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Antenna arrays for W-CDMA and UWB communications systems

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**ANTENNA ARRAYS FOR W-CDMA AND UWB COMMUNICATIONS
SYSTEMS**

A Thesis

Presented to

The Faculty of the Department of Electrical Engineering

San Jose State University

In Partial Fulfillment

of the Requirement for the Degree

Master

by

Haidong Wu

May 2005

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A handwritten signature in black ink, appearing to read 'R. Morelos-Zaragoza', written over a horizontal line.

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ABSTRACT

ANTENNA ARRAYS FOR W-CDMA AND UWB COMMUNICATIONS SYSTEMS

by Haidong Wu

This thesis addresses the adaptive antenna array techniques for the W-CDMA and UWB systems. For the third generation and future wireless communications systems with high bit-rate signals and variable environments, it is necessary to develop efficient schemes with fast and stable performances.

A modified lattice algorithm is proposed for the W-CDMA systems. Simulations show that this algorithm has fast and stable convergence as well as good multi-user interference suppressions.

The symbol-rate reception scheme for the W-CDMA system is also proposed. The antenna array processing is performed on the chip-level received signals over the symbol period. Analysis and simulation show that this scheme has very good BER and multi-user interference suppression performances.

The antenna array processing for the UWB systems is different from the conventional techniques. The antenna selection is proposed. Its easy operation and good performance make this technique applicable for the UWB systems with very high-rate signals under dense multipath environment.

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LIST OF ABBREVIATIONS

ASIC	Application Specific Integrated Circuit
AWGN	Additive White Gaussian Noise
BER	Bit Error Rate
Bps	Bits Per Second
CDMA	Code Division Multiple Access
DA	Data-Aided
DAC	Delay Adjustment Computer
DOA	Direction of Arrival
DS-CDMA	Direct Spread CDMA
DSP	Digital Signal Processing
EVM	Error Vector Magnitude
FIR	Finite Impulse Response
FPGA	Field Programmable Gate Array
GSM	Globe System for Mobile Communications
I&D	Integrate and Dump
i.i.d	Independently and Identically Distributed
IMT-2000	International Mobile Telecommunications-2000
LMS	Least Mean Square Algorithm
LPF	Low-Pass Filter
LOS	Line of Sight
LS	Least Square
LSL	Least Square Lattice Algorithm
Max SINR	Maximum SINR

Mcps	Mega Chips
MF	Matched Filter
ML	Maximum Likelihood
MMSE	Minimum Mean Square Error
MRC	Maximum-Ratio Combining
NLMS	Normalized LMS Algorithm
NLOS	Nonline of Sight.
OVSF	Orthogonal Variable Spread Factor
PARCOR	Partial Correlations
PDF	Probability Distribution Function
QoS	Quality of Service
RLS	Recursive Lease Square algorithm
RF	Radio Frequency
SC	Successive Channel
SDMA	Space Division Multiple Access
SF	Spreading Factor
SINR	Signal to Interference plus Noise Ratio
SIR	Signal to Interference power Ratio
SNOI	Signal Not of Interest
SNR	Signal to Noise Ratio
SOI	Signal of Interest
SW	Sliding Window
ULA	Uniformly Linear Antenna
UWB	Ultra Wide Band
VLSI	Very Large Scale Integration

VSS LMS	Variable Step Size LMS Algorithm
W-CDMA	Wideband CDMA
WMSA	Weighted Multislot Averaging
WPAN	Wireless Personal Area Network

LIST OF SYMBOLS

$\otimes$	convolution operation
$\lfloor \cdot \rfloor$	taking the integrate part
$(\cdot)^T$	transpose operation
$(\cdot)^*$	conjugate operation
$(\cdot)^H$	Hermitian transpose (conjugate transpose)
$(\cdot)^{-1}$	inverse operation
$E(\cdot)$	expectation operation
$\frac{\partial(\cdot)}{\partial(\cdot)}$	partial differential
$\ \mathbf{w}_1^i(n)\ $	norm operation, $\ \mathbf{w}_1^i(n)\ = \sqrt{\mathbf{w}_1^{iH}(n)\mathbf{w}_1^i(n)}$.
$\text{tr}(\cdot)$	trace operation
$C_{k,D}(i)$	orthogonal channelization code for W-CDMA data channel
$C_{k,P}(i)$	orthogonal channelization code for W-CDMA pilot channel
$S_{k,I}(i) + jS_{k,Q}(i)$	complex scrambling code
$D_k(m)$	data sequence of the data channel
E_C	chip energy
T_C	chip interval
T_P	pilot bit interval
L	total number of the resolvable multipaths

P_k	the total number of the resolvable multipaths
$h_{u,k}(t)$	channels time responses
$\alpha_{u,k,l}$	channel fading
$\varphi_{u,k,l}$	phase delay of the l -th path
$\tau_{u,k,l}$	time delay of the l -th path
λ_{carry}	carrier wavelength
θ_k	incident angle of the signal
f_{carry}	carrier frequency
K	total number of active users
$n_u(t)$	additive white gaussian noise with double-sided power spectral density of $N_0/2$.
$p(t)$	the UWB pulse
$\mathbf{w}_{1,U}^i$	vector of the weight factor
$\hat{X}_1^{iR}(n)$	locally generated reference signal
$\tilde{\mathbf{f}}_u^u(n)$	the u -th element in the vector $\tilde{\mathbf{f}}_u(n)$.
$\mathbf{f}_u^{1,u}(n)$	sub-vector with the first to u -th elements of vector $\mathbf{f}_u(n)$.
$\bar{\mathbf{J}}$	transformation matrix
$\hat{\mathbf{X}}_{U,1}^i(n)$	input signal vector
$\kappa_{f,u}(n-1)$	forward prediction reflection coefficient
$\kappa_{b,u}(n-1)$	backward estimation reflection coefficient
$\mathbf{I}$	$U \times U$ identity matrix
λ	positive, constant weighting factor

$G(\theta)$	beamforming pattern
$C_{ch,256}^0$	channelization code for pilot signal
f_D	Doppler frequency
$P_e(\cdot)$	bit error probability
$p(\cdot)$	probability density function
N_E	channel estimation period

CHAPTER 1

BACKGROUND AND OVERVIEW

1.1 Introduction

Smart antennas are now regarded by many within the wireless communications industry as a core system component in the third, fourth, and the future-generation mobile networks. However, antennas are not smart. It is the reception schemes and digital signal processing that make the system smart. Although it might seem that smart antennas are the new technology, the fundamental theory of smart antennas is not new. In fact, they have been applied in defense-related systems since World War II.

In recent years, with the demands for enhancing the system performances of the wireless communications as well as the emergence of powerful hardware instruments and innovative signal processing algorithms, smart antenna systems have become practical for commercial usage. As a result, motivated by these demands, in this thesis theoretical researches are carried out to propose and investigate new signal processing algorithm and reception schemes for the W-CDMA and UWB wireless communication systems. In this chapter, some backgrounds and overviews related to the smart antennas, including the adaptive antenna arrays as well as W-CDMA and UWB systems, are introduced. In Section 1.2 there are basic ideas of the smart antennas. Fundamental issues for the application of smart antennas are presented in Section 1.3. Brief introductions of W-CDMA and UWB systems are shown in Section 1.4. Finally, the motivation and

contribution of this thesis are shown in Section 1.5.

1.2 The Development of Smart Antennas

Dr. Andrew Viterbi has said that spatial filtering is the next step in improving the capacity of cellular systems. Smart antennas are just the approach to realize the spatial filtering. By exploiting the spatial domain, smart antennas can combat multipath fading of the signal from desired user and suppress signals from interfering users. Accordingly, smart antennas offer a broad range of benefits as capacity enhancement, coverage extension, high data rate transmission, improved power control, reduction of multipath, and co-channel interference. These benefits are very helpful and crucial for the current third generation mobile communications systems, such as W-CDMA, CDMA2000, fourth generation mobile communication systems such as UWB, and future generation systems, which are all aiming for the high data rate and multimedia wireless services. In these wireless systems, high data rate users due to their high transmission power may cause even larger interference to the low rate users and the multi-user and co-channel interference are main problems that affect the system performances. Smart antennas are efficient methods to suppress large multi-user interferences and provide substantial capacity improvements.

Several projects and field trials have emerged reinforcing the benefits of smart antennas systems. For examples, Ericsson-Mannesmann employed and tested smart antennas for the GSM, realizing the capacity of 100%-200% increase. Jack H. Winters and AT&T research lab tested a four-element antenna array in an IS-136 system and got a 40% range increase in a Raleigh fading environment. Mamoru Sawahashi et al from

NTT Docomo in Japan tested a smart antenna for the W-CDMA networks and showed a substantial improvement in average BER. Furthermore, the Commission of the European Community specially sponsored the TSUNAMI project to do further research on the smart antennas for the wireless communication systems.

Researches on smart antennas have focused on both the architecture of antenna elements and the signal processing. Work on the signal processing has concentrated on the design of efficient algorithms for DOA estimation and adaptive antenna arrays. Recent work on DOA estimation is presented by Swindlehurst & Kailath (1993), Zoltowshi, Haardt, & Mathews (1996), and Strobach (2001). While, researches on algorithms for adaptive antenna arrays range from the LMS (Widrow & Sterns, 1985), NLMS, RLS (Liberti & Rappaport, 1999) to some new algorithm: VSS LMS and some complicated algorithm (Foutz & Spanias, 2001).

In W-CDMA systems, smart antennas can be used to spatially separate signals so as to allow different users to share the same spectral resources, provided that they are spatially separable at the base station. Thus, more than one user can be allocated simultaneously to the same transmission channel in the same sector or cell with only the angular separation. Therefore, it greatly increases frequency reuse and system capacity, and reduces infrastructure cost. In UWB systems, although there are no cellular structures and not too many users, to enhance the transmission range and to further increase the maximum number of simultaneous users, and to mitigate the multi-user interferences, multiple antenna or antenna array techniques are promising approaches.

1.3 Smart Antenna Systems

1.3.1 *Switched-beam and Adaptive Antenna Array Systems*

Smart antenna systems are achieved by the use of antenna arrays. An antenna array is comprised by several antenna elements that are spatially arranged at specified locations. The linear, circular, and planar arrays are the three most commonly used geometries, but it has been found that the geometry of the array does not have a significant effect on the performance of the antenna array. The antenna arrays focus their radiation pattern (patterns defined in the form of the relative gain corresponding to different incident angles and comprised by several beams) toward the desired user while rejecting the unwanted interference. These systems are mainly classified as switched-beam and adaptive antenna array systems.

Switched-beam systems select the radiation patterns from the preset patterns in order to enhance the received signals. When the mobile user moves through the 120° sector (now cellular systems usually use 120° sectorization), the system detects the characteristic of the received signals and then chooses the appropriate preset patterns. Because these systems have the fixed patterns, the overall gain for the desired user and suppressions for the interferers are depended on the location of the users very much. If some interferer is closer to the active part of the pattern than the desired user, it may be got more enhancement than the desired user.

The concept of adaptive antenna array was introduced almost four decades ago to advance the performance of military radar (and possibly radio communications) operating

in a dense signal environment, where active interference can be present and pose a major threat. Adaptive antenna array electronically points a main beam toward a desired signal source without a priori knowledge of the source's angular position and at the same time, suppresses interference and noise arriving from other look directions. The operation of such an antenna system is controlled by an adaptive signal processing algorithm that is designed to achieve specific performance criteria for a desired signal. The systems applying adaptive antenna arrays are capable of modifying their time, frequency, and spatial response by means of amplitude and phase weighting and internal feedback control. Adaptive antenna arrays with U antenna elements can form up to $U-1$ nulls to cancel up to $U-1$ interferers or the SNOIs and direct the main beam with gain towards the desired user or SOI. Therefore, adaptive antenna array can customize an appropriate pattern for each individual user. Hence, performances of the adaptive antenna arrays are better than the performances of the switched-beam systems (Choi, Shim, & Sarkar, 1999). Also, when a high-level interference is present, the interference rejection capability of the adaptive antenna system provides significantly more coverage than switched-beam system. Adaptive antenna arrays can locate and track both the desired user and interferers, and can automatically adjust the antenna patterns to enhance the performances of the reception by using different the signal processing algorithms.

1.3.2 Adaptive Beamforming

The results of the signal processing algorithms are the complex weight factors of the antenna arrays, the functions of which are to amplify or attenuate signals from different antenna elements. Then summers are used to add all of the received signals in

order to tune out the SNOIs, while enhancing the SOI. Such a process is so-called the beamforming. It is sometimes referred to as the spatial filtering, because received signals from some incident direction are filtered out, while others are amplified. Actually, such a spatial filtering is analogous to designing an FIR filter for filtering a time signal.

The adaptive algorithms that generate the complex weight factors are generally classified as DOA-based and temporal-reference-based, or signal-structure-based (Boukalov & Haggman, 2000). In DOA-based algorithm, the algorithms pass the DOA information to the beamformer (consisted by antenna elements, complex weight factor, and the summer). This is used to design a beamforming patterns with the main beam directed to the SOI and with nulls in the directions of SNOIs.

On the other hand, temporal-reference-based algorithms use a known reference signals to adjust the weight factors of the adaptive antenna arrays and to form a beamforming pattern with maximum to the SOI and nulls to the SNOIs. In the realization of this algorithm, errors between the beamformer output and the reference signal is used to update the complex weight factors. Once the beamformer has locked onto the reference signal, the complex weight factors are maintained fixed and so are the beamforming patterns.

1.3.3 Antenna Selection

Although adaptive antenna array can improve the capacity and reliability of wireless communications, the multiple adaptive signal processing for multiple antennas is costly in terms of size, power, and hardware. As a result, antenna selection is a low-cost,

low-complexity alternative to capture many of the advantages of the antenna array (Sanayei & Nosratinia, 2004). Actually, antenna selection is different from the adaptive antenna array. It is the diversity via multiple receiver antennas and is a direct extension of traditional receiver diversity ideas. Received signals from different antennas are considered as several versions of the transmission signal, each experiencing a different channel response, including fading and delay. Antenna selection chooses the antenna with the highest SNR, and performs detection based on the signal from the selected antenna.

1.4 W-CDMA and UWB Systems

W-CDMA is short for Wideband Code Division Multiple Access. It is a technology for wideband digital radio communications of Internet, multimedia, video, and other capacity-demanding applications.

WCDMA is the dominating 3G technology, providing higher capacity for voice and data and higher data rates. WCDMA uses a new spectrum with a 5 MHz carrier, providing 50 times higher data rate than in present GSM networks (and 10 times higher data rate than in GPRS networks). WCDMA handles up to 2 Mbps for local area access or 384 Kbps for wide area access.

WCDMA is also known as UMTS and has been adopted as a standard by the ITU under the name IMT-2000 direct spread. The gradual evolution from today's systems is driven by demand for capacity, which is required by new and faster data based mobile services. WCDMA enables better use of available spectrum and more cost-efficient

network solutions. The operator can gradually evolve from GSM to WCDMA, protecting investments by re-using the GSM core network and 2G/2.5G services.

UWB is short for Ultra-Wideband. This technology has the key attributes necessary to add significant value for consumers of wireless home entertainment and mobile multimedia products. Media servers, set-top boxes, flat-panel screens, digital camcorders, and other multimedia applications need a high data-rate and high Quality of Service wireless connection to help ensure “wire-like” performance.

Ultra-Wideband technology brings the convenience and mobility of wireless communications to high-speed interconnects in devices throughout the digital home and office. Designed for short-range, wireless personal area networks, UWB is the leading technology for freeing people from wires, enabling wireless connection of multiple devices for transmission of video, audio, and other high-bandwidth data.

A traditional UWB transmitter works by sending billions of pulses across a very wide spectrum of frequencies several GHz in bandwidth. The corresponding receiver then translates the pulses into data by listening for a familiar pulse sequence sent by the transmitter. Specifically, UWB is defined as any radio technology having a spectrum that occupies a bandwidth greater than 20 percent of the center frequency, or a bandwidth of at least 500 MHz.

UWB’s combination of broader spectrum and lower power improves speed and reduces interference with other wireless spectra (Li & Rusch, 2002). In the United States, the Federal Communications Commission has mandated that UWB radio transmissions

can legally operate in the range from 3.1 GHz up to 10.6 GHz, at a limited transmit power of -41dBm/MHz . Consequently, UWB provides dramatic channel capacity at short range that limits interference.

1.5 The Motivation, Contribution, and Organization of this Thesis

In this thesis, thorough researches on the smart antenna technologies for W-CDMA and UWB systems are presented. There are three goals of this thesis.

The first goal of this thesis is to investigate the realization and performance of digital signal processing algorithms of the antenna arrays for the W-CDMA systems. At the present, mobile users may move rapidly so that the DOAs of their signals can change rapidly as well. Furthermore, because the wireless environments are time varying, the weight factors must be updated continuously. Consequently, for realization the adaptive antenna array technique, it is important to apply adaptive algorithms with both fast convergence and good performance in steady state for the antenna arrays so that the weight factors can be timely and correctly adjusted to maximize the quality of the signal from desired user. So motivated by this crucial part of the adaptive antenna arrays, research of this thesis is mainly focused on proposing, investigating, and verifying new algorithm with above advantages for the adaptive antenna arrays.

So far, LMS, RLS, and some other algorithms have been widely studied for the beamforming of antenna arrays. The main advantages of LMS are easy analysis and simple implementation. But, as well known, LMS algorithm is sensitive to the step size and the input signals. The smaller the step size is, the slower the convergence is. And the

step size must be selected right according to the input signals. As a result, if the step size is inappropriately selected, it is possible that the algorithm cannot converge so that the beamformer output is not correct. Since in the multimedia mobile systems, the data rates are high, for example 2Mbps or higher in the third generation wireless communications networks, the period to be used for adaptive updating the weight factors and beamforming is very short. Thus, LMS and some modified LMS algorithms are not the appropriate solution for the adaptive antenna array because of their slow and unstable convergent performances as well as sensitivity to the input signals. RLS algorithm is insensitive to the input signals and has a very good performance with both fast convergence and stable adaptive performance, but its computational complexity is a little higher. As a result, it costs high payload in the implementation especially when the number of the antenna elements is large. So, one of the contributions of this thesis is to design and verify a new fast-convergent and efficient algorithm: the lattice algorithm.

The lattice algorithm has been studied for different applications. It has been shown that lattice has the following good properties: efficient computation, modular structure, fast convergence, and linear increase of computation with the number of stages. These are results of its efficient and stage-by-stage structure. In the lattice algorithm, two main operations: the forward prediction and backward estimation are performed simultaneously and the operations are decoupling from stage to stage so that the inputs of the final update computation are uncorrelated (Haykin, 2001). This algorithm is designed to apply some the same principles for the RLS algorithm, so the lattice algorithm is also insensitive to the input signal (received signals from different antenna elements) and

converges very fast.

But because the traditional lattice algorithm is only design for the time sequences, it cannot be directly used for the adaptive antenna array with array signals. Therefore, some work of this thesis is to modify the traditional lattice algorithm. Thus, the innovated algorithm is appropriate for the antenna array and inherits the advantages of the traditional lattice algorithms.

The second goal of this thesis is to investigate the reception scheme of antenna arrays for the W-CDMA systems. In most researches, adaptive antenna array techniques are performed on the de-spread pilot symbols. As a result, several symbols are needed to obtain the convergent results from the adaptive algorithms. Under some conditions with not good channel estimation or large interference, it takes more symbols to get convergent results. Also with a long adaptation period, the channel information and incident angle information of all users maybe have already changed. Consequently, it is important to have a fast and efficient reception scheme to obtain more samples for the adaptation and get the convergent results in a short time in which the channel does not change. So, in this thesis the joint process of de-spreading and adaptive antenna array processing is proposed and investigated. In this process, the adaptive antenna array processing is never performed on the symbol-rate pilot samples of the desired user, but rather on the chip-rate pilot samples, the demodulated pilot samples out from the matched shape filters. The adaptation results, the antenna array weight vectors, can be updated symbol-by-symbol, or per several symbols. Thus, for example within a symbol time (256 chips), the adaptations are performed on 256 samples that are enough for some efficient

adaptive algorithms. Even though the convergent results need more than one symbol to complete the adaptation, the total adaptation time is in the order of a symbol period. In such a period, the channel is assumed unchanged. Then, the convergent antenna array vectors are used to combine the data inputs from different antenna elements to complete the beamforming. Beamforming can be performed on the demodulated chip-rate data samples or the de-spread symbol-rate data samples. Both methods give the same result. The weight vector adaptation and beamforming are performed for each resolved path of the desired user. Finally, the beamformed data symbols from each path are combined by a RAKE receiver and sent to a decision device or a soft-input decoder. Through the analysis, it is shown that because the adaptation is performed within the symbol period, it makes usage of the orthogonal properties of the spreading codes for different users. The pilot signals of desired user are assumed to be orthogonal to the interferences over the symbol period so that the cost function of the adaptive algorithms can get a minimum result over the symbol period.

The performance of the joint symbol-rate update adaptive antenna array processing with the aid of the traditional channel estimation is also investigated. The channel estimation is applied to re-construct the reference signals. It is obtained by multiplying the channel estimation results with the known pilot signals. There are some known complex channel estimation methods like weighted multislot averaging and modified WMSA. But in the research of this thesis, the performance of a WCDMA uplink adaptive antenna array processing is thoroughly analyzed and simulated. In this case, the coherent time of the channel is in the order of the slot period and also the total

phase change of the desired user due to the Doppler shift is small over the slot period. Therefore, it is assumed that the channel does not change within a slot and use a simple averaging estimator and not the WMSA estimation.

The third goal of this thesis is to investigate the method to apply antenna array technology for the UWB systems. Actually, applications of the antenna array for the UWB systems are different from the processing for W-CDMA systems, so it is necessary to find some way to realize the antenna array technologies for the UWB systems with the consideration of the tradeoffs between the performances and the computational complexities. The antenna selection together with the antenna combining is considered to realize the antenna array processing for the UWB systems. The antenna selection method assumes that the U antennas in the array are separated widely enough so that the transmitted channels are assumed independent. So, from the antenna array, there are U replicas of the received signal from the U different channels. Because these replicas are independent, they can be thought as the U independent samples of the received signals. By comparing the performances of the all received signals, receiver selects the received signal from the antenna from which the performance is the best. The comparisons are based on the SINR.

The remaining parts of this thesis are organized as follows. Chapter 2 describes the detailed W-CDMA and UWB models used for the analysis of the adaptive antenna array. In Chapter 3 the investigations of the modified lattice algorithm, as well as RLS algorithm for adaptive antenna array in W-CDMA systems are presented and the modified lattice algorithm is proposed with thorough theatrical explanations. Overall

simulation results are also presented. In Chapter 4, symbol rate processing of adaptive antenna array for W-CDMA system with traditional channel estimations is explained and with detailed theoretical analyses and simulations. In Chapter 5, a practical scheme on the realizations of antenna array for the UWB systems is proposed with comprehensive examinations and simulations. Finally, conclusion and summary are drawn in Chapter 6.

CHAPTER 2

ANALYTICAL MODELS OF THE ANTENNA ARRAYS, W-CDMA, AND UWB

The models for the analysis and investigation of the antenna array for the W-CDMA and UWB systems are introduced in this chapter. System model of the W-CDMA and UWB, wireless channel model and antenna array model are discussed respectively in Section 2.1, Section 2.3, Section 2.3, and Section 2.4. And the input signals for the adaptive antenna array processing under W-CDMA and UWB models are presented in Section 2.5 and 2.6 respectively.

2.1 The W-CDMA System Model

2.1.1 *Transmitter Structure*

W-CDMA system has been accepted as a wireless access scheme for IMT-2000 (Adachi, Sawahashi, & Suda, 1998; 3GPP, 2000). In this system, all the users share the same carrier under the DS-SS principle. The uplink of the W-CDMA system is further considered for the adaptive antenna arrays in this thesis. As shown in Figure 2-1 (Dahlman, Gudmundson, Nilsson, & Skold, 1998; *IEEE* 1999), there are two types of channels in the uplink: control (pilot) channel and data channel. The pilot channel is used for coherent reception. The data and pilot channels are mapped onto I/Q branches respectively in the transmitter (data-I branch; pilot-Q branch). Both branches are spread

by two different orthogonal channelization codes, $C_{k,D}(i)$ and $C_{k,P}(i)$. The resultant signals from both I and Q branches are added together to form one complex signal, which is subsequently scrambled by the complex scrambling code $S_{k,I}(i) + jS_{k,Q}(i)$. For simple analysis, only the baseband equivalent transmission is considered and the cases with only one desired user and several interference users are assumed to perform the adaptive antenna array processing after matched filtering. It is also assumed that perfect power control has been done and all users are modeled under the principle of the W-CDMA system frame format. Thus, the baseband signal of the k -th user can be expressed as

$$\begin{aligned} y_k(n) &= \sqrt{E_C} \{ D_k(m) \cdot C_{k,D}(n) + j\sqrt{\beta}C_{k,P}(n) \} \cdot \{ S_{k,I}(n) + jS_{k,Q}(n) \} \\ &= \sqrt{E_C} \{ \sqrt{\beta}\bar{S}_{k,P}(n) + D_k(m)\bar{S}_{k,D}(n) \} \end{aligned} \quad (2-1)$$

where $D_k(m)$ is the data sequence with bit interval T and $m = \lfloor n/N \rfloor$, where N is the spreading factor. E_C is the chip energy, β is the power ratio of pilot to data channels.

$\bar{S}_{k,P}(n)$ and $\bar{S}_{k,D}(n)$ are given by

$$\bar{S}_{k,P}(n) = jC_{k,P}(n) \{ S_{k,I}(n) + jS_{k,Q}(n) \} \quad (2-2)$$

$$\bar{S}_{k,D}(n) = C_{k,D}(n) \{ S_{k,I}(n) + jS_{k,Q}(n) \} \quad (2-3)$$

so the resultant baseband up-link signal of k -th user is

$$\begin{aligned} S_k(t) &= \sum_{n=-\infty}^{\infty} \bar{S}_k(n)h(t - nT_C) \\ &= \sqrt{E_C} \left\{ \sqrt{\beta} \sum_{i=-\infty}^{\infty} \bar{S}_{k,P}(n)h(t - nT_C) + D_k(m) \sum_{n=-\infty}^{\infty} \bar{S}_{k,D}(n)h(t - nT_C) \right\} \\ &= \sqrt{E_C} \{ \sqrt{\beta}\bar{S}_{k,P}(t) + D_k(m)\bar{S}_{k,D}(t) \} \end{aligned} \quad (2-4)$$

where $h(t)$ stands for the impulse response of the pulse shaping filter which in the WCDMA system is a root-raised cosine filter with a rolloff factor of 0.22. $\bar{S}_{k,P}(t)$ and $\bar{S}_{k,D}(t)$ are given as followings

$$\bar{S}_{k,P}(t) = \sum_{n=-\infty}^{\infty} \bar{S}_{k,P}(n)h(t - nT_C) \quad (2-5)$$

$$\bar{S}_{k,D}(t) = \sum_{n=-\infty}^{\infty} \bar{S}_{k,D}(n)h(t - nT_C) \quad (2-6)$$

So, according to Figure 2-1, the transmitted signal of the k -th user can be written as

$$\begin{aligned} y_k(t) = & \sqrt{E_{k,D}} \sum_{n=-\infty}^{\infty} [D_k(m)C_{k,D}(n)S_{k,I}(n) - \sqrt{\beta_k}C_{k,P}(n)S_{k,Q}(n)]h(t - nT_c) \cos(\omega_0 t + \phi_k) \\ & - \sqrt{E_{k,D}} \sum_{n=-\infty}^{\infty} [D_k(m)C_{k,D}(n)S_{k,Q}(n) + \sqrt{\beta_k}C_{k,P}(n)S_{k,I}(n)]h(t - nT_c) \sin(\omega_0 t + \phi_k) \end{aligned} \quad (2-7)$$

ϕ_k is the carrier phase.

2.1.2 Wireless Channel Model

The antenna arrays transmit or receive signals through a wireless propagation channel. The propagation channel comprises many paths with different time delays. The wireless channel is characterized by a multipath and fading model. The simplified tapped delay line multipath channel model is utilized. Therefore, The complex low-pass equivalent impulse response of the channel of the k -th user at the u -th antenna element can be expressed as,

$$h_{u,k}(t) = \sum_{l=1}^{P_k} \alpha_{u,k,l}(t) \delta(t - \tau_{u,k,l}(t)) e^{j\varphi_{u,k,l}(t)} \quad (2-8)$$

where P_k represents the total number of the resolvable multipaths for the k -th user, $\alpha_{u,k,l}$ stands for the channel fading, $\varphi_{u,k,l}$ and $\tau_{u,k,l}$ represent the phase and the time delay of the l -th path, respectively.

2.1.3 Antenna Array Model

The adaptive antenna arrays are implemented by capturing the RF signals from the all the antenna elements and converting them into two branches of binary baseband I and Q signals, which jointly represent the amplitudes and phases of the signals received at the elements of the array. Then antenna array weight factor is adaptively updated and adjusted. Thereby, these weight factor forms the desired pattern of gain and suppressions when weighting the received signals together.

An adaptive antenna array consists of an array of antenna elements. In general, the array may consist of a number of antenna elements distributed in any desired pattern as uniformly spaced linear array, uniform circular, or uniformly spaced planar array. A linear array with the distance of half a carrier wavelength ($d = 0.5\lambda_{\text{carry}}$) between two adjacent elements is assumed (Adachi, Sawahashi, & Suda, 1998; 3GPP, 2000; Liberti & Rappaport, 1999). Figure 2-2 shows DOAs on the antenna array. The same fading signal is considered to arrive at different antenna elements with the same amplitude but different phase lead. It is assumed that the user is located at a sufficiently long distance to the base station so that the DOA is the same at all antenna elements (Adachi, Sawahashi, & Suda, 1998). Thus, the phase lead of the k -th user signal at the u -th antenna element with respect to the first element is

$$\phi_{u,k} = (u - 1)\pi \sin \theta_k \quad (2-9)$$

where θ_k is the incident angle of the signal.

Because all antenna elements are placed very close and the number of elements is small, the distance between two ending elements is small. For example, when carrier frequency f_{carry} is 2GHz, the transmission time between two adjacent elements is about

$$2.5 \times 10^{-10} \text{ seconds } \left(\frac{0.5\lambda_{\text{carry}}}{c} = \frac{1}{2f_{\text{carry}}}, c = \lambda_{\text{carry}} \cdot f_{\text{carry}} = 3 \times 10^8 \text{ m/s} \right), \text{ or about}$$

9.6×10^{-4} of a chip interval (chip rate is 3.84 Mcps). When six elements are considered, the transmission time from the first element to the last element is only

$5 \times 9.6 \times 10^{-4} = 0.005$ of one chip interval. Thus, the power delay profile of each path signal is assumed to be the same at all elements, i.e.

$$\alpha_{1,k,l} = \alpha_{u,k,l} = \alpha_{k,l}, \quad u = 1, 2, \dots, U \quad (2-10)$$

It can be Rayleigh, Rician, or Nakagami random variable according to the specifically applied channel model. Moreover, it is assumed that all users have the same number of multipaths, L . As a result,

$$h_{1,k}(t) = h_{U,k}(t) = h_k(t) = \sum_{l=1}^L \alpha_{k,l}(t) \delta(t - \tau_{k,l}(t)) e^{j\varphi_{k,l}(t)} \quad (2-11)$$

But, the phase for the m -th element is different from element to element as:

$$\varphi_{u,k,l} = \varphi_{1,k,l} + \phi_{u,k} = \varphi_{k,l} + \phi_{u,k} \quad (2-12)$$

2.1.4 Input of the Adaptive Antenna Array Processing

The composite spread signal received on each antenna element was converted into baseband I and Q branches by a quadrature detector. After being sampled using an A/D converter and being filtered using a wave shaping filter, the processed signal sequences on I and Q branches are fed into the MF. Then the signal sequences on I and Q branches are despread by the MFs and resolved into multiple copies responding to the different propagation paths. The MFs are implemented as a bank of correlators. Each correlator consists of a multiplier and I&D filter. In each MF, the received signal is first multiplied with a locally generated spreading code (scrambling code multiplying channelization code). And the time delay of these local spreading codes are set to be same as the measured time delay of each resolvable propagation paths, and then integrated by the I&D filter over data symbol interval. Thus, the received signal at the m -th antenna element is given by

$$\mathbf{X}(t) = \mathbf{\Theta} \bar{\mathbf{Y}}(t) + \mathbf{N}_U(t) \quad (2-13)$$

where matrix $\mathbf{\Theta}$ represents the effect of the ULA with U antenna elements on the total K received signals and $\mathbf{\Theta}_k = [1 \ \phi_{2,k} \ \dots \ \phi_{U,k}]^T$ so,

$$\mathbf{\Theta} = [\mathbf{\Theta}_1 \ \mathbf{\Theta}_2 \ \dots \ \mathbf{\Theta}_K] = \begin{bmatrix} 1 & 1 & \dots & 1 \\ e^{j\pi \sin \theta_1} & e^{j\pi \sin \theta_2} & & e^{j\pi \sin \theta_K} \\ e^{j2\pi \sin \theta_1} & e^{j2\pi \sin \theta_2} & & e^{j2\pi \sin \theta_K} \\ \dots & & \dots & \\ e^{j(U-1)\sin \theta_1} & e^{j(U-1)\sin \theta_2} & & e^{j(U-1)\sin \theta_K} \end{bmatrix} \quad (2-14)$$

$\bar{\mathbf{Y}}$ is the transmitted signal vector formulated in (2-13) with the responses of the

transmitted channels as:

$$\bar{\mathbf{Y}}(t) = \begin{bmatrix} h_1(t) \otimes y_1(t) \\ h_2(t) \otimes y_2(t) \\ \dots \\ h_K(t) \otimes y_K(t) \end{bmatrix} \quad (2-15)$$

where $h_1(t), h_2(t), \dots, h_K(t)$ are the channels time responses for each user. With (2-8), (2-15) can be further modeled in the time-variant linear filter as:

$$\bar{\mathbf{Y}}(t) = \begin{bmatrix} \sum_{l=0}^{L-1} \alpha_{1,l}(t) y_1(t - \tau_{1,l}(t)) e^{j\varphi_{1,l}(t)} \\ \sum_{l=0}^{L-1} \alpha_{2,l}(t) y_2(t - \tau_{2,l}(t)) e^{j\varphi_{2,l}(t)} \\ \dots \\ \sum_{l=0}^{L-1} \alpha_{K,l}(t) y_K(t - \tau_{K,l}(t)) e^{j\varphi_{K,l}(t)} \end{bmatrix} \quad (2-16)$$

$\mathbf{N}_U(t) = [n_1(t), \dots, n_U(t)]^T$ is the AWGN from the U -element antenna array. $n_u(t)$ is assumed with zero-mean and variance σ_0^2 . So, the received signal on the u -th antenna is

$$X_u(t) = n_u(t) + \sum_{k=1}^K \sum_{l=0}^{L-1} \alpha_{k,l}(t) y_k(t - \tau_{k,l}(t)) e^{j\theta_{k,l}(t) + (u-1) \sin \theta_k} \quad (2-17)$$

Thus, the i -th path matched output of the desired user (the first user is assumed as the desired user) from the u -th antenna element is as:

$$\hat{X}_{u,1}^i(t) = \hat{n}_{u,1}^i(t) + \sum_{k=1}^K \hat{y}_{k,1}^i(t) e^{j(u-1) \sin \theta_k} \quad (2-18)$$

By sampled at nT_c (n for short in the following), the output vector of the match filter can be written as:

$$\hat{\mathbf{X}}_{U,1}^i(n) = \hat{\mathbf{Y}}_1^i(n) + \hat{\mathbf{N}}_{U,1}^i(n) \quad (2-19)$$

where $\hat{\mathbf{Y}}_1^i(n) = [\hat{y}_{1,1}^i(n), \hat{y}_{2,1}^i(n) \dots \hat{y}_{K,1}^i(n)]^T$ and $\hat{\mathbf{N}}_{U,1}^i(n) = [\hat{n}_{1,1}^i(n), \hat{n}_{2,1}^i(n) \dots \hat{n}_{U,1}^i(n)]^T$.

The outputs of the detected i -th path from the first user's data and pilot symbols are denoted as $\tilde{\mathbf{X}}_1^{iD}(mN)$ ($\tilde{\mathbf{X}}_1^{iD}(m)$ for short) and $\tilde{\mathbf{X}}_1^{iP}(m)$.

2.2 Analytical Models of DS-UWB

DS-UWB uses a combination of a single-carrier spread-spectrum design and wide coherent bandwidth. Unlike conventional wireless systems, which use narrowband modulated carrier waves to transmit information, DS-UWB transmits data by pulses of energy generated at very high rates: in excess of 1 billion pulses per second, providing support for data rates of 28M, 55M, 110M, 220M, 500M, 660M, and 1320Mbps. A fixed UWB chip rate in conjunction with variable-length spreading code words enables this scalable support.

DS-UWB provides four key advantages over legacy wireless technologies: quality of service, high data rates that scale to 1G bps or more, lower cost, and longer battery life. These attributes mean DS-UWB is well suited to be the physical layer for WPAN's.

A DS-UWB system is identical to a normal DS-SS system except that a different pulse than the raised cosine is used (Mielczarek, Wessman, & Svensson, 2003).

2.2.1 Transmitter Structure

The transmitter of the DS-UWB is shown in the paper (Li & Rusch, 2002). The equiprobable binary bit stream $s_k(i) \in [1, -1]$ is BPSK modulation by a spread waveform. The spread waveform can be obtained for the shaping pulse by

$$w_k(t) = \sum_i c_k(i) p(t - iT_c) \quad (2-20)$$

where the T_c is the period of the monocycle and $c_k(i)$ is the spread PN code with period N_c . The UWB pulse $p(t)$ is assumed to include the differential effects in the transmitter and the receiver antenna systems. A typical pulse employed in the thesis (Win & Scholtz, 2000; Ramirez-Mireles, 2001) is the second derivative of a Gaussian pulse given by

$$p(t) = \left[1 - 4\pi \left(\frac{t}{\tau_m} \right) \right] \exp \left[-2\pi \left(\frac{t}{\tau_m} \right)^2 \right] \quad (2-21)$$

where τ_m controls the width of the pulse. Thus, the modulated transmitter signal can be expressed as (where $T_s = N_c T_c$),

$$s_k(t) = \sum_{m=-\infty}^{\infty} b_k(m) w_k(t - mT_s) \quad (2-22)$$

2.2.2 IEEE UWB Channel Model

The IEEE UWB channel model is based on the Saleh-Valenzuela model where multipath components arrive in clusters (Intel, 2002; Molisch, Foerster, & Pendergrass,

2003). The differences are that the amplitude has a lognormal distribution and the phase is randomly set to $[1, -1]$. This multipath channel can be defined as:

$$h(t) = X \sum_{l=0}^L \sum_{k=0}^K \alpha_{k,l} \delta(t - T_l - \tau_{k,l}) \quad (2-23)$$

where the multipath gain, including amplitude and phase, is defined by $\alpha_{k,l}$ for cluster l and ray k . The l -th cluster arrives at T_l and its ray k arrives at $\tau_{k,l}$ which is relative to the first path in cluster l i.e. $\tau_{0,l} = 0$. The interarrival time between two clusters $T_{l+1} - T_l$ or two rays within one cluster $\tau_{k+1,l} - \tau_{k,l}$ is exponentially distributed. Clearly, the interarrival time between any two rays is not an integer multiple of the pulse duration. Log-normal shadowing is modeled with $X = 10^{n/20}$ where n has a normal distribution with the mean equal to 0 dB and the standard deviation equal to 3 dB.

There are four different models, CM1, CM2, CM3, and CM4, for different channel characteristics that are presented in Table2-1. NP is the number of paths within 10 dB of the strongest path. NP 85% gives the number of paths containing 85 percent of the energy. Note that the expected energy of a generated channel impulse response is not normalized to 1 but approximately 1.2695 due to the shadowing term X .

The above models assume that channel impulse responses are constant during transmission of one packet if it is shorter than $200 \mu s$. Moreover, channel realizations are assumed to be independent between packets.

2.2.3 Antenna Array Model

The uniformly spaced linear planar array is considered for the UWB systems (Hussain, 2002). It is similar to the antenna array in Figure 2-2. But, the effects and responds of the antenna array for the UWB signals are different from those for the W-CDMA signals because in UWB there is no carrier any more so that there are no phase changes defined in (2-9) also because the period of the UWB monocycle is very short, the responds of the antenna array are reflected by the temporal shifts of the received signals. So, the responds of the u -th antenna of the k -th user can be expressed as:

$$h_{u,k} = h_k(t) \otimes \delta(t - \tau_{u,k}(\theta_k)) \quad (2-24)$$

where θ_k is the incident angle of the signal and

$$\tau_u(\theta_k) = (u - 1) \frac{d}{c} \sin \theta_k, \quad u = 1, 2, \dots, U \quad (2-25)$$

It is the relative temporal-shifts to received signal from the first antenna because it is assumed $\tau_1(\theta_k)$ equal 0. So the received signal at the u -th antenna of the k -th users is

$$y_{1,k}(t) = h_{u,k}(t) \otimes s_k(t) \text{ and } y_{u,k}(t) = y_{1,k}(t - \tau_{u,k}(\theta_k)) \quad (2-26)$$

and the received signal from the u -th antenna from all the K users is

$$y_u(t) = \sum_{k=1}^K h_{u,k}(t) \otimes s_k(t) + n_u(t) \quad (2-27)$$

$n_u(t)$ is assumed with zero-mean and variance σ_0^2 .

Figure 2-1 Up-link Baseband Equivalent Transmission Model of the k -th User

Figure 2-2 Antenna Array Model

Target Channel Characteristic	CM1	CM1	CM1	CM1
Distance (m)	0-4	0-4	4-10	
(Non-)Line of sight	LOS	NLOS	NLOS	NLOS
Mean excess delay (ns)	5.05	10.38	14.18	
RMS delay spread (ns)	5.28	8.03	14.28	25
NP _{10dB}			35	
NP (85%)	24	36.1	61.54	

Table 2-1. The IEEE UWB Channel Characteristic for 4 Different Scenarios

CHAPTER 3

MODIFIED LATTICE AND RLS ALGORITHMS FOR ADAPTIVE ANTENNA ARRAY IN W-CDMA SYSTEMS

3.1 Introduction

In the Code-Division Multiple-Access (CDMA) system, multiple users use different code sequences to share the same frequency band. But due to the imperfect orthogonal among the different code sequences and the fading and multi-path distortion from the wireless environments, multiple-access interference is a major limitation to the channel capacity. Adaptive antenna array is a technique that can be used to constructively enhance the desired user and suppress the interferences so that the signal to interference plus noise ratio can be increased. While RAKE combining is a technique that effectively collects the multi-path energy so that fading effect can be reduced and the signal power can be enhanced. Thus, adaptive antenna array and RAKE combining can effectively increase channel capacity; lower the bit-error-rate when working together.

Adaptive antenna array technique makes use of adaptive algorithms to find and suppress the interfering signals in space processing. They let the nulls of the antenna beam steer to the interfering signals while the main beam to the desired signal. In fact, the basic processing in adaptive antenna array technique is the adaptation of the weighting vector for the antenna array. The weighting vector is adaptively adjusted, or adapted, to

maximize the quality of the desired signal (Liberti & Rappaport, 1999). Since the mobile channel is time varying, the weighting vector must be updated periodically. So far, LMS, RLS, QRD-RLS and some other algorithms (Sheikh, Gesbert, Gore, & Paulraj 1999; Tanaka, Harada, Sawahashi, & Adachi, 2000; Liberti & Rappaport, 1999; Razavilar, Rashid-Farrokhi, & Liu, 1999; Banister & Zerdler, 2002) have been widely studied for the adaptive antenna array technique.

Lattice algorithm has been studied for different applications (Wang & Prahatheesan, 2000; Chew, Soni, Zeidler, & Ku, 1994) such as linear predictions. Lattice algorithms together with Least-Square optimal criteria show good properties as efficient computation, modular structure, fast convergence, and linear increase of computation with the number of stages (Haykin, 2001). Also, adaptive filters based on lattice algorithms have the order-recursive structures that are easy to be realized by VLSI. But, the use of order-recursive adaptive filters is limited to temporal signal processing applications that permit the exploitation of the time-shifting property of the input data (Sayed & Kailath, 1994). So far, lattice algorithms have seldom applied for the adaptive antenna arrays.

There are three goals in this chapter. The first goal is to innovate a modified LSL algorithm that can be applied for the vector inputs from the antenna arrays. In the conventional LSL algorithm, the derivations are restricted in the literature to the case of shift-structured data (Sayed & Kailath, 1994; Merched & Sayed, 2001). As a result, the similarities between the vector data and the temporal shift structured data are analyzed at first. Then, the vector rotations are applied to introduce vector data into the LSL

algorithm as the input. Also, errors residuals and coefficients in the modified LSL algorithm are modified into the vector formats instead of the scale formats in the traditional LSL algorithm. Then, the lattice structure in the modified algorithm is constructed by following the Recursive-LSL algorithm based on *priori Residuals* with Error-Feedback. LSL algorithms have two main types, one is the QRD-LSL and the other is the Recursive-LSL. Recursive-LSL also can be reviewed in the format of the QRD-LSL, but the Recursive-LSL corresponds to the explicit recursions of the information form and compact order- and time-update (Merched & Sayed, 2001). Furthermore, the Recursive-LSL algorithms also have two main types, one is based on *posteriori Residuals* and the other is based on *priori Residuals* with Error-Feedback. It is shown that the later has better numerical accuracy and the numerical stability performances (Ling & Proakis, 1984; Honig & Messerschmitt, 1981). Therefore, the Recursive-LSL based on *priori Residuals* with Error-Feedback is considered for the adaptive antenna array.

The second goal in this chapter is to show the uplink 2-D Rake receiver for the W-CDMA system. In the CDMA system, space domain processing, such as the adaptive antenna arrays processing, is usually combined with path diversity obtainable with a RAKE receiver. Thus, for the data reception, it can make usage of the beamformer-RAKE receivers, *i.e.* 2-D RAKE receivers to exploit both the diversity of RAKE combining and the diversity from beamforming (Khalaj, Paulraj, & Kailath, 1994; Ma, Hsu, & Chiuch, 2002; Li & Liu, 2003; Choi, 1999). In this chapter, the receiver combines adaptive antenna array adopted the modified LSL algorithm together with the RAKE reception. Simulations in the form of average bit error probability show that the receiver

has very good interference mitigation performances and very good reception performances.

The third goal of this chapter is to completely present and investigate the comparisons between the lattice algorithm and the other algorithms in the usage of adaptive antenna array, such as RLS. From these simulations and comparisons, it is shown that the lattice algorithm has some comparable performances with RLS algorithm, while its computational complexities are low and its structure is efficient.

In Section 3.2, adaptive antenna arrays and signal processing algorithms are briefly introduced. The overall introduction and innovation of the modified LSL algorithm for the usage to adaptive antenna arrays is proposed and described Section 3.3. The 2-D receiver with both adaptive antenna array and RAKE combining is presented in Section 3.4. Simulations for the modified LSL algorithm and the 2-D RAKE receiver are shown in Section 3.5. And then the comparisons between modified LSL and RLS algorithms in the usages of adaptive antenna array are presented and discussed in Section 3.6.

3.2 Adaptive Antenna Arrays and Signal Processing Algorithms

In antenna array system, it is constructed by fixed, switched, or adaptive antenna arrays. Fixed and switched antenna arrays can generate fixed or switched beams to achieve suppressions and get antenna gains, but they can have problems with locking onto the wrong beam due to interference so that they provide limited interference suppressions and they cannot suppress interference if it is in the same beam of the desired

user. However, the adaptive antenna arrays adopt a very different approach. Adaptive antenna arrays use sophisticated signal-processing algorithms to continuously distinguish desired user from interfering users and thus adaptively estimate their DOA's. So, adaptive antenna arrays dynamically alter the weight factor of the antenna arrays and then update their beamforming patterns based on changes of both the desired user and the interfering users locations.

There are several methods in practical alterations of weight factor. The method with the aid of pilot reference signal (the temporal-reference-based method in Chapter 1) is considered. As discussed in Chapter 2, despread pilot signals are adopted for alterations of weight factor. Therefore, the weighted output from the antenna array is subtracted from the reference signal (locally re-constructed pilot signal) to generate an error signal. Then, this error signal is used to control the update of the weight factor of the antenna array.

Alterations of the weight factor are based on selected optimization criteria, such as MMSE, LS error, or the maximum SINR. The LS optimization criterion is considered because unlike the Max SNIR method, knowledge of the noise and DOA of the desired signal are not required in alterations based on LS criterion. Defining that

$\mathbf{w}_1^i(m) = [w_{1,1}^i(m), w_{1,1}^i(m), \dots, w_{U,1}^i(m)]^T$ is the vector of the weight factor with regard to the desired user (the first user is assumed as the desired user), the cost function under the LS criterion is defined as:

$$J(\mathbf{w}_1^i) = \sum_m \left| \mathbf{w}_1^{iH} \tilde{\mathbf{X}}_{U,1}^i(m) - \tilde{X}_1^{iR}(m) \right|^2 \quad (3-1)$$

where $\tilde{X}_1^{iR}(m)$ is the locally generated reference signal (Haykin, 2001). Generally, it is difficult to solve such a cost function directly because it has some the $U \times U$ matrix operations that are not easy to perform when U is large. And because the wireless environment is time-variable, it is also required that the solution of the weight factor should be updated periodically. Therefore, some adaptive signal processing algorithms are applied to update the weight factor of the antenna arrays. These algorithms either use the estimates of statistics of the most recently input signals to calculate the new solution periodically or by adding value to the current weight factor to generate the new weight factor. They are so-called the block mode and iterative mode of the adaptive algorithm. The most applied algorithms are LMS, VSS LMS, RLS algorithms, and so on. However, generally speaking, the convergence of LMS or VSS LMS is slow and sensitive to the characteristics of input signals, while the computational complexity of RLS is large. Thus, it is necessary to find or design some new algorithm with good performances.

3.3 Adaptive Antenna Array Processing Based on LSL Algorithm

Adaptive algorithms have been widely studied for adaptive antenna array technique. Here, the lattice algorithm is considered. However, in the conventional lattice algorithm, the input is limited only to the temporal-shifted data, whereas in the case of the antenna array, the input is a vector of time-varied data sequences from different antenna. Hence, some modifications are made to the conventional lattice algorithm so that the modified LSL algorithm is applicable for the adaptive antenna array technique.

In the conventional lattice algorithm (Haykin, 2001) two major operations, the

forward prediction and the backward estimation, run simultaneously. The forward prediction error at the u -th stage ($1 \leq u \leq U$) is defined as:

$$\begin{aligned}\tilde{e}_{f,u}(m) &= \hat{X}(m) + \sum_{k=2}^u \tilde{\mathbf{f}}_u^{i*}(m) \cdot \tilde{x}(m-k) \\ &= \tilde{\mathbf{f}}_u^H(m) \cdot \hat{\mathbf{X}}(m)\end{aligned}\quad (3-2)$$

where $\hat{\mathbf{X}}(m) = [\hat{X}(m), \dots, \hat{X}(m-u+1)]^T$ is the temporal data. $\tilde{\mathbf{f}}_u(m) = [1 \dots \tilde{f}_u^u(m)]^T$ is the complex vector of the forward prediction weights. This vector is optimized based on the LS criterion. The backward estimation error residual is defined as:

$$\begin{aligned}\tilde{e}_{b,u}(m) &= \hat{X}(m-u) + \sum_{i=1}^{u-1} \tilde{\mathbf{b}}_u^{i*}(m) \cdot \hat{X}(m-i) \\ &= \tilde{\mathbf{b}}_u^H(m) \cdot \hat{\mathbf{X}}(m)\end{aligned}\quad (3-3)$$

where $\tilde{\mathbf{b}}_u(m) = [\tilde{b}_u^1(m), \tilde{b}_u^2(m), \dots, 1]^T$ is the complex vector of the backward estimation weights. $\tilde{e}_{f,u}(m)$ and $\tilde{e}_{b,u}(m)$ are updated order-by-order as:

$$\tilde{e}_{f,u+1}(m) = \tilde{e}_{f,u}(m) + \tilde{\kappa}_{f,u+1}(m-1) \cdot \tilde{e}_{b,u}(m-1) \quad (3-4)$$

$$\tilde{e}_{b,u+1}(m) = \tilde{e}_{b,u}(m-1) + \tilde{\kappa}_{b,u+1}(m-1) \cdot \tilde{e}_{f,u}(m) \quad (3-5)$$

where $\tilde{\kappa}_{f,u+1}(m-1)$ and $\tilde{\kappa}_{b,u+1}(m-1)$ are the PARCOR's or the "reflection coefficients".

The lattice structure of the order-update is showing in Figure 3-1.

However in the usage of lattice algorithm for the vector data from antenna array, the forward prediction and backward estimation error residuals, $e_{f,u}(m)$ and $e_{b,u}(m)$, are defined as:

$$\mathbf{f}_u^H(m) \cdot \hat{\mathbf{X}}_{U,1}^i(m) = e_{f,u}(m) \quad (3-6)$$

$$\mathbf{b}_u^H(m) \cdot \tilde{\mathbf{X}}_{U,1}^i(m) = e_{b,u}(m) \quad (3-7)$$

$\mathbf{f}_u(m)$ and $\mathbf{b}_u(m)$ are the forward prediction and backward estimation weight vectors.

$\mathbf{f}_u(m)$ and $\mathbf{b}_u(m)$ are in the form of $U \times 1$ vector with the last $U - u$ elements equal zero as

$$\mathbf{f}_u(m) = \begin{bmatrix} \mathbf{f}_u^{1,u}(m) & \mathbf{0}_{U-u} \end{bmatrix}^T \text{ and } \mathbf{b}_u(m) = \begin{bmatrix} \mathbf{b}_u^{1,u}(m) & \mathbf{0}_{U-u} \end{bmatrix}^T. \text{ Elements in } \tilde{\mathbf{X}}_U^i(m) \text{ can}$$

be rewritten as $X_1(m) = \tilde{\mathbf{X}}_U^1(m)$, $X_2(m) = \tilde{\mathbf{X}}_U^2(m) = [\bar{\mathbf{J}} \cdot \tilde{\mathbf{X}}_U^i(m)]^1 \dots X_u(m) = \tilde{\mathbf{X}}_U^u(m) =$

$[\bar{\mathbf{J}}^{u-1} \tilde{\mathbf{X}}_U^i(m)]^1$ where $\bar{\mathbf{J}}^{u-1} = \underbrace{\bar{\mathbf{J}} \cdot \bar{\mathbf{J}} \dots \bar{\mathbf{J}}}_{u-1}$ and $\bar{\mathbf{J}}$ is defined as:

$$\begin{aligned} \bar{\mathbf{J}} &= \begin{bmatrix} \mathbf{J}_U & \mathbf{J}_1 & \mathbf{J}_2 & \dots & \mathbf{J}_{U-1} \end{bmatrix} \\ &= \begin{bmatrix} 0 & 1 & 0 & \dots & 0 \\ 0 & 0 & 1 & \dots & \vdots \\ \vdots & \vdots & 0 & \dots & \vdots \\ \vdots & \vdots & \dots & \dots & 1 \\ 1 & 0 & 0 & \dots & 0 \end{bmatrix}_{U \times U} \end{aligned} \quad (3-8)$$

$\mathbf{J}_1 = \begin{bmatrix} 1 & 0 & \dots & 0 \end{bmatrix}^T$, $\mathbf{J}_2 = \begin{bmatrix} 0 & 1 & \dots & 0 \end{bmatrix}^T \dots \mathbf{J}_U = \begin{bmatrix} 0 & 0 & \dots & 1 \end{bmatrix}^T$. The properties

of $\bar{\mathbf{J}}$ are $\bar{\mathbf{J}} \cdot \bar{\mathbf{J}}^T = \bar{\mathbf{J}}^T \cdot \bar{\mathbf{J}} = \mathbf{I}$ ($\mathbf{I}$ is the identity matrix) and

$$\bar{\mathbf{J}} \cdot \mathbf{X}_U(m) = \begin{bmatrix} \mathbf{X}_U^1(m) \\ \mathbf{X}_U^2(m) \\ \vdots \\ \mathbf{X}_U^{U-1}(m) \\ \mathbf{X}_U^U(m) \end{bmatrix} \begin{array}{c} \uparrow \\ \uparrow \\ \uparrow \\ \uparrow \end{array} \Leftrightarrow \begin{bmatrix} \mathbf{X}_U^2(m) \\ \mathbf{X}_U^3(m) \\ \vdots \\ \mathbf{X}_U^U(m) \\ \mathbf{X}_U^1(m) \end{bmatrix}$$

which is the forward cyclic shift of the vector.

$$\bar{\mathbf{J}}^H \cdot \mathbf{X}_U(m) = \begin{bmatrix} \mathbf{X}_U^1(m) \\ \mathbf{X}_U^2(m) \\ \vdots \\ \mathbf{X}_U^{U-1}(m) \\ \mathbf{X}_U^U(m) \end{bmatrix} \begin{array}{c} \leftarrow \\ \downarrow \\ \downarrow \\ \downarrow \\ \leftarrow \end{array} \Leftrightarrow \begin{bmatrix} \mathbf{X}_U^U(m) \\ \mathbf{X}_U^1(m) \\ \vdots \\ \mathbf{X}_U^{U-2}(m) \\ \mathbf{X}_U^{U-1}(m) \end{bmatrix}$$

which is the backward cyclic shift of the vector. Consequently the backward estimation error in (3-6) can be rewritten as $\mathbf{e}_{b,u}^1(m)$ and define $\mathbf{e}_{b,u}^2(m) \dots \mathbf{e}_{b,u}^u(m)$ as:

$$\mathbf{e}_{b,u}^1(m) = \mathbf{b}_u^H(m) \cdot \bar{\mathbf{J}}^0 \cdot \mathbf{X}_U(m) = \sum_{i=0}^U \mathbf{b}_u^{i*}(m) \cdot [\bar{\mathbf{J}}^i \cdot \mathbf{X}_U(m)]^1 \quad (3-9)$$

$$\mathbf{e}_{b,u}^2(m) = \mathbf{b}_u^H(m) \cdot \bar{\mathbf{J}}^1 \cdot \mathbf{X}_U(m) = \sum_{i=0}^U \mathbf{b}_u^{i*}(m) \cdot [\bar{\mathbf{J}}^i \cdot \mathbf{X}_U(m)]^2 \quad (3-10)$$

$\vdots$

$$\mathbf{e}_{b,u}^u(m) = \mathbf{b}_u^H(m) \cdot \bar{\mathbf{J}}^{u-1} \cdot \mathbf{X}_U(m) = \sum_{i=0}^U \mathbf{b}_u^{i*}(m) \cdot [\bar{\mathbf{J}}^i \cdot \mathbf{X}_U(m)]^u \quad (3-11)$$

Then the vector presentation of (3-9) ~ (3-11) can be shown compactly as:

$$\begin{bmatrix} \mathbf{e}_{b,u}^1(m) \\ \mathbf{e}_{b,u}^2(m) \\ \vdots \\ \mathbf{e}_{b,u}^M(m) \end{bmatrix} = \begin{bmatrix} \mathbf{b}_u^H(m) \cdot \bar{\mathbf{J}}^0 \cdot \mathbf{X}_U(m) \\ \mathbf{b}_u^H(m) \cdot \bar{\mathbf{J}}^1 \cdot \mathbf{X}_U(m) \\ \vdots \\ \mathbf{b}_u^H(m) \cdot \bar{\mathbf{J}}^{U-1} \cdot \mathbf{X}_U(m) \end{bmatrix} = \sum_{i=0}^U \mathbf{b}_u^{i*}(m) \cdot \bar{\mathbf{J}}^i \cdot \mathbf{X}_U(m) \quad (3-12)$$

where the backward estimation error $\mathbf{e}_{b,u}(m)$ is defined as $\left[\mathbf{e}_{b,u}^1(m) \mathbf{e}_{b,u}^2(m) \dots \mathbf{e}_{b,u}^U(m) \right]^T$.

Similarly, the forward prediction error as $\mathbf{e}_{f,u}^1(m)$ is rewritten and modified forward prediction error elements are defined in the same way in (3-9) ~ (3-11). As a result,

$$\mathbf{e}_{f,u}^1(m) = \mathbf{f}_u^H(m) \cdot \bar{\mathbf{J}}^0 \cdot \mathbf{X}_U(m) = \sum_{i=0}^U \mathbf{f}_u^{i*}(m) \cdot \left[\bar{\mathbf{J}}^i \cdot \mathbf{X}_U(m) \right]^1 \quad (3-13)$$

$$\mathbf{e}_{f,u}^2(m) = \mathbf{f}_u^H(m) \cdot \bar{\mathbf{J}}^1 \cdot \mathbf{X}_U(m) = \sum_{i=0}^U \mathbf{f}_u^{i*}(m) \cdot \left[\bar{\mathbf{J}}^i \cdot \mathbf{X}_U(m) \right]^2 \quad (3-14)$$

⋮

$$\mathbf{e}_{f,u}^u(m) = \mathbf{f}_u^H(m) \cdot \bar{\mathbf{J}}^{u-1} \cdot \mathbf{X}_U(m) = \sum_{i=0}^U \mathbf{f}_u^{i*}(m) \cdot \left[\bar{\mathbf{J}}^i \cdot \mathbf{X}_U(m) \right]^u \quad (3-15)$$

The forward prediction error vector $\mathbf{e}_{f,u}(m)$ is defined as $\left[\mathbf{e}_{f,u}^1(m), \dots, \mathbf{e}_{f,u}^U(m) \right]^T$ and

$$\mathbf{e}_{f,u}(m) = \begin{bmatrix} \mathbf{f}_u^H(m) \cdot \bar{\mathbf{J}}^0 \cdot \mathbf{X}_U(m) \\ \mathbf{f}_u^H(m) \cdot \bar{\mathbf{J}}^1 \cdot \mathbf{X}_U(m) \\ \vdots \\ \mathbf{f}_u^H(m) \cdot \bar{\mathbf{J}}^{U-1} \cdot \mathbf{X}_U(m) \end{bmatrix} = \sum_{i=0}^U \mathbf{f}_u^{i*}(m) \cdot \bar{\mathbf{J}}^i \cdot \mathbf{X}_U(m) \quad (3-16)$$

From the vector structure of $\mathbf{e}_{f,u}(m)$ and $\mathbf{e}_{b,u}(m)$, it is easily to construct the order-update of them as:

$$\begin{aligned} \mathbf{e}_{f,u+1}(m) &= \sum_{i=0}^{u+1} \mathbf{f}_u^{i*}(m) \cdot \bar{\mathbf{J}}^i \cdot \mathbf{X}_U(m) \\ &= a_1 \cdot \sum_{i=0}^u \mathbf{f}_u^{i*}(m) \cdot \bar{\mathbf{J}}^i \cdot \mathbf{X}_U(m) + a_2 \cdot \sum_{i=1}^{u+1} \mathbf{b}_{u-1}^{i*}(m) \cdot \bar{\mathbf{J}}^i \cdot \mathbf{X}_U(m) \\ &= a_1 \cdot \sum_{i=0}^u \mathbf{f}_u^{i*}(m) \cdot \bar{\mathbf{J}}^i \cdot \mathbf{X}_U(m) + a_2 \cdot \bar{\mathbf{J}} \cdot \sum_{i=0}^u \mathbf{b}_u^{i*}(m) \cdot \bar{\mathbf{J}}^i \cdot \mathbf{X}_U(m) \\ &= a_1 \cdot \mathbf{e}_{f,u}(m) + a_2 \cdot \bar{\mathbf{J}} \cdot \mathbf{e}_{b,u}(m) \end{aligned} \quad (3-17)$$

and

$$\begin{aligned}
\mathbf{e}_{b,u+1}(m) &= \sum_{i=0}^{u+1} \mathbf{b}_u^{i*}(m) \cdot \bar{\mathbf{J}}^i \cdot \mathbf{X}_U(m) \\
&= a_3 \cdot \sum_{i=1}^{u+1} \mathbf{b}_{u-1}^{i*}(m) \cdot \bar{\mathbf{J}}^i \cdot \mathbf{X}_U(m) + a_4 \cdot \sum_{i=0}^u \mathbf{f}_u^{i*}(m) \cdot \bar{\mathbf{J}}^i \cdot \mathbf{X}_U(m) \\
&= a_3 \cdot \bar{\mathbf{J}} \cdot \sum_{i=0}^u \mathbf{b}_u^{i*}(m) \cdot \bar{\mathbf{J}}^i \cdot \mathbf{X}_U(m) + a_4 \cdot \sum_{i=0}^u \mathbf{f}_u^{i*}(m) \cdot \bar{\mathbf{J}}^i \cdot \mathbf{X}_U(m) \\
&= a_3 \cdot \bar{\mathbf{J}} \cdot \mathbf{e}_{b,u}(m) + a_4 \cdot \mathbf{e}_{f,u+1}(m)
\end{aligned} \tag{3-18}$$

Subsequently by setting $a_1 = a_3 = 1$, a_2 and a_4 are optimized also as the PARCORs or the “reflection coefficients”. The *priori* error residuals are considered for the optimizations of PARCOR as $a_2 = \kappa_{f,u}(m-1)$ and $a_4 = \kappa_{b,u}(m-1)$. Therefore, (3-17) and (3-18) are rewritten as

$$\mathbf{e}_{f,u+1}(m) = \mathbf{e}_{f,u}(m) + \kappa_{f,u}(m-1) \cdot \bar{\mathbf{J}} \cdot \mathbf{e}_{b,u}(m) \tag{3-19}$$

$$\mathbf{e}_{b,u+1}(m) = \bar{\mathbf{J}} \cdot \mathbf{e}_{b,u}(m) + \kappa_{b,u}(m-1) \cdot \mathbf{e}_{f,u}(m) \tag{3-20}$$

Substituting (3-12) and (3-16) into (3-19), it is obtained that

$$\begin{aligned}
\mathbf{e}_{f,u+1}^u(m) &= \mathbf{e}_{f,u}^u(m) + \kappa_{f,u}(m-1) \cdot \mathbf{e}_{b,u}^{u+1}(m) \\
&= \mathbf{f}_u^H(m) \cdot \bar{\mathbf{J}}^{u-1} \cdot \mathbf{X}_u(m) + \kappa_{f,u}(m-1) \cdot \mathbf{b}_u^H(m) \cdot \bar{\mathbf{J}}^u \cdot \mathbf{X}_u(m) \\
&= [\mathbf{f}_u^H(m) + \kappa_{f,u}(m-1) \cdot \bar{\mathbf{J}} \cdot \mathbf{b}_u^H(m)] \cdot \bar{\mathbf{J}}^{u-1} \cdot \mathbf{X}_u(m)
\end{aligned}$$

With (3-6), it is shown that

$$\begin{aligned}
\mathbf{f}_{u+1}^H(m) &= \mathbf{f}_u^H(m) + \kappa_{f,u}(m-1) \cdot \mathbf{b}_u^H(m) \cdot \bar{\mathbf{J}} \Leftrightarrow \\
\mathbf{f}_{u+1}(m) &= \mathbf{f}_u(m) + \kappa_{f,u}^*(m-1) \cdot \bar{\mathbf{J}}^T \cdot \mathbf{b}_u^H(m)
\end{aligned} \tag{3-21}$$

From the definition of $\bar{\mathbf{J}}$ in (3-8), (3-21) can be expanded into the form of vector as:

$$\begin{bmatrix} \mathbf{f}_{u+1}^{1,u+1}(m) \\ \mathbf{0}_{U-u-1} \end{bmatrix} = \begin{bmatrix} \mathbf{f}_u^{1,u}(m) \\ 0 \\ \mathbf{0}_{U-u-1} \end{bmatrix} + \kappa_{f,u}^*(m-1) \cdot \begin{bmatrix} 0 \\ \mathbf{b}_u^{1,u}(m) \\ \mathbf{0}_{U-u-1} \end{bmatrix} \quad (3-22)$$

Also in the same way, it is easy to obtain that

$$\begin{aligned} \mathbf{e}_{b,u+1}^u(m) &= \mathbf{e}_{b,u}^{u+1}(m) + \kappa_{b,u}(m-1) \cdot \mathbf{e}_{f,u}^u(m) \\ &= \mathbf{b}_u^H(m) \cdot \bar{\mathbf{J}}^u \cdot \mathbf{X}_u(m) + \kappa_{b,u}(m-1) \cdot \mathbf{f}_u^H(m) \cdot \bar{\mathbf{J}}^{u-1} \cdot \mathbf{X}_u(m) \\ &= [\mathbf{b}_u^H(m) \cdot \bar{\mathbf{J}} + \kappa_{b,u}(m-1) \cdot \mathbf{f}_u^H(m)] \cdot \bar{\mathbf{J}}^{u-1} \cdot \mathbf{X}_u(m) \end{aligned}$$

Similarly,

$$\begin{aligned} \mathbf{b}_{u+1}^H(m) &= \mathbf{b}_u^H(m) \cdot \bar{\mathbf{J}} + \kappa_{b,u}(m-1) \cdot \mathbf{f}_u^H(m) \Leftrightarrow \\ \mathbf{b}_{u+1}(m) &= \bar{\mathbf{J}}^T \cdot \mathbf{b}_u(m) + \kappa_{b,u}(m-1) \cdot \mathbf{f}_u(m) \end{aligned} \quad (3-23)$$

and

$$\begin{bmatrix} \mathbf{b}_{u+1}^{1,u+1}(m) \\ \mathbf{0}_{U-u-1} \end{bmatrix} = \begin{bmatrix} 0 \\ \mathbf{b}_u^{1,u}(m) \\ \mathbf{0}_{U-u-1} \end{bmatrix} + \kappa_{b,u}^*(m-1) \cdot \begin{bmatrix} \mathbf{f}_u^{1,u}(m) \\ 0 \\ \mathbf{0}_{U-u-1} \end{bmatrix} \quad (3-24)$$

It is clear that in (3-22) and (3-24) the updates of weight vectors follow the *Levinson-Durbin* recursion that is the fundamental to the lattice algorithm (Haykin, 2001).

From above analysis, it is shown that the modified LSL algorithm with the vector-structure has some similarities with the conventional LSL. They are both based on the *Levinson-Durbin* recursion and are only different in the ways to realize the recursion for different types of inputs. The vector rotations or the element cyclic shift operations in the modified LSL are correspondent to the time-shift operations. Accordingly, from (3-4) ~ (3-5) and (3-18) ~ (3-19), the correspondences among them are summarized in Table 3-1.

Therefore, the modified LSL is constructed by mapping the correspondences in Table 3-1 into the conventional lattice algorithm. The modified algorithm has a vector input and the error residuals are also in vector formats. The vector rotation ($\bar{\mathbf{J}}$) replaces the time-shift operation (z^{-1}). The joint-estimation recursive LSL algorithm based on the *priori* residuals with error feedback is applied (Sayed & Kailath, 1994). The details are listed in Table 3-2 and its structure is presented in Figure 3-2. It shows that the shift-structured input and error residuals in the conventional LSL are currently in the forms of vector-structured rotations such as $\mathbf{e}_{b,u-1}^2(m) \Leftrightarrow \tilde{\mathbf{e}}_{b,u-1}(m-1)$. On the other hand, similar to the conventional LSL, the PARCORs ($\kappa_{f,u}(m-1) \Leftrightarrow \tilde{\kappa}_{f,u}(m-1), \kappa_{b,u}(m-1) \Leftrightarrow \tilde{\kappa}_{b,u}(m-1)$) are optimized to minimize the *priori* error residuals. But compared with the conventional LSL, there are two extra steps in Table 3-2 as:

$$\underline{\Psi}_{b,u}(m) = \underline{\gamma}_u(m-1) \cdot \left| \mathbf{e}_{b,u}^1(m) \right|^2 + \lambda \cdot \underline{\Psi}_{b,u}(m-1)$$

and

$$\underline{\gamma}_u(m) = \underline{\gamma}_{u-1}(m) - \underline{\gamma}_{u-1}^2(m) \cdot \frac{\left| \mathbf{e}_{b,u-1}^1(m) \right|^2}{\underline{\Psi}_{b,u}(m)}$$

because these two steps are the optimization of joint-estimation error residuals $\zeta_u(m)$ and reflection coefficients $\kappa_u(m-1)$ are updated based on $\mathbf{e}_{b,u}^1(m)$ while correspondent operations for $\kappa_{f,u}(m-1)$ and $\kappa_{b,u}(m-1)$ are based on $\mathbf{e}_{f,u-1}^1(m)$ and $\mathbf{e}_{b,u-1}^2(m)$. Without loss of generality, the normalized weights are used to complete beamforming.

3.4 2-D Rake Receiver for the Adaptive Antenna Array Processing

For the W-CDMA systems uplink data reception, the beamformer-RAKE receivers are applied and investigated their performance. The structure of the 2-D RAKE receivers is shown in Figure 3-3. As mentioned above, the signals received by each antenna element are first passed through correlators as the demodulation. Each correlator is matched to the desired user's code with a time shift equal to the time delay of the desired signal paths. The correlator outputs are then passed through the beamformer to further enhance the desired signal. The weight w_1^i for the adaptive antenna array is optimized with the modified LSL algorithm.

Beamforming process is the weighting process. The beamformed i -th path of desired signal is $w_1^{iH} \tilde{X}_{U,1}^i(m)$ with average SNR as ξ_i . It is assumed interferences as the colored Gaussian noises according to Central-Limit Theorem (Cheun, 1997; D'Amico, Mengali, & Morelli, 2003; Wang & Kwon, 2003; Guenach & Vandendorpe, 2001; Chen, Yu, & Wang, 2001) and channels experience Rayleigh fading, thus the exact PDF of SNR of the 2-D RAKE receiver output is given as (Mun, Choi, & Park, 2001):

$$p_{\Gamma}(\gamma) = \sum_{i=1}^{L_A} \frac{\pi_i}{\xi_i} e^{-\gamma/\xi_i} \quad (3-25)$$

where the available number of branches for the RAKE receiver is L_A and

$$\pi_i = \prod_{\substack{j=1 \\ j \neq i}}^{L_A} \frac{\xi_i}{\xi_j - \xi_i} \quad (3-26)$$

Consequently the average probability of error for the 2-D RAKE receiver is

$$\bar{P}_e = \int_0^\infty P_e(\gamma) \cdot p_\Gamma(\gamma) d\gamma \quad (3-27)$$

For the binary orthogonal signals (Proakis, 2001; Abbas & Sheikh, 1998),

$$\bar{P}_e = \frac{1}{2} \sum_{i=1}^{L_A} \pi_i \cdot \left[1 - \sqrt{\frac{\xi_i}{2 + \xi_i}} \right] \quad (3-28)$$

3.5 Simulations of Different Algorithms for the Adaptive Antenna Array

In the simulation, the uplink of the W-CDMA standards is assumed. And different cases are considered for different purposes. The pilot signals from different users are considered to complete the weights updating, because the pilot information is known both to the base station and to the mobile station after synchronization and without distortion of the system performance, all '1' pilot symbols are assumed. In the same cell, all users have the same pilot signals with different timing shifts and scrambling code, thus base station can use all the received users' pilot information to complete the adaptations along with a reference signal at the base station side. The pilot signals after demodulations are considered for these operations. Matched filters are applied to complete the demodulation. In the simulation, the reference signal is reconstructed from the demodulated received pilot signals. For the wireless transmission environment, IMT M.1225 channel model is applied. Subsequently, matched filters also are used to resolve different multipath of desired user and then adaptations are completed to resolved paths respectively.

3.5.1 *Parameters and Assumptions in Simulations*

The uplink asynchronous transmission of the W-CDMA system is considered. Both interfering users and desired users are assumed with the same data rate (30 kb/s, with spreading factor fixed to 256) for the pilot signal and different data rates (480kb/s, 960 kb/s or 1920 kb/s, with spreading factors 16, 8 or 4). And all users have the same power. As mentioned above, the pilot signal's formats are known to the base station, where all pilot symbols are with unitary value. The transmission power ratio of pilot to data signals is $\beta = 0.25$ (3GPP, 2000).

Received signal at all antenna elements are first demodulated with the spreading code for the desired user's pilot signals and then the results are taken into the weights updating block as the inputs. These input signals include both the pilot signals of the desired user and the pilot signals of other users, which are considered as interferences. Each user's spreading code is formed by two types of codes, one is the channelization codes, the other is the scrambling codes. The despreading is performed first by multiplying the pilot by the channelization codes and then complex multiplying the results with the scrambling codes. The pilot signals are set to Q branch and data signals are set to I branch. Because in W-CDMA system, all the users have the same channelization code $C_{ch,256}^0$ for pilot signals, the scrambling codes decided the orthogonal performance of the spreading codes among different users. The scrambling codes have two types, the long code and the short code. Long code has a period of 38400 *chips*, equal a frame, while the short code is a 256 *chips* (one symbol length) length codes.

Long scrambling codes are considered. Then, after demodulation, the lattice algorithm uses the demodulated signal to complete the adaptations. Thus, all users have the same pilot signals, but with different scrambling codes. Other parameters of the uplink transmission are given in Table 3-3.

Signals from one user at different antenna elements are assumed experiencing the same multipath transmission and have the same incident angles. It is assumed that each user experiences a 6-path Rayleigh fading channel in the IMT2000 M.1225 channel model (ITU M.1225) in which path delays are respectively $[0 \ 1.152 \ 34.176 \ 49.536 \ 65.664 \ 76.8] \times T_c$ and corresponding relative powers of taps are $[-2.5 \ 0 \ -12.8 \ -10.0 \ -25.2 \ -16.0]$ dB. Users are considered with a velocity equal to 250 km/h so that the maximum Doppler frequency f_D is 416.6Hz. Because of the co-existence of multipaths from desired and interfering users, matched filters are used to resolve different paths of the desired user.

3.5.2 *RLS Algorithm*

The performances of the LSL algorithm with the RLS algorithm are also compared. RLS is a well-known adaptive algorithm (Haykin, 2001). The detail of this algorithm is presented in Table 3-5.

3.5.3 *Overall Performances of the LSL and RLS Algorithms*

A five users case is considered first in which four interfering users and one desired user are assumed with different data rates (30 or 60 kbps) and powers. The

antenna array is assumed with five antenna elements. For easy simulations, it is assumed that all the l -th paths of a user at each antenna element have the same incident angle. All users have the same constant pilot signals, but with different scrambling codes (Truncated-Gold sequence). The power ratio of pilot to data channels is $\beta=0.25$.

The incident angle of the desired user is 60° and the incident angles of the interferers are 45° , 15° , -30° , and -60° . The power of desired user is $\frac{1}{4}$ to the interfering users. The data rate of desired use is 30 kbps and interfering users are assumed with data rate 60 kbps. The reference signal is considered reconstructing from channel estimation with the Additive White Gaussian Noise (Wang & Kwon, 2003; Chen, Yu, & Wang, 2001). It is followed the steps of the modified LSL algorithm in Table 3-2 at first to figure out the weight vectors for each path of desired user and then perform the 2-D RAKE receptions.

The performance of the LSL algorithm is first investigated. The averaged mean-square estimation errors $e_i = \frac{1}{M} \sum_{m=1}^M \left| \tilde{X}_1^{iR}(m) - \mathbf{w}_i^H(m) \cdot \tilde{\mathbf{X}}_U^i(m) \right|^2$ of different paths (1-st, 3-rd, and 6-th) are shown in Figure 3-4 and 3-5 with different reference signal performances. SNR's of the reference signals for the 1-th, 3-rd, and 6-th paths in the former are $[15 \ 1.47 \ 0.2745]$ dB and $[-5 \ -19.08 \ -20.28]$ dB in the latter because it is assumed that channel estimations have the same noise powers for each path. Simulations in this thesis are all *Monte-Carlo*-based and results are recorded after each an iteration (per pilot bit) and averaged. From these figures, it is shown that the error curves are in very good convergent performances. With a good reference, after 150 iterations (1 slot or

0.01ms) the estimation errors have reached their convergent values. Even with a not good reference, the convergences of the estimation errors are also good. After 600 iterations (4 slots or 0.04ms), all estimation errors almost converge.

In Figure 3-4 and Figure 3-5, the estimation errors from RLS algorithm are shown in dot lines. It is observed that LSL and RLS algorithms almost have the similar convergent performances. For some paths (the curves A and D in both Figure 3-4 and Figure 3-5), the estimation errors from the modified LSL algorithm converge fast than errors from RLS algorithm.

Figure 3-6 illustrates how performances of the average probability of error for the 5-user case are varied for different reference signals. Three conditions *A*, *B*, and *C* are listed in Table 3-4. It is observed that under the reference signals with high average SNR, 2-D RAKE receivers have very good performances. Consequently, from Figure 4-6, it can conclude that reference signals are very important to both the convergence of the adaptive antenna array processing and data reception of the 2-D RAKE. Actually without the information of the channel phase parameters, $e^{j\varphi_{1,l}}$, it is not easy to estimate the DOA parameters $e^{j\phi_{u,1}}$ out from the composite phase parameters $e^{j(\varphi_{1,l} + \phi_{u,1})}$. In the simulations, the channel is assumed known to the receivers. As a result, good channel estimations are essential for the adaptive antenna array processing.

Comparisons between the modified LSL and RLS algorithms are also done on performances of the average probability of error. From the curve C and E in Figure 3-6, it is observed that the modified LSL works better than RLS algorithm when the channel

estimation and reference signal are not good.

The cases with more than 5 users are also considered and incident angles are considered as random variables as well. The simulation conditions for this case are the same as the previous 5-user case and condition 4 except that incident angles are selected randomly as -59° , 14° , 18° , -38° , -10° , 51° , 5° , -21° , 45° for all the 9 interfering users while desired user's angle is still 60° . The number of the antenna elements is also taken into considerations. The results are shown in Figure 4-7. It is observed that with more users, 5-element antenna array antenna array is not able to distinguish all interferences and suppress them since the nulls of a 5-element antenna array can only have 4 nulls so the adaptive antenna array processing is just to suppress interferences as much as possible. With more antenna elements, the performances of the average probability of error become better and better. It is shown that with an 11-element antenna array, the average probability is around 10^{-5} which is much better than 10^{-3} from a 5-element antenna array. The beamforming pattern for the 11-element antenna array is shown in Figure 3-8.

Actually from Figure 3-7, when $U=8$ and 9, the performances of the average probability of error are also very good. Although $U=8$ and 9 are less than the number of the users (10 users), some users are so close to each other that they locate around the same nulls in the antenna beam. Thus, they are suppressed all to a level. Also the LSL algorithm is adaptive processing that tries to converge to the optimal results.

The final outputs' performances from adaptive antenna array based on RLS and the modified LSL algorithms are also compared. The results are presented in Figure 3-9

with considerations of different numbers of antenna elements and different numbers of active users. It is shown that two algorithms introduce almost same outputs' performances. But, when the adaptive antenna array has small numbers of antenna elements, *e.g.* 1-7 in the figure, the outputs' final performances from the modified LSL algorithm are better than performances from RLS algorithm.

3.6 Conclusion

In this chapter, a modified lattice algorithm applied for the adaptive antenna array in WCDMA system has been proposed. This algorithm is realized by the recursive lattice algorithm based on the Least-square criteria and based on the *priori* residuals with error feedback. Together with the RAKE combining, the 2-D RAKE receiver with the modified LSL algorithm for the spatial processing is also presented. Simulation results show that the modified LSL algorithm is applicable for the adaptive antenna array processing and has faster convergent performances. Moreover, the 2-D RAKE receivers present good performances of the average probability of error. It is also concluded that reference signals and channel estimation are very important to both the convergence of the adaptive antenna array processing and data reception of the 2-D RAKE.

Because lattice algorithm has a module structure that can be seen from Figure 3-2, all stages can share one set of adaptive structure. As a result, it is useful to take advantage of the efficient modular structure for the VLSI design for the adaptive antenna array processing. Also the recursive LSL algorithm with error feedback is considered with a good numerical behavior in a limited-precision environment (Merched & Sayed,

2001; Ling & Proakis, 1984; Honig & Messerschmitt, 1981). The modified lattice algorithm with above advantages is suitable for the adaptive antenna array processing and has the almost same performances when compared to RLS algorithm. Also, the computational complexity of the modified LSL is around U^2 while the computational complexity of RLS algorithm is around $3U^2$.

Figure 3-1. Order-Update of Conventional Lattice Algorithm

Figure 3-2. Overall Structure for the Modified LSL with Joint Estimation

($[\cdot]^1$ means the operation to get only the first element out of the vector for further processing)

	Shift-Structured	Vector-Structured
Input	$x(m)$	$\mathbf{X}_U^i(m)$
Operation for Input and Error Residuals	z^{-1}	$\bar{\mathbf{J}}$
Error Residuals	$\tilde{e}_{f,u}(m-i)$	$\mathbf{e}_{f,u}^{i+1}(m)$
	$\tilde{e}_{b,u}(m-i)$	$\mathbf{e}_{b,u}^{i+1}(m)$
Prediction and Estimation Weight	$\tilde{\mathbf{f}}_u^i(m)$	$\mathbf{f}_u^i(m)$
	$\tilde{\mathbf{b}}_u^i(m)$	$\mathbf{b}_u^i(m)$
Reflection Coefficients	$\tilde{\kappa}_{f,u}(m-1)$	$\kappa_{f,u}(m-1)$
	$\tilde{\kappa}_{b,u}(m-1)$	$\kappa_{b,u}(m-1)$

Table 3-1. Correspondences in Conventional and Modified LSL

Initialization

For $u=1$ to U , repeat:

$$\Psi_{f,u}(0) = \Psi_{b,u}(0) = \Phi_u(0) = \text{a very small positive value}$$

$$\lambda = \text{a value very close to 1 (e.g. 0.9999)}$$

$$\kappa_{f,u}(m) = \kappa_{b,u}(m) = \kappa_u(m) = 0$$

$$\zeta_0(m) = \text{Reference Signal}$$

$$\mathbf{f}_1^1(m) = \mathbf{b}_1^1(m) = 1$$

$$\gamma_0(m) = \underline{\gamma}_0(m) = 1$$

For $n>0$, repeat:

$$\mathbf{e}_{f,0}(m) = \mathbf{e}_{f,u}(m) = \mathbf{e}_u = \tilde{\mathbf{X}}_U^1(m)$$

Iterations

For $u=1$ to U , repeat:

$$\mathbf{e}_{f,u}(m) = \mathbf{e}_{f,u-1}(m) + \kappa_{f,u}^*(m-1) \cdot \bar{\mathbf{J}} \cdot \mathbf{e}_{b,u-1}(m)$$

$$\mathbf{e}_{b,u}(m) = \bar{\mathbf{J}} \cdot \mathbf{e}_{b,u-1}(m) + \kappa_{b,u}^*(m-1) \cdot \mathbf{e}_{f,u-1}(m)$$

$$\zeta_u(m) = \zeta_{u-1}(m) - \kappa_u^*(m-1) \cdot \mathbf{e}_{b,u}^1(m)$$

$$\Psi_{f,u}(m) = \gamma_u(m-1) \cdot |\mathbf{e}_{f,u}^1(m)|^2 + \lambda \cdot \Psi_{f,u}(m-1)$$

$$\Psi_{b,u}(m) = \gamma_u(m-1) \cdot |\mathbf{e}_{b,u}^2(m)|^2 + \lambda \cdot \Psi_{b,u}(m-1)$$

$$\underline{\Psi}_{b,u}(m) = \underline{\gamma}_u(m-1) \cdot |\mathbf{e}_{b,u}^1(m)|^2 + \lambda \cdot \underline{\Psi}_{b,u}(m-1)$$

$$\kappa_{f,u}(m) = \kappa_{f,u}(m-1) - \gamma_{u-1}(m-1) \cdot \frac{\mathbf{e}_{b,u-1}^2(m)}{\Psi_{b,u-1}(m)} \cdot \mathbf{e}_{f,u}^{1*}(m)$$

$$\kappa_{b,u}(m) = \kappa_{b,u}(m-1) - \gamma_{u-1}(m-1) \cdot \frac{\mathbf{e}_{f,u-1}^1(m)}{\Psi_{f,u-1}(m)} \cdot \mathbf{e}_{b,u}^{1*}(m)$$

$$\kappa_u(m) = \kappa_u(m-1) + \underline{\gamma}_{u-1}(m) \cdot \frac{\mathbf{e}_{b,u}^1(m)}{\underline{\Psi}_{b,u}(m)} \cdot \zeta_u^*(m)$$

$$\gamma_u(m) = \gamma_{u-1}(m) - \gamma_{u-1}^2(m) \cdot \frac{|\mathbf{e}_{b,u-1}^2(m)|^2}{\Psi_{b,u}(m)}$$

$$\underline{\gamma}_u(m) = \underline{\gamma}_{u-1}(m) - \underline{\gamma}_{u-1}^2(m) \cdot \frac{|\mathbf{e}_{b,u-1}^1(m)|^2}{\underline{\Psi}_{b,u}(m)}$$

$$\begin{bmatrix} \mathbf{f}_{u+1}^{1,u+1}(m) \\ \mathbf{0}_{U-u-1} \end{bmatrix} = \begin{bmatrix} \mathbf{f}_u^{1,u}(m) \\ 0 \\ \mathbf{0}_{U-u-1} \end{bmatrix} + \kappa_{f,u}^*(m-1) \cdot \begin{bmatrix} 0 \\ \mathbf{b}_u^{1,u}(m) \\ \mathbf{0}_{U-u-1} \end{bmatrix}$$

$$\begin{bmatrix} \mathbf{b}_{u+1}^{1,u+1}(m) \\ \mathbf{0}_{U-u-1} \end{bmatrix} = \begin{bmatrix} 0 \\ \mathbf{b}_u^{1,u}(m) \\ \mathbf{0}_{U-u-1} \end{bmatrix} + \kappa_{b,u}^*(m-1) \cdot \begin{bmatrix} \mathbf{f}_u^{1,u}(m) \\ 0 \\ \mathbf{0}_{U-u-1} \end{bmatrix}$$

$$\mathbf{T}_U(m) = \begin{bmatrix} 1 & 0 & \dots & 0 \\ \mathbf{b}_2^{1*}(m) & 1 & \dots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ \mathbf{b}_U^{1*}(m) & \mathbf{b}_U^{2*}(m) & \dots & 1 \end{bmatrix}$$

$$\mathbf{P}_U(m) = [\kappa_1(m), \kappa_2(m), \dots, \kappa_U(m)]^T$$

$$\mathbf{w}_U^i(m) = \mathbf{T}_U^H(m) \mathbf{P}_U(m) \quad \mathbf{w}_U^i = \|\mathbf{w}_U^i(m)\|$$

Table 3-2. Modified Recursive LSL Based on a *Priori* Errors with Feedback

Figure 3-3. Structure of the 2-D RAKE Receiver

Chip rate (Bandwidth)		3.84 Mcps (5 MHz)
Carrier frequency (uplink)		1900 MHz
Data rate		30 kbps, 60 kbps
Data spreading factor		128, 64
Pilot symbol rate		15 kbps
Pilot spreading factor		256
Frame structure	Slot	0.667 ms
	Frame	10 ms
Spreading code	Channelization code	OVSF sequence (256,128 chips)
	Scrambling code	Truncated-Gold sequence
Modulation		QPSK
Channel estimation for reference signal		Pilot symbol-assisted channel estimation

Table 3-3 Uplink Transmission Parameters

		Path 1	Path 1	Path 3	Path 4	Path 5	Path 6
<i>A</i>		15.43dB	1.35dB	1.47dB	0.93dB	3.25dB	0.2745dB
<i>B</i>		5 dB	-9.15dB	-9.11dB	-9.43dB	-6.63dB	-10.14dB
<i>C</i>		-5dB	-19.25dB	-19.08dB	-18.97dB	-16.77dB	-20.28dB

Table 3-4. SNR Conditions of the Reference Signals

Initialization	Weight Update
Adaptation of the desired user i -th path	<i>For $m > 0$, repeat:</i>
Initial weight vector $(1, 1, \dots, 1)^T$	$\mathbf{k}(m) = \frac{\lambda^{-1} \mathbf{P}(m-1) \tilde{\mathbf{X}}_U^i(m)}{1 + \lambda^{-1} \tilde{\mathbf{X}}_U^{iH}(m) \mathbf{P}(m-1) \tilde{\mathbf{X}}_U^i(m)}$
$\mathbf{P}(0) = \delta^{-1} \mathbf{I}$, $\mathbf{I}$ is a U -by- U identity matrix	$e_{1,U}^{(\text{RLS})}(m) = \tilde{X}_1^i(m) - \mathbf{w}_{1,U}^{(\text{RLS})H}(m) \tilde{\mathbf{X}}_U^i(m)$
δ = small positive constant (e.g. 0.004)	$\mathbf{w}_{1,U}^{(\text{RLS})}(m) = \mathbf{w}_{1,U}^{(\text{RLS})}(m-1) + \mathbf{k}(m) e_{1,U}^{(\text{RLS})}(m)$
λ = a value very close to 1 (e.g. 0.9999)	$\mathbf{P}(m) = \lambda^{-1} \mathbf{P}(m-1) - \lambda^{-1} \mathbf{k}(m) \tilde{\mathbf{X}}_U^i(m) \mathbf{P}(m-1)$
	At time m , $\mathbf{w}_{1,U}^{(\text{RLS})}(m)$ is derived
	$\mathbf{w}_{1,U}^{(\text{RLS})}(m) = \frac{\mathbf{w}_{1,U}^{(\text{RLS})}(m)}{\sqrt{U \cdot \ \mathbf{w}_{1,U}^{(\text{RLS})}(m)\ ^2}}$

Table 3-5. Adaptive Computations of RLS

Figure 3-4. Averaged Mean-Square Estimation Errors with Good Reference Signals

Figure 3-5. Averaged Mean-Square Estimation Errors with Not Good Reference Signals

Figure 3-6. Averaged Bit Error Probability under Different Conditions

Figure 3-7. Averaged Bit Error Probability with Different Number of Antenna

Elements, U

Figure 3-8. Beamforming Pattern from the 11-Element Antenna Array

Figure 3-9. Performances with Different Number of Antenna Elements U and Active Users K

CHAPTER 4

SYMBOL RATE PROCESSING OF ADAPTIVE ANTENNA ARRAY FOR W-CDMA SYSTEM WITH TRADITIONAL CHANNEL ESTIMATIONS

4.1 Introduction

In this chapter, the symbol rate processing of adaptive antenna array for the W-CDMA is introduced. It is different from the methods in Chapter 3 because the digital signal processing is never performed on the received signal symbols after demodulation. It is based directly on the signals just after matched filter with the chip level. This processing will take advantage of the orthogonal characteristics of the CDMA scheme. Analysis and simulation all show that this processing has a very good performance and fast convergence. Therefore, this processing can provide more samples for the adaptive antenna array to estimate the optimum antenna weight vector and it is also very fast since the decision is on the level of symbol rate.

The rest of the chapter is organized as follows. The Section 4.2 expands the descriptions for the received signals from the antenna array. Section 4.3 analyses the proposed symbol rate adaptive antenna array processing and its performance on uplink data receptions with the aid of an adaptive antenna array. Section 4.4 expresses the

scheme of the symbol-rate processing. Numerical performance results under the Nakagami channel model are presented in Section 4.5. Section 4.6 shows simulation results of a 5-element adaptive antenna array under the IMT2000 M.1225 channel model.

4.2 Details of the Received Signals

As shown in the Figure 4-1, in the first step of signal processing in the receivers shown in Chapter 3, the received signals at each antenna element are demodulated first and then filtered by the pulse matched filters. Consequently, the i -th path matched output of the desired user from the u -th antenna element sampled at nT_c (n for short in the following) is shown in (2-19). It is assumed that the 1-st user is the desired user and the paths of 1st user are deterministic. Both $\hat{n}_{u,1}^i(t)$ and $\hat{y}_{k,1}^i(t)$ can be expanded and presented equivalently into real and imaginary part respectively as:

$$\begin{aligned}\hat{y}_{k,1}^i(t) &= \hat{y}_{k,1}^{iI}(t) + j \cdot \hat{y}_{k,1}^{iQ}(t) \\ &= \int_{nT_c + \tau_{1,i}}^{(n+1)T_c + \tau_{1,i}} \left\{ \sum_{l=0}^{L-1} \alpha_{k,l}(t) y_k(t - \tau_{k,l}(t)) e^{j\theta_{k,l}(t)} \right\} \cos(\omega_0 t) h^*(t - nT_c - \tau_{1,i}) dt \\ &\quad - j \cdot \int_{nT_c + \tau_{1,i}}^{(n+1)T_c + \tau_{1,i}} \sum_{l=0}^{L-1} \left\{ \alpha_{k,l}(t) y_k(t - \tau_{k,l}(t)) e^{j\theta_{k,l}(t)} \right\} \sin(\omega_0 t) h^*(t - nT_c - \tau_{1,i}) dt\end{aligned}\quad (4-1)$$

Ignoring the high frequency components in (4-1),

$$\begin{aligned}\hat{y}_{k,1}^{iI}(n) &= \sum_{l=0}^{L-1} \rho_{k,l}^I(n) [D_k(m) S_{k,I}(n-l) C_{k,D}(n-l) - \sqrt{\beta_k} S_{k,Q}(n-l) C_{k,P}(n-l)] \\ &\quad - \sum_{l=0}^{L-1} \rho_{k,l}^Q(n) [D_k(m) S_{k,Q}(n-l) C_{k,D}(n-l) + \sqrt{\beta_k} S_{k,I}(n-l) C_{k,P}(n-l)]\end{aligned}\quad (4-2)$$

$$\begin{aligned}\hat{y}_{k,l}^{iQ}(n) = & \sum_{l=0}^{L-1} \rho_{k,l}^Q(n) [D_k(m)S_{k,I}(n-l)C_{k,D}(n-l) - \sqrt{\beta_k}S_{k,Q}(n-l)C_{k,P}(n-l)] \\ & + \sum_{l=0}^{L-1} \rho_{k,l}^I(n) [D_k(m)S_{k,Q}(n-l)C_{k,D}(n-l) + \sqrt{\beta_k}S_{k,I}(n-l)C_{k,P}(n-l)]\end{aligned}\quad (4-3)$$

where

$$\rho_{k,l}^I(n) = \frac{1}{2} \sqrt{E_{k,D}} \alpha_{k,l}(n) R[\hat{\tau}_{k,l}(n)] \cos \hat{\varphi}_{k,l}(n) \quad (4-4)$$

$$\rho_{k,l}^Q(n) = \frac{1}{2} \sqrt{E_{k,D}} \alpha_{k,l}(n) R[\hat{\tau}_{k,l}(n)] \sin \hat{\varphi}_{k,l}(n) \quad (4-5)$$

$\hat{\varphi}_{k,l}(n) = \phi_k + \theta_{k,l}(nT_c) - \omega_0 \tau_{k,l}(nT_c)$ is the carrier phase offset. $R[\hat{\tau}_{k,l}(n)]$ is the combined transmit and receive filter pulse shaping as an effective form at the output of the pulse matched filters (Fock, Baltersee, Schulz-Rittic, & Meyr, 2001)

$$R[\hat{\tau}_{k,l}(n)] = \int_{-\infty}^{\infty} h(t)h^*(\hat{\tau} - t) dt \quad (4-6)$$

where $\hat{\tau}_{k,l}(n) = \tau_{k,l}(nT_c) - \tau_{1,i} - \lfloor (\tau_{k,l}(nT_c) - \tau_{1,i}) / T_c \rfloor T_c$ and $|\hat{\tau}_{k,l}(n)| \leq T_c / 2$.

Instead of the convolution form of in (4-6), an equivalent Fourier transform form can be applied to decide $R[\hat{\tau}_{k,l}(n)]$.

$$R[\hat{\tau}_{k,l}(n)] = \int_{-\infty}^{\infty} |H(f)|^2 e^{j2\pi f \hat{\tau}_{k,l}} df = \int_{-\infty}^{\infty} |H(f)|^2 \cos(2\pi f \hat{\tau}_{k,l}) df \quad (4-7)$$

Thus, the mean of the ensemble-averaged correlation matrix of $\hat{\mathbf{X}}_{U,1}^i(n)$ over a window period $\tilde{N}$ ($\tilde{N}$ is set as multiple of N) can be presented as:

$$\begin{aligned}
E[\Phi_1^i(n)] &= E\left[\sum_{n=mN}^{mN+\tilde{N}} \hat{\mathbf{X}}_{U,1}^i(n) \cdot \hat{\mathbf{X}}_{U,1}^{iH}(n)\right] \\
&= \Theta E\left[\sum_{n=mN}^{mN+\tilde{N}} \hat{\mathbf{Y}}_1^i(n) \cdot \hat{\mathbf{Y}}_1^{iH}(n)\right] \Theta^H + E\left[\sum_{n=mN}^{mN+\tilde{N}} \hat{\mathbf{N}}_{U,1}^i(n) \cdot \hat{\mathbf{N}}_{U,1}^{iH}(n)\right]
\end{aligned} \tag{4-8}$$

Let $\Phi_1^{iY}(n) = \sum_{n=mN}^{mN+\tilde{N}} \hat{\mathbf{Y}}_1^i(n) \cdot \hat{\mathbf{Y}}_1^{iH}(n)$ and it is a $K \times K$ matrix. Consequently, the

element in $E[\Phi_1^{iY}(n)]$ can be expressed as:

$$\begin{aligned}
E[\phi_{k_1, k_2, 1}^{iY}(n)] &= E\left[\sum_{n=mN}^{mN+\tilde{N}} \hat{Y}_{k_1, 1}^i(n) \cdot \hat{Y}_{k_2, 1}^{i*}(n)\right] \\
&= E\left[\sum_{n=mN}^{mN+\tilde{N}} \hat{Y}_{k_1, 1}^{iI}(n) \hat{Y}_{k_2, 1}^{iI}(n)\right] + E\left[\sum_{n=mN}^{mN+\tilde{N}} \hat{Y}_{k_1, 1}^{iQ}(n) \hat{Y}_{k_2, 1}^{iQ}(n)\right] \\
&\quad + j \cdot E\left[\sum_{n=mN}^{mN+\tilde{N}} \hat{Y}_{k_1, 1}^{iQ}(n) \hat{Y}_{k_2, 1}^{iI}(n)\right] - j \cdot E\left[\sum_{n=mN}^{mN+\tilde{N}} \hat{Y}_{k_1, 1}^{iI}(n) \hat{Y}_{k_2, 1}^{iQ}(n)\right]
\end{aligned} \tag{4-9}$$

where $1 \leq k_1 \leq K; 1 \leq k_2 \leq K$. Substituting (4-2) and (4-3) into (4-9) and assuming

$\tilde{N} = N$ for analysis purpose, we can expand E^{II} and E^{QI} as:

E^H

$$\begin{aligned}
&= E[\rho_{k_1,l}^I(mN)\rho_{k_2,l}^I(mN)] \cdot E \left\{ \sum_{n=mN+1}^{(m+1)N} \sum_{l=0}^{L-1} [D_{k_1}(m)S_{k_1,I}(n-l)C_{k_1,D}(n-l) - \sqrt{\beta_{k_1}}S_{k_1,Q}(n-l)C_{k_1,P}(n-l)] \right. \\
&\quad \left. \cdot [D_{k_2}(m)S_{k_2,I}(n-l)C_{k_2,D}(n-l) - \sqrt{\beta_{k_2}}S_{k_2,Q}(n-l)C_{k_2,P}(n-l)] \right\} \\
&- E[\rho_{k_1,l}^I(mN)\rho_{k_2,l}^Q(mN)] \cdot E \left\{ \sum_{n=mN+1}^{(m+1)N} \sum_{l=0}^{L-1} [D_{k_1}(m)S_{k_1,I}(n-l)C_{k_1,D}(n-l) - \sqrt{\beta_{k_1}}S_{k_1,Q}(n-l)C_{k_1,P}(n-l)] \right. \\
&\quad \left. \cdot [D_{k_2}(m)S_{k_2,Q}(n-l)C_{k_2,D}(n-l) + \sqrt{\beta_{k_2}}S_{k_2,I}(n-l)C_{k_2,P}(n-l)] \right\} \\
&- E[\rho_{k_1,l}^Q(mN)\rho_{k_2,l}^I(mN)] \cdot E \left\{ \sum_{n=mN+1}^{(m+1)N} \sum_{l=0}^{L-1} [D_{k_1}(m)S_{k_1,Q}(n-l)C_{k_1,D}(n-l) + \sqrt{\beta_{k_1}}S_{k_1,I}(n-l)C_{k_1,P}(n-l)] \right. \\
&\quad \left. \cdot [D_{k_2}(m)S_{k_2,I}(n-l)C_{k_2,D}(n-l) - \sqrt{\beta_{k_2}}S_{k_2,Q}(n-l)C_{k_2,P}(n-l)] \right\} \\
&+ E[\rho_{k_1,l}^Q(mN)\rho_{k_2,l}^Q(mN)] \cdot E \left\{ \sum_{n=mN+1}^{(m+1)N} \sum_{l=0}^{L-1} [D_{k_1}(m)S_{k_1,Q}(n-l)C_{k_1,D}(n-l) + \sqrt{\beta_{k_1}}S_{k_1,I}(n-l)C_{k_1,P}(n-l)] \right. \\
&\quad \left. \cdot [D_{k_2}(m)S_{k_2,Q}(n-l)C_{k_2,D}(n-l) + \sqrt{\beta_{k_2}}S_{k_2,I}(n-l)C_{k_2,P}(n-l)] \right\}
\end{aligned} \tag{4-10}$$

It is assumed that the channel parameters of (4-4) and (4-5) remain constant over one symbol duration (that means $\rho_{k,l}^I(n) = \rho_{k,l}^I(mN)$ $\rho_{k,l}^Q(n) = \rho_{k,l}^Q(mN)$). As a result, the mean values $E[\rho_{k_1,l}^Q(n)\rho_{k_2,l}^Q(n)]$ and $E[\rho_{k_1,l}^I(n)\rho_{k_2,l}^I(n)]$ can be easily proved equal to 0 when $k_1 \neq k_2$, as well as $E[\rho_{k_1,l}^I(n)\rho_{k_2,l}^Q(n)] = 0$. The aperiodic correlations of different scrambling and channelization codes are also assumed with zero mean. The mean values of the synchronised correlations of the scrambling and channelization codes are equal to N . Thus, the resulting E^H is given by

$$\begin{aligned}
E^H &= \begin{cases} N(1 + \beta_k) \sum_{l=1}^L \{ E[\rho_{k,l}^I(n)]^2 + E[\rho_{k,l}^Q(n)]^2 \} & k_1 = k_2 = k \\ 0 & k_1 \neq k_2 \end{cases} \\
&= \begin{cases} \Re_{k,l} \frac{NE_k}{4} \sum_{l=0}^{L-1} E[\alpha_{k,l}^2(mN)] & k_1 = k_2 = k \\ 0 & k_1 \neq k_2 \end{cases}
\end{aligned} \tag{4-11}$$

where $\Re_{k,l} = E[R^2(\hat{\tau}_{k,l}(n))]$ and $E_k = (1 + \beta_k) E_{k,D}$ is the combined energy of both data channel and pilot channel of the k -th user. Similarly, E^{QI} has the following form

$$\begin{aligned}
E^{QI} &= E[\rho_{k_1,l}^Q(mN)\rho_{k_2,l}^I(mN)] \cdot E \left\{ \sum_{n=mN+1}^{(m+1)N} \sum_{l=0}^{L-1} [D_{k_1}(m)S_{k_1,I}(n-l)C_{k_1,D}(n-l) - \sqrt{\beta_{k_1}}S_{k_1,Q}(n-l)C_{k_1,P}(n-l)] \right. \\
&\quad \left. \cdot [D_{k_2}(m)S_{k_2,I}(n-l)C_{k_2,D}(n-l) - \sqrt{\beta_{k_2}}S_{k_2,Q}(n-l)C_{k_2,P}(n-l)] \right\} \\
&\quad - E[\rho_{k_1,l}^Q(mN)\rho_{k_2,l}^Q(mN)] \cdot E \left\{ \sum_{n=mN+1}^{(m+1)N} \sum_{l=0}^{L-1} [D_{k_1}(m)S_{k_1,Q}(n-l)C_{k_1,D}(n-l) + \sqrt{\beta_{k_1}}S_{k_1,I}(n-l)C_{k_1,P}(n-l)] \right. \\
&\quad \left. \cdot [D_{k_2}(m)S_{k_2,I}(n-l)C_{k_2,D}(n-l) - \sqrt{\beta_{k_2}}S_{k_2,Q}(n-l)C_{k_2,P}(n-l)] \right\} \\
&\quad + E[\rho_{k_1,l}^I(mN)\rho_{k_2,l}^I(mN)] \cdot E \left\{ \sum_{n=mN+1}^{(m+1)N} \sum_{l=0}^{L-1} [D_{k_1}(m)S_{k_1,Q}(n-l)C_{k_1,D}(n-l) + \sqrt{\beta_{k_1}}S_{k_1,I}(n-l)C_{k_1,P}(n-l)] \right. \\
&\quad \left. \cdot [D_{k_2}(m)S_{k_2,I}(n-l)C_{k_2,D}(n-l) - \sqrt{\beta_{k_2}}S_{k_2,Q}(n-l)C_{k_2,P}(n-l)] \right\} \\
&\quad - E[\rho_{k_1,l}^I(mN)\rho_{k_2,l}^Q(mN)] \cdot E \left\{ \sum_{n=mN+1}^{(m+1)N} \sum_{l=0}^{L-1} [D_{k_1}(m)S_{k_1,Q}(n-l)C_{k_1,D}(n-l) + \sqrt{\beta_{k_1}}S_{k_1,I}(n-l)C_{k_1,P}(n-l)] \right. \\
&\quad \left. \cdot [D_{k_2}(m)S_{k_2,Q}(n-l)C_{k_2,D}(n-l) + \sqrt{\beta_{k_2}}S_{k_2,I}(n-l)C_{k_2,P}(n-l)] \right\}
\end{aligned} \tag{4-12}$$

Applying the above assumptions to E^{II} , it can be proved that $E^{QI} = 0$. Also it can be proved that $E^{IQ} = E^{QI}$ and $E^{II} = E^{QQ}$, and therefore

$$E[\Phi_1^{iY}(n)] = \begin{bmatrix} \Re \frac{NE_1}{2} A_T & 0 & \dots & 0 \\ 0 & \Re \frac{NE_2}{2} A_T & \dots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \dots & \Re \frac{NE_K}{2} A_T \end{bmatrix} \tag{4-13}$$

where $A_T = \sum_{l=1}^L E[\alpha_{1,l}^2(mN)]$, since it is assumed that all users experience the same independent identical distribution. And $\Re_{k,l}$ is identical for different user and different

path as $\Re$, because $\hat{\tau}_{k,l}(n)$ is uniformly distributed in $[-T_c/2, T_c/2]$. Thus assuming

$$E_T = E_1 = E_2 = \dots = E_K,$$

$$E[\Phi_1^i(n)] = \Re \frac{NE_T}{2} A_T \Theta \Theta^H + N\sigma_0^2 \mathbf{I} \quad (4-14)$$

To minimize the interference of the undesired users, under the symbol-rate processing for the adaptive antenna array, after pulse matched filtering all received signals undergo the digital beamforming. Beamforming is performed by the weighted average for the i -th path of k -th user. The beamforming output of the desired user is

$$\begin{aligned} \tilde{X}_1^i(n) &= \mathbf{w}_1^{iH} \hat{\mathbf{X}}_{U,1}^i(n) \\ &= \mathbf{w}_1^{iH} \Theta \hat{\mathbf{Y}}_1^i(n) + \mathbf{w}_1^{iH} \hat{\mathbf{N}}_{U,1}^i(n) \\ &= \mathbf{G}_1^i \hat{\mathbf{Y}}_1^i(n) + \mathbf{w}_1^{iH} \hat{\mathbf{N}}_{U,1}^i(n) \end{aligned} \quad (4-15)$$

Let $\tilde{y}_1^i(n) = \mathbf{G}_1^i \hat{\mathbf{Y}}_1^i(n)$, $\tilde{n}_{U,1}^i(n) = \mathbf{w}_1^{iH} \hat{\mathbf{N}}_{U,1}^i(n)$. By taking $g_1^i(\theta_k) = g_1^{iI}(\theta_k) + jg_1^{iQ}(\theta_k)$

into consideration, then $\tilde{y}_1^i(n)$ can be expressed as the real and imaginary parts with

updated $\hat{\rho}_{k,l}^I(n)$ and $\hat{\rho}_{k,l}^Q(n)$ from $\rho_{k,l}^I(n)$ and $\rho_{k,l}^Q(n)$.

$$\begin{aligned} \tilde{y}_1^{iI}(n) &= \sum_{k=1}^K \sum_{l=0}^{L-1} \hat{\rho}_{k,l}^I(n) [D_k(m) S_{k,I}(n-l) C_{k,D}(n-l) - \sqrt{\beta_k} S_{k,Q}(n-l) C_{k,P}(n-l)] \\ &\quad - \sum_{k=1}^K \sum_{l=0}^{L-1} \hat{\rho}_{k,l}^Q(n) [D_k(m) S_{k,Q}(n-l) C_{k,D}(n-l) + \sqrt{\beta_k} S_{k,I}(n-l) C_{k,P}(n-l)] \end{aligned} \quad (4-16)$$

$$\begin{aligned}
\tilde{y}_{k,1}^{iQ}(n) &= \sum_{k=1}^K \sum_{l=0}^{L-1} \hat{\rho}_{k,l}^Q(n) [D_k(m)S_{k,I}(n-l)C_{k,D}(n-l) - \sqrt{\beta_k}S_{k,Q}(n-l)C_{k,P}(n-l)] \\
&\quad + \sum_{k=1}^K \sum_{l=0}^{L-1} \hat{\rho}_{k,l}^I(n) [D_k(m)S_{k,Q}(n-l)C_{k,D}(n-l) + \sqrt{\beta_k}S_{k,I}(n-l)C_{k,P}(n-l)]
\end{aligned} \tag{4-17}$$

where

$$\hat{\rho}_{k,l}^I(n) = \frac{1}{2} \sqrt{E_{k,D}} \alpha_{k,l}(n) g_1^{(i)I}(\theta_k) R(\hat{\tau}_{k,l}(n)) \cos \hat{\varphi}_{k,l}(n) \tag{4-18}$$

$$\hat{\rho}_{k,l}^Q(n) = \frac{1}{2} \sqrt{E_{k,D}} \alpha_{k,l}(n) g_1^{(i)Q}(\theta_k) R(\hat{\tau}_{k,l}(n)) \sin \hat{\varphi}_{k,l}(n) \tag{4-19}$$

The next process in detecting both data and pilot symbol is despreading and de-scrambling. $\tilde{X}_1^i(n)$ is multiplied with complex conjugate scrambling codes and then the resultant real and imaginary parts are multiplied by the data sequence or the pilot channelization codes respectively. The final real and imaginary results are accumulated over the symbol period. These codes are locally generated and the corresponding to the path of the desired user. The practical process of despreading and de-scrambling is shown in Figure.4.2. The outputs of the detected i -th path from the first user's data and pilot symbols are $\tilde{X}_1^{iD}(mN)$ ($\tilde{X}_1^{iD}(m)$ for short) and $\tilde{X}_1^{iP}(m)$ respectively which can be presented as:

$$\begin{aligned}
\tilde{X}_1^{iD}(m) &= \sum_{n=mN+1}^{(m+1)N} [\tilde{y}_1^i(n) + \tilde{n}_{Z,1}^i(n)] [S_{1,I}(n-i) - jS_{1,Q}(n-i)] \cdot C_{1,D}(n-i) \\
&= \tilde{y}_1^{iD}(m) + \tilde{n}_{U,1}^{iD}(m)
\end{aligned} \tag{4-20}$$

$$\begin{aligned}
\tilde{X}_1^{iP}(m) &= \sum_{n=mN+1}^{(m+1)N} [\tilde{y}_1^i(n) + \tilde{n}_{Z,1}^i(n)] [S_{1,I}(n-i) - jS_{1,Q}(n-i)] \cdot -jC_{1,P}(n-i) \\
&= \tilde{y}_1^{iP}(m) + \tilde{n}_{Z,1}^{iP}(m)
\end{aligned} \tag{4-21}$$

By applying (4-16) and (4-17) into (4-20), $\tilde{y}_1^{iP}(m)$ can be presented as:

$$\begin{aligned}
\tilde{y}_1^{iP}(m) &= \sum_{n=mN+1}^{(m+1)N} \tilde{y}_1^i(n) [-S_{1,Q}(n-i) - jS_{1,I}(n-i)] \cdot C_{1,P}(n-i) \\
&= \sum_{n=mN+1}^{(m+1)N} [\tilde{y}_1^I(n) + j \cdot \tilde{y}_1^Q(n)] [-S_{1,Q}(n-i) - jS_{1,I}(n-i)] \cdot C_{1,P}(n-i) \\
&= \underbrace{\sum_{n=mN+1}^{(m+1)N} [-\tilde{y}_1^I(n)S_{1,Q}(n-i) + \tilde{y}_1^Q(n)S_{1,I}(n-i)] \cdot C_{1,P}(n-i)}_{\tilde{y}_1^{iPI}(m)} \\
&\quad + j \cdot \underbrace{\sum_{n=mN+1}^{(m+1)N} [-\tilde{y}_1^Q(n)S_{1,Q}(n-i) - \tilde{y}_1^I(n)S_{1,I}(n-i)] \cdot C_{1,P}(n-i)}_{\tilde{y}_1^{iPQ}(m)} \tag{4-22}
\end{aligned}$$

where the real and imaginary parts are written respectively as followings:

$$\begin{aligned}
\tilde{y}_1^{iPI}(m) &= \\
&- \sum_{n=mN+1}^{(m+1)N} \sum_{k=1}^K \sum_{l=0}^{L-1} \hat{\rho}_{k,I}^I(n) [D_k(m)S_{k,I}(n-l)C_{k,D}(n-l) - \sqrt{\beta_k}S_{k,Q}(n-l)C_{k,P}(n-l)] S_{1,Q}(n-i)C_{1,P}(n-i) \\
&+ \sum_{n=mN+1}^{(m+1)N} \sum_{k=1}^K \sum_{l=0}^{L-1} \hat{\rho}_{k,I}^Q(n) [D_k(m)S_{k,Q}(n-l)C_{k,D}(n-l) + \sqrt{\beta_k}S_{k,I}(n-l)C_{k,P}(n-l)] S_{1,Q}(n-i)C_{1,P}(n-i) \\
&+ \sum_{n=mN+1}^{(m+1)N} \sum_{k=1}^K \sum_{l=0}^{L-1} \hat{\rho}_{k,I}^Q(n) [D_k(m)S_{k,I}(n-l)C_{k,D}(n-l) - \sqrt{\beta_k}S_{k,Q}(n-l)C_{k,P}(n-l)] S_{1,I}(n-i)C_{1,P}(n-i) \\
&+ \sum_{n=mN+1}^{(m+1)N} \sum_{k=1}^K \sum_{l=0}^{L-1} \hat{\rho}_{k,I}^I(n) [D_k(m)S_{k,Q}(n-l)C_{k,D}(n-l) + \sqrt{\beta_k}S_{k,I}(n-l)C_{k,P}(n-l)] S_{1,I}(n-i)C_{1,P}(n-i)
\end{aligned} \tag{4-23}$$

$$\begin{aligned}
\hat{y}_1^{iPQ}(m) = & \\
& - \sum_{n=mN+1}^{(m+1)N} \sum_{k=1}^K \sum_{l=0}^{L-1} \hat{\rho}_{k,l}^Q(n) [D_k(m)S_{k,I}(n-l)C_{k,D}(n-l) - \sqrt{\beta_k}S_{k,Q}(n-l)C_{k,P}(n-l)]S_{1,Q}(n-i)C_{1,P}(n-i) \\
& - \sum_{n=mN+1}^{(m+1)N} \sum_{k=1}^K \sum_{l=0}^{L-1} \hat{\rho}_{k,l}^I(n) [D_k(m)S_{k,Q}(n-l)C_{k,D}(n-l) + \sqrt{\beta_k}S_{k,I}(n-l)C_{k,P}(n-l)]S_{1,Q}(n-i)C_{1,P}(n-i) \\
& - \sum_{n=mN+1}^{(m+1)N} \sum_{k=1}^K \sum_{l=0}^{L-1} \hat{\rho}_{k,l}^I(n) [D_k(m)S_{k,I}(n-l)C_{k,D}(n-l) - \sqrt{\beta_k}S_{k,Q}(n-l)C_{k,P}(n-l)]S_{1,I}(n-i)C_{1,P}(n-i) \\
& + \sum_{n=mN+1}^{(m+1)N} \sum_{k=1}^K \sum_{l=0}^{L-1} \hat{\rho}_{k,l}^Q(n) [D_k(m)S_{k,Q}(n-l)C_{k,D}(n-l) + \sqrt{\beta_k}S_{k,I}(n-l)C_{k,P}(n-l)]S_{1,I}(n-i)C_{1,P}(n-i)
\end{aligned} \tag{4-24}$$

The noise term in (4-20) is given by

$$\tilde{n}_{U,1}^{iP}(m) = \sum_{n=mN+1}^{(m+1)N} \tilde{n}_{U,1}^i(m) [S_{1,I}(n-i) - jS_{1,Q}(n-i)]C_{1,P}(n-i) \tag{4-25}$$

Because the pilot symbol is constant, the pilots are applied to do the channel estimation. The error of this channel estimation is non-trivial and has direct contribution to system performance. A simple way to reduce the estimation error is to use a block average over N_E pilot symbols (Adachi, Sawahashi, & Suda, 1998) that can be considered as a low-pass filter as shown in Figure 4-2. Because the spreading factor N_1 of the pilot channel is fixed to be 256, while the spreading factor N_2 of data channel is varied from 4 to 256, the channel estimate that is derived from one long pilot symbol corresponds to N_1 / N_2 data symbols. The properties of channel estimation of the i -th path from the desired user at the first antenna element after block averaging can be written as:

$$\bar{X}_1^{iR}(m) = \frac{1}{N_E} \sum_{\tilde{m}=0}^{N_E-1} \bar{X}_1^{iP}(m + \tilde{m}) \tag{4-26}$$

Therefore, a corresponding reference signal is constructed by multiplying the result of channel estimation to the channelization and scrambling code for the desired user's pilot channel. The adaptation processing of the antenna array weights is given by,

$$\widehat{X}_1^{iR}(n) = \widehat{X}_1^{iR}(m) \cdot jC_{1,P}(n)[S_{1,I}(n) + jS_{1,Q}(n)] \quad (4-27)$$

Because the phase differences in different antenna elements are all related to the first antenna element (Liberti & Rappaport, 1999), the channel estimations from the first antenna element are not affected by incident angles $\theta_{k,l}(t)$.

4.3 Symbol Rate Processing Scheme of the Adaptive Antenna Array

It is clear that in the received signals the desired user is interfered by the other users. Thus, an adaptive antenna array can be applied to mitigate interferences from undesired users. With the different gains in G_1^i for different incident angles, an antenna array can maximize signal-to-noise ratio of the desired user and minimize that of interfering users. Accordingly, the weight vector of the antenna array is optimised to perform the augmentation and suppressions. Because both transmission channels and interfering users are unknown to the desired user, it is necessary for receivers to make the weight vector adaptive to variances from channels and interferences. The RLS algorithm (Haykin, 2001; Banister & Zeidler, 2002) is considered here for the adaptation of weight vector. The signal vector of the i -th path from the desired user at different antenna elements $\widehat{X}_1^i(n)$ is the input of the RLS algorithm. The output is the $w_1^i(n)$. $\widehat{X}_1^{iR}(n)$ is

considered as the reference signal. Thus, the general formulations of adaptation with RLS algorithm is as follows:

$$\mathbf{w}_1^i(n) = [\Phi_1^i(n)]^{-1} \mathbf{V}_1^i(n) \quad (4-28)$$

where $[\Phi_1^i(n)]^{-1}$ is the inverse of $\Phi_1^i(n)$ and

$$\begin{aligned} \mathbf{V}_1^i(n) &= \sum_{n=mN}^{mN+\tilde{N}} \hat{\mathbf{X}}_1^i(n) \hat{X}_1^{iR*}(n) \\ &= \oplus \sum_{n=mN}^{mN+\tilde{N}} \hat{\mathbf{Y}}_1^i(n) \hat{X}_1^{iR*}(n) + \sum_{n=mN}^{mN+\tilde{N}} \hat{\mathbf{N}}_{U,1}^i(n) \hat{X}_1^{iR*}(n) \quad (4-29) \end{aligned}$$

It is assumed that $[\Phi_1^i(n)]^{-1}$ and $\mathbf{V}_1^i(n)$ are independent because they both go through different average process. Thus the mean value of $\mathbf{w}_1^i(n)$ equals $E\{[\Phi_1^i(n)]^{-1}\}E[\mathbf{V}_1^i(n)]$

$[\Phi_1^i(n)]^{-1}$ can be described as a *complex Wishart distribution* (Haykin, 2001). It is

proved that $E\{[\Phi_1^i(n)]^{-1}\} = \{E[\Phi_1^i(n)]\}^{-1}/(\tilde{N} - K - 1)$. Also because $\hat{\mathbf{N}}_{U,1}^i(n)$ is

considered as i.i.d, the expectation of $\mathbf{V}_1^i(n)$ is given by

$$\begin{aligned} E[\mathbf{V}_1^i(n)] &= \oplus E\left\{ \tilde{X}_1^{iR*}(m) \sum_{n=mN}^{mN+\tilde{N}} \hat{\mathbf{Y}}_1^i(n) [-jC_{1,P}(n-i)][S_{1,I}(n-i) - jS_{1,Q}(n-i)] \right\} \\ &= \oplus E[\tilde{X}_1^{iR*}(m)] \cdot \underbrace{E\left\{ \sum_{n=mN}^{mN+\tilde{N}} \hat{\mathbf{Y}}_1^i(n) [-jC_{1,P}(n-i)][S_{1,I}(n-i) - jS_{1,Q}(n-i)] \right\}}_{E[\hat{\mathbf{Y}}_1^{iV}(n)]} \quad (4-30) \end{aligned}$$

where the components in $E[\hat{\mathbf{Y}}_1^{iV}(n)]$ can be further described by applying $\hat{y}_{k,1}^{iI}(n)$ and

$\hat{y}_{k,1}^{iQ}(n)$ in (4-2) and (4-3) as:

$$\begin{aligned}
E[\hat{y}_{k,1}^{iV}(n)] = & \\
E \left[\left\{ \sum_{l=0}^{L-1} \rho_{k,l}^I(n) [D_k(m)S_{k,I}(n-l)C_{k,D}(n-l) - \sqrt{\beta_k}S_{k,Q}(n-l)C_{k,P}(n-l)] \right\} C_{1,P}(n-i) [-S_{1,Q}(n-i) - jS_{1,I}(n-i)] \right. \\
& \left. - \left\{ \sum_{l=0}^{L-1} \rho_{k,l}^Q(n) [D_k(m)S_{k,Q}(n-l)C_{k,D}(n-l) + \sqrt{\beta_k}S_{k,I}(n-l)C_{k,P}(n-l)] \right\} C_{1,P}(n-i) [-S_{1,Q}(n-i) - jS_{1,I}(n-i)] \right] \\
& + jE \left[\left\{ \sum_{l=0}^{L-1} \rho_{k,l}^Q(n) [D_k(m)S_{k,I}(n-l)C_{k,D}(n-l) - \sqrt{\beta_k}S_{k,Q}(n-l)C_{k,P}(n-l)] \right\} C_{1,P}(n-i) [-S_{1,Q}(n-i) - jS_{1,I}(n-i)] \right. \\
& \left. - \left\{ \sum_{l=0}^{L-1} \rho_{k,l}^I(n) [D_k(m)S_{k,Q}(n-l)C_{k,D}(n-l) + \sqrt{\beta_k}S_{k,I}(n-l)C_{k,P}(n-l)] \right\} C_{1,P}(n-i) [-S_{1,Q}(n-i) - jS_{1,I}(n-i)] \right]
\end{aligned} \tag{4-31}$$

With the above assumptions that the mean value of aperiodic correlations of different scrambling and channelization codes are zero and the value of synchronised correlations is N , so $E[\hat{y}_{1,1}^{iV}(n)] = \tilde{N}\sqrt{\beta_1 E_{1,D}}\alpha_{1,i} \times [\cos \hat{\varphi}_{1,i}(mN) + j \sin \hat{\varphi}_{1,i}(mN)]$ and $E[\hat{y}_{k,1}^{iV}(n)]$ equal 0 ($k \neq 1$). Thus, $E[\mathbf{V}_1^i(n)]$ can finally be given by

$$E[\mathbf{V}_1^i(n)] = \tilde{N}\alpha_{1,i}\sqrt{\beta_1 E_{1,D}} [\cos \hat{\varphi}_{1,i}(mN) + j \sin \hat{\varphi}_{1,i}(mN)] E[\tilde{X}_1^{iR*}(m)] \mathbf{\Theta}_1 \tag{4-32}$$

With reference to the results in the paper (Wang & Chen, 2001), the statistics of $\tilde{X}_1^{iP}(m)$ and $\tilde{X}_1^{iD}(m)$ can equivalently obtained. For the real and imaginary parts, the expectations of them are

$$E[\tilde{X}_1^{iPI}(m)] = N\alpha_{1,i}(mN)g_1^{iI}(\theta_1)\sqrt{\beta_1 E_{1,D}} \cos \hat{\varphi}_{1,i}(mN) \tag{4-33}$$

$$E[\tilde{X}_1^{iPQ}(m)] = N\alpha_{1,i}(mN)g_1^{iQ}(\theta_1)\sqrt{\beta_1 E_{1,D}} \sin \hat{\varphi}_{1,i}(mN) \tag{4-34}$$

$$E[\tilde{X}_1^{iDI}(m)] = ND_1(m)\alpha_{1,i}(mN)g_1^{iI}(\theta_1)\sqrt{E_{1,D}} \cos \hat{\varphi}_{1,i}(mN) \tag{4-35}$$

$$E[\tilde{X}_1^{iDQ}(m)] = ND_1(m)\alpha_{1,i}(mN)g_1^{iQ}(\theta_1)\sqrt{E_{1,D}} \sin \hat{\varphi}_{1,i}(mN) \tag{4-36}$$

For easy analysis, it is assumed that all the users have the same power ratio and chip energy so that the variances of the interference and noise in $\tilde{X}_1^{iP}(m)$ and $\tilde{X}_1^{iD}(m)$ are identical and equal to

$$\begin{aligned} \text{var } IN_1^i &= \underbrace{\Re \frac{N}{2} \sum_{k=1}^K E_k |g_1^i(\theta_k)|^2 \sum_{\substack{l=0 \\ \text{if } k=1, l \neq i}}^{L-1} E[\alpha_{k,l}^2(mN)] + \Re \frac{N}{4} E_T E[\alpha_{1,i}^2(mN)]}_{\text{Variance of Interference}} \\ &\quad + \underbrace{E \left\| \mathbf{w}_1^{iH}(n) \hat{\mathbf{N}}_{Z,1}^i(n) \right\|^2}_{\text{Variance of Noise}} \\ &\approx \frac{\Re}{2} N E_T A_T \sum_{k=1}^K |g_1^i(\theta_k)|^2 + N \left\| \mathbf{w}_1^i(n) \right\|^2 \sigma_0^2 \end{aligned} \quad (4-37)$$

Assuming that the successive $\tilde{X}_1^{iP}(m + \tilde{m})$ in (4-26) for different $\tilde{m}$ are independent of each other, it is obtained

$$E[\tilde{X}_1^{iR}(m)] = E[\tilde{X}_1^{iPI}(m)] + jE[\tilde{X}_1^{iPQ}(m)] \quad (4-38)$$

$$\text{var}[\tilde{X}_1^{iR}(m)] = \frac{1}{N_E} \left\{ \text{var}[\tilde{X}_1^{iPI}(m)] + \text{var}[\tilde{X}_1^{iPQ}(m)] \right\} \quad (4-39)$$

Combining with (4-14), (4-30), (4-32 ~ 4-34) and (4-38), $E[\mathbf{w}_1^i(mN) | \alpha_{1,i}]$ can be

written as:

$$E[\mathbf{w}_1^i(mN) | \alpha_{1,i}] = \frac{\beta}{1 + \beta} \frac{N^2 E_T}{N - K - 1} g_1^i(\theta_1) \alpha_{1,i}^2 \left(\Re \frac{N E_T}{2} A_T \mathbf{\Theta} \mathbf{\Theta}^H + N \sigma_0^2 \mathbf{I} \right)^{-1} \mathbf{\Theta}_1 \quad (4-40)$$

$E[\mathbf{w}_1^i(mN) | \alpha_{1,i}]$ is considered as the final value of the weight vector, $\hat{\mathbf{w}}_1^i(n)$, over the

adaptation period from mN to $(m+1)N$. It is clear that in (4-40) the amplitude of the

weight vector is affected by the channel, previous adaptation, and other factors. The effect from these factors can be cancelled by normalization so that $\|\hat{\mathbf{w}}_1^i(n)\| = 1$ and

$$\hat{\mathbf{w}}_1^i(n) = \frac{\left(\Re \frac{NE}{2} A_T \Theta \Theta^H + N\sigma_0^2 \mathbf{I} \right)^{-1} \Theta_1}{\left\| Z \cdot \left(\Re \frac{NE}{2} A_T \Theta \Theta^H + N\sigma_0^2 \mathbf{I} \right)^{-1} \Theta_1 \right\|} \quad (4-41)$$

Consequently, the gain vector can be expressed as:

$$\mathbf{G}_1^i = \left[\frac{\Theta_1^H \left(\Re \frac{NE_T}{2} A_T \Theta \Theta^H + N\sigma_0^2 \mathbf{I} \right)^{-1} \Theta_1}{\left\| Z \cdot \left(\Re \frac{NE_T}{2} A_T \Theta \Theta^H + N\sigma_0^2 \mathbf{I} \right)^{-1} \Theta_1 \right\|}, \dots, \frac{\Theta_1^H \left(\Re \frac{NE_T}{2} A_T \Theta \Theta^H + N\sigma_0^2 \mathbf{I} \right)^{-1} \Theta_K}{\left\| Z \cdot \left(\Re \frac{NE_T}{2} A_T \Theta \Theta^H + N\sigma_0^2 \mathbf{I} \right)^{-1} \Theta_1 \right\|} \right]^T \quad (4-42)$$

The RAKE receiver that attempts to collect the signal energy from all the received signal paths is applied here. $\gamma_1^i(m) = \tilde{X}_1^{iDI}(m)\tilde{X}_1^{iRI}(m) + \tilde{X}_1^{iDQ}(m)\tilde{X}_1^{iRQ}(m)$ is the i -th path output. $C_{P,1}(n)$ and $C_{D,1}(n)$ are orthogonal, $\hat{\rho}_{k,l}^I(n)$ and $\hat{\rho}_{k,l}^Q(n)$ are independent, so $\tilde{X}_1^{iDI}(m)$ and $\tilde{X}_1^{iRI}(m)$ as well as $\tilde{X}_1^{iDQ}(m)$ and $\tilde{X}_1^{iRQ}(m)$ are independent. Thus the expectation of $\gamma_1^i(m)$ is given by

$$\begin{aligned} E[\gamma_1^i(m)] &= E[\tilde{X}_1^{iDI}(m)]E[\tilde{X}_1^{iRI}(m)] + E[\tilde{X}_1^{iDQ}(m)]E[\tilde{X}_1^{iRQ}(m)] \\ &= N^2 \sqrt{\beta} D_1(m) \|g_1^i(\theta_1)\|^2 \alpha_{1,i}^2(mN) E_{1,D} \end{aligned} \quad (4-43)$$

The variance of $\gamma_1^i(m)$ is

$$\begin{aligned}
& \text{var}[\gamma_1^i(m)] \\
&= E\left\{\left[\widetilde{X}_1^{iD/I}(m)\widetilde{X}_1^{iR/I}(m) + \widetilde{X}_1^{iD/Q}(m)\widetilde{X}_1^{iR/Q}(m)\right]^2\right\} - E^2\left[\widetilde{X}_1^{iD/I}(m)\widetilde{X}_1^{iR/I}(m) + \widetilde{X}_1^{iD/Q}(m)\widetilde{X}_1^{iR/Q}(m)\right] \\
&= \left(\beta + \frac{1}{N_E}\right)\alpha_{1,i}^2(mN)N^3E_{1,D}\left[\sigma_0^2 + \frac{\Re}{2}E_TA_T\sum_{k=1}^K\|g_1^i(\theta_k)\|^2\right] + \frac{2N^2}{N_E}\left[\sigma_0^2 + \frac{\Re}{2}E_TA_T\sum_{k=1}^K\|g_1^i(\theta_k)\|^2\right]^2
\end{aligned} \tag{4-44}$$

Practically, the available number of branches for the RAKE receiver is L_A . As a result,

the final RAKE for the decision of the data symbol of the desired user is

$\Upsilon(m) = \sum_{i=1}^{L_A} \gamma_1^i(m)$. Because results from each path are independent from each other and $\Upsilon(m)$ is an accumulation of i.i.d random variables, $\Upsilon(m)$ is considered as a Gaussian variable. The mean and variance of $\Upsilon(m)$ are as followings where $\alpha_1 = \sum_{i=1}^{L_A} \alpha_{1,i}^2(mN)$:

$$E[\Upsilon(m)|\alpha_1] = \sum_{i=1}^{L_A} E[\gamma_1^i(m)] = \sqrt{\beta}E_{1,D}\alpha_1N^2D_1(m)\sum_{i=1}^{L_A}\|g_1^i(\theta_1)\|^2 \tag{4-45}$$

and

$$\begin{aligned}
& \text{var}[\Upsilon(m)|\alpha_1] = \sum_{i=1}^{L_A} \text{var}[\gamma_1^i(m)] \\
&= \left(\beta + \frac{1}{N_E}\right)N^3E_{1,D}\alpha_1\left[\sigma_0^2 + \frac{\Re}{2}E_TA_T\sum_{k=1}^K\|g_1^i(\theta_k)\|^2\right] + \frac{2N^2}{N_E}\sum_{i=1}^{L_A}\left[\sigma_0^2 + \frac{\Re}{2}E_TA_T\sum_{k=1}^K\|g_1^i(\theta_k)\|^2\right]^2
\end{aligned} \tag{4-46}$$

Therefore, the average probability of error for the RAKE receiver is obtained as

$$P_e = \int_0^\infty P_e(\alpha_1)p(\alpha_1)d\alpha_1 \text{ for the desired user reception. Combining (4-45) and (4-46),}$$

$P_e(\alpha_1)$ is the conditional error probability of coherent detection with instantaneous SINR

equals $E^2[\Upsilon(m) | \alpha_1] / \text{var}[\Upsilon(m) | \alpha_1]$ of the RAKE output. This yields

$$\begin{aligned}
P_e(\alpha_1) &= \frac{1}{2} \text{erfc}(\sqrt{\text{SINR}}) = \frac{1}{2} \text{erfc} \left\{ \frac{\text{E}[\Upsilon(m) \mid \alpha_1]}{\sqrt{\text{var}[\Upsilon(m) \mid \alpha_1]}} \right\} \\
&= \frac{1}{2} \text{erfc} \left\{ \left[\frac{(1+\beta)}{\alpha_1} \left(1 + \frac{1}{\beta g_A^2 N_E} \right) \left(\frac{1}{\text{SNR}} + \frac{\Re}{2} A_T \frac{g_T}{N} \right) + \frac{1}{\alpha_1^2} \frac{2L_A(1+\beta)^2}{\beta g_A^2 N_E} \left(\frac{1}{\text{SNR}} + \frac{\Re}{2} A_T \frac{g_T}{N} \right)^2 \right]^{-1/2} \right\}
\end{aligned} \tag{4-47}$$

where $g_A = \sum_{i=1}^{L_A} \|g_1^i(\theta_1)\|^2$, $g_T = \sum_{k=1}^K \|g_1^i(\theta_k)\|^2$, $\text{SNR} = NE_T / \sigma_0^2$, and $\text{erfc}(x) = \frac{2}{\sqrt{\pi}} \int_x^\infty e^{-t^2/2} dt$.

4.4 Performances of the Symbol Rate Processing Scheme

A Nakagami distribution fading channel model is considered for the analysis of the reception performance with adaptive antenna array. This channel model is more general than the Rayleigh model and Nakagami distribution has the flexibility of modelling fading conditions that are more or less severe than Rayleigh fading and even special cases as Rayleigh, Rician distributions (Al-Hussaini & Al-Bassiouni, 1985; Aalo & Zhang, 2001).

According to the reference (Al-Hussaini & Al-Bassiouni, 1985), all received signals are assumed through a channel with i.i.d gains, where the PDF of α_1 can be treated as a Gamma distribution as:

$$p(\alpha_1) = \left(\frac{m_1}{\Omega_1} \right)^{m_1} \frac{\alpha_1^{m_1-1}}{\Gamma(m_1)} \exp \left(-\frac{m_1}{\Omega_1} \alpha_1 \right) \tag{4-48}$$

where m_1 is the Nakagami fading parameters and $\Omega_1 = \sum_{l=1}^{L_A} E \alpha_{1,l}^2 = A_T$. As a result,

the average bit error probability P_e is given by

$$\begin{aligned}
 P_e &= \int_0^\infty P_e(\alpha_1) p(\alpha_1) d\alpha_1 \\
 &= \int_0^\infty \frac{1}{2} \operatorname{erfc} \left\{ \left[\frac{(1+\beta)}{\alpha_1} \left(1 + \frac{1}{\beta g_A^2 N_E} \right) \left(\frac{1}{SNR} + \frac{\Re}{2} A_T \frac{g_T}{N} \right) + \frac{1}{\alpha_1^2} \frac{2M(1+\beta)^2}{\beta g_A^2 N_E} \left(\frac{1}{SNR} + \frac{\Re}{2} A_T \frac{g_T}{N} \right)^2 \right]^{1/2} \right\} \\
 &\quad \cdot \left(\frac{m_1}{\Omega_1} \right)^{m_1} \frac{\alpha_1^{m_1-1}}{\Gamma(m_1)} \exp \left(-\frac{m_1}{\Omega_1} \alpha_1 \right) d\alpha_1 \quad (4-49)
 \end{aligned}$$

The performances of the W-CDMA system with adaptive antenna array based on RLS algorithm are examined. The effects of average received SNR, the number of users K , the number of antenna elements, and channel estimation period N_E on the BER performance are presented in the following figures. The total number of multipaths is set equal to 8 and the number of resolvable paths of the RAKE receiver equal to 6. The incident angles of both desired user and interfering users are set randomly within $[-60^\circ \ 60^\circ]$. The spreading factors for pilot and data channel are 256 and 64 respectively. The channel estimation is performed over $N_E = 4$ continuous estimations, the same as estimation over the period of one pilot symbol (256 chips).

In Figure 4-3, the effects brought by the average received SNR with the different numbers of antenna elements are presented. It is clear that with more antenna elements the performance becomes better. Even in the case that the number of the antenna elements is less than the number of users, improvements in BER performances are achieved. Note that when $U=1$ the result is the same as the result in the reference (Wang

& Chen, 2001) with the same condition. With additional antenna elements and adaptive antenna array processing the error performances can be enhanced dramatically.

Figure 4-4 presents the effects of channel estimation in error performance. The reference signal for the adaptation is reconstructed by multiplying channel estimations to channelization and scrambling codes and the reference signal is very important to the adaptive algorithm, so the variable N_E affects system performance significantly. To illustrate the effects clearly, the number of total active users is $K=10$ while the antenna array has 8 elements. With long-time average, the channel estimations become better so that the performance improves. However, because these results are based on the assumption that the channel keeps constant statistical properties over long average time, N_E should be selected reasonably in practice.

The BER performances for different numbers of active users are shown in Figure 4.5 with different number of elements in the antenna array. It is assumed that the SNR of the received signal is 15dB. It is clear that with more active users the BER become higher. But it is also seen that with $U = 10$, BER can get a value of about 10^{-3} .

4.5 Simulations Results from Symbol Rate Processing Scheme

Simulations are also considered to investigate the performances with the symbol-rate adaptive antenna processing in the W-CDMA system uplink under the M.1225 (ITU-R M.1225) channel model. In the W-CDMA system, all the users have the same channelization code $C_{ch,256}^0$ for the pilot signals. Five users are considered here in which

four interfering users and one desired user are assumed with different data rates (30 or 60 kbps) and powers. The antenna array is assumed with five antenna elements. For easy simulations, it is assumed that all the l -th paths of a user at each antenna element have the same incident angle. All users have the same constant pilot signals, but with different scrambling codes (Truncated-Gold sequence). The power ratio of pilot to data channels is $\beta=0.25$. Other parameters of the uplink transmission are similar to the reference (ITU-R M.1225).

The channel, the incident angles of the interferers, and the desired user are the same as the definition in Chapter 3 and in the reference (Joham & Utschick, 2001). The average received signal to interference power ratio at each antenna of the desired user is -12 dB (Interfering users have quadruple of the desired user's power).

In the simulations, the RLS algorithm is applied to perform the adaptation of the weight. The adaptation is done over every one pilot symbol, or 256 chips. The initial weight vector is set as $(1,1,\dots,1)^T$. In the adaptive computation of the weight vector, after each adaptation, the resultant weight vector will be introduced into the next adaptation as the initial weight for the next 256 adaptations. As in the analysis of Section 4.2, the reference signal here is also reconstructed by using the result of channel estimation. The estimation is performed by a correlation operation of received signal according to the desired user's spreading codes (channelization and scrambling codes). The reference signal is constructed by multiplying the estimation results with the pilot signals. Because the received signals are multi-path signals, each path performs its own beamforming

including the antenna weight adaptations and weighted processing. Beamformed signals from each path then complete the RAKE combination and decisions.

Figure 4-6 is the result of BER performance of the received signals with the symbol-rate adaptive antenna processing compared to the reception performance under the traditional RAKE reception without adaptive antenna processing. It is shown that with the symbol-rate adaptive antenna processing the BER performances are enhanced dramatically. In Figure 4-7, the beamforming patterns after each adaptation representing the gain for different incident angles are shown for consecutive 400 adaptations. It is seen that most beamforming results are very good with the main lobe pointing to the desired user while the nulls point to the interfering users. A small amount of adaptation does not have a good beamforming performance.

Figure 4-1. Block Diagram of the W-CDMA Uplink Reception Applied Adaptive Antenna Array

Figure 4.2. Details of De-Scrambling, Despreading, Channel Estimation, and RAKE

Receiver

Figure 4.3. BER Performance with Different Number of Antenna Elements

Figure 4.4. BER Performance with Different Channel Estimation Periods

Figure 4.5. BER Performances with Different Number of the Active Users

Figure 4.6. Simulated BER Performances with and without Adaptive Antenna Processing

Figure 4-7. Simulated Beamforming Performance with Adaptive Antenna Processing

CHAPTER 5

ANTENNA ARRAY PROCESSING FOR UWB SYSTEMS

UWB is to be one of the key technologies to achieve highest data rates while in parallel keeping the hardware costs limited. A major design challenge of high data rate UWB systems is the equalization of the channel impulse response consisting of, for instance, 20-70 paths with a corresponding delay spread in indoor environments of typically 20 to 100 ns.

Not only the channel estimation is a real challenge, but also the equalizer may require an unjustifiable amount of numerical calculation power, which is in principal contradiction to the main objective of low to medium cost UWB systems. A further major limitation of UWB systems is its restricted range due to the extremely low transmit power.

To overcome these difficulties and in parallel to further increase the maximum number of simultaneous users, multiple antenna or antenna array techniques seem to be a promising approach. With the high power multi users, the reception performances of the UWB systems are limited by the short spreading codes. Antenna array technique is applied to mitigate multiple users interferences.

5.1 Introduction

Since applications of the antenna array for the UWB systems are different from the conventional processing in chapter 3 and 4 (Hussain, 2002), it is necessary to find efficient methods to realize the antenna array processing for the UWB systems while considering the tradeoffs between the performances and the computational complexities. Several methods are considered to realize the antenna array processing. They are antenna selection, antenna combining, space-time processing, and wideband processing. They are shown respectively in Figure 5.1 – 5.4. The antenna selection method in Figure 5-1 assumes that there are U antenna elements in the array and each of them is separated widely enough so that the transmitted channels (from the sources/transmitters to each receiver) are independent. As a result, from the U antenna array, U replicas of the received signal can be obtained. Because the U replicas are independent, they can be thought as the U independent samples of the received signals. The performances of the U received signal are compared and then select the received signal from the antenna from which the performance is the best. The comparisons are based on the EVM that defines the average constellation error with respect to the reference signals. Because it is assumed that these U replicas are independent, the probability that received signals is affected by the fading, multi-user interferences are much less than the case when only one antenna is used for the reception. Because the best signal for the U received signal is selected and it is possible that some received signals are with the high performances while some are not good enough, it can avoid getting the signals not good enough while just keep the best signal. The worst case of this method is that all the U replicas are all not good enough

and cannot get the right reception. The probability for this case is very small. On the other hand, when the receiver only adopts one antenna, it is unable to recover the signal when the received signal suffers fading or large interference. And there are no extra resources to get any enhancement.

Another consideration for the antenna selection is for the computational complexities. It is easy to understand that the computational complexities of the antenna selection are not very large because the computational complexities are just linear increase with the number of the antenna elements. Moreover, because in this method it is no necessary to do the complex computations like those for the adaptive algorithms, it is easily to be performed and is more applicable for the very high-speed UWB systems. Under the high-speed data-rate transmissions, it is also necessary to complete the data receptions with the same high-speed. Therefore, if the complex adaptive algorithms are applied, it may bring a better performance, but they need adaptive computational progresses so that the computations are intensive and time-consuming. For instance, in Figure 5-3, it is a space-time array processing. Because it is assumed that effects of antenna array are reflected by the relative temporal shifts to the desired user, it needs to estimate the temporal shifts of all users. The estimations are performed by detecting the reception errors introduced by different temporal shifts and then adjusting delays for the correlator inside MF and RAKE by DAC (Hussain, 2002). In Figure 5-4, it is a wideband adaptive processing for the antenna array. It is based on the concept of weighting. It applies the adaptive computations for the weighting coefficient for each path from the different antenna. The entire weighting coefficients are optimized with the reference

signals. The adaptive computations are very intensive (Liberti & Rappaport, 1999). It is also difficult to be realized in hardware with the very high speed and large hardware costs. As a result, the antenna selection method is a good way to act as the tradeoff between the performances and complexities.

In Section 5.2, the received signals in the UWB systems are briefly introduced. The overall coherent receiver, including RAKE, equalizer structure, is given in Section 5.3. Parameter estimations and the details of the antenna array processing for the UWB systems are presented in Section 5.4. Experimental results are described in Section 5.5.

5.2 Received Signals

The receiver structure of the proposed UWB antenna selection scheme is also shown in Figure 5-1. From (2-22), the received signals can be further expressed as

$$\begin{aligned} y_u(t) &= \sum_{k=1}^K h_{u,k}(t) \otimes s_k(t) + n_u(t) \\ &= \sum_{m=-\infty}^{\infty} \sum_{k=1}^K b_k(m) \hat{h}_{u,k}(t) + n_u(t) \end{aligned} \quad (5-1)$$

where

$$\hat{h}_{u,k}(t) = \sum_{l=1}^L \alpha_{l,k} w(t - \tau_{l,k} - \tau_{u,k}(\theta_k)) \quad (5-2)$$

Thus, after the matched filter, together with the pulse and PN matching, the output can be expressed as:

$$x_u(t) = y_u(t) \otimes w(-t)$$

$$= \sum_{m=-\infty}^{\infty} \sum_{k=1}^K b_k(m) \tilde{h}_{u,k}(\tau_{l,k} - \tau_{u,k}(\theta_k)) + \tilde{n}_u(t) \quad (5-3)$$

where $\tilde{h}_{u,k}(t) = \int \hat{h}_{u,k}(t - \tau) w_k(-\tau) d\tau$ and $\tilde{n}_u(t) = n_u(t) \otimes w(-t)$.

5.3 Coherent Receiver Algorithms

In the UWB system, the coherent receiver needs to estimate the delay, amplitude and phase of multipath components. There exist several optimal and suboptimal coherent detectors. If the channel introduces intersymbol interference, it is not optimal to make decisions on single symbols but rather on a sequence of symbols. An optimal maximum-likelihood sequence detector compares the received sequence with all possible sequences and selects the closest one.

5.3.1 RAKE Receiver

The coherent Rake receivers are applied where a subset of the strongest taps is used in a maximum-ratio combining fashion, which maximizes the signal to noise ratio. The basic version of the Rake receiver consists of multiple correlators (fingers) and each finger can detect/extract the signal from one of the multipath components provided by the channel. The outputs of the fingers are appropriately weighted and combined to obtain the benefits of multipath diversity. The MRC is based on the assumption that the spreading code gives a low crosscorrelation between successive chips (Cassoli, Win, & Molisch, 2003). In Figure 5-1, the match filter output is sampled by $n\Delta$ where $\Delta = T_s / N$ and N is the number of the fingers of the RAKE receiver and it can be 16, 32, or 64.

$$x_u(n\Delta) = \sum_{m=-\infty}^{\infty} \sum_{k=1}^K b_k(m) \tilde{h}_{u,k}(n\Delta - mN\Delta) + \tilde{n}_u(n\Delta) \quad (5-4)$$

The RAKE receiver collects signal energy from L_A fingers with the largest absolute values. Consequently, the output of the RAKE receiver can be expressed as:

$$\bar{b}_u(m) = \sum_{l=1}^{L_A} x_{u,l}^R x_u(mN + n_l) \quad (5-5)$$

where x_l^R is the weight and n_l is the tap delay for the l -th finger.

5.4 Parameter Estimation

In the DS-UWB system, the data-aided approach that is based on using a number of known pilot symbols is applied to estimate the channel impulse response. All pilot symbols are transmitted in the beginning of the data packet, the channel is estimated and the rest of the packet is decoded based on the acquired channel characteristics. It is shown in Figure 5-5. Although it is possible to use a maximum-likelihood approach to estimate the arrival time, amplitude and phase of the channel taps, two suboptimal estimation algorithms with reduced complexity are tested. The first is denoted successive channel estimation (D'Amico, Mengali, & Morelli, 2003) and the second is the well-known sliding window algorithm. The sliding window algorithm crosscorrelates the received pilot sequence with the known pilot sequence and calculates gains and delays for all taps at the same time. The sliding window is ML-optimal for a one-tap channel or if the shifted versions of the received signal are mutually orthogonal. The successive channel estimation improves the performance of the sliding window algorithm by using an iterative algorithm. The algorithm begins by using the sliding window algorithm to

find the parameters of the strongest finger. The delayed version of the transmitted signal corresponding to the estimated finger is then subtracted from the received sequence and the algorithm is repeated. The number of iterations is equal to the number of taps to be estimated. The channel estimations are then averaged over the pilot symbols. If the number of the pilot symbols is 4 and the number of data symbols is 8, thus from (5-5), the channel estimation for the l -th path for each 12 symbols packet can be express as:

$$x_{u,l}^R = \frac{1}{4} \sum_{m=0}^3 x_u (mN + n_l + m_0) \quad (5-7)$$

Then the channel estimations remain constant after 4 pilot symbols and they are applied for the detections of the 8 data symbols. After the reception of a packet of 12 symbols, the channel estimations are performed again for the next packet.

5.5 Antenna Array for UWB Systems

As mentioned in Introduction, the conventional adaptive antenna array processing is not applicable for the UWB systems because the antenna response can not reflected on the phase of the received signals. The antenna array responds are in the term of temporal delays of the received signals. Therefore, a temporal-space processing is better for the UWB systems. However, the temporal-space processing needs complicated and intensive calculations (Liberti & Rappaport, 1999), so it is necessary to find efficient approaches.

Antenna selection is one of the diversity receptions. It is a classical and effective technique for combating deep fades experienced on wireless communications (Bahceci, Duman, & Altunbasak, 2003).

In UWB systems, antenna array is applied to combat the multi user interferences. Although these interferences are not the fades, they are assumed independent so that they are similar to the fades that are also considered independent.

The conventional antenna selection specifies that of U antenna, the one providing the largest SNR be selected for data recovery. This is analogous to choosing the largest instantaneous received power when the average noise power is the same on each antenna (Annamalai & Bhargava, 1996). For practical implementations, however, measurements of SNR may be difficult or expensive, especially for high data rates. To be most effective, the system should be able to make its selection in a period of time equal to or less than the interval of the shortest signal that will be transmitted. For this reason, the antenna with the largest composite received signal (signal-plus-interference and noise, S+I in short) is chosen.

As a result the antenna selection is based on the measurements of the S+I of the average pilot output from the U antenna. This effectively results in improved receiver performance as well as implementation advantages over the conventional selection approach. The probability of error is given by

$$\begin{aligned}
 P_e &= 1 - \sum_{u=1}^U \text{Prob} \{ \bar{b}_u(m) > 0 | b_u(m) = 1; |\bar{b}_u(m)| > |\bar{b}_i(m)|, i \neq u \} \\
 &= 1 - \sum_{u=1}^U \int_0^\infty f_{\bar{b}_u}(\bar{b}_u) \left(\prod_{i \neq u} \int_{-\bar{b}_u}^{\bar{b}_u} f_{\bar{b}_i}(\bar{b}_i) \right) d\bar{b}_u
 \end{aligned} \tag{5-8}$$

The fades and the noises are assumed as i.i.d. The average probability in (5-8) can be simplified as:

$$P_e = 1 - U \int_0^\infty [F_{\bar{b}_u}(\bar{b}_u) - F_{\bar{b}_u}(-\bar{b}_u)]^{U-1} f_{\bar{b}_u}(\bar{b}_u) d\bar{b}_u \quad (5-9)$$

where $F_{\bar{b}_u}(b) = \int_{-\infty}^b f_{\bar{b}_u}(x) dx$

Another selection method is based on the EVM. It is the measure of the difference between the (ideal) waveform and the measured waveform. EVM is calculated at the symbol rate. The signal to be analyzed consists of a packet of M symbols, each symbol consists of a certain number of samples equally distributed over the symbol interval and the number of samples representing a symbol is given by the ratio of the sampling frequency over the symbol frequency. If the sampling frequency equals 4 times the symbol frequency, then the number of samples per symbol is 4.

The instantaneous error vector is obtained by subtracting the ideal reference from the modified version of the measured waveform. The root mean square EVM is defined with the pilot symbol s_p as:

$$EVM_{RMS} = \sqrt{\frac{\sum_m |x_{u,l}^R - s_p|^2}{|s_p|^2}} \quad (5-10)$$

Therefore, if the number of pilot symbol is 4 and $s_p = 1$, $EVM_{RMS} = \sqrt{\sum_4 |x_{u,l}^R - 1|^2}$

and calculated at the beginning of each packet. According to the reference (IEEE Std 802.16, 2004), the EVM shall be measured over the continuous portion of a packet occupying at least $\frac{1}{4}$ of the total transmission frame.

5.6 Simulation of the UWB Antenna Array Processing

The simulation of the UWB antenna array processing is done by the Matlab simulink. The details for the simulation are shown in Figure 5-6. The signal is based on the previous definition for the UWB signal. There are 4 pilot symbols at the beginning of each packet and 8 data in the left part of the packet. Each symbol is spread by 15 chips PN code. And the UWB monocycle is performed by the Hermitian pulse with the period of 2×10^{-9} s. The UWB signals experiences the UWB channel. In the simulation, the IEEE 802.15.3a Channel Model (Intel, 2002) is applied. It includes the CM1, CM2, CM3, and CM4 models mentioned in Chapter 2. AWGN is added at each antenna. Multi users reception is considered here. In the simulations, 3 interference users with different spread PN codes are considered. The received signals experience channel estimations, RAKE receptions, and antenna selection. The channel estimations are completed by averaging 4 continuous outputs from the pilot symbols. The RAKE receptions are realized by the 60-finger structure that has 4 oversampled replicas in each chip. The antenna selection algorithm in the Section 5.5 is applied and based on the EVM.

Figure 5-1. Antenna Selection for the UWB Systems

Figure 5-2. Antenna Beamforming for the UWB Systems

Figure 5-3. Space-Time Processing for the UWB Systems

Figure 5-4. Wideband Adaptive Antenna Array Processing for the UWB Systems

Figure 5-6. Block Diagram of the Simualtions for UWB Antenna Array

CHAPTER 6

SUMMARY AND CONCLUSIONS

The adaptive antenna array techniques for the W-CDMA and UWB systems are proposed and analyzed in this thesis. It is an important research issue since adaptive antenna arrays have been widely adopted in wireless communications systems. Moreover, because with the fast development of the third generation and beyond 3G wireless communications, high bit-rate signals, multi users, and dense multipath environments have been taken in considerations. The receptions of wireless communications have to combat the effects from both the multi-user interferences and variable channels.

In the W-CDMA systems, most previous developed algorithms for the adaptive update of the antenna array weight factor seldom include the lattice algorithm (Boukalov & Haggman, 2000; Tsoulos, Athanasiadou, & Piechocki, 2000). With regard to applying the least-square lattice algorithm to adaptive update the weight factors of the antenna array, some modifications have been made to the structure of the conventional lattice algorithm with temporal input so that it is suitable to be utilized in antenna arrays with vector inputs. The modified lattice algorithms are still based on the Levinson-Durbin algorithm and adopt the least-square optimization criteria. Simulations prove that the convergent results of this lattice algorithm are nearly the same as Winer-Hopf (Wiener & Hopf, 1931) solutions.

In the W-CDMA system, the bit-rate is high. Therefore it is necessary to find some efficient scheme to perform the adaptive antenna array timely and accurately. Most schemes currently applied are based on the received symbols. Consequently, it needs several symbols to complete the convergence of the adaptive processing. As the result, it has to wait for a long time to obtain the weighting factors for the antenna array. However, the channels and movement of the users have already changed. In this thesis, the adaptations are performed with received signals and the reference signal both based on chip rate level over one symbol period. Bit error rate performances are analyzed under the Nakagami fading channel model. Simulations are performed under practical parameters and the M.1225 channel model. It is shown that the adaptive antenna array with RLS algorithm performed over symbol period improves the system performance of the W-CDMA uplink reception dramatically.

The adaptive antenna array processing for the UWB systems is different from that for the W-CDMA system. The transmission bit-rate of the UWB system is very high and there is no carrier for the transmissions. As a result, the convention antenna processing based on the phase variations cannot be applied for the UWB systems. Moreover, because signal rate of the UWB systems is very high, it is impossible to achieve complex adaptive computations for the antenna array processing. It is necessary to apply some efficient scheme. In this thesis, the antenna selection scheme is proposed. Its simple processing and good performance make it applicable for the antenna array processing for the UWB systems.

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