vector is denoted as \( \overline{S}_{ik} \). Note that a bar on the top of a symbol represents a transmitter’s parameter. The transmit array manifold vector \( \overline{S}_{ik} \) is also defined as in Eqn. 2.1 which is a function of the channel parameters associated with the transmitter array, including the antenna array geometry of the transmitter and the direction of departures (DODs) of the channel. According
Figure 2.6: Vector-Input Scalar-Output (VISO) channel model

The channel impulse response of the VISO channel can be expressed as

\[
\text{VISO channel impulse response} = \sum_{k=1}^{K_i} \beta_{ik} \overline{S}^T_{ik} \delta(t - \tau_{ik})
\]  

(2.7)

and the output scalar-signal is given as

\[
x_i(t) = \sum_{k=1}^{K_i} \beta_{ik} \overline{S}^T_{ik} m_i(t - \tau_{ik})
\]

### 2.2.4 Vector-Input Vector-Output (VIVO) Channel

A representative example is the single link system where the antenna arrays are used at both ends of the transmission link. Based on whether channel state information (CSI) is available at the transmitter, the VIVO channel model can be further categorised as

- **VIVO channel model with CSI at the transmitter**: If the channel information is available at the transmitter, the spatial information at the transmitter can be characterised using the transmit array manifold vector. Such a VIVO channel model incorporating the transmit array manifold vector is shown in Fig. 2.7. This implies that the signals leaving the transmitting elements experience a common fading channel towards the receiving antenna array. The channel model is an extension of the SIVO channel model (Fig.
by incorporating transmit array manifold vector. The channel impulse response vector is thus given as

\[
\mathbf{VIVO \text{ with CSI at the transmitter: }} \sum_{k=1}^{K_i} \beta_{ik} \mathbf{F}_{ik}(t) \mathbf{S}_{ik} \mathbf{S}_{ik}^T \mathbf{x}(t - \tau_{ik}) \tag{2.8}
\]

The vector-signal output \( x_i(t) \) is formulated as

\[
x_i(t) = \sum_{k=1}^{K_i} \beta_{ik} \mathbf{F}_{ik}(t) \mathbf{S}_{ik} \mathbf{S}_{ik}^T m_i(t - \tau_{ik}) \tag{2.9}
\]

Note that \( \mathbf{F}_{ik}(t) = 1 \) for the flat fading channel and Eqn. 2.9 can be reduced to

\[
x_i(t) = \sum_{k=1}^{K_i} \beta_{ik} \mathbf{S}_{ik} \mathbf{S}_{ik}^T m_i(t - \tau_{ik}) \tag{2.10}
\]

which forms an important part of signal modelling in Chapter 4.

- **VIVO channel model without CSI at the transmitter:** This is the case when CSI is unavailable, or is not known at the transmitter. It is often seen in an open loop multiplex system where the input signals are demultiplexed, encoded and sent simultaneously from the \( N \) transmitting antennas. The signal received at each antenna is therefore a superposition of the transmitted signals from each antenna via multipath fading. One may view such
a system as providing \( \bar{N} \) SIVO channels between the transmitter and the receiver. This is best illustrated in Fig. 2.8. For convenience and without loss of generality, the number of multipaths are assumed to be \( K_i \) for all the SIVO channels. The channel impulse response can be expressed as a function of a collection of SIVO channels and is given as

\[
\text{VIVO without CSI at the transmitter} = \sum_{j=1}^{\bar{N}} \sum_{k=1}^{K_i} \mathcal{F}_{ik}(t) \beta_{ik}^{(j)} \sum_{l=1}^{j} \delta(t - \tau_{ik}^{(l)})
\]

(2.11)

The superscript \( j \) denotes the SIVO channel associated with the \( j \)th antenna of the transmitter array. Note that the fading coefficients \( \beta_{ik}^{(j)} \) in Eqn. 2.11 models the wave propagation from the \( j \)th antenna element to the reference point of the transmitter, in addition to the random phase shifts (due to the path losses and shadowing) between the reference points of the transmitter and the receiver. The two VIVO models can be seen as an unified model as \( \beta_{ik}^{(j)} \) in Eqn. 2.11 has absorbed the transmit array manifold vector \( \bar{S}_{ik} \) in Eqn. 2.8. Eqn. 2.11 can be rewritten in a compact format as

\[
\sum_{k=1}^{K_i} S_{ik} \text{diag} \left( \beta_{ik} \circ \mathcal{F}_{ik}(t) \right) \delta(t - \tau_{ik})
\]

(2.12)

where

\[
\left\{ \begin{array}{l}
\delta(t - \tau_{ik}) = \left[ \delta(t - \tau_{ik}^{(1)}), \delta(t - \tau_{ik}^{(2)}), \ldots, \delta(t - \tau_{ik}^{(\bar{N})}) \right]^T \in \mathbb{C}^{\bar{N} \times 1} \\
\tau_{ik} = \left[ \tau_{ik}^{(1)}, \tau_{ik}^{(2)}, \ldots, \tau_{ik}^{(\bar{N})} \right]^T \in \mathbb{C}^{\bar{N} \times 1} \\
\beta_{ik} = \left[ \beta_{ik}^{(1)}, \beta_{ik}^{(2)}, \ldots, \beta_{ik}^{(\bar{N})} \right]^T \in \mathbb{C}^{\bar{N} \times 1} \\
\mathcal{F}_{ik}(t) = \left[ \mathcal{F}_{ik}^{(1)}(t), \mathcal{F}_{ik}^{(2)}(t), \ldots, \mathcal{F}_{ik}^{(\bar{N})}(t) \right]^T \in \mathbb{C}^{\bar{N} \times 1} \\
S_{ik} = \left[ S_{ik}^{(1)}, S_{ik}^{(2)}, \ldots, S_{ik}^{(\bar{N})} \right] \in \mathbb{C}^{N \times \bar{N}}
\end{array} \right.
\]

(2.13)

which is formed by concatenating the parameters due to the \( k \)th path at each SIVO channel. The signal at the output of the channel thus is given as

\[
x_{ik}(t) = \sum_{k=1}^{K_i} S_{ik} \text{diag} \left( \beta_{ik} \circ \mathcal{F}_{ik}(t) \right) m_i(t - \tau_{ik})
\]

(2.14)

where \( m_i(t - \tau_{ik}) \) is the vector-signal defined as

\[
m_i(t - \tau_{ik}) = \left[ m_{i1}(t - \tau_{ik}^{(1)}), m_{i2}(t - \tau_{ik}^{(2)}), \ldots, m_{i\bar{N}}(t - \tau_{ik}^{(\bar{N})}) \right]^T \in \mathbb{C}^{\bar{N} \times 1}
\]

consisting the delayed copies of the vector-signal \( m_i(t) \) at the \( k \)th path of each SIVO channel.
2.3 Space-Time System Framework

An array communication system is composed of three basic functional units: transmitter, radio channel and receiver. Various types of wireless channel models have been presented in Section 2.2. With the space-time channel model, the transmitted signal $m_i(t)$ or $m_v(t)$ can hence be subsequently transformed to realise its corresponding continuous-time signal $x_i(t)$ or $x_v(t)$ at the front-end antenna array of the receiver. The channel models form part of the system framework as discussed in this section together with the transmitter and receiver block. The diagrammatical modelling of the transmitter and receiver blocks will be established to complement the description of the channel given in Section 2.2.

An $M$-user DS-CDMA multiplex system is shown in Fig. 2.9 referred herein as the MIMO arrayed system where antenna arrays are employed at both the base station (BS) receiver and the mobile station (MS) transmitters. At point A of the transmitter block, the $i^{th}$ user’s data stream is first demultiplexed into $\overline{N}$ substreams and each user’s signal is DS-CDMA modulated to form the baseband vector-signal $m_i(t)$, $i = 1, \ldots, M$. The signals are then transmitted via $\overline{N}$ antennas (Point B). The continuous transmitted signal propagates via a number of multipaths before arriving at the receiver employing an antenna array of $N$ antennas. The output of the antenna array is a vector-signal constituting the
signal output at each antenna elements \( (x(t) \text{ at Point C}) \). The physical communication medium is the VIVO MA-2 channel (defined as in Fig. 2.2) composing of a bank of VIVO channel as described in Fig. 2.8. The task of the receiver is to capture the data and process it accordingly so that the information signals can be recovered. The use of antenna array at the receiver can significantly increase the channel capacity by exploiting the spatial diversity [13], for example, to combat fading and to perform interference cancellation. The continuous signal \( x(t) \) is passed and stored in the bank of TDLs (shown in Fig. 1.7) to obtain the \( NL \)-dimensional discretised signal \( x[n] \) at Point D. The channel estimation and interference suppression block at the receiver will be studied extensively in Chapters 3. Each element of the vector-signal at point E represents a path associated with the desired user. The paths associated with different antennas needs be identified and grouped, i.e. \( y_{\text{Ant},i}[n] \; \forall \; i = 1, \ldots, M \). This is completed using the correlation analysis process which will be studied in Chapter 3. Having been assigned with the antenna ID, paths are combined and multiplexed (Point G) before passing through the decision device to form the recovered data symbols at Point H.

The architecture shown in Fig. 2.9 is most suitable when the transmitter does not have the channel state information and the receiver is based on more sophisticated channel estimation and reception techniques. When perfect/partial channel state information (CSI) is available at the transmitter, the advanced signal processing techniques can be applied at both the transmitter and receiver for jointly optimising the system performance. Such a system architecture is shown in Fig. 2.10 which depicts an \( M \) user DS-CDMA downlink system with \( N \) transmit and \( N \) receive antennas. Multiple users’ signals are first spreaded with their own spreading codes and then applied by a transmit beamformer before being transmitted together at Point A. The wireless channel is based on the downlink VIVO MA-2 channel model and is composed of a set of VIVO channels described in Fig. 2.7. At the \( i \)th MS receiver, the interference is removed from the discretised signal \( x_i[n] \) (Point E) by applying the receive beamformer \( w_i^H \; \forall \; i = 1, \ldots, M \). The problem under consideration is joint transmitter-receiver (Tx-Rx) beamforming over VIVO MA-2 channel, which will be carefully investigated in Chapter 4.

The incorporation of the antenna array in the OFDM-CDMA system is shown in Fig. 2.11. In this model, multiple mobiles transmit their information signals using a single antenna and they arrive at a base station that uses an antenna array to separate individual signals. The channel is viewed as a typical VIVO MA-1
Figure 2.9: Block diagram of the space-time system model as used in Chapter 3. Also note $\bar{N}K_1 = K$, and the notations will be defined in Chapter 3.
Figure 2.10: Block diagram of the space-time system model as used in Chapter 4.
channel composing of $M$ SIVO channels described in Fig. 2.4. This case will be studied in Chapter 5 where the cyclic spatial-temporal signature of the channel exploited by the antenna array will facilitate the development of more powerful channel estimation and interference cancellation techniques at the antenna array receiver.

2.4 Conclusion

The array manifold vector is first introduced which forms part of the constituents in the modelling of the space-time channel. This is then followed by the comprehensive mathematical modelling of space-time fading channels where the channels are categorised into four basic types with the channel impulse response derived for each channel type. It is shown that the space-time channel plays a significant role in transforming the signal from the transmitter-end to the receiver-end. The communication system framework is based on the integration of three fundamental blocks in a communication system: the transmitter, the propagation channels and the receiver. Each fundamental block may vary in the designing of a space-time system architecture. The antenna array frameworks that will be developed in the subsequent chapters are outline lastly in this chapter. Chapter 3 studies a multiplex MIMO arrayed system based on Fig. 2.9; Chapter 4 considers joint Tx-Rx beamforming in a multiple access system based on Fig. 2.10; In Chapter 5, arrayed OFDM-CDMA system as shown in Fig. 2.11 will be developed and analysed.
Figure 2.11: Block diagram of the space-time system model as used in Chapter 5
Chapter 3

Arrayed MIMO System

In this chapter, a multiple access Vector-Input Vector-Output (VIVO) channel is employed in a DS-CDMA communication system environment where both BS and MS are equipped with antenna arrays. In particular, based on the modelling presented in Chapter 2, a subspace based blind single-user receiver is proposed which requires no estimation of interfering users’ parameters and achieves a Signal-to-Noise-Plus-Interference-Ratio (SNIR) enhancement comparative to a multiuser decorrelator. As the subspace of unwanted signals is identified by removing the component of the desired signal from the overall signal subspace, a novel method for path power estimation is also proposed. The performance of the receiver is also robust to channel estimation errors in the event of any unidentified (incomplete) or erroneous (incorrect) channel parameter.
3. Arrayed MIMO System

3.1 Introduction

The increased use of antenna elements at both ends of the transmission link, giving rise to the research in Multiple Input and Multiple Output (MIMO) systems [53][54]. Blind channel estimation in MIMO systems is currently a very active research area and numerous blind and semiblind MIMO channel estimation and decoding techniques have been proposed [55]. For instance, blind decoding schemes that exploit the unique features of space-time codes have been proposed in [56][57][58] which, however, may not be applicable to a generalised space-time system. Furthermore, subspace based approaches have been proposed for MIMO channel estimation [41][59]. In [41], an adaptive subspace approach for channel estimation and a subspace space-time MMSE detector is proposed for an asynchronous MIMO CDMA system over multipath fading channels. However, this is not a superresolution technique and it is based on non-parametric channels models as in [60][61][62]. Note that, as it was stated in [63], the use of superresolution subspace techniques results in a substantially better performance in a multiuser system. The most representative superresolution subspace type technique is MUSIC (Multiple Signal Classification) [16] which was initially developed for direction finding and has evolved into a well-established technology [64]. ESPRIT [19] is another popular subspace technique with lower computation requirement than MUSIC but is more sensitive to array uncertainties (calibration errors). Both MUSIC and ESPRIT have been extended to Joint Angle and Delay Estimation (JADE –MUSIC [65] and JADE – ESPRIT [66]). The Joint Spread, Angle and Delay Estimation (SADE) method has recently been developed in [67][68]. However, all the above superresolution subspace techniques have not been employed in a MIMO type environment.

In this chapter, a blind subspace-type superresolution channel estimation algorithm based on the novel concept of Doppler-STAR manifold vector will be proposed for an arrayed MIMO system. The Doppler-STAR manifold vector will be used to describe, in a realistic manner, the dispersive behavior of the propagation channel in delay, direction, and Doppler frequency shift. In addition, an integrated estimator of the path gain of the multipath channel will be provided. The accurate parameterisation of the channel model using Doppler STAR manifold vector would result in a more accurate description of the channel, leading to the design of more powerful receivers. Note that the JADE and the proposed channel estimation approach belong to the same class of subspace techniques although the JADE method is not designed to handle the Doppler frequency shift.
The inclusion of Doppler frequency estimation in blind MIMO channel is not a trivial matter. Classic approaches to Doppler estimation typically use pilot signal [69][70] and most are developed for flat-fading channels [71][72]. In this chapter, the Doppler frequency shift has been incorporated in a novel way into the arrayed MIMO system and can be estimated over time-varying frequency selective MIMO fading channels. Note that the multipath Doppler spread, which is often regarded as one of the detrimental factors in degrading the performance of existing MIMO receivers, is being employed in the proposed receiver to provide an additional form of diversity [73][74].

The structure of the chapter is as follows. An arrayed MIMO channel model is first developed in Section 3.2 by introducing the Doppler-STAR array manifold vectors. Based on the model, a subspace type channel estimator is proposed in Section 3.3 which is able to provide a comprehensive estimation of DOA, TOA, Doppler frequency shift as well as path power of the MIMO channel. A blind single-user receiver is then proposed in Section 3.4 which requires only the knowledge of the desired user’s spreading code and is able to achieve (asymptotically) complete interference cancellation. The proposed framework is supported by a number of computer simulation studies presented and discussed in Section 3.5. The overall performance of the proposed arrayed MIMO system will be studied and compared with different types of receivers. Finally, the chapter is concluded in Section 3.6.

### 3.2 Arrayed MIMO System Model

With reference to Fig. 2.9 showing an $M$-user DS-CDMA arrayed MIMO system, the $i^{th}$ user’s data stream is first demultiplexed into $N$ substreams which are then differentially encoded and modulated using BPSK/QPSK DS-CDMA schemes. A unique spreading code is assigned to each user to be applied across its transmitting elements. An antenna-array of $N$ elements is used at the base station receiver for signal estimation and reception. According to Fig. 2.9, the overall wireless MIMO channel for $M$ active users can be modelled as a combination of $M$ VIVO channels described in Fig. 2.8.

#### 3.2.1 Spread Spectrum Transmission

With reference to Fig. 2.9, the BPSK/QPSK data stream is demultiplexed into a number of streams, one per antenna. This implies that during the $n^{th}$ time
interval of duration $T_{cs}$, a data vector associated with the $i^{th}$ user defined as

$$a_i[n] = [a_{i1}[n], a_{i2}[n], \ldots, a_{iN_c}[n]]^T$$ (3.1)

is transmitted (see point A). Then, the $i^{th}$ user’s information data is spreaded by its unique pseudo-noise (PN) code sequence of length $N_c$, i.e.

$$\alpha_i = [\alpha_i[0], \alpha_i[1], \ldots, \alpha_i[N_c - 1]], \text{ with } \alpha_i[k] = \pm 1$$

With reference to Fig. 2.9, the baseband signal $m_i(t)$ at point B can be modelled as

$$m_i(t) = \sum_{n=-\infty}^{+\infty} a_i[n] c_{PN,i}(t - nT_{cs}), \text{ } nT_{cs} \leq t < (n + 1)T_{cs}$$ (3.2)

where one period of the pseudo-noise spreading waveform associated with the $i^{th}$ user, $c_{PN,i}(t)$, is modelled as

$$c_{PN,i}(t) = \sum_{p=0}^{N_c-1} \alpha_i[p] c(t - pT_c), \text{ } pT_c \leq t < (p + 1)T_c$$ (3.3)

where $c(t)$ denotes the unit amplitude chip pulse-shaping waveform of duration $T_c$. For convenience and without loss of generality, certain features unimportant to the model but present in a practical system have been omitted, such as higher order source coding, multiple layer of channel coding and interleaving etc.

Prior to transmission, the baseband modulated signal $m_i(t)$ is upconverted to a bandpass signal using a carrier frequency $F_c$, i.e.

$$\sqrt{P_i} m_i(t) \exp(j(2\pi F_c t + \psi_i))$$

where $P_i$ is the transmitted power and $\psi_i$ is the random carrier phase uniformly distributed in $[0, 2\pi]$. At the receiver, it is then downconverted by the carrier $\exp(-j2\pi F_c t)$. For convenience the carrier will be ignored and baseband transmission will be assumed in this thesis. Thus the carrier is not shown in Fig. 2.9. The baseband received signal vector $\bar{x}(t)$ at point C is a composition of $M$ users’ signal components based on Eqn. 2.14. In the presence of additive white Gaussian noise, it is given as

$$\bar{x}(t) = \sum_{i=1}^{M} \sum_{k=1}^{K_i} S_{ik} \text{diag} \left( \beta_{ik} \odot \bar{x}_{ik}(t) \right) m_i(t - \tau_{ik}) + \eta(t)$$ (3.4)

The channel impulse response is defined in Eqn. 2.12. Note that the term $\sqrt{P_i} \exp(j(\psi_i - 2\pi F_c \tau_{ik}))$ has been absorbed in $\beta_{ik}$, i.e. $\beta_{ik} \triangleq \beta_{ik} \odot \sqrt{P_i} \exp(j(\psi_i -$
2\pi F_c (\tau_{ik})]. The symbol \( \mathbf{n}(t) \) is denoted as the complex white Gaussian noise vector of power \( \sigma_n^2 \). Eqn. 3.4 can be rewritten in a compact form as

\[
\mathbf{x}(t) = \sum_{i=1}^{M} \mathbf{S}_i \text{diag} \left( \mathbf{\beta}_i \odot \mathbf{F}_i(t) \right) \mathbf{m}_i(t - \tau_i) + \mathbf{n}(t)
\]  

(3.5)

where

\[
\begin{align*}
\mathbf{S}_i &= [S_{i1}, S_{i2}, \ldots, S_{iK_i}] \in \mathbb{C}^{N \times K_i} \\
\mathbf{\beta}_i &= \left[ \beta_{i1}^T, \beta_{i2}^T, \ldots, \beta_{iK_i}^T \right]^T \in \mathbb{C}^{K_i \times 1} \\
\mathbf{\tau}_i &= \left[ \tau_{i1}^T, \tau_{i2}^T, \ldots, \tau_{iK_i}^T \right]^T \in \mathbb{C}^{K_i \times 1} \\
\mathbf{F}_i(t) &= \left[ \mathbf{F}_{i1}(t), \mathbf{F}_{i2}(t), \ldots, \mathbf{F}_{iK_i}(t) \right]^T \in \mathbb{C}^{K_i \times 1} \\
\mathbf{m}_i(t - \tau_i) &= \left[ m_{i1}^T(t - \tau_{i1}), m_{i2}^T(t - \tau_{i2}), \ldots, m_{iK_i}^T(t - \tau_{iK_i}) \right]^T \in \mathbb{C}^{K_i \times 1}
\end{align*}
\]

is matrix (or vectors) formed by concatenating all the paths in the way specified above. Therefore, the total number of paths due to the \( i \)th user is \( K_i = NK_i \), which is the summation of the number of paths due to each SIVO channel. For convenience and without loss of generality, the following notations are used to denote one path component (say, the \( k \)th path out of a total number of \( K_i \) path) in Eqn. 3.5.

\[
\begin{align*}
S_{ik} &\triangleq \text{the } k \text{th column of } \mathbf{S}_i \quad \forall k = 1, \ldots, K_i \\
\beta_{ik} &\triangleq \text{the } k \text{th element of } \mathbf{\beta}_i \quad \forall k = 1, \ldots, K_i \\
\tau_{ik} &\triangleq \text{the } k \text{th element of } \mathbf{\tau}_i \quad \forall k = 1, \ldots, K_i \\
\mathbf{F}_{ik}(t) &\triangleq \text{the } k \text{th element of } \mathbf{F}_i(t) \quad \mathbf{F}_{ik}(t) = \exp(j 2\pi f_{ik} t) \quad \forall k = 1, \ldots, K_i
\end{align*}
\]

In the following discussion, these notations are used to denote the \( k \)th path of the \( i \)th user without differentiating the transmit antenna with which the path is associated.

### 3.2.2 Spatial Temporal ARray Manifold Vector (STAR)

Fig. 1.7, which for convenience is repeated here as Fig. 3.1 shows the input and output of a tapped-delay-line (TDL) structure. According to the figure, the continuous baseband signal \( x(t) \) is first discretised at the front-end of the receiver to obtain the discretised vector-signal \( \mathbf{x}[n] \). It is sampled at a rate of \( 1/T_s \) (where \( T_c = T_s \)) and passed through a bank of \( N \) tapped-delay lines (TDL), each of length \( L = 2N_c \) (to accommodate the lack of synchronisation due to maximum delay spread of \( T_{cs} \)).
A $2Nc$-dimensional discretised output frame $\tilde{x}[n]$ is thus formed and read for every $T_{cs}$ within the $n^{th}$ observation interval. However due to the lack of synchronization, the content of each TDL contains contributions from not only the current but also the previous and next symbols. To model such contributions as well, the $2Nc \times 2Nc$ time down-shift (or up-shift) operator matrix $J$ (or $J^T$) is defined as

$$J = \begin{bmatrix} 0 & 0 & \cdots & 0 & 0 \\ 1 & 0 & \cdots & 0 & 0 \\ 0 & 1 & \cdots & 0 & 0 \\ \vdots & \vdots & \ddots & \vdots & \vdots \\ 0 & 0 & \cdots & 1 & 0 \end{bmatrix} = \begin{bmatrix} 0_{2Nc-1}^T \\ \mathbb{I}_{2Nc-1} \\ 0_{2Nc-1} \end{bmatrix}$$  \hspace{1cm} (3.6)$$

The matrix provides a convenient way to model the delay of the $k^{th}$ path of the $i^{th}$ user as follows

$$J^l_{ik} \zeta_i \quad \forall k = 1, \ldots, K_i$$

where $l_{ik} = \lceil \tau_{ik}/T_c \rceil$ is the discretised multipath delay and $\zeta_i$ is a vector containing
the $i^{th}$ user’s spreading code sequence padded with zeros of length $N_c$ defined as

$$
\xi_i = \left[ \begin{array}{c} \alpha_i \\ 0_{N_c} \end{array} \right] \in \mathbb{C}^{2N_c \times 1}
$$

(3.7)

Upon concatenating the contents of the TDLs, the spatially and temporally sampled vector-signal can be represented using the *Spatial Temporal AArray (STAR)* manifold vector

$$
S_{ik} \otimes \mathbb{J}^{i_{ik}} \xi_i \quad \forall k = 1, \ldots, K_i
$$

(3.8)

### 3.2.3 Doppler-STAR Manifold Vector

By taking into account the discretised Doppler effect associated with the $n^{th}$ symbol interval, the STAR manifold vector in Eqn. 3.8 is extended to the *Doppler-STAR manifold vector* corresponding to the $k^{th}$ path of the $i^{th}$ user, given as

$$
\mathbf{h}_{ik} \triangleq S_{ik} \otimes (\mathbb{J}^{i_{ik}} \xi_i \otimes \mathcal{F}_{ik})
$$

(3.9)

and the vector $\mathcal{F}_{ik}$ is modelled as

$$
\mathcal{F}_{ik} \triangleq \mathcal{F}(f_{ik}) = \begin{bmatrix}
1 \\
\exp(j2\pi f_{ik} T_c) \\
\exp(j2 \cdot 2\pi f_{ik} T_c) \\
\vdots \\
\exp(j(2N_c - 1) \cdot 2\pi f_{ik} T_c)
\end{bmatrix} = \exp\left( j2\pi f_{ik} T_c \begin{bmatrix} 0 \\ 1 \\ 2 \\ \vdots \\ 2N_c - 1 \end{bmatrix} \right)
$$

(3.10)

representing the chip level Doppler frequency shift due to the Doppler frequency $f_{ik}$ at the $k^{th}$ path of the $i^{th}$ user’s. Based on Eqn. 3.9, the overall Doppler effect of the channel is the product of the chip-level Doppler component (modelled by the Doppler-STAR vector) and the symbol level Doppler component given as

$$
\mathbf{h}_{ik} \exp(j2n\pi f_{ik} T_{cs})
$$

It is worth mentioning that each Doppler-STAR vector is unique for each multipath of each user. Thus, in a multipath multiple access environment, each single path of each user is distinct from that of another user. This is the key in parametric channel estimation developed in the next section. The uniqueness of Doppler-STAR manifold vector provides a means of distinguishing each individual path during the channel estimation process.
3.2.4 Discrete-Time Signal Model

Based on the Doppler-STAR manifold vector of Eqn. 3.9, each user’s discretised signal (the $i^{th}$ user’s signal containing $K_i$ path components, say) can now be formed as

$$x_i[n] = \begin{cases} \mathbb{H}_i^{\text{prev}} \left( \beta_i \odot \mathbf{f}_i[n-1] \otimes \left( 1_{K_i} \otimes \mathbf{a}_i[n-1] \right) \right) & \text{Previous} \\ + \mathbb{H}_i \left( \beta_i \odot \mathbf{f}_i[n] \otimes \left( 1_{K_i} \otimes \mathbf{a}_i[n] \right) \right) & \text{Current} \\ + \mathbb{H}_i^{\text{next}} \left( \beta_i \odot \mathbf{f}_i[n+1] \otimes \left( 1_{K_i} \otimes \mathbf{a}_i[n+1] \right) \right) & \text{Next} \end{cases} \quad (3.11)$$

where the vector-symbol $\mathbf{a}_i[n]$ is defined in Eqn. 3.1. For the symbol level Doppler components associated with the $i^{th}$ user, the symbol $\mathbf{f}_i[n]$ is defined as

$$\mathbf{f}_i[n] = \begin{bmatrix} \exp(j2\pi f_{i1} T_{cs}) \\ \exp(j2\pi f_{i2} T_{cs}) \\ \vdots \\ \exp(j2\pi f_{iK_i} T_{cs}) \end{bmatrix} \in \mathbb{C}^{K_i \times 1}$$

The matrix $\mathbb{H}_i$ is constructed as

$$\mathbb{H}_i = \begin{bmatrix} \mathbf{h}_{i1}^T \\ \mathbf{h}_{i2}^T \\ \vdots \\ \mathbf{h}_{iK_i}^T \end{bmatrix} \in \mathbb{C}^{2N_c \times K_i}$$

containing the invariant Doppler-STAR manifold vector $\mathbf{h}_{ik} \forall k = 1, \ldots, K_i$. Furthermore, the two matrices associated with the previous and next symbols $\mathbb{H}_i^{\text{prev}}$ and $\mathbb{H}_i^{\text{next}}$, respectively, can be written as a function of $\mathbb{H}_i$ as follows:

$$\mathbb{H}_i^{\text{prev}} = (I_N \otimes (J^T)^{N_c}) \mathbb{H}_i \in \mathbb{C}^{2N_c \times K_i}$$

and

$$\mathbb{H}_i^{\text{next}} = (I_N \otimes J^{N_c}) \mathbb{H}_i \in \mathbb{C}^{2N_c \times K_i} \quad (3.14)$$

The discretised received signal-vector $\mathbf{x}[n]$ is a composition of multiple users’ signals and is given as

$$\mathbf{x}[n] = \sum_{i=1}^{M} x_i[n] + \mathbf{n}[n] \quad (3.15)$$

Now, by substituting Eqn. 3.15 with Eqn. 3.11, the received signal vector is in the form of

$$\mathbf{x}[n] = \sum_{i=1}^{M} \begin{bmatrix} \mathbb{H}_i^{\text{prev}} \left( \beta_i \odot \mathbf{f}_i[n-1] \otimes \left( 1_{K_i} \otimes \mathbf{a}_i[n-1] \right) \right) \\ + \mathbb{H}_i \left( \beta_i \odot \mathbf{f}_i[n] \otimes \left( 1_{K_i} \otimes \mathbf{a}_i[n] \right) \right) \\ + \mathbb{H}_i^{\text{next}} \left( \beta_i \odot \mathbf{f}_i[n+1] \otimes \left( 1_{K_i} \otimes \mathbf{a}_i[n+1] \right) \right) \end{bmatrix} + \mathbf{n}[n] \quad (3.16)$$
Let’s consider the 1st user as the transmitter of interest. By rearranging the terms in Eqn. 3.16, the discretised signal vector $x[n]$ can be rewritten as

$$x[n] = \underbrace{H_1}_{2Nc \times K_1} \begin{bmatrix} \beta_1 \circ f_1[n] \circ (1_{K_1} \otimes a_1[n]) \\ \beta_1 \circ f_1[n-1] \circ (1_{K_1} \otimes a_1[n-1]) \\ \beta_1 \circ f_1[n+1] \circ (1_{K_1} \otimes a_1[n+1]) \end{bmatrix}$$

$$+ \underbrace{[H_1^{\text{prev}}, H_1^{\text{next}}]}_{2Nc \times 2K_1} \begin{bmatrix} \beta_1 \circ f_1[n] \circ (1_{K_1} \otimes a_1[n]) \\ \beta_1 \circ f_1[n-1] \circ (1_{K_1} \otimes a_1[n-1]) \\ \beta_1 \circ f_1[n+1] \circ (1_{K_1} \otimes a_1[n+1]) \end{bmatrix}$$

$$+ \sum_{i=2}^{M} \underbrace{[H_i^{\text{prev}}, H_i, H_i^{\text{next}}]}_{2Nc \times 3K_i} \begin{bmatrix} \beta_i \circ f_i[n] \circ (1_{K_i} \otimes a_i[n]) \\ \beta_i \circ f_i[n-1] \circ (1_{K_i} \otimes a_i[n-1]) \\ \beta_i \circ f_i[n+1] \circ (1_{K_i} \otimes a_i[n+1]) \end{bmatrix}$$

$$+ \underbrace{n[n]}_{2Nc \times 1}$$ \hspace{1cm} (3.17)

For convenience, the dimension of each matrix is denoted in Eqn. 3.17. Hence, the discretised signal $x[n]$ can be identified to contain four components, namely the desired signal, ISI, MAI and noise as laid out in that order in Eqn. 3.17. Assuming "ergodicity", the theoretical covariance matrix $R_{xx}$ of the received signal $x[n]$ is

$$R_{xx} = \mathcal{E}\{x[n]x[n]^H\}$$

which, in practice, over a finite observation interval of $L_s$ snapshots, can be approximated by the "sample" covariance matrix as follows

$$R_{xx} = \frac{1}{L_s} \sum_{n=1}^{L_s} x[n]x[n]^H$$

In this case, the theoretical covariance matrix $R_{xx}$ can be written as

$$R_{xx} = \mathcal{E}\{x[n]x[n]^H\}$$

$$= \underbrace{H_1 \text{diag}(P_1)H_1^H}_{R_{\text{desired}}} + \underbrace{[H_1^{\text{prev}}, H_1^{\text{next}}][I_2 \otimes \text{diag}(P_1)][H_1^{\text{prev}}, H_1^{\text{next}}]^H}_{R_{\text{ISI}}}$$

$$+ \sum_{i=2}^{M} \underbrace{[H_i^{\text{prev}}, H_i, H_i^{\text{next}}][I_3 \otimes \text{diag}(P_i)][H_i^{\text{prev}}, H_i, H_i^{\text{next}}]^H}_{R_{\text{MAI}}}$$

$$+ \underbrace{\sigma_n^2 I_{2Nc}}_{R_{\text{noise}}},$$

(3.18)
where $P_i$ is the average power of each multipath given as

$$P_i = [|\beta_{i1}|^2, |\beta_{i2}|^2, \ldots, |\beta_{iK_i}|^2]^T$$

Note that $E \left\{ (\beta_i \odot f_i[n] \odot (1_{K_i} \otimes a_i[n])) (\beta_i \odot f_i[n] \odot (1_{K_i} \otimes a_i[n]))^H \right\}$ \forall i becomes (theoretically) the diagonal matrix $\text{diag}(P_i)$ due to the vector $f_i[n]$ and the vector $\beta_i$.

### 3.3 Blind Channel Estimation

#### 3.3.1 Angle, Delay and Doppler Estimation

By observing the covariance matrix $\mathbb{R}_{xx}$ given by Eqn. 3.18, it is clear that the observation space of dimensionality $2N_cN$ (dimensionality of $\mathbb{R}_{xx}$) can be partitioned into two complementary subspaces which are:

1. The “overall signal subspace” spanned by the invariant Doppler-STAR manifold vector $\{h_{ik}, \forall i, k\}$ associated with all the paths of all users. That is,
   
   (a) the desired user’s path from all $N$ antennas (columns of $H_1$);
   (b) the desired user’s ISI (columns of $\{H_1^{\text{prev}}, H_1^{\text{next}}\}$);
   (c) MAI (columns of $\{H_i^{\text{prev}}, H_i, H_i^{\text{next}} \forall i > 1\}$).

2. The noise subspace with contributions only of the additive white Gaussian noise.

This implies that the dimensionality of the signal subspace (i.e. the total number of multipaths) is $\mathcal{K} = \sum_{i=1}^{M} \mathcal{K}_i$ which should be less than the total dimensionality $2N_cN$, i.e. $\mathcal{K} < 2N_cN$. However, the paths of the desired user, in addition of belonging to the overall signal subspace, also belong to another nonlinear subspace (Doppler-STAR manifold) defined as follows

$$\mathcal{M} = \{s(\theta) \otimes (J^l \xi_l \odot \mathcal{E}(f)) \in C^{2N_cN}, \forall \theta \in \Omega_\theta, l \in \Omega_l, f \in \Omega_f\}$$

where

$$\left\{ \{\Omega_\theta, \Omega_l, \Omega_f\} = \text{parameter spaces of } \theta, l \text{ and } f \right\}$$

$$\mathcal{E}(f) = \exp(j2\pi ftc[0, 1, \ldots, (2N_c - 1)]^T)$$
3. Arrayed MIMO System

The intersection of the manifold $\mathcal{M}$ with the “overall signal subspace” will only provide the desired user’s parameters which are the directions, delays and Doppler frequencies of its paths. This intersection can be found from the following minimisation problem

$$(\theta_1, l_1, f_1) = \arg \min_{\theta, l, f} \xi(\theta, l, f)$$

where

$$\xi(\theta, l, f) = (\mathcal{S}(\theta) \otimes (\mathcal{J}_l \otimes \mathcal{E}(f)))^H \mathbb{P}_{\mathbb{E}_n} (\mathcal{S}(\theta) \otimes (\mathcal{J}_l \otimes \mathcal{E}(f)))$$

(3.19)

In Eqn. 3.19, $\mathbb{P}_{\mathbb{E}_n}$ is the projection operator of the noise subspace $\mathcal{L}[\mathbb{E}_n]$ obtained by the eigendecomposition of covariance matrix $\mathbb{R}_{xx}$ and $\mathbb{P}_{\mathbb{E}_n} = \mathbb{E}_n (\mathbb{E}_n^H \mathbb{E}_n)^{-1} \mathbb{E}_n^H = \mathbb{E}_n \mathbb{E}_n^H$. Please note that the number of signals are assumed to be known or it can be detected using methods such as in [75] based on Akaike Information Criterion (AIC)[76] and Minimum Description Length (MDL)[77][78], etc. As the signal subspace and the noise subspace are orthogonal to each other, the cost function $\xi(\theta, l, f)$ would be expected to be approximately zero for parameters at which desired paths do exist and non-zero where paths do not exist.

However, the minimisation process of $\xi(\theta, l, f)$ in Eqn. 3.19 is prohibitively complex. Nevertheless, it can be proved that Eqn. 3.19 can be split into a minimisation of a two-dimensional cost function for space-time channel parameters $(\theta, l)$ estimation followed by an extra $K_1$ (the total number of multipaths corresponding to the desired user) one-dimensional searches, (i.e. one per path) performed in parallel over the Doppler frequency shift $f$ as follows

$$(\bar{\theta}_1, \bar{l}_1) = \arg \min_{\theta, l} \left\{ (\mathcal{S}(\theta) \otimes (\mathcal{J}_l \otimes \mathcal{E}(f)))^H \mathbb{P}_{\mathbb{E}_n} (\mathcal{S}(\theta) \otimes (\mathcal{J}_l \otimes \mathcal{E}(f))) \right\}$$

(3.20)

$$f_{1k} = \arg \min_f \left\{ h_{1k}(f)^H \mathbb{E}_n \mathbb{E}_n^H h_{1k}(f) \right\} \big|_{(\bar{\theta}_1, \bar{l}_1), \forall k}$$

(3.21)

where

$$\left\{ \begin{array}{l}
(\bar{\theta}_1, \bar{l}_1) = \{(\theta_{11}, l_{11}); \ldots; (\theta_{1K_1}, l_{1K_1}); \ldots \}

h_{1k}(f) = \mathcal{S}(\theta_{1k}) \otimes (\mathcal{J}_{l_{1k}} \otimes \mathcal{E}(f))
\end{array} \right.$$  

Proof of Eqn. 3.19-3.21 can be found in the Appendix 3.7.1. Note that by denoting

$$[\mathbb{H}_1(f) = [\mathbb{h}_{11}(f), \mathbb{h}_{12}(f), \ldots, \mathbb{h}_{1K_1}(f)] \in \mathbb{C}^{2Nc \times K_1}$$

$$[\mathbb{f}_1 = [f_{11}, f_{12}, \ldots, f_{1K_1}]^T$$

2Note that theoretically the dimension of $\mathbb{R}_{xx}$ minus the multiplicity of the minimum eigenvalue of $\mathbb{R}_{xx}$ is used to specify the number of incident signals. However, for a finite observation interval, the number of signals is estimated using Akaike Information Criterion (AIC), Minimum Description Length (MDL), etc., while the power of the noise is the mean of the smaller eigenvalues, rather than the minimum eigenvalues.
Eqn. 3.21 can also be written in a compact matrix form as

\[ f_1 = \arg \min_f \text{diag} (H_1(f)H^nE_nE^nH_1(f)) \]  

Thus a number of peak searches of the spectrums obtained by the evaluation of Eqn. 3.20 and 3.21 (or Eqn. 3.22) will provide the set of direction, delay and Doppler frequency shift estimates

\[ \Xi = \{(\theta_{11}, l_{11}, f_{11}); (\theta_{12}, l_{12}, f_{12}); \ldots ; (\theta_{1K_1}, l_{1K_1}, f_{1K_1})\} \in \mathcal{C}^{K_1 \times 3} \]

Note that the set \( \Xi \) contains, in non-specific order, the path parameters of the desired user for all transmitting antennas. Based on the set \( \Xi \) of estimated parameters, the channel matrix of the desired user can be constructed as

\[ H_1 = [\mathbf{h}_{11}, \mathbf{h}_{12}, \ldots, \mathbf{h}_{1K_1}] \in \mathcal{C}^{2N_n \times K_1} \]  

with

\[ \mathbf{h}_{1k} = \mathcal{S}(\theta_{1k}) \otimes (\mathcal{J}_{l_{1k}} \otimes \mathcal{F}(f_{1k})) \]  

Paths associated with different antennas (or SIVO channels) will later be identified and grouped using Correlation Analysis Assignment described in Section 3.4.3.

### 3.3.2 Path Power Estimation

Once the set \( \Xi \) of the desired user’s multipath parameters has been found, the path power \( |\beta_{1k}|^2 \) \( \forall k = 1, \ldots, K_1 \) (i.e. for every path) associated with the \( k^{th} \) subset of \( \Xi \) can be found by performing a one-dimensional minimisation search as follows

\[ |\beta_{1k}|^2 = \arg \min_P \xi_{\text{power}}(P) \]  

where \( \xi_{\text{power}}(P) \) is a specially designed cost function defined as follows

\[ \xi_{\text{power}}(P) = \gamma \left[ \sum_{i=1, \text{eig}_i > 0} \frac{\text{eig}_i(R_{1k}(P))}{\text{trace}(R_{1k}(0))} \right] + \log_{10} \left[ \frac{\sum_{i=1, \text{eig}_i < 0} |\text{eig}_i(R_{1k}(P))|}{\text{trace}(R_{1k}(0))} \right] \]  

with

\[ R_{1k}(P) = \tilde{R}_{xx} - \sigma_n^2 \tilde{R}_{2N_n} - P \bar{h}_{1k} \bar{h}_{1k}^H \quad \forall k = 1, \ldots, K_1 \]  

and \( \gamma \) is the scaling factor and is in the order of

\[ \gamma = \mathcal{O} \left( \left. \frac{\text{step size} \cdot \|h_{1k}\|^2}{\text{trace}(R_{1k}(0))} \right) \right) \]
and $\text{eig}_i(\mathbb{R}_{1k}(P))$ represents the $i^{th}$ eigenvalue of $\mathbb{R}_{1k}(P)$ and the denominator $\text{trace}(\mathbb{R}_{1k}(0))$ is used for normalization.

From Eqn. 3.27, it is clear that the effect of noise is removed from the data covariance matrix $\mathbb{R}_{xx}$ and the effects of the $k^{th}$ path can only be completely (asymptotically) removed from $\mathbb{R}_{xx}$ when $P$ equals to $|\beta_{1k}|^2$. In this case, $\mathbb{R}_{1k}(P)$ does not contain the $k^{th}$ path statistics and the rank of $\mathbb{R}_{1k}(P)$ falls by one\textsuperscript{1}. An illustrative example and more details regarding path power estimation is presented in Section 3.5 and the complexity analysis of the proposed algorithm in Appendix 3.7.2.

### 3.4 Reception

#### 3.4.1 Space-Time Multiuser Receiver

The discretised signal vector is passed through an integrated channel estimator to obtain the channel parameters which are used to reconstruct channel matrix $[\mathbb{H}_{\text{prev}}, \mathbb{H}_i, \mathbb{H}_{\text{next}}], i = 1, \ldots, M$. With the knowledge of the channel, space-time receivers can be devised to recover the transmitted information symbols. Linear receivers are considerably less complex to implement than many non-linear designs while producing satisfactory BER/SNIR performance. In the thesis only linear receivers are considered. According to Fig. 2.9, the output of the linear receiver is a vector of decision variables at point E, i.e.

$$y[n] = \mathbb{W}^H \mathbb{z}[n]$$

The mathematical expressions for various types of space-time receivers which will be used in the subsequent chapters for simulation studies and performance comparison are given below:

1. **Rake Rx**: The RAKE receiver is a single-user (SU) receiver which requires only the knowledge of the desired user’s spreading sequence and its associated channel parameters. This is an optimum Rx but for white Gaussian channels. Its main drawback is its susceptibility to the near-far problem effect, which necessitates the application of open-loop or closed-loop power control. Its expression extended to include the Doppler-STAR is given as follows

$$\mathbb{W}_{\text{RAKE}} = \mathbb{H}_1 \in \mathbb{C}^{2N_c \times K_1}$$

\textsuperscript{1}This method, however, will give incorrect estimate if paths are partially or fully correlated and the data covariance matrix constructing $\mathbb{R}_{xx}$ has non-zero off diagonal elements.
The weight matrix is derived by constructing $H_1$ based on Eqn. 3.23.

2. **Decorrelating Rx**: The formulation of the decorrelating receiver is to eliminate the MAI and ISI interferences, thus making it more tolerant to the near-far problem effect. It is a suboptimal multiuser (MU) linear receiver and by denoting the composite channel matrix $H$ as

$$H = \begin{bmatrix} H_{1,1}, & H_{1,2}, & \cdots, & H_{1,n} \\ H_{2,1}, & H_{2,2}, & \cdots, & H_{2,n} \\ \vdots & \vdots & \ddots & \vdots \\ H_{M,1}, & H_{M,2}, & \cdots, & H_{M,n} \end{bmatrix} \in C^{2NN_c \times 3K}$$

The decorrelating weight can be expressed as

$$W_{Dec} = \text{col}_{K_1+1:2K_1} \{ \left[ H(H^H H)^{-1} \right] \} \in C^{2NN_c \times K_1} \tag{3.29}$$

where the operator $\text{col}_{i:j} \{ \cdot \}$ selects columns $i$ to $j$ of a matrix. Depending on the type of the manifold vector used to construct the weight, there are *Doppler-STAR decorrelating Rx* with weights constructed using Doppler-STAR manifold vector as in Eqn. 3.9 and *STAR decorrelating Rx* without Doppler compensation by using STAR manifold vector defined in Eqn. 3.8.

3. **MMSE Rx**: It is also a suboptimal MU linear receiver with its weight given below

$$W_{MMSE} = \text{col}_{K_1+1:2K_1} \{ \left[ H(H^H H + \sigma_n^2 I_{3K})^{-1} \right] \} \in C^{2NN_c \times K_1} \tag{3.30}$$

4. **Subspace MMSE receiver**: A subspace type MMSE MU detector is proposed in [41]. It is implemented in multiple antenna MIMO CDMA systems and is chosen for performance comparison in simulation studies in this chapter. The channel is estimated using a method proposed in [60] which is also a subspace approach. Please refer to [60] and [41] for details of this approach.

### 3.4.2 The Subspace Based Single-User Rx for Interference Cancellation

In this section, a subspace based single-user receiver is proposed which achieves the performance close to the multiuser receivers. Channel estimates obtained using algorithms proposed in Section 3.3 will be incorporated with the proposed subspace based receiver for MAI and ISI suppression. The data flow diagram of the system is shown in Fig. 3.2 where the MAI/ISI suppression (interference cancellation) blocks and the correlation analysis assignment process will be studied in this section. As discussed previously, the desired signal component, ISI and
MAI components all lie in the same signal subspace. However, in order to suppress interferences from the received signals, the interference subspace spanned by ISI and MAI components should be identified and isolated from the overall signal subspace.

Using the desired user’s channel estimates and power estimates, $R_{\text{unwanted}}$ can be constructed by removing desired user’s multipath component from $R_{xx}$ as

$$R_{\text{unwanted}} = R_{xx} - H_1 \text{diag} \left( \left[ |\beta_{11}|^2, |\beta_{12}|^2, \ldots, |\beta_{1K_1}|^2 \right]^T \right) H_1^H$$

(3.31)

where $|\beta_{1k}|^2 \forall k = 1, \ldots, K_1$ has been estimated using Eqn. 3.26. By performing the eigendecomposition of $R_{\text{unwanted}}$, the matrix $E_{\text{unwanted}}$ is formed, which has $(2K_1 + \sum_{i=2}^{M} 3K_i)$ signal eigenvectors as its columns, and is a basis for the unwanted signal subspace $\mathcal{L}[E_{\text{unwanted}}]$.

Having found the unwanted signal subspace, the complement projection operator $P_{\text{unwanted}}$ can be formed. By applying this operator on $\mathbf{z}[n]$, the received signal is projected on to the orthogonal complement subspace of unwanted signals, i.e. $P_{\text{unwanted}} \mathbf{z}[n]$. Thus based on Eqn. 3.17, it is clear that both MAI and ISI from the received signals can be successfully suppressed leaving only the transformed terms of desired signals and noise:

$$P_{\text{unwanted}} \mathbf{z}[n] = P_{\text{unwanted}} \mathbf{H}_1 \text{diag}(\beta_i) \left( f_1[n] \odot \left( I_{K_1} \otimes \mathbf{a}_1[n] \right) \right)$$

transformed desired signal

$$+ P_{\text{unwanted}} \mathbf{n}[n]$$

transformed noise

(3.33)

Then, by pre-multiplying both sides of Eqn. 3.33 with $(H_1^H P_{\text{unwanted}} H_1)^{-1} H_1^H$, Eqn. 3.33 is further transformed to

$$\left( H_1^H P_{\text{unwanted}} H_1 \right)^{-1} H_1^H P_{\text{unwanted}} \mathbf{z}[n]$$

(3.34)

$$= \left( H_1^H P_{\text{unwanted}} H_1 \right)^{-1} H_1^H P_{\text{unwanted}} \mathbf{H}_1 \text{diag}(\beta_i) \left( f_1[n] \odot \left( I_{K_1} \otimes \mathbf{a}_1[n] \right) \right)$$

$$+ \left( H_1^H P_{\text{unwanted}} H_1 \right)^{-1} H_1^H P_{\text{unwanted}} \mathbf{n}[n]$$
3. Arrayed MIMO System

Spatial Smoothing (Optional)

Space-Time Estimation

Path Power Estimation

Channel Reconstruction

Figures 3.2: Data flow diagram of the proposed space-time single user detector
$$\mathbf{W} = \text{diag} (\beta_1) \left( \mathbf{F}_1[n] \odot (\mathbf{L}_{K_1} \otimes \mathbf{a}_1[n]) \right) + \mathbf{W}_{H}[n] \in \mathbb{C}^{K_1 \times 1} \quad (3.35)$$

The weight matrix $\mathbf{W}$ is a zero-forcing (ZF) type and is given as

$$\mathbf{W} = \mathbf{P}^\perp_{\text{unwanted}} \mathbf{H}_1 (\mathbf{H}^H_1 \mathbf{P}^\perp_{\text{unwanted}} \mathbf{H}_1)^{-1} \quad (3.36)$$

### 3.4.3 Correlation Analysis Assignment

The vector signal (point E in Fig. 2.9) is of dimension $\mathcal{K}_1 = K_1 \mathcal{N}$ (representing $\mathcal{K}_1$ unique paths) which is still contaminated with the phase shifts caused by complex fading coefficients and symbol level Doppler components, plus an additive noise component. The phase differences between paths must be first compensated. This can be done by using differential decoding techniques. If $y[n]$ denotes the vector-signal at the output of the differential decoder (that is after phase compensation), the signals now need to be partitioned to differentiate those branches belonging to the different transmitting antennas of the desired user. This is completed by performing the Correlation Analysis Assignment illustrated in Fig. 3.3 where an example is given with total number of paths $\mathcal{K}_1 = 6$ to be separated. The correlation matrix of the output $\underline{y}[n]$ is first constructed as

$$\mathbb{R}_{yy} = \mathcal{E}\{y[n]y[n]^H\} \quad (3.37)$$

The lower triangular matrix contains enough information needed for grouping the paths. A $6 \times 6$ correlation matrix is shown in Fig. 3.3 where $\text{corr}_{ij} = [\mathbb{R}_{yy}]_{ij}$ denotes the entry at position $(i, j)$ of $\mathbb{R}_{yy}$ and is the correlation of the $i^{th}$ and $j^{th}$ path of $\underline{y}[n]$. It is easy to see

$$\begin{cases} 
\text{corr}_{ij} \simeq 1 & \text{if path } i, j \text{ are associated with the same signal source} \\
\text{corr}_{ij} \ll 1 & \text{if path } i, j \text{ are associated with different signal source}
\end{cases}$$

Based on the covariance matrix, the grouping algorithm is carried out and the steps are described as follows

1. Set column number $j = 1$ and antenna number $n = 1$;

2. Move to column $j$. Select $i \forall i = j, \ldots, \mathcal{K}_1$ such that the cross-correlation at position $(i, j)$ is above a prespecified threshold and group them into set $\Xi_n$, i.e. $\Xi_n = \{i, \text{corr}_{ij} \text{ > Threshold} \forall i = j, \ldots, \mathcal{K}_1\}$. Set $n = n + 1$;

3. If $j < \mathcal{K}_1$, set $j = j + 1$ (the next column), go to Step 4. If $j = \mathcal{K}_1$ (the last column), exit;
Figure 3.3: An illustration of the Correlation Analysis Assignment. Paths belonging to different antennas are represented using different colours.
4. If path $j$ doesn’t belong to any sets that have been created, i.e. $j \notin \{\Xi_1 \cup \Xi_1 \ldots \cup \Xi_n\}$, go to Step 2. Otherwise, go to Step 3.

The procedure is briefly illustrated in the example shown in Fig. 3.3. Note that if the channel parameters of any particular branch are erroneous (incorrect channel estimation) or unidentified (incomplete channel estimation), the Correlation Analysis Assignment will leave that branch unassigned, thus inducing robustness to the receiver. At the output of the correlation analysis, the vector-signal $y[n]$ has been sorted into $N$ groups, denoted as

$$
\begin{bmatrix}
    y_{\text{Ant},1}[n]^T, y_{\text{Ant},2}[n]^T, \ldots, y_{\text{Ant},N}[n]^T
\end{bmatrix}^T
$$

These branches are then maximum ratio combined (MRC) followed by the decision device yielding

$$
\hat{a}_{1j}[n] = \text{sign}(\text{Re}\{w_{\text{MRC},j}^HY_{\text{Ant},j}[n]\})
$$

where $\hat{a}_{1j}[n]$ is the $n^{th}$ recovered (differentially decoded) symbol transmitted from the $j^{th}$ antenna of the desired user, whereas $w_{\text{MRC},j}$ is the combining weight vector obtained using the principal eigenvector of the covariance matrix of $y_{\text{Ant},j}[n]$.

With reference to Fig. 3.2, a summary of the complete procedure covering the main steps of the proposed arrayed MIMO estimator and receiver is briefly described as follows:

1. Sample the array output and concatenate the tapped-delay lines (TDL) contents to form the discretised signal vector $\underline{x}[n]$ at point A;

2. Form the (smoothed) covariance matrix $R_{xx}$ of the discretised signal vector and apply the cost function in Eqn. 3.20 and 3.21 to jointly estimate the angle, delay (Point B) and Doppler frequency shift (point C);

3. Having obtained the space-time channel parameters, the path power is then estimated according to Eqn. 3.25 (point D);

4. Based on the estimated channel parameters, compute weight $W$ in Eqn. 3.36 and apply it onto the discretised signal vector as seen in Eqn. 3.35 (point E);

5. Compensate phase shift by differential decoding (point F) and apply Correlation Analysis Assignment to assign the output signal streams to different transmitting antennas (point G);
6. Combine branches that belong to the same transmitting elements using Eqn. 3.39 and multiplex the data streams to recover the data stream at point H.

### 3.5 Simulation Studies

Several representative examples are presented in this section to highlight the key benefits of using array manifold concept in MIMO systems. Consider a uniform $N = 5$ element linear array of half-wavelength spacing operating in the presence of $M = 3$ co-channel BPSK DS-CDMA users, each having $\bar{N} = 2$ transmitting antenna elements, employing data demultiplexer at the transmitter to transmit a vector of symbols that maximise the transmission data rate. Each user is assigned a unique Gold sequence of length $N_c = 31$ with rectangular chip pulse-shaping. The chip rate is set at $1/T_c = 1.2288$ Mchips/s with a carrier frequency of $F_c = 2$GHz. The array is assumed to collect 400 data symbols for processing at each time.

The 1st user is the desired user having an input signal-to-noise-ratio (SNR) of 20dB. The maximum Doppler spread is set at 200Hz which corresponds to a maximum speed of 108km/h. The Interference-to-Noise Ratio (ISR) is assumed 20dB (i.e. near-far problem) and all 3 users have 10 multipaths each, with their parameters listed in Table 3.1. Fig. 3.4 shows in both the (a) 3D surface plot and (b) 2D contour plot that all the 10 multipaths, associated with the two transmitting elements of the desired user, can be identified/estimated successfully using the proposed algorithm\(^2\).

Having estimated the space-time channel parameters, $K_1 (= 10)$ one-dimensional searches are performed over Doppler frequency domain. By setting the search step size as 1Hz, the Doppler-MUSIC spectrum due to the desired user is plotted as depicted in Fig. 3.5. It is therefore apparent that the Doppler frequency shifts associated with all 10 multipaths, can be correctly estimated.

Fig. 3.6 demonstrate the capability of the conventional MUSIC and the STAR-MUSIC in resolving closely spaced paths with identical delays. It is a single user system having two paths coming from direction $70^\circ$ and $70^\circ$ with

---

\(^2\)Notice that the algorithm can still operate even when the desired user’s paths are arriving at the same time (co-delay) or arriving from the same direction (co-directional). The former case cannot be resolved for a general array geometry but for a uniform linear array where spatial smoothing can be overlaid on top of $R_{zz}$ to form the smoothed covariance matrix $R_{\text{smooth}}$ [11] (see the optional block in Fig. 3.2). In the simulation studies, the array is partitioned into 2 overlapping 4-element subarrays for spatial smoothing.
### 3. Arrayed MIMO System

User 1 with code vector $\Omega_1$

| Path | $\theta_{1k}$ | $l_{1k}T_c$ | $f_{1k}$ | $|\beta_{1k}|^2$ | Path | $\theta_{2k}$ | $l_{2k}T_c$ | $f_{2k}$ | $|\beta_{2k}|^2$ | Path | $\theta_{3k}$ | $l_{3k}T_c$ | $f_{3k}$ | $|\beta_{3k}|^2$ |
|------|---------------|-------------|---------|----------------|------|---------------|-------------|---------|----------------|------|---------------|-------------|---------|----------------|
| 1st  |               |             |         |               | 2nd  |               |             |         |               | 3rd  |               |             |         |               |
| $k = 1$ | 40°          | $8T_c$     | 30Hz    | 0.1334        | $k = 1$ | 30°          | $10T_c$    | $-90$Hz  | 22.7490       | $k = 1$ | 20°          | $15T_c$    | 40Hz     | 3.0420        |
| $k = 2$ | 50°          | $18T_c$    | 100Hz   | 0.0633        | $k = 2$ | 70°          | $25T_c$    | 60Hz     | 45.9155       | $k = 2$ | 60°          | $4T_c$     | $-80$Hz  | 19.4572       |
| $k = 3$ | 70°          | $25T_c$    | $-160$Hz| 0.1569        | $k = 3$ | 80°          | $20T_c$    | 170Hz    | 6.5390        | $k = 3$ | 60°          | $7T_c$     | $-10$Hz | 40.6202       |
| $k = 4$ | 90°          | $18T_c$    | 100Hz   | 0.3316        | $k = 4$ | 80°          | $21T_c$    | 30Hz     | 2.8903        | $k = 4$ | 60°          | $10T_c$    | 5Hz      | 12.9542       |
| $k = 5$ | 100°         | $12T_c$    | 0Hz     | 0.3148        | $k = 5$ | 100°         | $10T_c$    | $-110$Hz | 21.9062       | $k = 5$ | 100°         | $3T_c$     | 180Hz    | 23.9264       |

2nd Antenna element

| $k = 6$ | 60°          | $15T_c$    | 150Hz   | 0.0379        | $k = 6$ | 110°         | $11T_c$    | $-100$Hz | 23.0314       | $k = 6$ | 110°         | $20T_c$    | $-2$Hz    | 28.5003       |
| $k = 7$ | 90°          | $5T_c$     | $-80$Hz | 0.2418        | $k = 7$ | 80°          | $5T_c$     | $-60$Hz  | 16.2399       | $k = 7$ | 130°         | $11T_c$    | 110Hz    | 3.1565        |
| $k = 8$ | 90°          | $20T_c$    | 100Hz   | 0.3072        | $k = 8$ | 90°          | $5T_c$     | $-60$Hz  | 51.1615       | $k = 8$ | 130°         | $12T_c$    | $-50$Hz  | 14.3002       |
| $k = 9$ | 120°         | $10T_c$    | 0Hz     | 0.0955        | $k = 9$ | 120°         | $28T_c$    | 150Hz    | 2.0691        | $k = 9$ | 140°         | $8T_c$     | 90Hz     | 35.0292       |
| $k = 10$| 130°         | $15T_c$    | $-120$Hz| 0.3177        | $k = 10$| 150°         | $25T_c$    | 0Hz      | 7.4982        | $k = 10$| 160°         | $18T_c$    | $-170$Hz| 19.0138       |

Note: (1) The first user is assumed to be the desired user with ISR=20dB and SNR=20dB;

(2) ISR is defined as $10 \log_{10}(\sum_{k=1}^{K_i} |\beta_{ik}|^2 / \sum_{k=1}^{K_i} |\beta_{1k}|^2) \forall i \neq 1, k$;

(3) Path delays are assumed multiple of $T_c$. Delays in fractional $T_c$ are absorbed by the fading coefficients in this thesis.

Table 3.1: User’s parameters
3. Arrayed MIMO System

Figure 3.4: (a) 3D surface plot and (b) 2D contour plot of the joint angle (DOA) and delay (TOA) estimation of all the multipath components of the desired user based on Eqn. 3.20. Taking position \( k = 3 \) as an example, it represents the estimates of the desired user’s 3\(^{rd} \) path of the 1\(^{st} \) antenna that arrives from direction 70° with delay 25\( T_c \) as can be found in Table 3.1.
identical delays (TOA = 5Tc) and zero Doppler frequency. It is simulated using a linear array of N = 5 and the user employs a Gold sequence of length Nc = 31. The result is obtained based on the theoretical covariance matrix of the signal (i.e. the number of snapshots approaches infinity) with zero noise. It has shown that the STAR-MUSIC (the solid line) provides sharp peaks, much better than the conventional MUSIC (the dotted line), at directions where paths exist. This is because the extension of the conventional array manifold to the Doppler-STAR manifold concept has provided the system with better detection and resolution capabilities.

Fig. 3.7 plots a one dimensional search over the fading coefficients. The fading coefficients of all the multipaths are listed in Table 3.1. Taking (User 1, Antenna 1, Path 3) for example, as the value of P increases, toward |β13|2, a portion of the signal power arriving from the 3rd path is gradually removed leading to a monotonic decrease of the first term of RHS of Eqn. 3.26. When P = |β13|2, all the effects of the signal coming from the path of interest have been removed and a deep drop can be seen at |β13|2 = 0.1569. The logarithm of the eigenvalue that is close to zero produces a sharp drop in the cost function which creates the global minimum. A further increase of P will create negative eigenvalues of ℜ13(P) and

Figure 3.5: Doppler frequency shift estimation of all the multipath components due to the desired user based on Eqn. 3.22. For example, peak k = 10 at f = −120Hz is the estimated Doppler frequency of the desired user’s 10th path.
3. Arrayed MIMO System

Figure 3.6: The performance comparison of (i) the conventional MUSIC and (ii) the STAR-MUSIC algorithms in resolving closely spaced sources ($70^\circ$ and $70.1^\circ$) with identical delays ($5T_c$) and without Doppler frequency shift.

the second term of RHS of Eqn. 3.26 starts operating. As $P$ continues growing, the negative eigenvalues are utilized constructively by the second term of the cost function to produce a monotonic increase of the cost function when the effect of the $3^{rd}$ path has been completely removed. It is important to point out that the searching range of parameter $P$ is within $\left[0, \frac{1}{2N_c} \text{trace}(R_{xx} - \text{eig}_{\text{min}}(R_{xx}) ||_{2N_c}) \right]$. The search for the fading coefficients of all the multipaths of the desired user can be efficiently realized in MATLAB by using the FMINBND function.

The channel estimates (directions, delays, Doppler Frequencies and path powers) can be found in Table 3.2. Paths associated with different antennas remain to be identified. The Correlation Analysis Assignment is thus performed at the output of processor $\mathcal{W}$. The lower triangular part of the cross-correlation matrix $R_{yy}$ is shown in the table. The procedure starts from the 1$^{st}$ column. By setting a prespecified threshold value of say 0.95 for comparison, it is clear that one set of the paths, i.e. $\Xi_1 = \{1, 2, 4, 6, 8\}$ has been successfully identified. Paths that belong to set $\Xi_1$ are identified and highlighted in column 1. The procedure now proceeds to the 2$^{nd}$ column. As path 2 belongs to the same group as path 1 thus column 2 is ignored and the procedure continues to the 3$^{rd}$ column. By looking at Column 3 (which doesn’t belong to $\Xi_1$), set $\Xi_2 = \{3, 5, 7, 9, 10\}$ is identified.
Now, let us consider a hypothetical case that there exists parameter estimation error in one of the estimated paths (say the 1st path). If the estimation error is directional of +5° (i.e. $\theta_1 = 45^\circ$ rather than $40^\circ$), then the correlation analysis results will be almost identical with the only difference being that cross correlation values at $(2, 1), (4, 1), (6, 1), (8, 1)$ fall below the threshold ($0.95$). Thus this path will no longer be recognized as a path associated with the first antenna and will be left unassigned. In this way, the proposed receiver is robust to channel estimation errors\(^3\).

As the proposed estimation algorithm is a searching algorithm, its accuracy is first studied as the resolution of the searching grid improves. Without loss of generality, the standard deviation between the channel estimates of the desired user’s 1st path (see Table 3.1) and its true channel vector is calculated in Fig. 3.8. The searching stepsize is set to be the same for both DOA and Doppler shift estimation. The stepsize for TOA estimation is fixed at $1T_c$ as oversampling is not

---

\(^3\)Pilot symbols (antenna IDs) can be inserted after the demultiplexer at the transmitter (see Fig. 2.9). This ID is exploited at the receiver before the multiplexer converter to combine branches in a correct order. The correlation analysis assignment developed in this section has further enhanced the system performance by introducing robustness against channel estimation errors.
### Table 3.2: Estimated set and correlation analysis

<table>
<thead>
<tr>
<th>Path</th>
<th>DOA/TOA/Doppler</th>
<th>Power</th>
<th>$\mathbf{R}_{yy}$</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>$(\theta_{1k}, l_{1k}T_c, f_{1k}Hz)$</td>
<td>$</td>
<td>\beta_{1k}</td>
</tr>
<tr>
<td>1</td>
<td>$(40^\circ, 8, 30)$</td>
<td>0.1334</td>
<td>1.000</td>
</tr>
<tr>
<td>2</td>
<td>$(50^\circ, 18, 100)$</td>
<td>0.0633</td>
<td>0.9981</td>
</tr>
<tr>
<td>3</td>
<td>$(60^\circ, 15, 150)$</td>
<td>0.0379</td>
<td>0.101</td>
</tr>
<tr>
<td>4</td>
<td>$(70^\circ, 25, -160)$</td>
<td>0.1569</td>
<td>0.9951</td>
</tr>
<tr>
<td>5</td>
<td>$(90^\circ, 5, -80)$</td>
<td>0.2418</td>
<td>0.1034</td>
</tr>
<tr>
<td>6</td>
<td>$(90^\circ, 18, 100)$</td>
<td>0.3316</td>
<td>0.9977</td>
</tr>
<tr>
<td>7</td>
<td>$(90^\circ, 20, 100)$</td>
<td>0.3072</td>
<td>0.1012</td>
</tr>
<tr>
<td>8</td>
<td>$(100^\circ, 12, 0)$</td>
<td>0.3148</td>
<td>0.9946</td>
</tr>
<tr>
<td>9</td>
<td>$(120^\circ, 10, 0)$</td>
<td>0.0955</td>
<td>0.1001</td>
</tr>
<tr>
<td>10</td>
<td>$(130^\circ, 15, -120)$</td>
<td>0.3177</td>
<td>0.1061</td>
</tr>
</tbody>
</table>

Note: $\mathbf{R}_{yy}$ is calculated using Eqn. 3.37 and the lower triangular elements of $\mathbf{R}_{yy}$ are listed in the table. Elements of $\mathbf{R}_{yy}$ at highlighted positions (path 1, 2, 4, 6, 8) have values above threshold (i.e. 0.95), thus can be identified to belong to the same transmitting antenna.
Figure 3.8: The standard deviation of the channel estimates to the true channel vector in the presence of noise as the searching resolution improves. The composite channel estimation method proposed in [60] is also plotted for comparison.

considered in the chapter. The solid curve in Fig. 3.8 is the composite channel vector reconstructed using the proposed algorithm. It is compared to the channel vector (dotted line) obtained using the non-parametric method proposed in [60]. It is clear that the parameter estimation based method has a superior accuracy over the composite channel estimation approach. It can be found that, for the proposed algorithm, the searching stepsize of 0.1 for both DOA and Doppler is a good compromise between the system performance and complexity. The performance degradation can hardly be observed when the searching stepsize is smaller than 0.1.

The overall performance of the proposed arrayed MIMO system as the Doppler frequency shift increases is examined through comparison with the Doppler-STAR RAKE Rx, Doppler-STAR Decorrelating Rx and Rx proposed in [41]. Fig. 3.9 depicts the performance of the proposed Doppler-STAR receiver as compared with the above receiver types. By keeping all the other simulation parameters constant, the maximum Doppler spread varied from 0 to 200Hz. The Doppler-STAR decorrelating receiver assumes full knowledge of the Doppler-STAR manifold vectors of all users and thus serves as a comparative upper bound for the performance of single receivers. It can be seen that our proposed receiver, which
requires only the Doppler-STAR manifold vectors of the desired user, seems to be very tolerant to the Doppler spread and achieves the performance as good as the Doppler-STAR decorrelating receiver. The receiver proposed in [41] experiences a performance degradation as the Doppler spread increases. This is because the front-end estimator is based on the non-parametric channel model and it estimates the overall channel response vector without estimating each of the channel parameters. Although the algorithm has lower computational cost, the price paid for the efficiency is a considerable reduction in the receiver’s ability to combat interferences and Doppler spread. The performance of the STAR decorrelating receiver which is formed using STAR manifold vector and thus ignores the Doppler component degraded significantly with the increase of the Doppler spread. In addition, the Doppler-STAR RAKE (using Doppler-STAR manifold vector) receiver and the STAR RAKE (using STAR manifold vector) receiver are also plotted which break down at the onset of the Doppler spread. This is because the magnitude of the interferences is not suppressed and remains very large.

As stated previously, the number of multipaths that can be resolved by the algorithm is much greater than the number of antenna elements available at the
Figure 3.10: The behaviour of the proposed receiver is investigated as the total number of multipaths increases. PN codes of length $N_c = 31$ and a linear array of $N = 5$ are used in the simulation which make the dimension of the observation space $2N_cN = 310$. ISR is kept constant at $-20$ dB for a fair comparison. The performance of the proposed receiver degrades sharply when the number of multipaths exceeds the dimension of the overall space.

array. However, the maximum number of resolvable multipaths (including ISI) $K_{\text{max}}$ should not exceed the dimension of the overall observation space which in other words, must fulfill the condition $K_{\text{max}} < 2N_cN$. Fig. 3.10 is thus plotted to show the SNIR performance of the system in the presence of MAI. Using the same system parameters as listed in Table 3.1, the limit $K_{\text{max}} < 2N_cN = 310$. The simulation starts from 60 multipaths of all users and increases up to 310 which reaches the dimension of the observation space. It can be seen that the proposed receiver can work in a fully loaded system. The SNIR level falls below 0 dB when the number of total multipaths exceeds 310. Rx in [41], however, experiences a performance downgrade even before the number of total multipaths reaches the system limit. This is because the Doppler spread has deteriorated the performance of the channel estimator. The decorrelating receiver is not a subspace type and thus does not suffer as much as the proposed receiver.

Furthermore, the near-far resistant capability of the proposed receiver is investigated in greater detail in Fig. 3.11. The power level of the other two interferers is varied with respect to that of the desired user, with each constituting a near-far
Figure 3.11: The SNIR performance of the proposed receiver in the presence of near-far problem as compared with i) Receiver proposed in [41]; ii) Doppler-STAR decorrelating receiver and iii) Doppler-STAR RAKE receiver.

ratio (NFR) as measured in ISR of between −20dB and 40dB. It is evident that the proposed receiver is insusceptible to the near-far problem and it has the same performance as the decorrelating receiver which however, requires the full knowledge of all the users’ channel parameters. Receiver proposed in [41] is subspace type and thus is also near far resistant, whereas the RAKE receiver is observed to be deteriorating with the NFR shortly after the interference level rises beyond the desired user’s signal power.

3.6 Conclusions

In this chapter, the potential benefit of incorporating the array manifold concept in typical MIMO systems is demonstrated by the proposed blind space-time channel estimation and interference cancellation scheme. The estimation algorithm can provide a complete estimation of angles, delays, Doppler frequency shifts, as well as the power of multipath coefficients. The subspace type receiver is a single user receiver which requires only the channel estimation of the desired user. However, it completely suppresses channel interferences and in the simulation studies,
it is shown that the proposed receiver achieves an SNIR enhancement comparative to a multiuser receiver. Furthermore, the proposed subspace type receiver exploits the signal structure in the joint space-time domain. The number of multipaths that can be resolved by the algorithm is much greater than the number of antenna elements available at the array. As a result, the proposed subspace receiver has superior capability of interference cancellation and can support more users for a given performance threshold. In addition, due to the inclusion of the Correlation Analysis Assignment in the detection process, the receiver is robust to erroneous or incomplete channel estimation.

3.7 Appendix

3.7.1 Proof of Eqn. 3.19-3.21:

Eqn. (21) is rewritten here as

$$\xi(\theta, l, f) = (\mathcal{S}(\theta) \otimes (\mathcal{J} \mathcal{C}_1 \otimes \mathcal{E}(f)))^H \mathbb{P}_{\mathbb{E}_n} (\mathcal{S}(\theta) \otimes (\mathcal{J} \mathcal{C}_1 \otimes \mathcal{E}(f)))$$

(3.40)

which can be rearranged into

$$\xi(\theta, l, f) = \text{trace}(\text{diag}(\mathcal{J} \mathcal{C}_1)(\mathcal{S}(\theta) \otimes \mathcal{I}_{2N_c})^H \mathbb{P}_{\mathbb{E}_n} (\mathcal{S}(\theta) \otimes \mathcal{I}_{2N_c})\text{diag}(\mathcal{J} \mathcal{C}_1)\mathcal{E}(f)\mathcal{E}(f)^H)$$

(3.41)

where \(\mathcal{E}(f)\) is defined as the chip level Doppler component with the form of

$$\mathcal{E}(f) = \begin{bmatrix} 1 \\ \exp(j2\pi fT_c) \\ \exp(j2 \cdot 2\pi fT_c) \\ \vdots \\ \exp(j(2N_c - 1) \cdot 2\pi fT_c) \end{bmatrix}$$

and \(\mathcal{E}(f)\mathcal{E}(f)^H\) is therefore a Toeplitz matrix with expression as

$$\mathcal{E}(f)\mathcal{E}(f)^H = \begin{bmatrix} 1 & \exp(-j2\pi fT_c) & \cdots & \exp(-j(2N_c - 1)2\pi fT_c) \\ \exp(j2\pi fT_c) & 1 & \cdots & \vdots \\ \vdots & \vdots & \ddots & \exp(-j2\pi fT_c) \\ \exp(j(2N_c - 1)2\pi fT_c) & \cdots & \cdots & 1 \end{bmatrix}$$
which can be split into two terms expressed as
\[
\mathcal{F}(f)\mathcal{F}(f)^H = \mathbf{1}_{2N_c} \mathbf{1}_{2N_c}^T + \mathbf{M}
\] (3.42)

where \(\mathbf{1}_{2N_c}\) is defined as a vector of length \(2N_c\) with all elements ones and
\[
\mathbf{M} = \begin{bmatrix}
0 & \exp(-j2\pi f T_c) - 1 & \cdots & \exp(-j(2N_c - 1)2\pi f T_c) - 1 \\
\exp(j2\pi f T_c) - 1 & 0 & \cdots & \\
\vdots & \vdots & \ddots & \exp(-j2\pi f T_c) - 1 \\
\exp(j(2N_c - 1)2\pi f T_c) - 1 & \cdots & 0 & 
\end{bmatrix}
\] (3.43)

which is the term that remains. Plug Eqn. 3.42 into Eqn. 3.41 to obtain
\[
\xi(\theta, l, f) = \text{trace}(\text{diag}(J^\dagger \xi_1)(S(\theta) \otimes \mathbf{I}_{2N_c})^H \mathbb{P}_{\mathbb{S}_n}(S(\theta) \otimes \mathbf{I}_{2N_c}) \text{diag}(J^\dagger \xi_1)(\mathbf{1}_{2N_c} \mathbf{1}_{2N_c}^T + \mathbf{M}))
\]

If \(N_c f_D T_c \ll 1\) which can be easily satisfied in a practical communication system, the second term in Eqn. 3.44 is close to 0 and can be ignored. In a similar fashion as we transformed Eqn. 3.40 to Eqn. 3.41, the first term of Eqn. 3.44 can be rearranged as
\[
\text{trace}(\text{diag}(J^\dagger \xi_1)(S(\theta) \otimes \mathbf{I}_{2N_c})^H \mathbb{P}_{\mathbb{S}_n}(S(\theta) \otimes \mathbf{I}_{2N_c}) \text{diag}(J^\dagger \xi_1)(\mathbf{1}_{2N_c} \mathbf{1}_{2N_c}^T))
\]

Rewrite Eqn. 3.20 in Section 3.3 as
\[
(\theta_1, l_1) = \arg \min_{\theta, f} \left\{ (S(\theta) \otimes J^\dagger \xi_1)^H \mathbb{E}_n \mathbb{E}_n^H (S(\theta) \otimes J^\dagger \xi_1) \right\} \]

(3.45)

It can thus be proved that the minimisation of Eqn. 3.40 shares the same minimum points as Eqn. 3.45. By searching over \(\theta\) and \(l\) using Eqn. 3.45, the estimations \((\theta_1, l_1)\) will be obtained. By defining
\[
h_{1k}(f) = S(\theta_{1k}) \otimes (J^\dagger \xi_1 \otimes \mathcal{F}(f))
\]

where \((\theta_{1k}, l_{1k})\) is the \(k\)th set of the estimates \((\theta_1, l_1)\), Eqn. 3.40 is reduced to a \(K_1\) one dimensional search over the Doppler frequency shift written as
\[
f_{1k} = \arg \min_f \left\{ h_{1k}(f)^H \mathbb{E}_n \mathbb{E}_n^H h_{1k}(f) \right\} \mid_{(\theta_{1k}, l_{1k}), y_k}
\]

(3.46)
Consequently, it has been proved that the minimisation of Eqn. 3.40 can be split into a sequential minimisation of a two-dimensional cost function for space-time channel parameters \((\theta, l)\) estimation (Eqn. 3.45) followed by an extra \(K_1\) (the total number of multipaths corresponding to the desired user) one-dimensional searches for Doppler frequency shift estimation \(f\) (Eqn. 3.46).

### 3.7.2 Complexity Analysis

The computational complexity of the proposed algorithm will be studied by counting the number of “floating-point operations” (FLOPs) required by the algorithms. Every operation is treated as complex multiplication and addition. The complexity calculation of the algorithms can be decomposed into several steps with the common operations listed as follows:

1. According to Eqn. 3.18, the calculation of the covariance matrix \(R_{xx}\) requires \(1/2 \left[ 2NN_c + (2NN_c)^2 \right] (8L - 1)\);

2. The flop count of the eigenvalue decomposition of \(R_{xx}\) is \((2NN_c)^3\);

3. The calculation of the projection operator \(P_{\mathbb{E}_s}\) requires \(1/2 \left[ 2NN_c + (2NN_c)^2 \right] (24K - 1)\);

4. The form of STAR and Doppler-STAR have approximately equal complexity as \(40N + 80N_c + 2NN_c\) (note that expectation is equivalent to 40 multiplications);

5. The total throughput calculation of Eqn. 3.19 has the complexity \(8(2NN_c)^2 + 6(2NN_c) - 2\);

The computational expenses of the three-dimensional minimisation process, the reduced dimensional minimisation process and the power estimation process will be briefly summarised as follows

a) **Three-Dimensional Minimisation Process (Eqn. 3.19):**

The three-dimensional cost function as in Eqn. 3.19 can be carried out with a brute-force search over the joint direction, time and Doppler frequency domain. However, for implementation purpose, all the possible STAR and Doppler-STAR manifold vectors can be precalculated (Step 4) and stored, thus will not be counted into the real time processing cost.

By defining \(N_\theta\), \(N_l\) and \(N_f\) as the number of searches over angular spread, time spread and Doppler frequency spread, steps 1-3 and 5 are repeated \(N_\theta N_l N_f\).
times. Hence, the flop count of the three-dimensional minimisation process can be calculated as in the order of

\[ \mathcal{O}(8N^3N_c^3 + 16N^2N_c^2L + 48N^2N_c^2K + 32N^2N_c^2N_\theta N_lN_f) \]

given the STAR and Doppler-STAR manifold vector sets are precalculated and stored for reuse.

b) Reduced-Dimensional Minimisation Process (Eqn. 3.20 and Eqn. 3.21):

The reduced-dimensional cost function is composed of a two-dimensional search over joint direction and time domain followed by a one-dimensional search over Doppler frequency domain. Hence, the flop count of the search process is reduced to

\[ \mathcal{O}(8N^3N_c^3 + 16N^2N_c^2L + 48N^2N_c^2K + 32N^2N_c^2N_\theta N_l + 32N^2N_c^2N_f) \]

c) Power Estimation Process:

Having obtained the estimated channel parameters such as DOA, TOA and Doppler frequency shift, the power estimation can be performed based on Eqn. 3.25. For each path of the desired user, the flop count for constructing the residual matrix \( \mathbb{R}_{1k}(P) \) as in Eqn. 3.27 is \((2NN_c)^3 + 2(2NN_c)^2 + 4(2NN_c - 1)\). By defining \( N_P \) as the number of searches over the range of power, the total complexity cost of the power estimation process is

\[ \mathcal{O}(8NPN_c^3) \]
Chapter 4

Joint Transmitter-Receiver Beamforming

In this chapter, the objective is to provide more flexibility and enhancement of the system capabilities and performance using a joint transmitter-receiver (Tx-Rx) beamformer in the downlink of a DS-CDMA system over multipath fading channels. The proposed investigation is based on the array manifold concept and thus, the space-time properties of the channel can be fully exploited, playing a crucial role in the operation of blind channel estimation and interference suppression techniques. The beamforming weights are designed to minimise the overall mean-squared-error (MSE) of the system. An iterative solution to the optimisation problem is firstly proposed under the system framework. A closed-form solution based on channel eigendecomposition is then presented. The convergence of the iterative method to the closed-form solution and their equivalence is examined using computer simulation studies. The performance of the proposed approach is also supported by some illustrative examples.
4.1 Introduction

The explosive demand for wireless internet and multimedia services requires more efficient techniques to be employed to enable high data rate downlink transmission in DS-CDMA systems. The use of beamforming is considered as an effective way to eliminate both multiple-access interference (MAI) and inter-symbol interference (ISI). While receiver-beamforming (Rx-beamforming) has long been an active area of research, the physical limitation of a mobile station (MS), such as the space and battery power prohibits more advanced array processing at the MS. Transmitter-beamforming (Tx-beamforming) from the base station (BS) to MS serves as a powerful alternative for increasing downlink capacity and it is based on criteria such as maximum signal-to-noise ratio (MSNR) \[44\][47] and maximum pseudo signal-to-noise plus interference ratio (PSNIR) \[47\]. In \[79\], joint transmit beamforming and power control is studied in a downlink wireless system where the transmit weight vector and the power allocation are calculated jointly so that the transmit power is minimised while the SNIR at each mobile is maintained above a certain threshold.

Currently more attention has been directed to joint transmitter-receiver beamforming (Tx-Rx beamforming) that offers a performance advantage over Tx-beamforming, or Rx-beamforming. Fig. 2.10 shows a joint Tx-Rx beamforming system where both mobiles and base stations are equipped with antenna arrays. When perfect/partial channel state information (CSI) is available at the transmitter, further signal processing can be applied prior to transmission. However, joint Tx-Rx beamforming affects the interference at all receivers at different locations. Therefore, the beamformer calculation cannot be done independently at each link. Joint Tx-Rx beamforming are developed under certain performance criteria for different systems. In \[80\], an iterative optimisation scheme that minimises the overall mean squared error (MSE) of the system is proposed over multipath channels in a synchronous multiuser system. In \[81\], joint beamforming is considered in MIMO channels arising from the use of multiple antennas at both the transmitter and the receiver and the channel is considered to be flat fading. In \[82\], the optimisation design is studied in the multicarrier MIMO system covering many desirable criterions such as MSE, SNIR and BER. \[83\] studies joint Tx-Rx beamforming in MIMO channels in a frequency selective DS-CDMA system. The signal processor at the transmitter and the receiver has a decoupled structure and the optimum Tx-Rx beamforming weights are optimised iteratively. Interestingly, in \[83\], it is pointed out that the system performance is not particularly sensitive
to channel estimation errors occurring at the transmitter because an optimal linear receiver is employed which is designed to be matched to the channel plus the transmit processing at the transmitter. If the Tx-beamformer is not optimum but the information of the channel plus the transmit processing at the transmitter is estimated correctly at the receiver, then the receiver can still reduce the influence of the transmitter mismatching to a certain extent and optimise the overall link quality.

However, early results on joint Tx-Rx beamforming in MIMO systems were developed based on non-parametric models. The array geometry is ignored in all the above mentioned works and therefore limits considerably the ability of antenna arrays in exploiting efficiently the inherent space-time structure of the multipath channels. The study presented in this chapter will be based on the array manifold concept for designing a joint Tx-Rx beamforming framework for “downlink" with significantly enhanced multiple-access interference suppression capabilities over multipath channels. In Section 4.2, the antenna array system over space-time fading channels is modelled. Small aperture arrays with half wavelength spacing are employed at both sides of the transmission link. In Section 4.3, two approaches for joint Tx-Rx beamforming are proposed under the MMSE criteria. The optimisation problem is first solved iteratively. Then a closed-form solution is derived based on the channel eigendecomposition. In Section 4.4 numerical simulations are presented. The proposed approaches are also supported by a number of illustrative examples showing a considerable performance improvement of the proposed algorithm relative to other Rx- and Tx-beamforming algorithms. The chapter is concluded in Section 4.5.

4.2 Downlink Signal Modelling

Let us consider an M user QPSK or BPSK DS-CDMA system with \( N \) transmit and \( N \) receive antennas. Fig. 2.10 shows a typical arrayed system for joint Tx-Rx beamforming. The \( i^{th} \) user’s data symbols \( a_i[n], \forall n \in \mathcal{Z} \) are first spreaded with a unique spreading code \( \alpha_i[p] \in \pm 1, p = 0, \ldots, N_c - 1 \) of period \( N_c \) and are then pre-processed with a unity-norm beamforming weight vector \( \overline{w}_i \in \mathbb{C}^{N \times 1} \) before transmission. The baseband transmitted signal is given by

\[
m(t) = \sum_{i=1}^{M} \sqrt{P_i} \overline{w}_i \sum_{n=-\infty}^{+\infty} a_i[n] \sum_{p=0}^{N_c-1} \alpha_i[p] c(t - nT_{cs} - pT_c) \tag{4.1}
\]

with \( nT_{cs} + pT_c \leq t < nT_{cs} + (p + 1)T_c \)
where

\[
\begin{aligned}
P_i &= \text{transmission power of the } i\text{th user} \\
T_{cs} &= \text{data symbol period} \\
T_c &= \text{chip period} \\
c(t) &= \text{unit amplitude chip pulse-shaping waveform of duration } T_c
\end{aligned}
\]

According to Fig. 2.10, the channel model is a composition of many single link VIVO channels shown in Fig. 4.1. Note that in this chapter, the slow fading channel is considered and the Doppler component is excluded from the model. Let us assume the transmitted signal arrives at the \( i\)th user via \( K_i \) multipaths.

Consider that the \( k\)th path of the \( i\)th user has the direction-of-departure (DOD) \((\bar{\theta}_{ik}, \bar{\phi}_{ik})\) and the direction-of-arrival (DOA) \((\theta_{ik}, \phi_{ik})\). Please note that a bar (\( \bar{A}\) or \( \bar{a}\)) represents a transmitter’s parameter. The symbols \( \beta_{ik} \) and \( \tau_{ik} \) denote respectively the complex fading coefficient and path delay due to the \( k\)th path of the \( i\)th user. Based on this model, the received baseband vector-signal of the \( i\)th MS (user) can be written explicitly as

\[
x_i(t) = \sum_{k=1}^{K_i} \beta_{ik} \overline{S}_{ik} \overline{S}_{ik}^T m(t - \tau_{ik}) + \mathbf{n}_i(t) \tag{4.2}
\]

where \( \mathbf{n}_i(t) \) is the complex white Gaussian noise vector of covariance matrix \( \sigma_n^2 \mathbb{I}_N \).

The array manifold vector \( \overline{S}_{ik} \) is defined in Eqn. 2.1 and the transmitter manifold vector \( \overline{S}_{ik} \triangleq \overline{S}(\bar{\theta}_{ik}, \bar{\phi}_{ik}) \) can also be defined in the same fashion.

![Figure 4.1: The Vector-Input Vector-Output (VIVO) slow fading channel](image-url)
The discretised baseband signal vector $\mathbf{x}[n]$ is formed by sampling $x(t)$ with period $T_s = T_c$ which is then passed through a bank of $N$ tapped-delay lines (TDL). To model the discretised signal $\mathbf{x}_i[n]$ at the output of the TDL, let us define the following Spatio-Temporal Array (STAR) manifold matrix

$$H_{ik,j} = \left(\frac{S_{ik}}{\varsigma_{ik}}\right) \otimes (J^T) \in \mathbb{C}^{2NN_c \times N}$$

(4.3)

associated with the $j^{th}$ user’s signal and arriving at the $i^{th}$ user via the $k^{th}$ path; $l_{ik} = \lceil \tau_{ik}/T_c \rceil$ is the discretised multipath delay and the matrix $J$ (or $J^T$) is defined as in Eqn. 3.6.

By defining the vector $\mathbf{a}[n] = [a_1[n], \ldots, a_M[n]]^T$ which denotes the data symbols of all users, the discretised received vector-signal $\mathbf{z}_i[n]$ at the $i^{th}$ user (point E in Fig. 2.10) can be modelled as

$$\mathbf{z}_i[n] = \left[ H_{i;1}^{\text{prev}} \mathbb{H}_i \mathbb{H}_i^{\text{next}} \right] (I_N \otimes \overline{\mathbb{W}}) \left[ \begin{array}{c} \mathbf{a}[n-1] \\ \mathbf{a}[n] \\ \mathbf{a}[n+1] \end{array} \right] + \mathbf{n}_i[n]$$

(4.4)

where $\mathbf{n}_i[n]$ represents the discretised noise vector. The matrix $\overline{\mathbb{W}}$ contains all the users’ Tx-beamforming vectors and is defined as follows

$$\overline{\mathbb{W}} = \left[ \begin{array}{cccc} \mathbb{W}_1 & 0_N & \cdots & 0_N \\ 0_N & \mathbb{W}_2 & \cdots & \vdots \\ \vdots & \vdots & \ddots & \vdots \\ 0_N & 0_N & \cdots & \mathbb{W}_M \end{array} \right] \in \mathbb{C}^{MN \times M}$$

(4.5)

The composite channel matrix

$$\mathbb{H}_i = [\mathbb{H}_{i,1}, \mathbb{H}_{i,2}, \ldots, \mathbb{H}_{i,M}] \in \mathbb{C}^{2NN_c \times MN}$$

contains the channel matrices of all users’ signals that arrive at the $i^{th}$ user with $\mathbb{H}_{i,j}$ $\forall j = 1, \ldots, M$ modelled as

$$\mathbb{H}_{i,j} = \sqrt{P_j} \sum_{k=1}^{K_i} \beta_{ik} \mathbb{H}_{ik,j} \in \mathbb{C}^{2NN_c \times N}$$

Furthermore, the two matrices associated with the previous and next symbols $\mathbb{H}_i^{\text{prev}}$ and $\mathbb{H}_i^{\text{next}}$, respectively, can be written as a function of $\mathbb{H}_i$ as follows:

$$\mathbb{H}_i^{\text{prev}} = (I_N \otimes (J^T)^{N_c}) \mathbb{H}_i$$

$$\mathbb{H}_i^{\text{next}} = (I_N \otimes J^{N_c}) \mathbb{H}_i$$
4.3 Joint Tx-Rx Beamforming

Joint Tx-Rx beamforming is characterised by the interaction of the optimisation at both transmitter and receiver sides. The interaction procedure can be realised by an iterative Tx-Rx optimisation approach in this chapter. It is assumed that both the transmitter and receiver have perfect knowledge of the channel, though subspace based blind methods can be applied at either the Tx and Rx to retrieve channel state information. With the channel and weight information available, the Rx-beamforming weight is first adapted according to the MMSE criterion. By denoting the receiver weight as $w_i$, the MMSE receiver of the $i^{th}$ user (MS) can be constructed as

$$w_i = \left[\bar{\mathcal{C}}_i \bar{\mathcal{C}}_i^H + \sigma_I^2 \mathbb{I}_{2NN_c}\right]^{-1} \bar{\mathcal{C}}_i \bar{\mathcal{E}}_{M+i} \tag{4.6}$$

where $\bar{\mathcal{C}}_i$ has been defined in Eqn. 4.4 and $\bar{\mathcal{E}}_k$ is a $3M \times 1$ column selection vector defined as

$$\bar{\mathcal{E}}_k = \begin{bmatrix} 0 \\ \vdots \\ 0 \\ 1 \\ 0 \\ \vdots \\ 0 \end{bmatrix} \quad \text{the } k^{th} \text{ element } \in \mathbb{C}^{3M \times 1}$$

which selects the $k^{th}$ column of the matrix $\left[\bar{\mathcal{C}}_i \bar{\mathcal{C}}_i^H + \sigma_I^2 \mathbb{I}_{2NN_c}\right]^{-1} \bar{\mathcal{C}}_i$. The output signal of the $i^{th}$ user (MS) is given as

$$y_i[n] = w_i^H \bar{x}_i[n]$$

After the weight vector $w_i$ at the $i^{th}$ receiver (MS) side has been constructed, each user’s Tx-beamforming vector $\bar{\mathcal{W}}$, $\forall i$ is optimised based on the minimisation of the MSE of all users. That is

$$J = \min_{\bar{\mathcal{W}}} \left\{ \sum_{i=1}^{M} \mathcal{E} \left\{ |a_i[n] - y_i[n]|^2 \right\} \right\} \tag{4.7}$$

with the constraint that

$$\|\bar{\mathcal{W}}\| = 1 \quad \forall i = 1, \ldots, M$$
4. Joint Transmitter-Receiver Beamforming

4.3.1 An Iterative Method Based On Lagrange Multipliers

To solve the above constrained optimisation problem, the method of Lagrange multipliers is employed so as to convert the constrained minimisation problem into an unconstrained minimisation problem. Therefore, the augmented unconstrained cost function is

\[ J = \min_{\mathbf{w}, \lambda_i, \forall i} \left\{ \sum_{i=1}^{M} \mathcal{E} \left\{ \left| \mathbf{a}_i[n] - \mathbf{w}_i^H \mathbf{z}_i[n] \right|^2 \right\} + \sum_{i=1}^{M} \lambda_i (\|\mathbf{w}_i\| - 1) \right\} \]  

(4.8)

It can be proved (see Appendix 4.6.1) that, for given receivers’ weight vector \(\mathbf{w}_i\), the optimal transmitter associated with the \(i^{th}\) user is given as follows

\[ \mathbf{w}_i = (\mathbf{G}_i^H \mathbf{G}_i + \mathbf{I}_N \lambda_i)^{-1} \mathbf{H}_{i,i}^H \mathbf{w}_i \]  

(4.9)

where

\[ \mathbf{G}_i = (\mathbf{I}_3 \otimes \mathbf{W}^H) [\mathbf{H}_i^{\text{prev}}^T, \mathbf{H}_i^T, \mathbf{H}_i^{\text{next}}^T]^T \in \mathbb{C}^{3M \times N} \]  

(4.10)

In Eqn. 4.10, the receivers’ weight matrix \(\mathbf{W}\) is defined as follows

\[ \mathbf{W} = \begin{bmatrix} \mathbf{w}_1 & 0_{2N_c} & \cdots & 0_{2N_c} \\ 0_{2N_c} & \mathbf{w}_2 & \cdots & \vdots \\ \vdots & \ddots & \ddots & \vdots \\ 0_{2N_c} & \cdots & \cdots & \mathbf{w}_M \end{bmatrix} \in \mathbb{C}^{2N_c \times M} \]

with \(\mathbf{w}_i, \forall i = 1, \ldots, M\) obtained using Eqn. 4.6. Furthermore, the matrix \(\mathbf{H}_i = [\mathbf{H}_{1,i}^T, \mathbf{H}_{2,i}^T, \ldots, \mathbf{H}_{M,i}^T]^T \in \mathbb{C}^{2N_c \times M} \) represents the signal components of the \(i^{th}\) user and received by all \(M\) users in the system. Given both the Tx- and Rx-beamforming vectors, it can be proved (see Appendix 4.6.1) that the Lagrange multiplier is given by

\[ \lambda_i = \mathbf{w}_i^H \mathbf{H}_{i,i}^H \mathbf{w}_i - \mathbf{w}_i^H \mathbf{G}_i^H \mathbf{G}_i \mathbf{w}_i \]  

(4.11)

From the above analysis, it can be seen that the Tx-beamformer and the Rx-beamformer as well as the Lagrange multiplier are functions of each other, thus an iterative algorithm is suggested to obtain the solution. Thus, the proposed joint Tx-Rx optimisation can be accomplished via steps listed below

1. Initialise transmitter weight vectors \(\overline{\mathbf{w}}, \forall i\) as \(\overline{\mathbf{w}}_i = \frac{1}{\sqrt{N}} \forall i\);  
2. Calculate the MMSE Rx-beamforming vectors \(\overline{\mathbf{w}}_i, \forall i\) using Eqn. 4.6;
3. Determine the corresponding Lagrange multiplier $\lambda_i \forall i$ using Eqn. 4.11, given both the transmitter and receiver weight vectors;

4. Using the receiver weight vector $w_i$ of step 2 and the Lagrange multiplier $\lambda_i$ obtained in Step 3, update the transmitter weight $\overline{w}_i$ via Eqn. 4.9;

5. Normalise transmitter weight vectors $\overline{w}_i$ to have unity norm;

6. Repeat Steps 2-5 until the convergence is reached.

### 4.3.2 A Closed-Form Solution Based On Channel Eigen-decomposition

In this section, an approximate closed-form solution to the minimisation problem of Eqn. 4.7 is derived by performing the eigendecomposition of the channels. It can be proved that when users’ spreading codes maintain good orthogonality, such as in a lightly or medium loaded CDMA system, the joint Tx-Rx beamforming can be obtained in one step. Simulation results also show that the iterative approach converges to the closed-form solution in such scenarios.

From Eqn. 4.7, the MSE of all users can be derived as

$$\sum_{i=1}^{M} E \left[ |a_i[n] - \overline{w}_i^H \tilde{x}_i[n]|^2 \right]$$

$$\begin{align*}
\quad & = \sum_{i=1}^{M} \left\{ \overline{w}_i^H \left( \overline{G}_i \overline{G}_i^H + \sigma_i^2 I_{2Nc} \right) \overline{w}_i + 1 - \overline{w}_i^H \overline{G}_i \varepsilon_{M+i} \\
& \quad - \varepsilon_{M+i} \overline{G}_i \overline{w}_i \right\} \\
& \quad = \sum_{i=1}^{M} \varepsilon_{M+i}^H \left( I_{3M} + \sigma_i^{-2} \overline{G}_i \overline{G}_i^H \right)^{-1} \varepsilon_{M+i} \quad (4.12)
\end{align*}$$

Note that Eqn. 4.12 is obtained by substituting $\varepsilon_i[n]$ with Eqn. 4.4. Then substituting $\overline{w}_i$ with Eqn. 4.6 and using the matrix inversion lemma$^1$, Eqn. 4.13 can be obtained. Full details of the derivation of Eqn. 4.13 can be found in the appendix 4.6.2. Furthermore, if the columns of $\overline{H}_{i}^{\text{prev}}, \overline{H}_{i}$ and $\overline{H}_{i}^{\text{next}}$ in $\overline{G}_i$ are orthogonal to each other$^2$, then $\sigma_i^{-2} \overline{G}_i \overline{G}_i$ approximates a diagonal dominated

---

$^1$Matrix Inversion Lemma: $(A + BCD)^{-1} = A^{-1} - A^{-1}B(DA^{-1}B + C^{-1})^{-1}DA^{-1}$. By substituting $A = I_{3M}, B = \overline{G}_i^H, C = \sigma_i^{-2}I_{2Nc}$ and $D = \overline{G}_i$, Eqn. 4.13 can be obtained.

$^2$The autocorrelation between paths with different delays is small due to small autocorrelation property of the PN sequence. However, for a heavily loaded system, although the interfering signals are weak, the number of interfering signals is high, and the orthogonality between columns of the channel matrix is not well maintained.
matrix and the matrix inversion in Eqn. 4.13 can be simply realised by taking the reciprocal of the diagonal elements of \((I_{3M} + \sigma_i^{-2} \mathbf{G}_i \mathbf{G}_i^H)\). Bearing in mind that \(\mathbf{G}_{M+i}\) is the selecting operator, Eqn. 4.13 can be simplified as

\[
\sum_{i=1}^{M} (1 + \mathbf{w}_i^H \mathbf{R}_{i,i} \mathbf{w}_i)^{-1} \tag{4.14}
\]

where \(\mathbf{R}_{i,i} = \sigma_i^{-2} \mathbf{H}_i \mathbf{H}_i^H\). Obviously, to minimise Eqn. 4.14 is equivalent to maximise \(\mathbf{w}_i^H \mathbf{R}_{i,i} \mathbf{w}_i\), \(\forall i = 1, \ldots, M\). Based on the Rayleigh-Ritz Theorem (see Chapter 4 in [84]), the upper bound of \(\mathbf{w}_i^H \mathbf{R}_{i,i} \mathbf{w}_i\) is achieved if \(\mathbf{w}_i\) is the principal eigenvector of \(\mathbf{R}_{i,i}\), i.e.

\[
\mathbf{w}_i = \text{principal eigenvector of } \mathbf{R}_{i,i} \tag{4.15}
\]

The procedure of implementing the closed-form solution is summarised below:

1. Construct channel matrix \(\mathbf{H}_{i,i}\) using the channel state information;
2. Form covariance matrix \(\mathbf{R}_{i,i} = \sigma_i^{-2} \mathbf{H}_{i,i} \mathbf{H}_{i,i}^H\);
3. Obtain \(\mathbf{w}_i\) as the principle eigenvector of \(\mathbf{R}_{i,i}\);
4. Calculate the MMSE Rx-beamforming vectors for each user \(\mathbf{w}_i\), \(\forall i\) using Eqn. 4.6.

### 4.4 Numerical Studies

In this section, the performance of the proposed joint Tx-Rx beamforming scheme is investigated through a number of numerical simulations. Consider that both the transmitter (or base station) and the receiver (or the mobile user) employ uniform linear arrays of half-wavelength spacing. The system is operating in the presence of \(M = 3\) co-channel BPSK DS-CDMA users. Signals are transmitted to each user via 5 multipaths. The DOD and DOA of each multipath are uniformly distributed within \([0, 180^\circ]\) and TOA of each path is randomly selected from a uniform distribution between \([0, 30 T_c]\). The path fading coefficients are complex Gaussian random variables and the multipath power sum for each user has unit magnitude. Each user employs a unique Gold sequence of length \(N_c = 31\) with rectangular chip pulse-shaping. The array is assumed to collect a block of 200 data symbols for processing at each time. The channel is assumed to be stationary.
during each data block and the channel state information is available to both transmitter and receiver.

Firstly, the MSE and output SNIR convergence curve of the proposed iterative method is studied for various SNR levels. The transmitter (BS) is equipped with an antenna array of 5 elements \((\overline{N} = 5)\) while the receivers (MS) are equipped with 3-element arrays \((N = 3)\). The Tx-beamforming weights are initialised as \(\overline{w_i} = \frac{1}{\sqrt{\overline{N}}} \forall i\). Fig. 4.2 shows that the channel eigendecomposition method (closed-form solution) serves very well as a lower bound for the iterative method for a 3-user system (i.e.\(M = 3\), that is a lightly loaded system). Both methods achieve the equivalent MSE level, however, it takes the iterative method a number of iterations to converge. Thus, in this environment, the closed-form solution is more efficient than the iterative method. In a similar fashion to the MSE criterion, Fig. 4.3 shows the SNIR\(_{out}\) as a function of the number of iterations indicating that the iterative method converges to the closed-form solution too.

Next, the CDMA system is extended to scenarios where the number of users is very large. Three scenarios are considered,

- \(M = 5\) (lightly loaded system)
- \(M = 15\) (medium loaded system)
- \(M = 31\) (heavily loaded system)

For these systems, the overall MSE criterion and output SNIR criterion of the system are plotted respectively in Fig. 4.4 and Fig. 4.5. It is observed that, for systems with \(M = 5\) and \(M = 15\), the convergence of the iterative method to the closed-form solution is very well maintained. In these cases, the closed-form approach instead of the iterative method can be used at the transmitter. As the number of users in the system increases, the orthogonality of users’ spreading sequences degrades. A mismatch occurs when \(M = 31\). In this case, the iterative method still converges to the global minimum of the cost function while the closed-form approach results in an MSE level slightly higher than the iterative method. Thus, in fully loaded systems, the iterative method is slightly better than the closed-form solution. Without loss of generality, in the following discussion, a system of 3 users will be considered and only the performance of closed-form solution will be plotted and compared with other existing methods.

Fig. 4.6 and Fig. 4.7 show respectively the BER and SNIR performance of the proposed channel eigendecomposition scheme as the number \(\overline{N}\) of transmit
4. Joint Transmitter-Receiver Beamforming

Figure 4.2: Convergence study for the iterative method in terms of total MSE with input $E_b/N_0 = 0, 5, 10$dB. It is bounded by the closed-form solution.

Figure 4.3: Convergence study for the iterative method in terms of output SNIR with input $E_b/N_0 = 0, 5, 10$dB. It is bounded by the closed-form solution.
Figure 4.4: Overall MSE performance of proposed methods in systems with varied number of users ($M = 5, 15, 31$).

Figure 4.5: Output SNIR performance of proposed methods in systems with varied number of users ($M = 5, 15, 31$).
antennas increases while the number $N$ of receive antennas is fixed. In particular, the two figures present the performance of the $i^{th}$ user (taken as the desired user) as a function of $E_b/N_0$. The near-far problem is considered by setting the signal to interference ratio equal to $-10$dB (i.e. SIR$= -10$dB). For the proposed method, the number of receive antennas is kept constant at $N = 2$ while the number of transmit antenna elements is $\overline{N} = 4$ or 8. The performance of Tx-beamformers, namely MSNR and PSNIR criterions [44][47] (extended to the space-time channel model presented in this chapter) are also present in Figs. 4.6 and 4.7. Note that these two methods are designed as follows

$$\overline{w}_i = \begin{cases} \arg \max_{\overline{w}} \left\{ \text{SNR}_i \triangleq \frac{E[H_{\overline{w}_i}^H \overline{H}_{\overline{w}_i}^H \overline{w}_i]}{\sigma_i^2} \right\} : \text{MSNR} \\ \arg \max_{\overline{w}} \left\{ \text{PSNIR}_i \triangleq \frac{E[H_{\overline{w}_i}^H \overline{H}_{\overline{w}_i}^H \overline{w}_i]}{E[H_{\overline{w}_i}^H \overline{H}_{\overline{w}_i}^H \overline{w}_i + \sigma_i^2 I_N]} \right\} : \text{PSNIR} \end{cases}$$

The transmit weight vector $\overline{w}_i$ that maximises SNR$_i$ of the $i^{th}$ user (MS) is equal to the principal eigenvector of $H_{i,i}^H H_{i,i}$. In the case of maximising PSNIR, $\overline{w}_i$ is found to be the generalised eigenvector of $(H_{i,i}^H H_{i,i}, (G_i^H G_i - H_{i,i}^H H_{i,i} + \sigma_i^2 I_N))$ where $G_i$ is defined in Eqn. 4.10.

The STAR-2D-RAKE receiver (Eqn. 3.28) and STAR-MMSE receiver (Eqn. 3.30) with $N = 2$ array elements employed at the MS are also evaluated and their performance is shown in Figs. 4.6 and 4.7. From these figures, it is clear that for Tx-beamforming schemes, the PSNIR outperforms MSNR because of PSNIR’s capability in suppressing co-channel interferences. For Rx-beamforming schemes, the MMSE receiver performs much better than the RAKE receiver. It can be seen that the BER and SNIR performances of the proposed method outperform all the other methods in the given scenario. Note that the transmit beamformer that forms part of the closed-form solution is actually an MSNR beamformer. Indeed, the performance enhancement and the capability of interference suppression seen by the closed-form solution is due to the employment of the MMSE beamformer at the receiver end.

Next, let us consider the situation when the number of transmitter antenna elements $\overline{N}$ at the base station is kept constant while the number of receiver antenna elements $N$ at the MS receiver increases. It can be seen from Figs. 4.8 and 4.9 that the proposed joint optimisation method improves the BER and SNIR performance of the system. Note that the array set of $(\overline{N}, N) = (4, 4)$ has a comparable performance to the MMSE Rx-beamformer with an array set of $(\overline{N}, N) = (1, 6)$. However, as there is a size restriction at the mobiles, by using
4. Joint Transmitter-Receiver Beamforming

Figure 4.6: BER performance of joint Tx-Rx beamforming, Tx-beamforming and Rx-beamforming as different number of transmitting antennas are employed ($N = 1, 4, 8$).

Figure 4.7: Output SNIR performance of joint Tx-Rx beamforming, Tx-beamforming and Rx-beamforming as different number of transmitting antennas are employed ($N = 1, 4, 8$).
4. Joint Transmitter-Receiver Beamforming

Figure 4.8: BER performance of joint Tx-Rx beamforming and Rx-beamforming as the number of receiver elements are chosen as $N = 2, 4, 6$.

Tx-beamforming at the BS the number of receive antenna element at the MS can be reduced while a comparable performance can still be achieved.

4.5 Conclusions

In this chapter, the problem of joint Tx-Rx beamforming in a multipath fading environment is studied. It has been shown that by utilizing the array manifold concept, the channel can be properly structurised in both the space and time domain. The beamforming weights are jointly optimised over the whole network under the MMSE criteria. Both an iterative solution and a closed-form solution to the optimisation problem have been provided. Numerical simulations have shown the superior performance of the proposed method over many existing beamforming algorithms.

4.6 Appendix

4.6.1 Derivation of the Iterative Solution:

By taking expectation w.r.t. $a[n]$ and $n[n]$ and by applying the Lagrange multiplier, the constraint cost function (Eqn. 4.7) can be reduced to an unconstraint
function as

$$J = \min_{\mathbf{w}_1, \ldots, \mathbf{w}_M} \left\{ \sum_{i=1}^M \left\{ 1 + \mathcal{E} \{ \mathbf{n}[n]^H \mathbf{w}_i \mathbf{w}_i^H \mathbf{n}[n] \} \right\} \right.$$  

$$+ \text{tr} \left( \mathbf{W}^H \left[ \mathbb{H}^\text{prev}_i, \mathbb{H}^H_i, \mathbb{H}^\text{next}_i \right] \left( \mathbb{I}_3 \otimes \mathbf{w}_i \mathbf{w}_i^H \right) \left[ \begin{array}{cc} \mathbb{H}^\text{prev}_i & \mathbb{H}^H_i \\ \mathbb{H}^H_i & \mathbb{H}^\text{next}_i \end{array} \right] \mathbf{W} \right)$$

$$- \text{tr}(\mathbf{w}_i^H \mathbb{H}_{i,i}^H \mathbf{w}_i) - \text{tr}(\mathbf{w}_i^H \mathbb{H}_{i,i} \mathbf{w}_i) + \lambda_i (\mathbf{w}_i^H \mathbf{w}_i - 1) \right\} \right. \right.$$  

(4.16)

Based on Eqn. 4.16, the weight vectors $\mathbf{w}_i \ \forall i = 1, \ldots, M$ are optimised individually by taking the partial derivative of $J$ w.r.t. $\mathbf{w}_i \ \forall i$. The gradient is readily obtained as

$$\frac{\partial J}{\mathbf{w}_i} = 2 \mathbb{G}_i^H \mathbb{G}_i \mathbf{w}_i + 2 \lambda_i \mathbf{w}_i - 2 \mathbb{H}_{i,i} \mathbf{w}_i$$

where $\mathbb{G}_i$ is defined as in Eqn. 4.10.

Letting the gradient equal to zero, i.e.

$$\frac{\partial J}{\mathbf{w}_i} = \mathbb{G}_i^H \mathbb{G}_i \mathbf{w}_i + \lambda_i \mathbf{w}_i - \mathbb{H}_{i,i} \mathbf{w}_i = 0$$  

(4.17)

the solution to the above equation is easily given by

$$\mathbf{w}_i = (\mathbb{G}_i^H \mathbb{G}_i + \lambda \mathbb{I}_N)^{-1} \mathbb{H}_{i,i} \mathbf{w}_i$$
The transmit weight $\mathbf{w}_i$ is then normalised in order to satisfy the constraint $\|\mathbf{w}_i\| = 1$. The Lagrange Multiplier $\lambda_i$ can then be derived from Eqn. 4.17 as

$$\lambda_i \mathbf{w}_i^H \mathbf{w}_i = \mathbf{w}_i^H \mathbf{H}_i^H \mathbf{w}_i - \mathbf{w}_i^H \mathbf{G}_i^H \mathbf{G}_i \mathbf{w}_i$$  (4.18)

### 4.6.2 Derivation of the Closed-Form Solution:

Based on Eqn. 4.7, the MSE of all users can be derived as

$$J = \min_{\mathbf{w}_1, \ldots, \mathbf{w}_M} \left\{ \sum_{i=1}^{M} \mathcal{E} \left\{ |a_i[n] - \mathbf{w}_i \mathbf{x}_i[n]|^2 \right\} \right\}$$  (4.19)

$$= \min_{\mathbf{w}_1, \ldots, \mathbf{w}_M} \left\{ \sum_{i=1}^{M} \left[ 1 - \mathcal{E} \{ \mathbf{w}_i \mathbf{x}_i[n] a_i[n] \} - \mathcal{E} \{ a_i[n] \mathbf{x}_i^H[n] \mathbf{w}_i \} + \mathcal{E} \{ \mathbf{w}_i \mathbf{x}_i[n] \mathbf{x}_i^H[n] \mathbf{w}_i \} \right] \right\}$$

From Eqn. 4.4, it is known that

$$\mathbf{x}_i[n] = \mathbf{G}_i \begin{bmatrix} a[n-1] \\ a[n] \\ a[n+1] \end{bmatrix} + \mathbf{n}_i[n]$$  (4.20)

By substituting $\mathbf{x}_i[n]$ in Eqn. 4.19 with Eqn. 4.20, it can be seen that

$$J = \min_{\mathbf{w}_1, \ldots, \mathbf{w}_M} \left\{ \sum_{i=1}^{M} \left[ \mathbf{w}_i^H \left( \mathbf{G}_i \mathbf{G}_i^H + \sigma_t^2 \mathbb{I}_{2NN_c} \right) \mathbf{w}_i - \mathbf{w}_i^H \mathbf{H}_i \mathbf{w}_i - \mathbf{w}_i^H \mathbf{H}_i^H \mathbf{w}_i + 1 \right] \right\}$$  (4.21)

Further substituting the receiver weight vector $\mathbf{w}_i$ using Eqn. 4.6, Eqn. 4.21 can be derived as

$$J = \min_{\mathbf{w}_1, \ldots, \mathbf{w}_M} \left\{ \sum_{i=1}^{M} \left[ \mathbf{w}_i^H \mathbf{H}_i^H \left( \mathbf{G}_i \mathbf{G}_i^H + \sigma_t^2 \mathbb{I}_{2NN_c} \right)^{-1} \mathbf{H}_i \mathbf{w}_i \\
- \mathbf{w}_i^H \mathbf{H}_i^H \left( \mathbf{G}_i \mathbf{G}_i^H + \sigma_t^2 \mathbb{I}_{2NN_c} \right)^{-1} \mathbf{H}_i \mathbf{w}_i \\
- \mathbf{w}_i^H \mathbf{H}_i^H \left( \mathbf{G}_i \mathbf{G}_i^H + \sigma_t^2 \mathbb{I}_{2NN_c} \right)^{-1} \mathbf{H}_i \mathbf{w}_i + 1 \right] \right\}$$  (4.22)

$$= \min_{\mathbf{w}_1, \ldots, \mathbf{w}_M} \left\{ \sum_{i=1}^{M} \left[ 1 - \mathbf{w}_i^H \mathbf{H}_i^H \left( \mathbf{G}_i \mathbf{G}_i^H + \sigma_t^2 \mathbb{I}_{2NN_c} \right)^{-1} \mathbf{H}_i \mathbf{w}_i \right] \right\}$$  (4.23)

Especially, when users’ spreading codes maintain good orthogonality, such as in a lightly or medium loaded system, the joint transmitter-receiver can be
obtained in one step. Bearing in mind that $\xi_k$ is the selection operator, Eqn. 4.23 can also be expressed as

$$J = \min_{\mathbf{w}_1, \ldots, \mathbf{w}_M} \left\{ \sum_{i=1}^{M} \left[ 1 - \xi_{M+i}^H \mathbf{G}_i^H \left( \mathbf{G}_i \mathbf{G}_i^H + \sigma_i^* \mathbb{I}_{2 \times N_c} \right)^{-1} \mathbf{G}_i \xi_{M+i} \right] \right\}$$

$$= \min_{\mathbf{w}_1, \ldots, \mathbf{w}_M} \left\{ \sum_{i=1}^{M} \xi_{M+i}^H \left[ \mathbf{I}_{3M} - \mathbf{G}_i^H \left( \mathbf{G}_i \mathbf{G}_i^H + \sigma_i^* \mathbb{I}_{2 \times N_c} \right)^{-1} \mathbf{G}_i \right] \xi_{M+i} \right\} \quad (4.24)$$

Using the Matrix Inversion Lemma $(A + BCD)^{-1} = A^{-1} - A^{-1}B(DA^{-1}B + C^{-1})^{-1}DA^{-1}$ and by defining

$$\begin{align*}
A &= \mathbf{I}_{3M} \\
B &= \mathbf{G}_i^H \\
C &= \sigma_i^* \mathbb{I}_{2 \times N_c} \\
D &= \mathbf{G}_i
\end{align*}$$

The objective function can then be rewritten as

$$J = \min_{\mathbf{w}_1, \ldots, \mathbf{w}_M} \left\{ \sum_{i=1}^{M} \xi_{M+i}^H \left( \mathbf{I}_{3M} + \sigma_i^{-2} \mathbf{G}_i^H \mathbf{G}_i \right)^{-1} \xi_{M+i} \right\} \quad (4.25)$$

Since columns of $\mathbb{H}_i^\text{prev}$, $\mathbb{H}_i$ and $\mathbb{H}_i^\text{next}$ in $\mathbf{G}_i^H$ are orthogonal to each other, $\sigma_i^{-2} \mathbf{G}_i^H \mathbf{G}_i$ thus approximates a diagonal matrix and the matrix inversion in Eqn. 4.25 can be simply realised by taking the reciprocal of the diagonal elements of $(\mathbf{I}_{3M} + \sigma_i^{-2} \mathbf{G}_i^H \mathbf{G}_i)$ and Eqn. 4.25 can be simplified as

$$\sum_{i=1}^{M} \left( 1 + \frac{\mathbf{w}_i^H \mathbf{G}_i^H \mathbf{G}_i \mathbf{w}_i}{\mathbb{R}_{i,i}} \right)^{-1} \quad (4.26)$$

Obviously, to minimise Eqn. 4.26 is equivalent to maximise $\mathbf{w}_i^H \mathbf{R}_{i,i} \mathbf{w}_i \forall i = 1, \ldots, M$ where $\mathbf{R}_{i,i}$ is defined in the equation. Based on the Rayleigh-Ritz Theorem (see Chapter 4 in [84]), the upper bound of $\mathbf{w}_i^H \mathbf{R}_{i,i} \mathbf{w}_i$ is achieved if $\mathbf{w}_i$ is the principle eigenvector of $\mathbf{R}_{i,i}$.
Chapter 5

Arrayed OFDM-CDMA System

Among many modulation technologies, Orthogonal Frequency Division Multiplexing (OFDM) has been widely implemented in high speed digital communications due to the recent advances of digital signal processing (DSP) and very large scale integrated (VLSI) circuit technologies. In this Chapter, the transceiver design in an uplink OFDM-CDMA system operating in a frequency-selective environment and using an antenna array at the receiver’s front-end will be studied. The OFDM-CDMA modulation scheme is an effective approach in combating frequency selective fading thus increasing the system capacity. A blind near-far resistant channel estimator is first devised which provides the path angle, delay and fading coefficient estimation. With the estimator integrated at the front-end, both post-FFT type and pre-FFT type space-time receivers have been proposed for interference cancellation. The effectiveness of the proposed approach is demonstrated even in the presence of strong interferences by computer simulation studies.
5. Arrayed OFDM-CDMA System

5.1 Introduction

The ever-increasing demand for high data rates in wireless networks requires the efficient utilisation of the limited bandwidth available, while supporting a high grade of mobility in diverse propagation environments. Orthogonal Frequency Division Multiplexing (OFDM) techniques [85] are capable of satisfying these requirements, since they are capable of coping with highly time-variant wireless channel characteristics. OFDM technique dates back to 1970s and its idea was to use parallel data and frequency division multiplexing with overlapping subchannels to combat multipath distortion as well as to fully use the available bandwidth.

Recently, the combination of Orthogonal Frequency Division Multiplexing (OFDM) with Code Division Multiple Access (CDMA) has been of significant interest as a means to take such advantages as bandwidth efficiency, fading resilience, and interference suppression capability which are crucial in future broadband data transmission. Three types of OFDM-CDMA schemes have been proposed which are multicarrier (MC-) CDMA, multicarrier DS-CDMA and multi-tone (MT-) CDMA [86][87][88]. As an OFDM based system, the symbol duration is increased on the parallel subchannels, thus reducing or eliminating the inter-symbol interference (ISI) caused by the multipath environments. In addition, the signals can be easily transmitted and received using the Fast Fourier transform (FFT) device without increasing the transmitter and receiver complexities. On the other hand, as a CDMA based system, the use of orthogonal spreading codes enables the multiple access capability of the system by maintaining orthogonality among different users’ transmissions. Unlike the pure DS-CDMA systems, the OFDM-CDMA transmitter spreads the original data stream over frequency domain which also provide optimal frequency diversity in a time-varying multipath channel.

In practical OFDM systems operating over a dispersive channel, a cyclic prefix (CP) longer than the anticipated multipath channel spread is usually inserted in the transmitted sequence. It is well known that this converts the linear (time-domain) convolution between the channel and the input into cyclic convolution or equivalently a (complex) multiplicative factor on each sub channel in the frequency domain. This naturally facilitates computationally simple frequency domain channel estimation by inserting a training sequence to estimate the factor on each subchannel [89][90][91]. However, systems using training sequences incur a price: significant loss of channel utilization that may be the overriding constraint.
for future high-speed services. Additionally, due to the time-varying nature of the channel in some wireless applications (i.e., those that seek to provide mobility support), the training sequence needs to be transmitted periodically, causing further loss of channel throughput.

The above concerns naturally lead to the design of blind and semi-blind channel estimation methods that avoid the need for training sequences. The capacity and the achievable integrity of communication systems is highly dependent on the system’s knowledge concerning the channel conditions encountered. Thus, the provision of an accurate and robust channel estimation strategy is a crucial factor in achieving a high performance. Some approaches exploit the inherent cyclostationary structure of an OFDM symbol induced by the Cyclic Prefix (CP) [92][93]. Other than the CP, the spreading sequence in an OFDM-CDMA system is another resource that can be exploited for purposes of channel estimation. The equivalence between OFDM-CDMA and DS-CDMA with respect of spreading code design enables the implementation of MC-CDMA signals with an OFDM spectrum as equivalent DS-CDMA systems [61][26][94][95]. The MC-CDMA system can be interpreted as an equivalently DS-CDMA system which employs a transformed version (inverse Fourier Transform) of the user’s specific PN-code. Blind channel estimation and detection algorithms that have been derived for DS-CDMA system can be adapted to MC-CDMA system. Since conventional DS-CDMA receivers are developed to cope with ISI in the detection process, the MC-CDMA mobile radio system exploiting an equivalent DS-CDMA receiver does not necessarily require the absence of ISI and, hence, of interchannel interference (ICI) in the detection process. Thus, the loss in bandwidth efficiency due to the guard interval can be avoided when accepting the higher receiver complexity [88].

Combining antenna array techniques with OFDM-CDMA transmission is undoubtedly advantageous in further increasing the system capacity without allocating additional frequency spectrum. The employment of antenna arrays is always beneficial in terms of mitigating the effects of MAI, since with the aid of beam-steering, it becomes possible to focus the receiver antenna beam on the served user while attenuating the co-channel interferers [96]. This is of particularly high importance in OFDM-CDMA, which exhibits a high sensitivity against co-channel interference.

In this chapter, an uplink arrayed OFDM-CDMA system will be studied. An equivalent DS-CDMA interpretation for the OFDM-CDMA system will be derived over frequency-selective fading channels in multiuser scenario. The Circular Spatial-Temporal ARray (C-STAR) manifold vector will be derived and incorpo-
rated into the model which results in a more accurate description of the channel, leading to the design of more powerful receivers. Using the antenna array, the blind subspace based superresolution estimation algorithm based on the C-STAR manifold vector will be proposed which fully exploits the space-time properties of the channel and jointly estimates the directions and delays of the multipaths. Once the channel estimates have been correctly obtained at the receiver, two types of blind space-time receivers, the post-FFT and pre-FFT receiver are proposed. For the post-FFT receiver, the DFT devices are used at the front-end of the receiver for channel diagonalisation. An MMSE receiver derived based on the DFT transformed signature of the channel is then used to further remove MAI components from the signal. The pre-FFT receiver is an integrated receiver which doesn’t require the DFT transformation but can completely suppress ISI and MAI interferences and is optimal with respect to the MMSE criteria. The effectiveness of the proposed approach, even in the presence of strong interferences, will be demonstrated by computer simulation studies. The MMSE multiuser detector and the 2D RAKE receiver are also employed for the performance analysis.

5.2 OFDM-CDMA Signal Modelling

The block diagram of an $M$ user OFDM-CDMA array system is shown in Fig. 2.11. Among various OFDM-CDMA schemes, the MC-CDMA technique is chosen to be employed at the transmitter. However, the adaptation of other OFDM-CDMA techniques is straightforward.

Unlike the DS-CDMA modulation, the user’s data symbols are first replicated into $N_{sc}$ parallel streams, each stream is multiplied with one chip of the assigned spreading code and then modulated onto a subcarrier. The continuous baseband OFDM-CDMA transmitting signals of the $i^{\text{th}}$ user during the $n^{\text{th}}$ data symbol period is given by

$$m_i(t) = a_i[n] \sum_{k=0}^{N_{sc}-1} \alpha_i[k] \exp(j2\pi \frac{k}{T_{cs}} t) \quad t \in [nT_{cs}, (n+1)T_{cs}] \quad (5.1)$$

where $a_i[n] = \pm 1, \forall n \in Z$ is the $i^{\text{th}}$ user’s BPSK symbol with symbol duration $T_{cs}$, $\{\alpha_i[k] \in \{-1, +1\}, k = 0, 1, \ldots, N_c - 1\}$ corresponding to the $k^{\text{th}}$ chip of the $i^{\text{th}}$ user’s spreading code sequence $\alpha_i = [\alpha_i[0], \alpha_i[1], \ldots, \alpha_i[N_c - 1]]$ of period $N_c$. In this chapter, it is assumed that the number of subcarriers and the spreading gain are the same (i.e. $N_c = N_{sc}$).
By sampling the modulated signal with a sampling rate of $1/T_s = N_{sc}/T_{cs}$ during an OFDM symbol interval $T_{cs}$, the discretised transmitted signal at instant $t = (n + g/N_{sc})T_{cs}$ (see point A in Fig. 5.1) can be formed as

$$m_i((n + g/N_{sc})T_{cs}) = a_i[n] \sum_{k=0}^{N_{sc}-1} \alpha_i[k]\exp(j2\pi(nk + k g/N_{sc})) \quad (5.2)$$

Let’s define $m_i[n, g] \triangleq m_i((n + g/N_{sc})T_{cs})$, Eqn. 5.2 can be written in a more compact form as

$$m_i[n, g] = a_i[n] \sum_{k=0}^{N_{sc}-1} \alpha_i[k]\exp(j2\pi k g/N_{sc}) \quad (5.3)$$

for $g = 0, 1, \ldots, N_{sc} - 1$

Notice that Inverse Discrete Fourier Transform (IDFT) is actually performed in Eqn. 5.3. The structure of the MC-CDMA modulator of the $i$th user using the IDFT operator is shown in Fig. 5.1. By denoting

$$m_i[n] = [m_i[n, 0], m_i[n, 1], \ldots, m_i[n, N_{sc} - 1]]^T$$

Eqn. 5.3 can be further simplified as

$$m_i[n] = a_i[n]\mathbb{F}\alpha_i \quad (5.4)$$

where $\mathbb{F}$ is the IDFT matrix defined as follows

$$\mathbb{F} = [\rho^0, \rho^1, \rho^2, \ldots, \rho^{(N_{sc}-1)}] \quad (5.5)$$

and

$$\rho = [1, \rho^1, \rho^2, \ldots, \rho^{(N_{sc}-1)}]^T$$

with

$$\rho = \exp(j\frac{2\pi}{N_{sc}}) \quad (5.6)$$

It is interesting to notice that an OFDM-CDMA signal which is spread in the frequency domain with a spreading sequence $\alpha_i$ as shown in Eqn. 5.4 can also be interpreted as being spread in the time domain with the transformed spreading sequence $\mathbb{F}\alpha_i$. The channel estimation and reception techniques that are traditionally used in DS-CDMA systems can be also applied to OFDM-CDMA systems.

In practical OFDM systems operating over a dispersive channel, a cyclic prefix (CP) longer than the anticipated multipath channel spread is usually inserted in
Figure 5.1: Block diagram of the OFDM-CDMA transmitter
the transmitted sequence to combat the hostile ISI. If \( L_{cp} \) chips have been inserted in front of each OFDM symbol, by denoting \( T_{cp} = L_{cp}T_s \), the symbol duration \( T_{cs} \) has now absorbed the CP duration and a complete OFDM symbol at point B in Fig. 5.1 is in the form of

\[
\tilde{m}_i[n] = [m_i[n, N_{sc} - L_{cp}], \ldots, m_i[n, N_{sc} - 1], m_i[n]^T]^T
\] (5.7)

After CP insertion and Parallel-to-Serial (PS) conversion, an OFDM-CDMA signal suitable for transmission is obtained by first low pass filtering to generate the baseband continuous signal and then upconverting to the carrier frequency \( F_c \). Suppose the transmitted signal is narrowband limited and an antenna array of \( N \) elements is employed at the receiver, the signal from the \( i^{th} \) user arrives at the receiver via \( K_i \) multipaths. Based on the structure/modelling of the SIVO channel as shown in Fig. 2.4, by denoting \( \tilde{m}_i(t) \) as the baseband continuous signal generated by low pass filtering \( \tilde{m}_i[n] \) (point C, Fig. 5.1), the received continuous signal after downconversion is given as

\[
x(t) = \sum_{i=1}^{M} \sum_{k=1}^{K_i} \beta_{ik} S_{ik} \tilde{m}_i(t - \tau_{ik}) + n(t)
\] (5.8)

Note that \( \beta_{ik} \) has absorbed both the path fading coefficient and a complex factor \( \exp(-2\pi F_c \tau_{ik}) \); the Doppler frequency shift is not considered. In addition, \( n(t) \) represents the complex additive white Gaussian noise (AWGN) of power \( \sigma^2 \).

Taking the multipath delay spread to lie within the range \([0, T_{cs})\), the output of each array element is sampled at a period of \( T_s = T_c \) and passed through a Tapped Delay Line (TDL) of length \( L = N_{sc} \) to form the discretised baseband signal. After CP is discarded, the received discretised signal has a circulant structure as shown in Fig. 5.2 and is free of interferences from the adjacent OFDM symbols\(^1\). However, each OFDM symbol still contains chip level interferences from its own circulated copies.

To model such interferences, let’s first introduce the following Circulant Spatial-Temporal ARray (C-STAR) manifold vector

\[
h_{ik} = S_{ik} \otimes (F_{c}^{\text{if}} \mathcal{F}_{\Omega_c})
\] (5.9)

\(^1\)In an uplink MC-CDMA system, the guard interval (CP) used for ISI prevention can also be used to contain small synchronisation offsets (less than the CP interval) between users. In this quasi-synchronous case, the signals from different users can be processed on a symbol-by-symbol basis, without interference from adjacent symbols. However, this does not apply to a fully asynchronous scenario. Without loss of generality, it is assumed that the system is perfectly synchronised in this chapter.
Figure 5.2: The circulant structure of received symbols and the intersymbol interferences (after CP removal).
where \( l_{ik} = [\tau_{ik}/T_s] \) is the discretised multipath delay of the \( i \)th user's \( k \)th path and the matrix \( \mathbf{J}_c \) is the circulating operator matrix given as

\[
\mathbf{J}_c = \begin{bmatrix}
0 & 0 & \cdots & 0 & 1 \\
1 & 0 & \cdots & 0 & 0 \\
0 & 1 & \cdots & 0 & 0 \\
\vdots & \vdots & \ddots & \vdots & \vdots \\
0 & 0 & \cdots & 1 & 0 \\
\end{bmatrix}
\]

(5.10)

The cyclic nature of Eqn. 5.9 is due to the fact that the insertion of CP converts the linear (time-domain) convolution between the channel and the input into cyclic convolution (see Fig. 5.2). By taking into account the chip level interference within current symbols as well as the MAI constituents, the discretised representation of the received signal vector \( \mathbf{z}[n] \) can hence be expressed in a compact matrix form as

\[
\mathbf{z}[n] = \sum_{i=1}^{M} \mathbf{H}_i \beta_i a_i[n] + \mathbf{n}[n]
\]

(5.11)

where \( \beta_i = [\beta_{i1}, \beta_{i2}, \ldots, \beta_{iK_i}]^T \) and \( \mathbf{n}[n] \) is the sampled noise vector. The matrix \( \mathbf{H}_i \) has columns the C-STAR manifold vectors and is given as

\[
\mathbf{H}_i = \begin{bmatrix}
\mathbf{h}_{i1}, \mathbf{h}_{i2}, \ldots, \mathbf{h}_{iK_i}
\end{bmatrix} \in \mathbb{C}^{N_{sc} \times K_i}
\]

(5.12)

Eqn. 5.11 can be further simplified as

\[
\mathbf{z}[n] = \mathbf{H} \begin{bmatrix}
\mathbf{1}_{K_1} a_1[n] \\
\mathbf{1}_{K_2} a_2[n] \\
\vdots \\
\mathbf{1}_{K_M} a_M[n]
\end{bmatrix} + \mathbf{n}[n] \in \mathbb{C}^{N_{sc} \times 1}
\]

(5.13)

where \( \mathbf{H} \) is the composite channel matrix given as

\[
\mathbf{H} = [\mathbf{H}_1, \mathbf{H}_2, \ldots, \mathbf{H}_M] \in \mathbb{C}^{N_{sc} \times K}
\]

(5.14)

with \( K = \sum_{i=1}^{M} K_i \), and \( \mathbb{B} = \text{diag}(\beta_{1}^T, \beta_{2}^T, \ldots, \beta_{M}^T) \) which contains fading coefficients of all the users on its diagonal line.
5. Arrayed OFDM-CDMA System

5.3 Angle, Delay and Path Fading Coefficient Estimation

5.3.1 Angle and Delay Estimation

As can be seen in Eqn. 5.11, the C-STAR manifold vectors associated with the $K_i$ paths of the $i$th user are linearly combined by the fading coefficient vector $\beta_i$ (i.e. $H_i\beta_i$), and this will make these paths indistinguishable in their contribution to the signal subspace. This "coherence problem" makes the estimation of the spatio-temporal channel parameters (i.e. delays and directions) using subspace-based superresolution techniques fail. To overcome this problem, take user $i$ as the user to be estimated and define matrix $C_l$

$$C_l = [J_1^c F_i, \ldots, J_{l-1}^c F_i, J_l^c F_i, \ldots, J_{N_{sc}}^c F_i]$$

which is a composite matrix containing the circularly shifted vector $F_i$ with the exclusion of the $l$th column (i.e. $J_l^c F_i$). By defining the preprocessor matrix $P_l^\perp$ as

$$P_l^\perp = I_N \otimes (I_{N_{sc}} - C_l (C_l^H C_l)^{-1} C_l^H)$$

(5.15)

and by applying the preprocessor $P_l^\perp$ to the received signal $x[n]$, i.e.

$$P_l^\perp x[n]$$

(5.16)

There will be no contributions to the preprocessed signal other than the path with delay equal to $l$. This is because paths that arrives with different delays will lie in the null space of the preprocessor matrix producing

$$P_l^\perp h_{ik} = 0, \quad \forall k = 1, \ldots, K_i \text{ if } l_{ik} \neq l$$

Let’s denote $R_l$ as the covariance matrix of the output signal of the preprocessor, i.e.

$$R_l = \mathcal{E}\{P_l^\perp x[n] x[n]^H P_l^\perp\}$$

It is clear that after the preprocessor (transformation), only the multipath with delay $l$ belongs to the transformed array manifold curve defined as follows

$$\mathcal{M} = \{P_l^\perp (\Sigma(\theta) \otimes (J_l^c F_i)) \in C^{N_{sc} \times 1}, \forall \theta \in \Omega_\theta, l \in \Omega_l\}$$

(5.17)

where

$$\{\Omega_\theta, \Omega_l\} = \text{parameter spaces of } \theta \text{ and } l$$
5. Arrayed OFDM-CDMA System

This curve is the transformed manifold curve of the \(i^{th}\) user with delay \(l\) and any other multipaths of the same user or interfering users are described by different manifold curves. The intersection of the manifold \(\mathcal{M}\) with the “overall signal subspace” lies in the overall signal subspace and is orthogonal to the noise subspace. Therefore, this intersection can be found from the following minimisation problem

\[
(\theta, l) = \arg \min_{\theta, l} \xi(\theta, l)
\]

with

\[
\xi(\theta, l) = \left[ \mathbb{P}_l^+ \left( \mathcal{S}(\theta) \otimes (\mathbb{J}[\mathbb{F}_i]) \right) \right]^H \mathbf{E}_{n,l} \mathbf{E}_{n,l}^H \left[ \mathbb{P}_l^+ \left( \mathcal{S}(\theta) \otimes (\mathbb{J}[\mathbb{F}_i]) \right) \right]
\]

In Eqn. 5.18, \(\mathbf{E}_{n,l}\) corresponds to the noise subspace \(\mathcal{L}[\mathbb{E}_{n,l}]\) spanned by noise eigenvectors of the covariance matrix \(\mathbb{R}_l\). The intersection will only provide the desired user’s parameters which are the directions and delays of its paths.

As the signal subspace and the noise subspace are orthogonal to each other, the cost function \(\xi(\theta, l)\) would be expected to be approximately zero for parameters at which desired paths do exist and non-zero where paths do not exist. Thus a number of peak searches of the spectrums obtained by the evaluation of Eqn. 5.18 will provide the set of directions and delays as

\[
\Xi_i = \{(\theta_{i1}, l_{i1}); (\theta_{i2}, l_{i2}); \ldots; (\theta_{iK_i}, l_{iK_i})\} \in \mathbb{C}^{K_i \times 2}
\]

Based on the parameter set \(\Xi_i\), the channel matrix of the \(i^{th}\) user can be constructed as

\[
\mathbb{H}_i = [\mathbb{h}_{i1}, \mathbb{h}_{i2}, \ldots, \mathbb{h}_{iK_i}] \in \mathbb{C}^{N_{sc} \times K_i}
\]

with \(\mathbb{h}_{ik}\) defined in Eqn. 5.9. Therefore, the composite channel matrix \(\mathbb{H}\) can be constructed according to Eqn. 5.14 based on the channel estimates of all the users.

### 5.3.2 Estimation of Complex Fading Coefficients

The estimation of complex fading coefficients is a research problem by itself and an assumption may be made that fading coefficients are known (i.e. have been estimated). However, here a simple approach will be used for their estimation in the case of high SNR. Indeed, the estimated multipath delays and angles can now be used to further estimate the complex channel fading coefficients. Substituting the channel matrix in Eqn. 5.13 with the estimated channel matrix \(\mathbb{H}\) and by
denoting $|\mathbb{B}|$ and $\angle \mathbb{B}$ the element by element modulus and phase angle of the fading coefficients respectively$^2$, Eqn. 5.13 can be rewritten as

\[ x[n] = \mathcal{H} |\mathbb{B}| \exp(j \angle \mathbb{B}) \begin{bmatrix} 1_{K_1}a_1[n] \\ 1_{K_2}a_2[n] \\ \vdots \\ 1_{K_M}a_M[n] \end{bmatrix} + n[n] \quad (5.20) \]

Based on Eqn. 5.20, by using the properties of the second order statistics of the received signal, the modulus matrix $|\mathbb{B}|$ thus can be estimated as

\[ |\mathbb{B}| = \sqrt{\mathcal{H}^\dagger (\mathbb{R}_{xx} - \sigma_n^2 \mathbb{I}_{NN}) (\mathcal{H}^\dagger)^H} \quad (5.21) \]

Multiplying matrix $(\mathcal{H}|\mathbb{B}|)^\dagger$ on both sides of Eqn. 5.20, it is easy to obtain

\[ \exp(j \angle \mathbb{B}) \begin{bmatrix} 1_{K_1}a_1[n] \\ 1_{K_2}a_2[n] \\ \vdots \\ 1_{K_M}a_M[n] \end{bmatrix} = (\mathcal{H}|\mathbb{B}|)^\dagger x[n] \quad (5.22) \]

Note that the noise term is ignored in Eqn. 5.22 thus the method is used in the environment of low noise level. Performing Hadamard-product operation on the 5.22 and owing to the fact that $a_i[n] = \pm 1 \ \forall i = 1, \ldots, M$, that is, $a_i[n]a_i[n] = 1 \ \forall i = 1, \ldots, M$, the phase angle matrix $\angle \mathbb{B}$ is thus given by

\[ \angle \mathbb{B} = \frac{\angle \text{diag} \left((\mathcal{H}|\mathbb{B}|)^\dagger x[n] \odot (\mathcal{H}|\mathbb{B}|)^\dagger x[n] \right)}{2} \quad (5.23) \]

Having identified both the modulus and the phase angles of the ambiguous complex fading coefficients, the overall fading coefficients $\mathbb{B}$ can be found as

\[ \mathbb{B} = |\mathbb{B}| \exp(j \angle \mathbb{B}) \quad (5.24) \]

Inspection of the solution shows that the phase angle returned by the numerator of Eqn. 5.23 is limited within the range $[-\pi, \pi]$. This will result in a phase ambiguity of $\pi$ in $\angle \mathbb{B}$. The ambiguity issue, however, can be resolved by using the differential encoder/decoder in the transceiver.

A final note is that the channel estimator proposed in this section operates in a similar fashion to the DS-CDMA estimator at the front-end of the receiver prior to DFT transformation. However, in the OFDM case, the channel ISI has been eliminated with the help of the cyclic prefix.

$^2$\(\mathbb{B}\) is a diagonal matrix.
5.4 Blind Reception for Interference Suppression

For an OFDM system, signal processing can be applied to either post-FFT (i.e. signal processing is carried out after the DFT transformation at each antenna element) [97][98] or pre-FFT (i.e. the DFT transformation is integrated into the receiver weights) [99][100] at the receiver. The post-FFT array processing is widely believed to transform the multipath channel into the flat channel using the FFT properties (i.e. the diagonalisation on cyclic channel structure), whereas the pre-FFT array processing has lower complexity as the DFT transformation, the interference cancellation and demodulation are integrated into the receiver weights. In this section, a post-FFT approach is first derived and followed by the development of a pre-FFT type receiver.

5.4.1 Post-FFT Signal Reception

In this section, a blind post-FFT receiver will be proposed where the received signals will be first transformed by using DFT operator and multiuser detection is performed at the DFT output. The block diagram of the receiver is shown in Fig. 5.3.

The channel parameters obtained using the channel estimator proposed in Section 5.3 will be used to construct the receiver weight matrix. Given Eqn. 5.11, the frequency domain signals at the DFT output (point B) is written as

\[ \tilde{z}[n] = (I_N \otimes F^{-1}) \tilde{x}[n] \]  

(5.25)

By substituting \( \tilde{x}[n] \) with Eqn. 5.13, the vector-signal \( \tilde{z}[n] \) can be rewritten as

\[ \tilde{z}[n] = (I_N \otimes F^{-1}) \begin{bmatrix} \mathcal{H}B \begin{bmatrix} 1_{K_1}a_1[n] \\ 1_{K_2}a_2[n] \\ \vdots \\ 1_{K_M}a_M[n] \end{bmatrix} + \mathcal{H}[n] \end{bmatrix} \]  

(5.26)

Bearing in mind that \( \mathcal{H} \) has its column components, the C-STAR manifold vector, i.e. \( \mathcal{H}_{ik} \forall i = 1, \ldots, M, k = 1, \ldots, K_i \), the inverse Fourier transform of the C-STAR manifold vector is given as

\[ (I_N \otimes F^{-1}) \mathcal{H}_{ik} = S_{ik} \otimes (F^{-1} \mathcal{H}^{ik} F \alpha_i) \]  

(5.27)
Figure 5.3: Block diagram of the post-DFT equalizer
Due to the cyclic structure of $J_{ik}$ (Eqn. 5.10), the following property of DFT/IDFT is used

$$\mathbf{F}^{-1}J_{ik}\mathbf{F} = \begin{bmatrix}
1 & 0 & \cdots & 0 \\
0 & \exp\left(-j2\pi \frac{t_{ik}}{N_{sc}}\right) & \ddots & \\
\vdots & \ddots & \ddots & 0 \\
0 & \cdots & 0 & \exp\left(-j2\pi \frac{t_{ik}}{N_{sc}}(N_{sc} - 1)\right)
\end{bmatrix}$$

(5.28)

indicating that by pre-multiplying matrix $J_{ik}$ with $\mathbf{F}^{-1}$ and post-multiplying with $\mathbf{F}$, the matrix $J_{ik}$ is diagonalised. Using this property, the transformed C-STAR manifold vector is derived as

$$(\mathbb{I}_N \otimes \mathbf{F}^{-1}) \mathbf{h}_{ik} = S_{ik} \otimes \left( \alpha_i \otimes \begin{bmatrix}
1 \\
\exp\left(-j2\pi \frac{t_{ik}}{N_{sc}}\right) \\
\vdots \\
\exp\left(-j2\pi \frac{t_{ik}}{N_{sc}}(N_{sc} - 1)\right)
\end{bmatrix} \right)$$

(5.29)

where $\rho$ is defined in Eqn. 5.6. By defining

$$\mathbf{h}_{ik} = S_{ik} \otimes (\alpha_i \otimes \rho^{-t_{ik}})$$

and

$$\mathbb{G} = \left[ \mathcal{H}_1, \mathcal{H}_2, \ldots, \mathcal{H}_M \right]$$

with

$$\mathcal{H}_i = [\mathbf{h}_{i1}, \mathbf{h}_{i2}, \ldots, \mathbf{h}_{IK}]$$

the post-FFT MMSE receiver is obtained as

$$\mathbf{w}_{\text{MMSE}} = \text{col} \left\{ \mathbb{G}(\mathbb{G}^H \mathbb{G} + \sigma^2 \mathbb{I}_K)^{-1} \right\}$$

(5.30)

As compared to the C-STAR manifold vector formed in Eqn. 5.9, the DFT operator has transformed the chip-level circular shift of an OFDM symbol (after CP removal) into a multiplicative phase shift corresponding to the path delay. The MMSE weight vector $\mathbf{w}_{\text{MMSE}}$ is then applied to further suppress the channel MAIs based on the transformed C-STAR manifold vector in Eqn. 5.29. The output of the receiver at point C is found to be

$$y[n] = \mathbf{w}_{\text{MMSE}}^H \hat{z}[n]$$

(5.31)
which is then diversity combined and passed through the decision device to obtain the recovered data symbols, i.e.

\[ \hat{a}_1[n] = \text{sign}(y[n]) \]

### 5.4.2 Pre-FFT Signal Reception

The data flow diagram of a pre-FFT receiver is shown in Fig. 5.4, where the DFT operation at the receiver can be viewed as being integrated into the weight construction block that is directly applied on to the received signals. Such a subspace receiver is a multiuser receiver based on the subspace projection technique in order to suppress the channel ISI and MAI, and hence near-far resistant. It is derived based on the estimation of the composite channel of all users to construct the subspace of the interference signal components, that is

\[ H_{\text{intf}} = [H_2, H_3, \ldots, H_M] \]  

(5.32)

defined as the composite channel matrix with the exclusion of the desired user’s multipaths \( H_1 \). Apparently, the columns of matrix \( H_{\text{intf}} \) spans the same subspace as that of the interferences occupied by the ISI and MAI. The receiver weight matrix thus is developed as

\[
\mathbf{w}_{\text{subspace}} = \mathbf{P}_{H_{\text{intf}}}^\perp H_1 (H_{\text{intf}}^H \mathbf{P}_{H_{\text{intf}}}^\perp H_1)^{-1} H_1 \]  

(5.33)

where \( \mathbf{P}_{H_{\text{intf}}}^\perp = \mathbf{I} - H_{\text{intf}} (H_{\text{intf}}^H \mathbf{P}_{H_{\text{intf}}}^\perp H_{\text{intf}})^{-1} H_{\text{intf}}^H \) is the complementary projection operator of matrix \( H_{\text{intf}} \). This method requires the knowledge of all the users’ spreading code sequences and the estimation of all the users’ channel vectors which is most appropriate for the uplink transmission.

From Eqn. 5.11, it is known that the received signal at point A contains three components, namely the desired signals, MAIs and noise, i.e.

\[ x[n] = H_1 \beta_1 a_1[n] + \sum_{i=2}^{M} H_i \beta_i a_i[n] + n[n] \]

By applying the subspace weight matrix onto the received signal and using the complementary projection operator \( \mathbf{P}_{\text{H_{intf}}}^\perp \) nulls the MAI component, the output
Figure 5.4: Block diagram of the subspace based pre-DFT channel equalizer.
of the receiver at point B is found to be

\[ y[n] = w_{\text{subspace}}^H x[n] \]
\[ = \beta_1 (H^H H_1^{-1} H_1^H \tilde{H}_1 H_1 \beta_1 a_1[n] + w_{\text{subspace}}^H n[n] \]
\[ = |\beta_1|^2 a_1[n] + w_{\text{subspace}}^H n[n] \]

Thus, the recovered data symbols after the decision device (at point C) is obtained as

\[ \hat{a}_1[n] = \text{sign}(y[n]) \] (5.34)

### 5.5 Simulation Studies

Computer simulations are presented in this section to highlight the key benefits of introducing antenna-array technology in typical OFDM-CDMA systems. The array used in the simulations is a 5 element uniform linear array with half wavelength element spacing. For all the results presented in this section, Gold codes have been used for PN-sequences and BPSK modulation has been used throughout. Consider an OFDM-CDMA system in the presence of \( M = 3 \) co-channel users, the first user is assumed to be the desired user with SNR = 20dB and the power of the interferers is considered to be 20dB higher than the desired user (Near-Far problem). Each user is assigned a unique Gold sequence of length \( N_c = 15, 31 \) or \( 63 \) with rectangular chip pulse-shaping. The transmission bandwidth is fixed at 800KHz with a carrier frequency of \( F_c = 2 \)GHz. The array is assumed to collect a frame of 200 OFDM symbols each time for processing.

All 3 users are assumed to have 5 multipaths each with their parameters as listed in Table 5.1 and User 1 is considered to be the desired user. It is seen from Fig. 5.5(a) and (b) that all the 5 multipaths associated with the desired user, can be identified/estimated successfully using the proposed algorithm. Notice that the algorithm can still operate even when the desired user’s paths are co-located in either space or time domain\(^3\).

The estimation of complex multipath fading coefficients is performed when the space-time channel parameters are available. Both the accurate and the estimated fading coefficients of the desired user are listed in Table 5.2. It can be seen clearly that our algorithm can successfully estimate both the modulus

\(^3\)For the co-delay case, spatial smoothing is performed by partitioning the array into 2 overlapping 4-element subarrays.
Figure 5.5: (a) Surface plot and (b) Contour plot of the space-time estimation of all the desired user’s multipath.
5. Arrayed OFDM-CDMA System

Table 5.1: User’s parameters

<table>
<thead>
<tr>
<th>Path</th>
<th>$\theta_{1k}$</th>
<th>$l_{1k}T_c$</th>
<th>$\theta_{2k}$</th>
<th>$l_{2k}T_c$</th>
<th>$\theta_{3k}$</th>
<th>$l_{3k}T_c$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$k = 1$</td>
<td>70°</td>
<td>$18T_c$</td>
<td>67°</td>
<td>$28T_c$</td>
<td>29°</td>
<td>$7T_c$</td>
</tr>
<tr>
<td>$k = 2$</td>
<td>70°</td>
<td>$25T_c$</td>
<td>89°</td>
<td>$5T_c$</td>
<td>71°</td>
<td>$25T_c$</td>
</tr>
<tr>
<td>$k = 3$</td>
<td>130°</td>
<td>$18T_c$</td>
<td>74°</td>
<td>$22T_c$</td>
<td>66°</td>
<td>$20T_c$</td>
</tr>
<tr>
<td>$k = 4$</td>
<td>90°</td>
<td>$3T_c$</td>
<td>118°</td>
<td>$11T_c$</td>
<td>79°</td>
<td>$29T_c$</td>
</tr>
<tr>
<td>$k = 5$</td>
<td>100°</td>
<td>$12T_c$</td>
<td>134°</td>
<td>$16T_c$</td>
<td>103°</td>
<td>$10T_c$</td>
</tr>
</tbody>
</table>

Table 5.2: Complex fading coefficients estimation of the desired user

<table>
<thead>
<tr>
<th>Parameters</th>
<th>$\beta_{1k}$</th>
<th>$\beta_{1k}$</th>
<th>RMSE</th>
</tr>
</thead>
<tbody>
<tr>
<td>Path $k = 1$</td>
<td>0.1721 + 0.4128j</td>
<td>0.1723 + 0.4129j</td>
<td>5.952 x 10^{-4}</td>
</tr>
<tr>
<td>Path $k = 2$</td>
<td>0.2360 - 0.3799j</td>
<td>0.2358 - 0.3797j</td>
<td></td>
</tr>
<tr>
<td>Path $k = 3$</td>
<td>0.2779 - 0.3504j</td>
<td>0.2778 - 0.3505j</td>
<td></td>
</tr>
<tr>
<td>Path $k = 4$</td>
<td>0.1802 + 0.4093j</td>
<td>0.1804 + 0.4092j</td>
<td></td>
</tr>
<tr>
<td>Path $k = 5$</td>
<td>0.4464 + 0.0277j</td>
<td>0.4465 + 0.0277j</td>
<td></td>
</tr>
</tbody>
</table>

and the phase angles of channel fadings and at SNR= 20dB, the estimation error measured in RMSE is as low as 5.952 x 10^{-4}.

Having estimated all the channel parameters, Fig. 5.6 depicts the performance of the proposed receivers as compared with the STAR MMSE detector and STAR RAKE receiver formed in Eqn. 3.30 and Eqn. 3.28 respectively. It can be seen that the pre-FFT subspace receiver has a performance as good as the STAR MMSE receiver. Note that the STAR MMSE receiver is also a pre-FFT type which performs multiuser signal detection jointly in space and time domain. The STAR RAKE receiver performs badly because it is a single user receiver and is incapable of resisting the “near-far” problem. The post-FFT receiver shows a performance degradation especially in low SNR range. This is because the normalization factor multiplying the DFT and IDFT is chosen as 1 and $\frac{1}{N_{sc}}$ in the simulations. In that case, the noise at the receiver input is AWGN with zero mean and variance $\sigma_n^2$. The noise after FFT is still AWGN but has variance $\sqrt{N_{sc}}\sigma_n^2$. However, the difference in the low SNR area can be mitigated by choosing a normalization of $\frac{1}{\sqrt{N_{sc}}}$ for both the DFT and IDFT. This would make DFT become a unitary transformation.

Now, let’s consider the performance of the systems with varied number of subcarriers. The proposed post-FFT and pre-FFT receiver are used to generate the plot. Fig. 5.7 and 5.8 compare respectively the output SNIR and BER performance of the system when the number of subcarriers is $N_{sc} = 15, 31$ and
Figure 5.6: Output SNIR performance versus input SNR with the employment of (i) post-FFT receiver, (ii) pre-FFT receiver and (iii) STAR MMSE receiver and (iv) STAR RAKE receiver respectively.
63 respectively. Note that the total signalling bandwidth has been kept equal for a fair comparison. Clearly, the performance increases significantly as the number of subcarriers increases. The pre-FFT receiver again has a better performance over the post-FFT receiver due to the noise being completely suppressed.

Next, the performance of the proposed receivers are evaluated by varying the number of antenna elements at the receiver array. Fig. 5.9 and 5.10 show respectively the SNIR and BER performance of the system when the number of receiver antenna elements increases from $N = 1$ (single antenna employment) to $N = 7$. It clearly shows the performance improvement as the number of antenna elements increases.

## 5.6 Conclusion

In this chapter, an arrayed OFDM-CDMA system focusing on the uplink channels for mobile communications has been carefully designed. A blind channel estimator has been developed followed by the proposal of a pre-FFT and a post-FFT type multiuser linear receiver based on the C-STAR array manifold vector. Superior to the OFDM-CDMA system with a single antenna at base station, the combination of antenna array techniques with OFDM-CDMA transmission is...
Figure 5.8: BER performance versus input SNR as the number of subcarriers increases.

Figure 5.9: Output SNIR performance versus input SNR as the number of antenna elements increases.
undoubtedly advantageous in suppressing channel ISI and MAI.
Chapter 6

Conclusions and Further Work

In this thesis, the issue of transceiver design over space-time fading channels has been addressed. In brief, the overall objectives of the research are as following

- To improve spectrum efficiency by blindly estimating channel state information and to provide superresolution and accuracy capability
- To enhance link quality by exploiting the diverse multipath contributions inherently present in space and time domains
- To increase system capacity by optimal beamforming towards the desired direction while providing complete interference cancellation in a multiuser system

To achieve the above objectives, three types of space-time transceiver systems are developed and studied. The Spatial-Temporal ARray (STAR) architecture lays the basic framework for the works proposed in the study. A summary of the technical work contained in previous chapters now follows, as well as some suggestions for areas of continuing research.

6.1 Thesis Summary

The physical layer design of the wireless network directly determines the channel structures between which the signals are transmitted and received. The Scalar-Input Vector-Output (SIVO) channel model is used when antenna array is equipped at the base station while single antenna is employed at the mobile station. The Vector-Input Vector-Output (VIVO) channel can be applied to scenarios where antenna arrays are used at both the transmitter and receiver. The
6. Conclusions and Further Work

study of the spatial and temporal dispersiveness of the multipath propagation channel results in the development of the space-time channel models thus allowing more advanced space-time processing and optimisation criteria developed in reducing ISI/MAI generation and offering higher SNIR ratio through array gain. The proposed architectural frameworks and the inherent space-time fading channel model has put forward the groundwork in the mathematical modelling for the subsequent chapters.

In Chapter 3, the space-time array processing technique is used in an arrayed MIMO DS-CDMA system to demonstrate the substantial performance enhancement due to the employment of the antenna arrays. The Doppler-STAR manifold vector is used to exploit conjointly the spatial and temporal signatures of the VIVO MA-2 channel which provides greater degree of freedom in signal isolation and MAI/ISI cancellation. A subspace type single-user (SU) receiver is proposed which is resistant to the effect of Doppler frequency and performs complete interference cancellation over multipath fast fading channels. The proposed receiver requires only the knowledge of the desired user’s spreading code but achieves an SNIR enhancement comparative to a multiuser receiver. It is also observed that the proposed receiver is robust against erroneous channel estimation due to the inclusion of the Correlation Analysis Assignment in the detection process. Simulation results have shown that the proposed receiver achieves much better performance than the conventional receivers in time-varying frequency selective environment.

The focus switches to the joint Tx-Rx beamforming design in Chapter 4. In this study, the beamforming weights for the transmitter and receiver are jointly optimised over the whole network under the MMSE criterion. Tx-beamforming and Rx-beamforming are substantially different in nature. Rx-beamforming can be implemented independently at each receiver, without affecting the performance of other links, while Tx-beamforming at each transmitter will change the interference to all other receivers. As a result, Tx-beamforming has to be done jointly in the entire network. An iterative solution and a closed-form solution to the optimisation problem have been proposed. The iterative solution is based on the method of Lagrange multipliers. As the Tx-beamforming and Rx-beamforming weights and the Lagrange Multiplier are functions of each other, thus an iterative algorithm is devised to obtain the solution. An approximate closed-form solution based on channel eigendecomposition is later derived for the lightly loaded CDMA system where code orthogonality is usually well maintained. The numerical simulations show that the performance improvement of the proposed algorithm relative
to other Tx/Rx-beamforming algorithms is considerable.

In Chapter 5, the potential benefit of incorporating the antenna array in a multiuser OFDM-CDMA system is demonstrated by the blind joint DOA, TOA and fading coefficients estimator. The novel channel estimator is based on the C-STAR manifold vector which captures the cyclic structure of the space-time channel due to the CP inserted. The algorithm belongs to the same class of subspace superresolution techniques as MUSIC; however, the inclusion of C-STAR is not a trivial extension but enhances the receiver’s detection capability and improves the estimation accuracy and resolution. With the channel estimator employed as the front-end, two types of array receivers are proposed. The post-DFT receiver requires DFT transformation at the receiver front-end and performs interference suppression in two steps while the pre-DFT receiver is an integrated receiver which exhibits superior performance due to its strong capability in suppressing interference and noise.

6.2 List of Contributions

The list below states the main novel contributions in the thesis and is supported by the publications detailed at the front of the thesis:

- Classification of space-time channel models based on the scalar input/output and vector inputs/outputs of the channel. The key feature of the proposed models is its superiority in modelling the space-time fading channels due to the incorporation of the array manifold vector and its competence in handling the multiple access multipath fading environment.

- Proposal of the blind subspace based superresolution channel estimator in an arrayed MIMO DS-CDMA communication system. The algorithm is based on a three-dimensional search over a MUSIC-type cost function in finding the intersection of the Doppler STAR manifold with the overall signal subspace which provides the desired user’s estimates i.e. the directions, delays and frequency offset of its paths.

- Reduction in the complexity of the proposed three-dimensional estimation algorithm by splitting it into a minimisation of a two-dimensional cost function for time and direction estimation followed by parallel one-dimensional searches over Doppler frequency shift.
6. Conclusions and Further Work

- Blind estimation of the power of fading coefficients using the Doppler-STAR manifold vector. It is a subspace type algorithm and provides high accuracy estimation of the path power.

- Proposal of a novel single-user Doppler–STAR receiver for complete interference cancellation. It requires only the knowledge of the desired user’s code sequence but achieves a performance comparative to a multiuser decorrelator and is very sustainable to the Doppler spread.

- Proposed use of Correlation Analysis Assignment in identifying and grouping multipath components, providing robustness to channel estimation errors.

- Development of the system framework for joint Tx-Rx beamforming in a downlink arrayed DS-CDMA systems. An iterative algorithm is proposed under the criteria of minimising the overall MSE of the entire network. The Tx-beamforming vector, Rx-beamforming vector and the Lagrange multiplier are iteratively updated until convergence is reached.

- Derivation of a closed-form solution of the joint Tx-Rx optimisation for lightly/medium loaded CDMA systems. The method is based on channel eigendecomposition and has been shown to be a good approximation to the iterative method.

- Development of the system framework for an arrayed OFDM-CDMA system, exploiting the cyclic structure of the space-time channel and providing blind TOA and DOA estimation over frequency selective channels. A fast method for blind estimation of the channel fading coefficients in a high SNR scenario is also proposed.

- Proposal of a post-DFT array receiver for OFDM-CDMA system. It is shown that the chip-level delay in the OFDM symbol is compensated by performing the DFT transform. Based on the transformed spatial-temporal channel signature, an MMSE receiver is then derived for MAI suppression.

- Proposal of a pre-DFT array receiver for OFDM-CDMA system where ISI and MAI are suppressed in one step. This approach has provided a better performance over the post-DFT approach especially when the level of noise is high (i.e. low SNR levels).
6.3 Suggestions for Future Work

A number of different problems associated with mobile communication systems have been investigated in this thesis, but many topics are still available which can benefit from additional research effort.

1. Reconfigurable MIMO Transceiver Design

In Chapter 3, a multiplexing system is studied where the user’s data stream is demultiplexed into substreams and transmitted via multiple antennas in order to maximise the network data rate. In Chapter 4, however, the approach of joint Tx-Rx beamforming is considered. In this approach, antenna arrays are used to enhance the signal reliability of the channel where data symbols are repeated over all the antenna elements and a beamformer is applied to form the strongest beam towards the desired direction.

In fact, in MIMO systems, it is desirable to have a mobile network infrastructure which is capable of supporting services with different QoS requirement (i.e. bit-error-rate and data-rate). As a result, there exists a trade-off between signal reliability and bandwidth efficiency [101], inspiring the investigation into reconfigurable MIMO systems. When channel SNR is high, spatial multiplexing techniques (by transmitting multiple data streams for each user) can be employed at the transmitter which achieves higher transmission data rate. When channel SNR is low, the beamformer can be applied to ensure a more reliable data transmission.

Some initial research has been done in developing the reconfigurable MIMO system to meet such requirement [102] - [104]. In [103], a multimode MIMO system is proposed where the number of transmitting data substreams for each user and the data rate per substreams are adapted according to channel state information. In a high SNR scenario, the system tends to accommodate more data substreams to support the broadband services whereas in a low SNR scenario, only one substream is allocated for each user and the precoder and decoder reduce to beamforming vectors thus enhancing the output SNR after the reception. [103] considers a point-to-point MIMO system and the problem is addressed over flat fading channels. However, in this thesis, a multiuser MIMO system framework over frequency selective channels is developed which serves as a good infrastructure to build the reconfigurable scheme upon.
The multiplexing order of all users plays a vital role in ensuring different data-rate services coexist. One of the key problems in system reconfiguration is to determine the multiplexing order according to the channel conditions and the different network QoS requirements. Therefore, criterion needs to be developed to identify the optimum multiplexing order for a particular user. Also, it is intuitive that there is a so-called inter-user multiplexing trade-off, i.e. multiplexing order of one user can always be increased by sacrificing other users’ multiplexing orders. In such a case, joint design criterion is considered by varying the ranks of precoding/decoding matrices of each user.

2. Robust Transceiver Design

The performance of high resolution antenna array algorithms is limited by model errors and uncertainties. Assumptions of the models might include the knowledge of array manifold vectors, the narrowband signals and the constant direction of arrivals during the processing period. If any of the assumptions fails in a practical situation, the performance of the algorithms might suffer greatly. It is thus important to consider the robust design of the array transmitter and receiver that is

- Robust to pointing errors: Pointing errors occur if the desired DOA is not known exactly, or if there is mismatch between the actual DOA of the signal and the looking direction of the beamformer. If the beamformer has MAI suppression capability, it is also sensitive towards the pointing errors associated with the interferers. The problem can cause a significant degradation in performance and makes the beamformer incapable of handling the real world problems. It is thus important to consider the robust design to minimise the sensitivity of the beamformer against pointing errors.

- Robust to moving sources: Array processing algorithms have been developed in this thesis for non-stationary environment. Doppler frequency effect caused by the source motion has been studied extensively. An assumption made in this study is that angular motion due to the source motion is negligible during the observation of the array data. However, the usefulness of superresolution estimation techniques such as MUSIC is limited by modelling errors. If any of the assumptions in the model is not valid, their estimation accuracy will suffer. It is thus
desirable to assess the effects of source motion on signal subspace and analyse the factors that determine the motion induced errors.

3. **Arrayed MIMO System with Space-Time Blocking Coding (STBC)**

The advantage of STBC can be demonstrated by the fact that, among various kinds of ST coding, only STBC has been implemented in practical wireless communication standards. Several diversity schemes have been proposed and adopted in 3G standards which are essentially based on the Space-Time Block Coding (STBC) transmission strategy suggested by Alamouti [52]. Space-Time Transmit Diversity (STTD) scheme first proposed by Texas Instruments has been accepted by 3GPP as the open loop transmit diversity proposal to WCDMA [105]. Other two open loop transmit diversity techniques submitted and adopted by CDMA 2000 [106] are the Space-Time Spreading (STS) scheme proposed by Lucent Technologies and the orthogonal transmit diversity (OTD) by Motorola.

Transmit antenna diversity as a means of further enhancing the system capacity and performance can be incorporated into the arrayed MIMO communication system. By using these transmit schemes, additional diversity can be gained without having to exhaust the number of spreading codes available. Many reported research works on STBC, including the conventional STTD, STS and OTD schemes, often assume a simplified channel which is stationary and flat fading. Through the incorporation of the space-time channel model and the subspace based array technology, the transmit diversity scheme can be generalised to the multiuser communication scenario with the most general form of propagation channel in terms of Doppler shift, time delay and direction of arrival for each multipath ray.

4. **Blind/Semi-Blind Implementation of Joint Tx-Rx Beamforming**

For both of the proposed joint Tx-Rx beamforming techniques in Chapter 4, knowledge of each user’s downlink channel is needed to formulate the solution. Since information about the downlink channel is usually unavailable to the base station, some recent works have suggested the use of uplink information in designing downlink beamforming weights. Those proposed approaches can be generally classified into two categories: non-DOA based approaches and DOA based approaches. Non-DOA based approaches can be found in [44][45] which is usually designed for TDD systems. For systems that operate under FDD mode, calibration must be carried out in either
spatial or frequency domain to compensate the array response mismatch due to the frequency spread. The DOA based approach offers the advantage of higher accuracy and does not pose any restrictions on the array geometry nor requires array calibration. In fact, long-term parameters depend on the large-scale physical properties of the radio terrain, thus remain invariant for the uplink and downlink transmission. This motivates the construction of downlink channel matrix using uplink parameters, such as the number of significant paths, path delays and directions as well as the slow path fadings. Blind algorithms that have been proposed in this thesis can be well adapted to the joint beamforming system. Also note that, as the transmit array manifold vector also forms part of the space-time channel modelling, algorithms for estimating the direction of departure (DOD) of the multipath also needs be developed to aid the beamforming weight construction.

5. The Centralised and Distributed Tx-Rx Weight Optimisation

The joint Tx-Rx optimisation in a downlink multiuser wireless system requires each user to have the channel state information and the beamforming weights of all users. Transceivers are designed in a centralised way and requires a significant user cooperation. Using the blind estimation methods proposed in this thesis, the transmission and the base station may be able to obtain, in a totally blind way, the associated channel parameters (i.e. DOA, DOD, TOA, Doppler frequency and path power) of its own. However, base station and mobile stations still rely heavily on the feedback channels to pass on the weight updates of the whole network. It seems more natural, however, to consider distributed solutions that do not require significant user cooperation. Such an approach fits naturally within the framework of game theory in which each user could be considered as a player in a game with multiple players. In any case, both centralized and distributed problem formulations are currently open problems.
References


REFERENCES


REFERENCES


[98] G. Leus and M. Moonen, “Per-tone equalization for mimo ofdm systems,” *Signal Processing, IEEE Transactions on* [see also Acoustics, Speech, and


