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Smart Antennas for 3G and Future Wireless Communications: Algorithms and Implementations

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School of Electrical & Electronic Engineering

A thesis submitted to the Nanyang Technological University
in fulfillment of the requirement for the degree of
Doctor of Philosophy

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Statement of Originality

I hereby certify that the work embodied in this thesis is the result of original research done by me and has not been submitted for a higher degree to any other University or Institute.

05/15/2005

Date

Sun Chen

SUN CHEN

*To
my supervisor*

*For
my beloved grandmother and parents*

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Summary

The increasing demand for spectral efficiency and new services of the mobile communications systems motivate our quest for more powerful techniques. Exploiting spatial domain with smart antennas or adaptive arrays has gained great interest among researchers. The main goal of the thesis is to investigate algorithms for interference suppression and direction finding for smart antennas. We study the state-of-art smart antenna technologies for both mobile stations (MSs) and base stations (BSs), and propose new solutions in responding to their limitations.

Various beamforming and direction-of-arrival (DOA) estimation algorithms have been designed. Most of these algorithms are designed based on the digital beamforming (DBF) array antennas. However, commercial implementations of smart antennas have been hindered by the fabrication and operation cost and high power consumption of the DBF antennas. To realize the mass implementation of the smart antennas at both BSs and MSs, especially at battery operated mobile terminals such as laptops, a beamforming algorithm and a high resolution eigenstructure-based direction finding algorithm based on single RF-port parasitic array antennas that feature low power consumption, low cost and ease of fabrication are proposed in this thesis. Experiments are carried out to justify the proposals.

The evolution of wireless communication systems also demands an increase of the system capacity. The space division multiplexing (SDM) scheme increases the channel capacity by frequency reuse. Employing beamforming algorithms, a BS array antenna forms distinct beam patterns directed towards MSs. Multiple MSs served under different beam patterns can be allocated to the same physical channel (time and frequency slot). However, this beamforming based space division multiple access (SDMA) scheme is not effective in an indoor or outdoor blocked scenario because of the difficulty in accurate DOA estimation. The inaccuracy results in nulls pointing to the desired MSs. Furthermore, large angular spread also precludes the scheme. From the antenna side, overlapping side lobes of antenna beam patterns induce multiple access interference (MAI).

To realize SDM in multipath environments, we propose the channel responses correlation coefficient as the channel duplication criterion for a null-forming based SDMA scheme. Different users are distinguished based on their spatial uniqueness (channel responses). The multipath wireless channel is modelled using a two dimensional (2D) ray-tracing model. The BS estimates the channel responses of the MSs. The study proves the increase of cell capacity with SDM in multipath environments where the conventional beamforming based SDMA scheme fails to function. The mutual coupling influence is also considered for practical performance evaluation.

Recently, antenna arrays that provide uncorrelated signal outputs are installed at both transmitters and receivers to build multiple-input multiple-output (MIMO) wireless communication channels. Transmitting space-time block codes (STBCs) provides an economic way to increase the data rate over wireless channels. It is promising to combine the beamforming and the MIMO STBC to satisfy the demands for the evolution of the wireless communication systems and bring a further improvement on the spec-

trum efficiency. Therefore, we investigate the combination of beamforming with STBC transmission. Two receiver structures that combine beamforming with Alamouti STBC decoders are proposed.

Abbreviations

1G	first-generation
2D	two dimensional
3D	three dimensional
2G	second-generation
3G	third-generation
4G	fourth-generation
ADC	analog-to-digital converter
AMPS	Advanced Mobile Phone Service
AOA	angle-of-arrival
AWGN	additive white Gaussian noise
BER	bit-error-rate
BPS	band pass filter
BPSK	binary phase-shift keying
BS	base station
CCC	cross correlation coefficient
CDMA	code division multiple access
CCI	co-channel interference
C/I	carrier to interference
CRCC	channel responses correlation coefficient

CSI	channel state information
DAC	digital-to-analog converter
DBF	digital beamforming
DCS	Digital Cellular System
DECT	Digital European Cordless Telecommunications
DOA	direction-of-arrival
DPA	dual-polarized antenna
DSP	digital signal processing
DECT	digital European cordless telephone
EDCT	Enhanced Digital Codeless Telephone
EFIE	electric-field integral equation
EGC	equal gain combining
ESPAR	electronically steerable parasitic array radiator
FDMA	frequency division multiple access
FPGA	field programable gate array
GLONASS	Global Navigation Satellite System
GNSS	global navigation satellite system
GPRS	general packet radio service
GPS	Global Positioning System
GSM	Global System for Mobile Communications
HPBW	half-power beamwidth
I	in-phase
IC	interference cancellation
IF	intermediate frequency
IMS	international mobile telecommunication
INR	interference-to-noise-ratio
IS	interim standard

LMS	least mean square
LNA	low noise amplifier
LOB	line-of-bearing
LOS	line-of-sight
MAC	media access control
MAI	multiple access interference
MBF	microwave beamforming
MCCC	maximum cross correlation coefficient
MIMO	multiple-input multiple-output
ML	maximum likelihood
MMIC	monolithic microwave integrated circuit
MMSE	minimum mean square error
MRC	maximal ratio combining
MS	mobile station
MSE	mean square error
MU	multiuser
MUSIC	multiple signal classification
NEC	Numerical Electromagnetic Code
NMSE	normalized mean square error
NTT	Nippon Telephone and Telegraph
OFDM	orthogonal frequency division multiplexing
PCS	personal communication services
PDC	Personal Digital Cellular
pdf	probability density function
PIFA	planar-inverted F antennas
PL	position location
RB	reversely biased

Q	quadrature
QoS	quality of service
QPSK	quadrature phase-shift keying
RB	reversely biased
RF	radio-frequency
SARLPA	switched and reactively loaded parasitic array
SC	switching combining
SCC	spatial correlation coefficient
SDMA	space division multiple access
SGA	steepest gradient algorithm
SINR	signal-to-interference-plus-noise ratio
SIR	signal-to-interference ratio
SNR	signal-to-noise ratio
SPSA	simultaneous perturbation stochastic approximation
SSC	switch and stay combining
STBC	space-time block code
STC	space-time code
SU	single user
TDMA	time division multiple access
TDOA	time-difference-of-arrival
UMTS	universal mobile telecommunications system
VBLAST	vertical Bell Lab layered space time
WLAN	wireless local-area networks
XPD	cross-polarization discriminaiton
ZF	zero-forcing

Symbols

A	a constant
$a(n)$	updating step size at n th iteration stage
a	a constant
$ \cdot $	absolute value
α	a constant
$\alpha(\phi_k)$	array factor of k th signal source with impinging azimuthal angle ϕ_k
$\tilde{\alpha}(\phi_k)$	$\mathbf{w}_{ESP}^T \alpha(\phi_k)$
$\tilde{\alpha}_m(\phi_k)$	$\tilde{\alpha} \left(\phi_k - \frac{2\pi}{36}(m-1) \right)$
$\bar{\alpha}(\phi_k)$	$\left[\tilde{\alpha}_1(\phi_k) \quad \tilde{\alpha}_2(\phi_k) \quad \dots \quad \tilde{\alpha}_6(\phi_k) \right]^T$
	steering vector in angular domain MUSIC
$\underline{\alpha}(\phi)$	steering vector in reactance domain MUSIC
$\beta_k^{(m)}$	propagation delay of the waves from MS- k
	$m = 0$: direct path; $m = 1, 2$ reflected paths
C	a constant
C_M	achieved MIMO channel capacity
C_Γ	coverage capacity given a channel duplication threshold Γ
$c(n)$	perturbation size at the n th iteration stage
d	array antenna inter-element spacing

$\text{Det}(\mathbf{A})$	determinant of matrix $\mathbf{A}$
D	number of array elements
$\text{diag}[\cdot]$	diagonal matrix
$\mathbf{e}_i$	eigenvector
ϵ	randomness and the inevitable error in observing the loss function
E_s	symbol energy
f_c	carrier frequency
g_0	array amplitude gain
$\mathbf{g}_n$	gradient of the objective function at the n th iteration stage
$(\cdot)^H$	conjugate-transpose
γ	a constant
γ_b	output SINR at a beamformer
Γ	channel duplication threshold
$\mathbf{H}$	MIMO channel
$\mathbf{H}(:, 1)$	the first column of $\mathbf{H}$
$\mathbf{H}(:, i - j)$	the i th to the j th column of $\mathbf{H}$
h_{ij}	channel from the j th transmit antenna to the i th receive antenna
$\mathbf{i}$	currents on monopoles
$\mathbf{I}_d$	an $d \times d$ dimensional identity matrix
$i_m, m = 1, \dots, 6$	current on the m th monopole
j	$\sqrt[3]{-1}$
k	index of signal sources
K	the number of total signal sources or mobile stations
$\parallel$	Ricean- K factor
L	length of the monopole
$L(\cdot)$	objective function
$\widehat{L_n^{(\pm)}}$	estimated objection function at n th iteration stage with perturbation

λ	wavelength
λ_{max}	maximum eigenvalue
λ_i	eigenvalues
M	the number of parasitic array subsets
μ	step size of the adaptation
N	number of total iterations
$n_i(t)$	noise at an array antenna the i th element or the i th sampling period
$\mathbf{n}$	noise sample vector
ν	a constant
$\propto$	proportional to
P_T	transmit power
ϕ	azimuthal angle
$\phi_k^{(m)}$	DOA of impinging waves from MS- k . $m = 0$: direct path; $m = 1, 2$: two reflected paths
R	radius
r	reference signals
R_X	the number of receive antennas
$\mathbf{R}_s$	desired signal correlation matrix
$\mathbf{R}_{ni}$	interference plus noise correlation matrix
$\mathbf{R}_{uu}$	correlation of $\mathbf{u}$
$\mathbf{R}_{ru}$	cross-correlation matrix between $\mathbf{r}$ and $\mathbf{u}$
$\text{span}[e_1, e_2, \dots, e_M]$	linear space spanned by $e_1, e_2, \dots, e_M$
S_{11}	return loss
$s_k(t)$	baseband signal
σ^2	noise power
$(\cdot)^*$	complex-conjugate
P_{MU}	MUSIC power spectrum

p_{sig}	signal power
π	3.1415926535
$(\cdot)^T$	transpose
$T_r(\cdot)$	trace of a matrix
T_X	the number of transmit antennas
Δ_n	perturbation direction at the n th iteration stage
θ	elevation angle
U_1	$\begin{bmatrix} 1 & 0 & 0 & 0 & 0 & 0 & 0 \end{bmatrix}^T$
$u_i(t), u_i$	signal at an array antenna i th element
$\mathbf{u}$	signal sample vector at an array antenna
ε	error between the desired signal and the reference signal
$\mathbf{V}_N$	noise subspace
$\mathbf{v}$	voltages on monopoles
v_s	transmitted voltage at the RF port
v_0	voltage signal at the center monopole
$v_m, m = 1, \dots, 6$	voltage signal at the m th monopole
$V_{cc_i}, i = 1, \dots, 6$	digital control voltage on the m th tunable reactance
$\mathbf{w}_k$	beamforming weight vector for the k th user
w_i	beamforming weight at the array i th element
$\mathbf{w}_{opt}$	optimum beamforming weight vector
$\mathbf{w}_{ESP}$	ESPAR equivalent weight vector
$\mathbf{W}_{ESP}$	ESPAR equivalent weight matrix
$\widehat{(\cdot)}$	denotes the estimated value
$\overline{(\cdot)}$	fixed component
$\widetilde{(\cdot)}$	diffuse component
$X(n)$	reactance values $(x_1, x_2, \dots, x_6)$ at the n -th iteration stage

$x_m, m = 1, \dots, 6$	reactance loaded at the m th parasitic monopole
$\mathbf{X}_L$	impedance matrix each diagonal entity being the impedance
y	array output
$\mathbf{Y}_c$	mutual admittance matrix
y_{ij}	mutual admittance between the i th and j th monopoles
$y_{far}(\phi)$	far field signal at azimuthal direction ϕ
ζ	distance space
$(\cdot)^{-1}$	matrix inverse
$E[\cdot]$	expected value operator

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Chapter 1

Introduction

1.1 Motivation

Since James Clerk Maxwell (1831-1879) predicted the existence of electromagnetic waves in 1867 and Guglielmo Marconi (1847-1937) sent first transatlantic wireless signals in 1901, global wireless communications have become a reality. Over the past 137 years, the wireless communications have evolved from first-generation mobile communication systems in 1970s, via second-generation (2G) systems in the late 20th century, to the third-generation (3G) systems. While the 3G system begins to take its role in the 21st century, technologies have ushered in the fourth-generation (4G) wireless communications era [1, 2].

The 3G systems promise high data rate (up to 2 Mbits/s) multimedia services [1-8]. New standards are under considerations. At the same time, the population of wireless subscribers is growing amazingly to a universal extent. It is estimated that the worldwide cellular subscribers will grow from 1.1 billion in 2002 to 2.0 billion in 2006

1.1. Motivation

2

at 13.6 percent rate [9]. The wireless system is so ubiquitous that it is shaping the way we communicate. As shown in Fig. 1.1, the ubiquitous wireless networks, which include satellite communications, high speed data rate transmissions, position location (PL) services, intelligent transportation system, and ad hoc networks, will give wireless access to anyone at anytime anywhere.

Figure 1.1: Upcoming wireless communication networks.

The emergence of new-generation wireless services confronts engineers with mainly three challenging demands [10]: 1) increasing the system capacity to accommodate more mobile subscribers; 2) developing new technologies to include value-added services to the current wireless communications services; and 3) improving the quality of services (QoS) over the wireless channels, whose capacity is limited by severe multipath fading, and realizing higher data rate wireless transmissions.

So far efforts have been focused on developing the coding, the modulation and protocols on the 2G systems [7, 10–14]. These efforts are not sufficient to meet the demands for the evolution of new generation value-added multimedia services and the exponential growth of the number of wireless subscribers. Recently, smart antennas or adaptive array antennas cast light in the tunnel to meet both demands. Smart antennas can adaptively control main beams and nulls and offer diversity freedom to mitigate multipath fading, hence increasing the channel capacity and improving the QoS [6, 7, 15–25].

The smart antennas can be classified, according to their structures and mechanism, into five categories: 1) switched beam antennas [12, 15, 17, 21, 23, 26, 27]; 2) phased array antennas [15, 26–30]; 3) digital beamforming (DBF) array antennas [15, 18, 20–22, 26, 27, 30, 31]; 4) parasitic array antennas [30, 32–34]; and diversity antennas [16, 17, 19]. Over the past few years, switched beam antennas and phased array antennas have played important roles in satellite communications and military scenarios [12]. In comparison to switched beam antennas and phased array antennas, a better beam pattern direction tuning ability can be achieved with DBF array antennas that have high speed digital signal processing (DSP) chips and high resolution analog to digital converters (ADCs) [27].

Various beamforming and direction-of-arrival (DOA) or angle-of-arrival (AOA) estimation algorithms have been designed for smart antennas [5, 7, 8, 18, 24, 27, 31, 35–38]. The simulation and experiments carried out by many researchers have demonstrated the abilities of these algorithms [18, 39–43]. Most of these algorithms are designed based on the DBF array antennas. Signals received by individual antenna elements are down-converted into baseband signals. These signals are digitized and fed into DSP chips where the algorithms reside in. However, radio-frequency (RF) circuit branches connected to the array elements, ADCs and the baseband DSP chip consume a considerable amount of dc power. Furthermore, each channel connected to the array sensor

has the same structure, so the fabrication cost increases with the number of array elements [30]. Thanks to the recent advancement of GaAs monolithic microwave integrated circuit (MMIC) technologies, the beamformer can be integrated into a single chip at the RF front end (microwave beamforming (MBF)) [44], instead of the baseband. The advantages are the reduced quantization errors and the increased dynamic range. However, their high fabrication cost still limit the range of implementations. All these factors make DBF and MBF antennas unsuitable for low power consumption and low cost systems and hinder the mass applications of the smart antenna technologies. *For example, it could be too costly to equip DBF antenna arrays at battery powered lap-tops or mobile computing terminals within a wireless network.*

Parasitic array antennas circumvent the DBF and MBF deployment problems. It is a reactively controlled directive array antenna [45, 46], in which one central element connected to the sole RF port and a number of surrounding parasitic elements form the array. Beam steering is achieved by switching ON and OFF parasitic elements or by adjusting the load reactances at parasitic elements surrounding the central active element. The switching circuitry and the load reactances are realized by reversely biased (RB) varactor diodes. Because there is only a small RB leakage current, the power consumption of the parasitic array antennas is comparatively small. Furthermore, the system has only one RF port and one consequent down-conversion circuit. From the viewpoint of power consumption and fabrication cost, it is obvious that the parasitic array antennas are suitable for mass implementations of smart antennas, especially for battery powered portable mobile terminals. *But the reported beamforming algorithms [47–52] are still not efficient enough for real time implementations in the wireless communications systems, where a fast beam tuning ability is required.*

Smart antennas also have applications at mobile stations. Employed at mobile terminals

(notebooks, PDAs *etc.*) in ad hoc networks or wireless local-area networks (WLAN) the direction finding ability avails the design of the packet routing protocol [53, 54], which decides the manner of the packet relaying. However, installing DBF at each mobile terminal could be an expensive approach in terms of the power consumption of the DBF antennas, the battery life of the mobile terminals, and the fabrication cost of the antennas and digital signal processing units. *To make the aforementioned single-port smart antennas a suitable candidate for mobile terminals at WLAN, high-resolution direction finding techniques should be implemented.*

The evolution of wireless communication systems also demands for an increase of the system capacity [10]. DBF array antennas have extensive applications at base stations (BSs) in cellular mobile communications. Installing DBF array antennas, the space division multiple access (SDMA) scheme increases the channel capacity through efficient frequency reuse [31, 42]. With beamforming algorithms, a BS array antenna forms distinct beam patterns directed to mobile stations (MSs) at each physical channel (time and frequency slot). Multiple MSs served under different beam patterns could be allocated to the same physical channel. However, beamforming based SDMA is not effective in an indoor and outdoor blocked scenarios because of the difficulty in accurate DOA estimation. The inaccuracy results in nulls pointing to the desired MSs. Furthermore, large angular spread also precludes the scheme. From the antenna side, overlapping side lobes of antenna beam patterns induce multiple access interference (MAI). *New methods should be employed to realize SDMA in multipath rich environments.*

Recently, diversity antennas that provide uncorrelated signal outputs are installed at both transmitters and receivers to build multiple-input multiple-output (MIMO) wireless communication channels [55–57]. The multipath phenomenon achieves a channel

capacity that increases with the number of array antenna elements [56]. Transmitting space-time codes (STCs) over a MIMO system dramatically increases the data rate over wireless channels [55, 58]. It is promising to combine the interference suppression ability of the beamforming algorithms with the STC transmission. *With interference suppression techniques, multiuser MIMO communications can be realized to achieve a higher system capacity.*

From the above discussion, it is understood that although tremendous developmental works have been advancing with smart antennas to develop new algorithm for low powered parasitic arrays, capacity improvement in SDMA scheme, DOA estimation and high data rate MIMO transmissions, there are still plenty of room for improvement in these topics. The thesis addresses these problems and develops new techniques to improve in these areas. This is the motivation of the PhD thesis.

1.2 Objectives

In responding to the aforementioned open problems, the goal of this thesis is to investigate algorithms and implementations of both analog (parasitic array antennas) and digital (DBF array antennas) smart antennas for higher data rate and better link quality wireless communications and direction based services. To fulfil the goal, the following tasks have been envisaged.

1. We begin our study with a comprehensive literature review of the applications, structures and algorithms of smart antennas. This exercise brings a new classification of smart antennas based on their implementation architecture and creates a strong foundation and background for further exploration.

2. Parasitic array antennas are found to be a potential candidate for low battery powered mobile ad hoc networks and similar applications such as automated/intelligent golf court trolley cars, shopping carts, robotic dogs, *etc.* we begin our study with developing efficient beamforming and direction finding algorithms for parasitic array antennas.
3. Satisfied with the developed fast beamforming algorithm, we proceed to solve some fundamental issues of DOA estimation problems using parasitic array antennas.
4. We also extend the study on smart antennas from parasitic array to DBF antennas. The multipath propagation phenomenon is detrimental to the performance of a conventional beamforming based SDMA scheme. In respond to that problem, we develop a new technique of null forming based SDMA for multipath environment applications.
5. In the context of MIMO communications, the meaning of “beamforming” has been extended to include diversity combining in recent years. To conjoin the interference suppression ability of beamforming techniques with the high data rate transmission supported by MIMO wireless systems, we combine beamforming algorithms with STC transmissions and examine the improved bit-error-rate (BER) performance and the achieved MIMO channel capacity.

The integrity of various topics in this study is illustrated in Fig. 1.2. The outcomes of the PhD study are outlined in the next section.

Figure 1.2: Topics covered in the thesis to achieve the goal of the study.

1.3 Major Contributions of the Thesis

The major contributions of the thesis are listed as follow:

- We have carried out a thorough literature review on the current development and the trend of smart antenna technologies, through which we conclude that smart antennas are playing important roles in the migration to the future wireless communications systems. A novel classification of smart antennas based on their algorithms, hardware and implementations was presented. The smart antennas are categorized into switched beam antennas, phased array antennas, DBF array antennas, parasitic array antennas and diversity array antennas. Among them, the parasitic array antennas are strong candidates for realizing commercial implementations at power consumption limited mobile terminals. Whereas DBF array antennas are suitable for applications at cellular BSs where the main limiting factor is the performance rather than the cost and the power consumption. Diversity array antennas have been implemented for many years to combat multipath fading. Recently, diversity array antennas are installed at both transmitters and receivers to build up MIMO communication systems. Transmitting STCs through the MIMO systems dramatically increases the data rate over wireless channels.
- Designed a novel adaptive beamforming algorithm for a low power consumption reactively controlled directive array antenna called as “electronically steerable parasitic array radiator (ESPAR)” antenna. It is an analog beamformer. The beamforming of an ESPAR antenna is based on strong mutual coupling among neighboring array elements. This is different from that for a DBF antenna array where the optimum beamforming weight is obtained by solving a linear equation. An adaptive beamforming algorithm based on simultaneous perturbation stochastic approximation theory was proposed.

- To justify the beamforming ability of the reactively controlled directive array antennas and test the convergency of the proposed adaptive beamforming algorithm, experiments was carried out during my internship at Advanced Telecommunications Research Laboratories International (ATR), Japan. The results show that the algorithm converges within 5000 training bits. Beam patterns can be formed adaptively to suppress interferences. The proposed adaptive beamforming algorithm makes the parasitic array antenna a strong candidate for WLAN applications.
- Proposed a novel parasitic array antenna structure, which has only one RF port. Being different from ESPAR antennas, the surrounding parasitic elements are not only reactively loaded but controlled by switches. We call the antenna "switched and reactively loaded parasitic array (SARLPA)". Since the system has only one RF port connected to the active central element and one subsequent down conversion circuit, it obviously consumes much less power than a DBF array antenna. A high-resolution direction finding using multiple signal classification (MUSIC) algorithm with periodic signals is employed for incoherent waves. Simulation justifies the proposal. Our heuristic study demonstrates the feasibility of employing eigenstructure-based direction-finding algorithms on a parasitic array antenna that has only one RF port.
- Proposed the channel responses correlation coefficient as channel duplication criterion for a null-forming based SDMA scheme. Different users are distinguished based on their spatial uniqueness (channel responses) at the BS antenna array. The multipath wireless channel is modelled using a two dimensional (2D) ray-tracing method. The study proves the increase of cell capacity with space division duplexing (SDM) in multipath environment where the conventional DOA or AOA assisted beamforming based SDMA scheme fails to function.

- Proposed two receiver structures that combine optimum receive beamforming with space-time block code (STBC) decoders for MIMO communications in the presence of interferences. Applied to the Alamouti STBC, the beamforming is combined with the conventional Alamouti decoder and the minimum mean square error (MMSE) decoder. The beamforming process maximizes the signal-to-interference-plus-noise ratios (SINRs) before the STBC decoders. The STBC decoding processes are modified to adopt beamforming. The properties of Alamouti STBC are preserved after the optimum beamforming to support higher data rate transmissions. We also found that the beamforming algorithm is environment-oriented, in that a directive pattern is formed to the direction of the impinging wave in a high Riccan- K factor.

We begin our study with a thorough literature review. Based on the understanding of the existing smart antenna technologies and the latest development trend, we investigate and test various beamforming and direction finding algorithms for parasitic array antennas, DBF array antennas and diversity array antennas, through which a considerable amount of original contributions have been produced.

1.4 Organization of the Thesis

The remainder of the thesis is organized as follows. Chapter 2 presents a literature review of the state-of-art smart antenna technologies. Firstly, both current and future applications of smart antennas are described. Then, we examine various implementations of smart antennas, including switched beam antennas, phased array antennas, DBF array antennas, parasitic array antennas and diversity antennas. Finally, some beamforming and direction finding algorithms are introduced.

We begin our study with investigating adaptive beamforming algorithms for parasitic array antennas. In Chapter 3, the theory of a reactively controlled directive array antenna is given. After that, a beamforming algorithm based on simultaneous perturbation stochastic approximation theory is proposed. Quick convergence of the algorithm substantiates the validity of the proposed algorithm.

Experiments are also carried out to probe into the beamforming ability of the reactively controlled directive array. In Chapter 4, the experiment testbed is briefly described. Experimental results are presented.

The direction finding ability of the parasitic array antenna is presented in Chapter 5. Two modified MUSIC algorithms, angular-domain MUSIC and reactance-domain MUSIC, using single RF port parasitic array antennas are presented. A calibration method is also described. Experimental evaluations of the DOA estimation techniques are also carried out in an anechoic.

Chapter 6 investigates the SDMA scheme in a multipath rich environment. Instead of using the angular separation, the channel responses correlation coefficient is proposed as the channel allocation criterion. The influence of mutual coupling is also considered.

Chapter 7 presents the two receiver structures that combine beamforming with the conventional and the MMSE Alamouti STBC decoders, respectively. The BER performance and the achieved MIMO channel capacities with different numbers of receive antennas are investigated. Furthermore, the performances of the two combinations are compared.

1.4. Organization of the Thesis

Finally, Chapter 8 concludes the work and provides recommendations for further research.

Chapter 2

Overview of Smart Antenna Technologies for Mobile Communications

2.1 Introduction

Over the decade, demands for better quality and new value-added services on the existing wireless communication infrastructures have risen beyond all expectations. It is estimated that half a billion handsets will be put into the context of the third generation system (UMTS, IMT 2000), which will provide an up to 2 MB bandwidth for each user [1]. The most challenging demands are: the need to increase the spectrum efficiency and the system capacity of the current wireless networks. These demands have brought technological challenges to service providers. Due to the ability of suppressing interference, combating against fading and providing new services, the smart antennas have become one of the key technologies to realize 3G and even 4G wireless

communications.

In this chapter, we start with an overview of the evolution of wireless communication systems. The evolution calls for new technologies to support new services and the improved quality. The smart antennas turn out to be a potential technology. In order to give an overview of the current development of smart antenna technologies, we first present their applications in wireless communications. After that, we compare different architectures of smart antennas and their suitability to various applications. A new classification of smart antennas is given. Finally, some of the algorithms are presented.

2.2 Evolution of Mobile Communications

A short article in the January 23, 1897 issue *Scientific American*, *Telegraphy Without Wires*, noted that “a young Italian, a Mr. Marcon” had recently demonstrated to the London post office the ability to transmit radio signals across three-quarters of a mile (one kilometer), and that “if the invention was what he believed it to be, our mariners would have been given a new sense and a new friend which would make navigation infinitely easier and safer than it now was” [59]. Intrigued by that, scientist and engineers had been investigating new wireless communications methods and services. The era of wireless communications begun.

Since then, wireless communications have experienced a phenomenal technological advancement manifested in various application scenarios including wireless remote control, codeless telephony, paging, wireless networks, terrestrial cellular telephone systems, and satellite communications [8, 10]. Among them, the cellular telephone systems have been developing from the first-generation (1G) systems deployed firstly by Nippon Telephone

and Telegraph (NTT) [10]; through the 2G systems including the Global System for Mobile Communications (GSM) and Digital Cellular System 1800 (DCS 1800) in Europe, Personal Communication System 1900 (PCS 1900), interim standard-95 (IS-95) and IS-136 in north America and Personal Digital Cellular (PDC) in Japan [8]; to the 3G system (W-CDMA, CDMA 2002) [7], and will approach the 4G systems in the near future [2].

The 3G systems are universal, multi-functional, globally compatible digital mobile radio system in that they allow users to communicate anyone anywhere, at anytime. The system integrates paging, cordless, cellular, wireless network and satellite communication systems into a single mobile terminal. The user can use the 3G terminal for conventional voice telephony, broadband access to the Internet, high speed packet data services and position tracking [1–8, 60, 61].

Some main features of 3G communications systems are:

- *High data rate communications:* The minimum requirements of the user data rates in different environments are: 144 kb/s for vehicular applications; 384 kb/s for pedestrian; 2 Mb/s for indoor office; 9.6 kb/s for satellite communications. These goals require an increased frequency efficiency that is measured by the bits/s/Hz.
- *New services:* These services include: intelligent transportation, PL services, ad hoc networks, *etc.* These location-aware services require new techniques to provide position information.
- *More subscribers:* The mobile communications will be so ubiquitous that more subscribers or users are accommodated. This feature demands a dramatic increase in the mobile radio system capacity measured by the number of users within a

given frequency spectrum.

During the migration from the 2G system to the 3G system, great efforts have been directed toward the development of the modulation, coding, and protocols for the 2G systems, such as the code division multiple access (CDMA) with IS-95, the time division multiple access (TDMA) with IS-136 and the GSM. These techniques have been exhausted to the limit due to the large number of subscribers within the limited channel capacity. Antennas have gained much interest among researchers. The smart antenna systems, in the form of adaptive arrays, can mitigate three major impairments caused by wireless channels: *fading*, *delay spread*, and *co-channel interferences*. It can also bring many benefits to the existing wireless communication systems, such as the enhancement of coverage for portable mobile applications, an improved signal-to-interference ratio (SIR) and an increased channel capacity [17, 21], a lower transmitted power, and a higher data rate. Employing smart antennas into current wireless communication infrastructures also brings value-added services, such as General Packet Radio Service (GPRS), users' PL and the intelligent transportation system [17, 62].

2.3 Current Applications of Smart Antennas

Along with the development of smart antenna technologies, the applications of smart antennas have extended from the early forms such as diversity and sectoring to the recent employment such as SDMA [31] and MIMO communications [58]. In this section, we give a brief introduction to some applications of smart antennas in the wireless mobile communications. Figure 2.1 illustrates the various applications of smart antennas. There are three major applications of smart antennas. They are direction finding, beamforming and diversity combining. The direction finding is critical to the realiza-

Figure 2.1: Applications of smart antennas.

tion of position location services, law enforcement, ad hoc networks *etc.* Beamforming has been viewed as an efficient technique for interference suppression. It has applications such as SDMA, sectoring and ad hoc networks. Whereas diversity techniques are utilized for combating against fading. Recently, diversity has been exploited to realize MIMO communications. In the following sections, we will describe these smart antenna applications.

2.3.1 Diversity Scheme

The diversity scheme has found many applications over the years and has been in wide use in a variety of present-day microwave systems [26, 63–66]. The aim of this scheme is to develop information from several independent signal paths to combination the multiple signals and reduction the effect of excessively high signal fade [67]. Because the probability of two independent signals having a deep fade at the same time is rare, combining them reduces the effect of the fades.

There are many mechanisms to provide diversity. They are space diversity (with spaced

array elements), frequency diversity (multicarrier CDMA, orthogonal frequency division multiplexing (OFDM)), AOA diversity (impinging waves with different directions), polarization diversity (transmitting signals with different polarized waves), time diversity (signal-repetition) and multipath diversity (Rake receiver technique) [63]. Among them, the space diversity is the easiest one to be implemented in that it only needs an array antenna with sufficient inter-element spacing. However, in considering the limited space for installing antennas at a mobile phone, the polarization diversity provides a compact installation [68–72].

There are two kinds of diversity schemes, one is the macroscopic diversity scheme; and another is the microscopic diversity scheme. A macroscopic diversity scheme is used to combat against random shadowing effects, where the signal power may drop to a point lower than the value predicted by a long-distance path loss model. The undulation of terrain contours determines the duration of the envelope variations during a long-term fading. The diversity scheme is employed by selecting signals from two separated base stations. On the other hand, in the microscopic cells, clusters surround mobile users. Time dispersion caused by multipath propagation delays and the random frequency modulation due to varying Doppler shifts on different multiple signals result in fluctuations of the signal power at both base stations and mobile terminals. This is commonly known as the Rayleigh fading (See Figure 2.2) [67]. Fig. 2.3 shows the envelope correlation of two signal samples at distance space ζ , where λ is wavelength of the carrier signal [66]. The model assumes that the DOAs of multipath signals are uniformly distributed over 360° . A spacing of 0.13λ achieves a correlation lower than 0.7. As for the aforementioned space diversity, array antennas with elements spacing of 0.38 wavelengths for a MS or tens of wavelength for a BS could receive independent signals at separate elements [64]. Thus, the microscopic diversity can be employed.

Figure 2.2: Rayleigh fading envelope [64].

Figure 2.3: Envelope correlation spacing in a 2-dimensional Clarke scenario [64].

The uplink performance improvement with diversity was proved by the testbed built by the international team of Ericsson at Stockholm for study of smart antenna receiver performance for digital Advanced Mobile Phone Service (AMPS) [73, 74]. The uplink receiver uses space and polarization diversities. The inter-elements spacing is 15λ . The system provides a 3.5 dB gain in signal-to-noise ratio (SNR).

2.3.2 Sectoring

The ultimate goal of implementing a wireless cellular system is to achieve high capacity while satisfying QoS expectation. Cell splitting is a simple way to reduce co-channel interference (CCI). And, currently most existing cellular and personal communication services (PCS) systems [8] make use of the sectoring technique in addition to the spatial diversity. Each cell is divided into radial sectors with wide-beam directional antennas installed at the BS [7, 8, 10, 26, 75]. Typically, those cells are divided into three sectors, which are 120° wide or six sectors that are 60° wide. Let us consider a cellular system with cluster size of seven. The 120° cell sectoring reduces the number of the first-tier CCIs from six to two and brings a 6.8 dB SIR gain [10]. In addition to CCI suppression, another attractive benefit of sectoring lies in that frequency resources could be reused more frequently, thus each cell could accommodate a larger traffic density without adding extra base stations.

Higher-order sectorization can be employed via beamforming or employing directional panel antennas [75] to increase the frequency reuse and thus further improve the system capacity. Figure 2.4 illustrates an 18-sector site, which is implemented by three multibeam antennas. Sectoring via adaptive beamforming even allows more flexible allocation of transmit power at BS [75]. However, the handover will be initiated more frequently. Therefore, the system complexity is also increased.

Figure 2.4: Example of an 18-sector site implemented using beamforming with three uniform linear antenna arrays.

2.3.3 Interference Suppression with Adaptive Beamforming

Adaptive antennas date back to 1959. The original work was attributed to L. C. Van Atta. (Electromagnetic Reflection, U.S. Patent 29080002, October 6, 1959). Since then, adaptive beamforming techniques have been employed to remove unwanted noise and jamming from the output, mainly in military applications [7]. With the thriving commercial wireless communication industry and the advancing microprocessor technologies, the adaptive beamforming techniques have found their applications in commercial wireless communications. With powerful DSP hardware at the base band, algorithms could control antenna beam patterns adaptively to the real signal environment, forming beams towards the desired signals while forming nulls to co-channel interferers. Thus, the system performance is optimized in terms of link quality and system capacity.

Field trials of both multibeam antennas and adaptive arrays have demonstrated the performance improvements. Promising results substantiate the potential benefits of the adaptive beamforming technologies [16, 18, 39–43, 76]. Metawave[©] has extensively studied the range increase of multibeam antennas [76]. Employing a 24-beam 15° beamwidth system, switching two strongest directional beams achieves a 5 dB gain in a typical cellular environment. An increased interference tolerance of 9 dB in the IS-136 system with a four-element adaptive antenna is demonstrated by Ericsson [18, 39–41]. Lucent/AT&T has demonstrated the operation with an equal-power interferer next to the desired mobile terminal several miles from the base station in an IS-136 system with a four-element adaptive antenna [41]. Field trials have also been done for Digital European Cordless Telecommunications (DECT) system under the European TSUNAMI project [42]. Commercial products include a four-beam smart antenna incorporated into a GSM base station product by Nortel, and adaptive array processing using two base station antennas incorporated into an IS-136 base station product by Ericsson [41, 43].

However, the extensive implementation of adaptive antennas has been a problem. Due to the system complexity, the fabrication cost and the operational expenditure, adaptive antennas are limited to base stations or military applications. Researchers are now endeavoring to make low profile and low power consumption antennas for mobile terminals. Recently, parasitic array antennas have become an active research topic [32]. The antennas normally have one active element connected to the RF port, and beam pattern adaptation is achieved by switching among parasitic elements or by adjusting the loaded reactance at the passive elements, as we will show in the following sections. Being a less complicated structure, it offers an economical way to design low power consumption smart antennas.

2.3.4 Space Division Multiple Access

Filtering in the space domain separates spectrally and temporally overlapping signals from multiple MSs. Thus, the spatial dimension can be exploited as a hybrid multiple access technique complementing existing multiple access schemes such as frequency division multiple access (FDMA), time division multiple access (TDMA), and code division multiple access (CDMA). This approach is usually referred to as SDMA. The realization of this filtering technique is accomplished by using a smart antenna, which is capable of modifying its time, frequency, and spatial responses by means of the amplitude and phase weighting and the internal feedback control (See Fig. 2.5). SDMA scheme enables multiple users within the same radio cell to be accommodated on the same frequency or time slot. Thus, it will increase the system capacity significantly [17, 31, 77–81]. The concept of the SDMA scheme is demonstrated by TSUNAMI project. A uniform linear array consists of eight elements with a beamforming card developed by ERA Technology [82] is used at the BS. The MUSIC algorithm as will be described in the following sections is used to find the direction of the MS. Kalman filter is used to track the MS [42]. Figure 2.5 illustrates such an SDMA scheme with spot beams at the BS.

In addition to applications at terrestrial cellular systems, the SDMA scheme has also been employed in satellite communications where each spot beam services a geographical area [31]. For L-band mobile satellite communication system, to increase the frequency reuse in the spatial domain, multiple beam satellite antennas are used for the coverage of Canada and the United States. In comparison with a single beam antenna, the frequency reuse is increased by a factor of 4/3 using a four-beam antenna and of 12/3 by a 12-beam antenna [31]. In Japan, a six-beam satellite antenna is used for 22 GHz band satellite broadcasting [83].

Figure 2.5: Space division multiplexing with spot beams at the BS.

2.4 Incoming Applications of Smart antennas

In addition to the aforementioned applications, new applications will be incorporated into the coming wireless communications standards of the 3G and 4G systems. Extensive investigations have been carried out to bring these new applications into reality.

2.4.1 Ad-Hoc Networks

An ad hoc network is a dynamic wireless network established by peer-to-peer communications among mobile nodes (e.g. laptops, PDAs) without the help of any infrastructure such as a wire/wireless backbone [6, 53, 54, 84–86]. The schematic concept is shown in Fig. 2.6, in which each mobile terminal directs its beam towards the neighboring node to form a mesh where each node is a potential router. Changing together a sequence of neighboring nodes dynamically changes the routing of an information packet from an

Figure 2.6: Ad hoc networks with directional antenna at each mobile terminal.

information source to a destination. Researchers are investigating media access control (MAC) protocols of ad hoc networks. Smart antennas are a critical part, as many MAC protocol designs include the antenna beam pattern control [53, 54]. With a properly oriented beam (towards the desired node), each node can resolve signals from neighboring nodes. The idea is the same as for an SDMA system. The DOA information from the array antenna is used in routing selection. Another important issue of the ad hoc network is the power consumption. Because mobile terminals for an ad hoc network have a limited battery life (around 3-4 hours for a laptop), so smart antennas for this application are also required to have easy beam pattern control and consume as less power as possible.

2.4.2 Position Location Based Services

Safety is the primary motivation for Position Location (PL) services. In the United States, Federal Communications Commission has mandated PL services within 125 meters accuracy on wireless emergency calls [17,26,87]. Using adaptive antennas, direction finding systems provide a mobile user's position information. This location information aids the emergency rescue, law enforcement, emergency calls and so on. With the position location information, other promising applications such as accident reporting, navigational services, automated billing, fraud detection, roadside assistances, and cargo tracking are also coming into reality.

Currently, there are many existing PL systems: the Global Positioning System (GPS), the Global Navigation Satellite System (GLONASS), the signpost navigation and the cellular geolocation [17]. Among them, the cellular geolocation is the most attractive one. It only needs to implement the smart antenna system, which has a DOA estimation ability, into the existing wireless communication systems. The realization is mainly based on three mechanisms [87]: beacon location methods, DOA estimation techniques, and time-difference-of-arrival (TDOA) techniques.

The concept of geolocation technique is shown in Fig. 2.7. The BSs located near the MSs estimate the DOA of the impinging wave from the terminal. From the DOA estimate, a line of bearing (LOB) from the base station to the mobile transmitter can be drawn. Multiple LOBs, drawn from different base station locations, intersect at the estimated location of the mobile terminal. As we will show in Chapter 5, the estimation of DOA can be accomplished with either an electrically steerable array antenna or an array antenna with a DOA estimation algorithm embedded at the base band DSP.

Figure 2.7: Concept of geolocation.

2.4.3 High Data Rate Transmission over MIMO Systems

The goal of the 3G communications systems is to provide high data rate transmission [1–5, 88]. The goal urges a significant increase of the channel capacity or say spectrum efficiency. Recently, MIMO systems shed a light in the tunnel to higher data-rate transmissions. It is also concluded that in commercial and financial aspects MIMO system is the most promising candidate that offers such a significant capacity upgrade [25, 58, 88–93].

Multipath fading has long been viewed as an impairment to the wireless communications link. Techniques such as diversity combining, equalization and adaptive beamforming have been investigated to reduce the fading. A different view is that exploiting multipath fading realizes higher data-rate transmissions. In a multipath rich environment, antenna arrays that provide uncorrelated signal outputs are installed at both transmitters and receivers to build MIMO wireless communication channels [27, 56, 57, 94–96].

2.4. Incoming Applications of Smart antennas

It is assumed that the transmitter and receiver antennas are within the far-field of each other. The Multipath phenomenon achieves an information-theoretic channel capacity that increases with the number of array antenna elements. In Fig 2.8, T_X and R_X are the numbers of transmit antennas and receive antennas, respectively. We observe that the capacity grows significantly when T_X and R_X increase.

Figure 2.8: MIMO information-theoretic channel capacities with different numbers of transmit antenna and receive antennas [56].

Transmission STC through MIMO systems takes the advantage of spatial and temporal diversities to increase the data-rate over fading channels [58]. Field tests and experiment results substantiate the capacity gain achieved by MIMO systems [55,97]. Lucent Technologies has achieved a transmission data rate of 40 bits/s/Hz at a 30 dB SNR indoor flat fading environment using 8 transmit and 12 receive antennas vertical Bell Lab layered space time (V-BLAST) architecture [98,99].

2.5 Architectures and Implementations of Smart antennas

In the previous section, we have introduced the applications of the smart antenna technologies. Various applications require different functions from smart antennas. For example, SDMA requires fine beam forming ability, PL services demand for DOA estimation and MIMO systems require decorrelation among signal samples. In this section, we examine various realizations and implementations of smart antennas. A new classification according to their architectures is presented. Figure 2.9 depicts the classification of various smart antenna implementations. They are classified based on their algorithms, hardware and architecture implementations into switched beam antennas, phased array antennas, digital beamforming array antennas, parasitic array antennas and diversity array antennas. In the following section, we will explicitly describe these smart antenna architectures.

Figure 2.9: Architectures of various smart antennas.

Figure 2.10: Functional block diagram of a switched beam antenna. The pattern patterns are produced in the far-field of the antenna.

2.5.1 Switched Beam Antennas

A switched beam antenna system consists of several highly directive, fixed, pre-defined beams [12, 15, 17, 21, 23, 26, 27]. In practice, multiple beams can be formed by a beam-forming network, which consists of an involved feeding system and multiple input ports. The functional block diagram of a switched beam antenna is shown in Fig. 2.10. An example of the beamforming network is the Butler matrix [12, 21, 26, 100], which consists of power splitters and phase shifters. It has been used extensively over the years in radar, electronic warfare (electronic support measures), and satellite systems. The receiver chooses from the one of the several beams to give the best performance, usually in terms of the maximum received power or the lowest BER [101].

In a sense, a switched beam is an extension of the conventional sector beam antenna in that it divides a sector into several microsectors [27]. Therefore, the switched beam antenna system is the easiest to be upgraded from the existing systems that employ

120° sector antennas and dual diversity per sector. However, switched beam antennas also have certain drawbacks. The received signal power from the mobile terminal drops greatly when the mobile terminal moves into the margin of a beam or the area between two beams. Furthermore, owing to the inability to distinguish a desired user from interferers, switched beam antennas are not effective in combating against CCI. If a strong interfering signal is at the center of a selected beam and the desired user is away from the center of that beam, the interfering signal can be enhanced far more than the desired signal. One solution of the aforementioned problems has been presented in [102–104], where phase shifters are added to the switching system. The structure achieves a smoother transition from one beam to another.

2.5.2 Phased Array Antennas

The research and development of phased array antennas have started for radar systems in USA since 1950s [105]. Since then, phased array antennas have been extensively utilized in radar, military scenarios, and satellite communications. In Japan, the application of phased array antenna began when Japanese Defense developed a passive array antenna consisting of element antennas equipped with phase shifters and power divider/combiner in 1960s [105]. In Australia a 12-element land mobile terminal phased array antenna is designed for keeping constant tracking of the MobilesatTM satellite while the terminal is moving [106].

Beam pattern steering of phased array antennas is achieved by implementing phase shift of signals from array elements [15, 26, 27, 29, 30, 102, 106]. Dynamic phased arrays make use of the DOA information from a desired user and steer a beam maximum towards that user, thus improving the capabilities of a switched beam antenna [27]. Tracking is needed to continuously steer the beam towards the desired user. However,

the disadvantage of the phased array antenna is the relatively high cost as expensive PIN diodes are used in phase shifters of beam forming networks [102].

2.5.3 Digital Beamforming Array Antennas

The technological advancement of DSP has brought in the concept of beamforming in the digital stage [15, 18, 20–22, 26, 27, 30, 31]. In comparison with switched beam antennas and phased array antennas, smoother pattern-direction tuning and higher resolution direction finding can be achieved with high-speed DSP and high resolution ADC. The functional block diagram of a DBF array antenna is shown in Fig. 2.11. Each down-conversion channel has a low-noise amplifier, a frequency converter, and an ADC. Signals received by individual antenna elements are down converted to the base band and digitized. Then, they are fed into a DSP chip where algorithms are carried out. A fine steering is achievable by employing the high-resolution sampling and quantization in an A/D conversion process. Many beamforming and DOA estimation algorithms have been studied [5, 7, 8, 18, 24, 27, 31, 35–38]. Simulations and tests show that they are fruitful in improving the system capacity and enhancing the system performance [18, 31, 37]. The DBF distinguishes itself from other counterparts in that it does not physically form any directional pattern. Rather, the beamforming is done by post-processing the data sample at the digital domain. It is due to the complex combination of the signal samples from the array element, the array responses to the impinging waves changes over different directions [107].

However, the high-resolution A/D conversion dissipates a substantial amount of dc power. Furthermore, since a DBF antenna needs the same number of A/D converters as the number of antenna elements, the fabrication cost increases with the number of antenna elements. In addition, because the A/D conversion cannot be carried out di-

Figure 2.11: Functional block diagram of DBF antenna.

rectly in the RF stage, the same amount of power are also consumed at the frequency converters [21, 30, 34]. Due to these reasons, the use of DBF antennas has been restricted to military applications or at the base station for terrestrial mobile wireless communications.

2.5.4 Parasitic Array Antennas

To circumvent the disadvantages of the aforementioned systems, adaptive beam forming based on parasitic array antennas has been explored [30, 32–34, 45, 108–120]. In a parasitic antenna, the active element, which is excited by the transmitted or the received signal source, is surrounded by a number of parasitic elements, which are not connected to the signal source. In the presence of neighboring elements, each one’s performance depends on not only its own current but also the currents on neighboring elements. For resonant elements with no current excitation of their own, there could be a substantial current induced by radiation from another source [121]. The far field radiation pattern is formed by superposition of the radiation of all antenna array elements [122]. Thus,

changing the radiation of each element is possible to direct the antenna beam towards desired directions.

Figure 2.12: Structure of a switched parasitic array.

The use of parasitic elements for controlling antenna beam patterns goes back to the work of Yagi and Uda during the 1930s [121] and has been patented in various forms for numerous applications. A variety of electronically steered Yagi-Uda arrays have been designed. To electronically change the direction of the beam pattern using parasitic elements, these monopole arrays are constructed on a ground plane with an active central element surrounded by rings of parasitic elements. The parasitic elements are switched using PIN diodes connected to the ground plane. A reflector formed by short-circuited elements is arranged behind the active element. Fig. 2.12 gives an example where beam patterns directed to three directions can be selected among three sets of parasitic monopoles. By controlling the states (ON or OFF) of PIN diodes at the parasitic elements, different sets of short-circuited monopoles function as directors and reflectors around the active central monopole. Therefore, directive beams are steered towards the desired directions [108–112, 115, 117, 118]. Instead of using one active element at a fixed position, employing multiple active elements produces multibeam antennas and

gives more choices of active elements' positions [112, 113, 116, 120]. Similar to those wire antenna elements switched by the diodes, it is also possible to use switches to control the radiation pattern of parasitic patch elements and thereby change the beam direction of a parasitic patch array antenna [114]. These aforementioned antennas are categorized as "switched parasitic antennas" [32]. Using switched parasitic antennas, beam directions can only be selected among the predefined directions. Analogous to switched beam antennas, this limits the performance.

Another category of parasitic array antennas is "reactively controlled directive array antenna". It dates back to Harrington's work in 1978 [45]. The functional block diagram is shown in Fig. 2.13. Changing reactance of the parasitic elements alters the radiation of each element. Therefore, directive beam patterns are formed by adjusting the reactance loaded on the parasitic elements. With this configuration, it is possible to achieve a smoother beam pattern tuning. However, it is until recently that extensive investigations on adaptive beamforming and DOA estimation using this kind of structure are carried out. One developed seven-element antenna is called "electronically steerable parasitic array radiator (ESPAR)" [34, 119, 123–125]. We developed a fast beamforming algorithm for ESPAR antennas and tested the convergence of the algorithm. In chapters 3 to 5, we will describe the developed algorithm and trial test results in detail.

In parasitic array antennas, the load reactances and switches are realized by the diodes. There is only a small leakage current. Therefore, the power consumption of the parasitic array antennas are very small. Furthermore, the system normally has only one RF port and one subsequent down-conversion circuit. From the viewpoints of the power consumption and the fabrication cost, it is obvious that the parasitic array antennas are suitable for mass implementations of smart antennas, especially for battery powered portable mobile terminals.

Figure 2.13: Functional block diagram of reactively controlled directive array antennas.

2.5.5 Diversity Antennas

Diversity techniques combat against multipath fading and improve the channel quality for a given radio spectrum in comparison with that from a single port antenna [16, 17, 19, 66]. Because the correlation among signals from the multiple ports of array antennas undermines the diversity combining performance [63, 66], the challenge of diversity antenna design is how to achieve signal decorrelation, especially at mobile terminals such as handphones where antenna elements are mounted within small spacing.

Colburn *et al.* [126] showed three candidate dual-diversity antenna configurations: two side-mounted planar-inverted F antennas (PIFAs), a back-mounted PIFA with a top-mounted helix, and a top-mounted PIFA with a flip monopole. Experiments results show sufficient signal decorrelation of these configurations. The diversity gain obtained with polarization diversity was demonstrated in [70], where dual-polarization diversity is obtained by an aperture coupled patch antenna or by two slanted dipoles over an infinite ground plane. Vaughan [117] presented diversity antennas using various para-

parasitic array antenna structures [32]. In one of the designs, three directional patterns are formed with three subsets of parasitic array as shown in Fig. 2.12. By selecting one of patterns pointing to different directions, angular diversity advantage is achieved. In another design, sixteen parasitic monopoles are placed uniformly in the circle of radius of 0.25 wavelength (λ), and an active monopole is located at the center of the circle. The directional beam pattern with 90° half-power beamwidth (HPBW) is rotated, $360/16^\circ$ each step, by shifting switch setting of parasitic elements. For adjacent beam patterns with 22.5° angular spacing, the envelope correlation of complex beam patterns is 0.91. For patterns with larger angular separation, say 45° , the envelope correlation of 0.7 is achieved. Virginia Tech Antenna Group has evaluated the envelope correlation, power imbalance and diversity gain with space diversity, polarization diversity and pattern diversity at handheld terminals [127]. The spatial diversity is achieved with two parallel half-wavelength dipoles with inter-element spacing variable from 0.1 to 0.5 wavelength. The dual polarization diversity antenna is constructed with a dipole and a printed wheel antenna separated vertically by 0.3 wavelength. As for the pattern diversity, it is achieved with two dipoles that have 0.25 wavelength inter-element spacing, connected to a Quadrature (90°) hybrid. The structure provides two orthogonal beams at the two ports. Measurement results in various multipath environments have clearly justified the feasibility of diversity at handheld terminals.

So far we have discussed various smart antennas available in open literature. Every antenna has its merits and demerits based on its design complexity, performance, cost and applications. Figure 2.14 generalizes the trade-off between the performance and complexity of various forms of smart antennas and some of their applications. For the applications where the cost and power consumption are the main limited factors, parasitic array antennas and switched beam antennas are suitable. While DBF array antennas and phased array antennas dominate in the applications such as satellite

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communications, BS applications, where high performance is desired. In addition to combating fading, diversity antennas can also be employed in the high-speed MIMO wireless transmissions.

Figure 2.14: Relation of various smart antenna structures and their applications

2.6 Algorithms for Smart Antennas

The wireless communications system is evolving with a fast pace. Better quality wireless communications, new services and applications will become available with the technological advances. Smart antennas play important roles in the evolution of wireless communications systems. In the previous section, we have described various smart antenna architectures. These physical structures map signals from the spatial-temporal domain to the time domain [107]. There are algorithms that process the received signals and make the antenna “smart”. Adaptive beamforming algorithms can be implemented on switched beam antennas, phased array antennas, DBF array antennas and parasitic array antennas. The aim of the adaptive beamforming algorithms is to suppress interferences by controlling the antenna’s spatial responses adaptively according to a given communication environment. Direction finding algorithms analyze signal samples obtained from the antenna and give direction information of the impinging waves. Diversity combining algorithms have been utilized to combat against fading by providing redundancy. Recently, various space-time codes have been proposed to provide transmit diversity. Due to the limited space we describe some of the algorithms that we have compared with and improved upon in the following chapters. A sheer volume of algorithms can be found in the following public literature [5, 7, 8, 18, 24, 27, 31, 35–38]. Figure 2.15 classifies some popular algorithms. According to the applications of smart antennas, these algorithms are classified into three categories: beamforming, direction finding and diversity combining. The adaptive beamforming algorithm we proposed for reactively controlled directive array falls into the analog beamforming class, whereas, DBF algorithms include the popular maximum SINR algorithm and so on. High-resolution direction finding can be achieved with the MUSIC algorithm, and low resolution can be realized by measuring the received signal power. Diversity techniques include receive diversity and transmit diversity. In this section, we review a few of them.

Figure 2.15: Various algorithms for smart antennas.

2.6.1 Beamforming Algorithms

Here we demonstrate a beamforming algorithm using a five-element linear array antenna with inter-element spacing d . Let us assume that there are totally K signal sources. The k th impinging plane wave is represented by $s_k(t)e^{j2\pi f_c t}$. Where $s_k(t)$ represents the base band modulated signal and f_c is the carrier frequency of the impinging plane waves. After passing the RF/IF branches, signals are converted down to the base band signal. Under the narrow-band assumption [37], the base band signal at the i th array element is presented as:

$$u_i(t) = \sum_{k=1}^K s_k(t) e^{-j \frac{2\pi}{\lambda} (i-1)d \cos(\phi_k)} + n_i(t) \quad (2.1)$$

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where $n_i(t)$ is the noise component with power σ^2 on the i th array element, which includes back ground noise and electronic noise generated in the i th channel [37]. In the rest of the thesis, we assume that the wave impinges the array with AOA ϕ in the azimuth plane, the elevation angle θ is 90° . Let $\alpha(\phi_k)$ represent the array factor of the k th impinging wave on the linear array antenna as shown in Fig. 2.16. Defined based on the array geometry [128], it is expressed as:

$$\alpha(\phi_k) = \begin{bmatrix} 1 \\ e^{-j\frac{2\pi}{\lambda}d\cos(\phi_k)} \\ e^{-j\frac{2\pi}{\lambda}2d\cos(\phi_k)} \\ e^{-j\frac{2\pi}{\lambda}3d\cos(\phi_k)} \\ e^{-j\frac{2\pi}{\lambda}4d\cos(\phi_k)} \\ e^{-j\frac{2\pi}{\lambda}5d\cos(\phi_k)} \end{bmatrix}, \quad (2.2)$$

Let

$$\mathbf{u} = \begin{bmatrix} u_1(t) & u_2(t) & u_3(t) & u_4(t) & u_5(t) \end{bmatrix}^T, \quad (2.3)$$

and

$$\mathbf{n} = \begin{bmatrix} n_1(t) & n_2(t) & n_3(t) & n_4(t) & n_5(t) \end{bmatrix}^T, \quad (2.4)$$

we have

$$\mathbf{u} = \begin{bmatrix} \alpha(\phi_1) & \alpha(\phi_2) & \dots & \alpha(\phi_K) \end{bmatrix} \begin{bmatrix} s_1(t) & s_2(t) & \dots & s_K(t) \end{bmatrix}^T + \mathbf{n}. \quad (2.5)$$

Superscript T represents transpose. The beamforming process is carried out at the base band. In the equations, the variables without a time argument represent base band signal digital samples. These samples from the array elements are multiplied by a complex weight and summed to form the array output y . Let $\mathbf{w} = \begin{bmatrix} w_1 & w_2 & w_3 & w_4 & w_5 \end{bmatrix}^T$

be the beamforming weight vector, the array output is expressed as:

$$y = \sum_{i=1}^5 w_i^* u_i = \mathbf{w}^H \mathbf{u}$$

(2.6)

where superscripts $*$ and H denote complex-conjugate and conjugate-transpose, respectively.

Figure 2.16: Digital beamforming with a five-element linear array antenna.

From the above equations, we can see that the beamforming process produces a linear combination of the array sensor outputs. The process is analogous to a linear filtering problem. The beamforming is a process of finding an optimal weight vector $\mathbf{w}$ based on certain criteria, such as the lowest BER, the signal-to-interference-plus-

noise ratio (SINR), the minimum-mean-square-error (MMSE), the minimum variance, the maximum cross correlation coefficient (MCCC) [31, 34–37, 128–131], ML [132], and algorithms that exploit signal higher statistics such as cyclostationary [133] *etc.*

MMSE Criterion

Beamforming algorithms based on MMSE criterion minimize the mean square error (MSE) between array outputs y and reference signals r :

$$\varepsilon^2 = [r - \mathbf{w}^H \mathbf{u}] [r - \mathbf{w}^H \mathbf{u}]^H. \quad (2.7)$$

Minimizing ε^2 with respect to $\mathbf{w}$ we obtain the optimum beamforming weight vector as:

$$\mathbf{w}_{opt} = \mathbf{R}_{uu}^{-1} \mathbf{R}_{ru} \quad (2.8)$$

where $\mathbf{R}_{uu}$ is the self-correlation matrix of $\mathbf{u}$ and $\mathbf{R}_{ru}$ is the cross-correlation matrix between $\mathbf{r}$ and $\mathbf{u}$. The equation is referred to as the Wiener-Hopf equation or the Wiener Solution [31, 134]. The reference signals in the algorithm can be realized with the 28-bit synchronization sequence in IS-95 TDMA standard, the training sequence in GSM standard or the dedicated training sequence in digital European cordless telephone (DECT) system [31].

To avoid matrix inversion operation in (2.8) and approach the Wiener solution in a non-stationary environment where the input self-correlation $\mathbf{R}_{uu}$ is not known exactly, least mean square (LMS) adaptation algorithm can be applied [134]. The procedure iteratively corrects the beamforming weight vector in the direction of the negative of the gradient that eventually lead to the MMSE. The procedure is described below.

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1. Obtain array output at the n th iteration:

$$y(n) = \mathbf{w}(n)^H \mathbf{u}(n). \quad (2.9)$$

2. Estimate the error:

$$\varepsilon(n) = r(n) - y(n). \quad (2.10)$$

3. Beamforming weight adaptation:

$$\mathbf{w}(n+1) = \mathbf{w} + \mu \mathbf{X}(n) \varepsilon(n)^*, \quad (2.11)$$

where μ is the step size of the adaptation.

Maximum SINR Criterion

Let us assume that $s_k(t)$ is the desired signal. The beamforming output SINR is expressed as:

$$\gamma_b = \frac{\mathbf{w}^H \{\alpha(\phi_k) s_k(t)\} \{\alpha(\phi_k) s_k(t)\}^H \mathbf{w}}{\mathbf{w}^H \left\{ \sum_{l=1, l \neq k}^K \alpha(\phi_l) s_l(t) + \mathbf{n} \right\} \left\{ \sum_{l=1, l \neq k}^K \alpha(\phi_l) s_l(t) + \mathbf{n} \right\}^H \mathbf{w}} = \frac{\mathbf{w}^H \mathbf{R}_s \mathbf{w}}{\mathbf{w}^H \mathbf{R}_{ni} \mathbf{w}}, \quad (2.12)$$

where $\mathbf{R}_s$ and $\mathbf{R}_{ni}$ are the desired signal correlation matrix and the interference plus noise correlation matrix, respectively. The problem of finding the optimum beamforming weight vector $\mathbf{w}_{opt}$ is a generalized Rayleigh quotient problem [135]. To maximize the output SINR γ_b , the optimum beamforming weight vector $\mathbf{w}_{opt}$ is chosen as the eigenvector corresponding to the maximum eigenvalue λ_{max} of $\mathbf{R}_{ni}^{-1} \mathbf{R}_s$ as:

$$\lambda_{max} \mathbf{w}_{opt} = \mathbf{R}_{ni}^{-1} \mathbf{R}_s \mathbf{w}_{opt}. \quad (2.13)$$

If g_0 is a constant represents amplitude gain of the array to the desired signal [136], we have

$$\lambda_{max} \mathbf{w}_{opt} = \mathbf{R}_{ni}^{-1} \mathbb{E} [s_k^2(t)] \boldsymbol{\alpha}(\phi_k) g_0 \quad (2.14)$$

where $\mathbb{E}(\cdot)$ is the expectation operator. Therefore, the scaled optimum beamforming weight vector is

$$\mathbf{w}_{opt} = \mathbf{R}_{ni}^{-1} \boldsymbol{\alpha}(\phi_k). \quad (2.15)$$

If the desired signal is uncorrelated with the noise plus interference correlation matrix, we have

$$\mathbf{R}_{uu} = \mathbf{R}_s + \mathbf{R}_{ni}. \quad (2.16)$$

Applying Woodbury's identity on $\mathbf{R}_{uu}$ in (2.8) can be expressed as:

$$\mathbf{R}_{uu}^{-1} = \left\{ \frac{1}{1 + \mathbb{E} [s_k^2(t)] \boldsymbol{\alpha}(\phi_k)^H \mathbf{R}_{ni}^{-1} \boldsymbol{\alpha}(\phi_k)} \right\} \mathbf{R}_{ni}^{-1} = \nu \mathbf{R}_{ni}^{-1}. \quad (2.17)$$

The wiener solution in (2.8) can be modified as:

$$\mathbf{w}_{opt} = \mathbb{E} [s_k^2(t)] \nu \mathbf{R}_{ni}^{-1} \boldsymbol{\alpha}(\phi_k). \quad (2.18)$$

Hence, the MMSE beamforming and the maximum SINR beamforming yield the same output SINR [31].

2.6.2 Direction Finding Algorithms

Direction finding or DOA estimation techniques can be classified into four groups: conventional approach, subspace method, ML method and other techniques that exploit signal spectral property [26]. Here, we briefly introduce one of the subspace methods, MUSIC. Developed by Schmidt [137, 138], it is a relatively simple and high-resolution

eigenstructure approach and has been widely used as a model-based method for DOA estimation and source location problems. There are also many improved forms, such as the Root-MUSIC algorithm and the beamspace-MUSIC algorithm [37].

Let us assume that a linear array antenna used for DOA estimation has $R_X = D$ inter-elements with element spacing d . The array factor expressed in (2.2) is a $D \times 1$ dimensional vector. Applying (2.5), the receive signal samples $\mathbf{u}$ at the array elements is a $D \times 1$ vector too. Here we assume $D > K$. After eigendecomposing the correlation matrix $\mathbf{R}_{uu}$, the $D \times (D - K)$ dimensional noise subspace matrix $\mathbf{V}_N$ is constructed with $D - K$ eigenvectors that correspond to $D - K$ smallest eigenvalues. Based on the property that the signal subspace is orthogonal to the noise subspace [26, 137, 138], the MUSIC power spectrum is plotted by computing

$$P_{MU}(\phi) = \frac{1}{\boldsymbol{\alpha}(\phi)^H \mathbf{V}_N \mathbf{V}_N^H \boldsymbol{\alpha}(\phi)}. \quad (2.19)$$

The K largest peaks in the MUSIC spectrum correspond to the DOAs of impinging waves.

It is worth to mention here that the aforementioned algorithms are methods based on signal sample covariance matrix. These methods cannot handle the coherent signals unless further processing such as spatial smoothing methods are applied. This is because the signal coherence leads to a rank deficiency of the signal sample correlation matrix. An new approach which is based on direction data of the signal sample instead of the covariance matrix has been proposed in [107]. The beamforming and DOA estimation methods therein are efficient in coping with blinking interference and coherent signals.

2.6.3 Space-Time Block Code to Achieve Diversity Advantages

As aforementioned, there are various diversity techniques such as space diversity, polarization diversity, frequency diversity, time diversity *etc.* Various methods for achieving the diversity advantages include switch and stay combining (SSC), selection combining (SC), equal gain combining (EGC) and maximum ratio combining (MRC) [10, 63, 65]. The performances of these methods have been presented in many literatures. Therefore, we will concentrate on a recently developed technique for realizing transmit diversity through space time block code transmission.

The aim of transmit diversity techniques is to provide the receiver with multiple uncorrelated replicas of the same signal using multiple transmit antennas [10]. At the receiver, diversity gain can be obtained with just a single antenna.

The transmit diversity can also be provided with transmitting STBC over multiple antennas. In [139], Alamouti has introduced a simple STBC with ML detection. The receiver with a single antenna can achieve the same diversity order as that provided by MRC using one transmit and two receive antennas. Due the receiving simplicity of the Alamouti STBC, it has been adopted in the 3G standard to provide transmit diversity [89, 140]. In this subsection, we give a brief introduction of Alamouti STBC transmission.

Let us assume a scenario where one MS with two transmit antennas ($T_X = 2$) communicates with a BS that has one receive antenna ($R_X = 1$). At symbol epoch NT_s , two symbols $\{c_1, c_2\}$ are transmitted in group over MS's two transmit antennas. c_1 and c_2 are transmitted over the first transmit antenna and the second transmit antenna of the

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MS, respectively. In the next symbol period $(N + 1)T_s$, $-c_2^*$ and c_1^* are transmitted over the first and the second transmit antennas of the MS, respectively. Let h_{11} and h_{12} be the fading channels from the first and the second MS transmit antennas to the BS receive antenna. Under the assumption that the channel fading state is constant over two symbol epochs, NT_s and $(N + 1)T_s$, the two received scalar outputs at the BS receive antenna over two symbol epochs are:

$$u(NT_s) = h_{11}c_1 + h_{12}c_2 + n(NT_s), \quad (2.20)$$

and

$$u((N + 1)T_s) = -h_{11}c_2^* + h_{12}c_1^* + n((N + 1)T_s), \quad (2.21)$$

where $n(NT_s)$ is the noise component. The above equation can be written in a matrix form as:

$$\begin{bmatrix} u(NT_s) \\ u((N + 1)T_s)^* \end{bmatrix} = \begin{bmatrix} h_{11} & h_{12} \\ h_{12}^* & -h_{11}^* \end{bmatrix} \begin{bmatrix} c_1 \\ c_2 \end{bmatrix} + \begin{bmatrix} n(NT_s) \\ n((N + 1)T_s)^* \end{bmatrix}, \quad (2.22)$$

where the channel matrix $\mathbf{H}$ is defined as:

$$\mathbf{H} = \begin{bmatrix} h_{11} & h_{12} \\ h_{12}^* & -h_{11}^* \end{bmatrix}. \quad (2.23)$$

The matrix is called “Alamouti virtual MIMO channel matrix”. It has time dimension in column and space dimension in row. Furthermore the matrix is orthogonal as:

$$\mathbf{H}\mathbf{H}^H = (|h_{11}|^2 + |h_{12}|^2) \mathbf{I}_2. \quad (2.24)$$

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The ML decoder produces two symbol $\hat{c}_1$ and $\hat{c}_2$ and for the detection of signal c_1 and c_2 .

$$\begin{aligned} \begin{bmatrix} \hat{c}_1 \\ \hat{c}_2 \end{bmatrix} &= \begin{bmatrix} h_{11}^* & h_{12} \\ h_{12}^* & -h_{11}^* \end{bmatrix} \begin{bmatrix} u(NT_S) \\ u((N+1)T_S)^* \end{bmatrix} \\ &= \begin{bmatrix} (|h_{11}|^2 + |h_{12}|^2) c_1 + h_{11}^* n(NT_S) + h_{12} n((N+1)T_S)^* \\ (|h_{11}|^2 + |h_{12}|^2) c_2 - h_{11} n((N+1)T_S)^* + h_{12}^* n(NT_S) \end{bmatrix}. \end{aligned} \quad (2.25)$$

From the above equation we can see that the ML decoding process is equivalent to a dual-branch MRC [10, 139]. If the BS has two receive antennas, the decoding process realizes a four-branch MRC.

Figure 2.17: BER performance comparison of BPSK MRC and Alamouti STBC transmission scheme

Fig 2.17 illustrates the performance comparison of Alamouti STBC transmission and MRC for different transmit power. For various diversity scheme or STBC transmission, the diversity order is exhibited by the slope of BER vs SNR curves [141–143]. It can be seen from the figure that the diversity orders of $T_X = 2, R_X = 1$ Alamouti scheme and $T_X = 2, R_X = 2$ are equal to that of dual-branch MRC and four-branch MRC, respectively. The 3-dB gaps are because of the splitting of transmit power at the MS when it has two transmit antennas. In Chapter 7, we will elaborate more on the ML decoding process and combine beamforming with the Alamouti STBC transmission scheme.

2.7 Proposed Smart Antenna Techniques

In the previous sections, we have reviewed the evolution of wireless communications and surveyed various smart antenna systems including their architectures and algorithms. As illustrated in Fig. 2.9, each smart antenna system has its suitability according to the implementation and application scenarios, the trade-off between system implementation complexity, cost and the performance. The main goal of this thesis is to design and develop various antenna algorithms for different implementations. Through the PhD study, we select different smart antenna architectures for different applications and develop algorithms and techniques for these applications. The developed techniques include: 1) a fast beamforming algorithm for a reactively controlled directive array antenna which features low power consumption and fabrication cost; 2) a high-resolution direction finding technique for single RF-port smart antennas. These two techniques aim at implementations at low power consumption mobile terminals in WLAN, ad hoc networks, *etc.*; 3) a base station array processing technique for implementing the SDMA scheme in a multipath rich environment; and 4) two receiver structures that combine

beamforming techniques with STBC receivers to achieve better BER performance and higher channel capacities. The reasons behind choosing these different specific smart antenna implementation architectures and designing the corresponding techniques are:

- After a comprehensive literature review, we learn that parasitic array antennas possess strong potentials in low power consumption applications, such as ad hoc networks, golf court car, shopping cart, *etc.* As we have described in the previous chapter, the switched parasitic array antenna selects one of the predefined beam to improve the performance. To achieve a smoother beam tunability, steepest gradient algorithm (SGA) has been applied to reactively controlled directive arrays. However, the SGA beamforming technique for parasitic array antennas requires long training time. Thus it is not efficient for practical implementations. Therefore, we develop a fast beamforming algorithm based on the reactively controlled directive array antenna structure. The algorithm requests less training time and brings a faster convergence, thus making the system suitable for commercial applications.
- For installing smart antennas at battery-powered mobile terminals in ad hoc networks, we select parasitic array antenna again as the implementation structure. In this application scenario, DOA information is helpful in settling up a communication link. Since the antenna has only one RF-port, which precludes direct utilization of the conventional MUSIC direction finding algorithms, we modify the MUSIC algorithm for the implementation on parasitic array antennas. The technique achieves a high-resolution direction finding performance comparable with the conventional DBF array antenna based MUSIC method.
- For the cellular BS applications the main limiting factor is the performance rather than the cost. Therefore, we choose DBF as the smart antenna implementation

structure. Employing smart antenna techniques at BSs realizes an efficient frequency reuse by virtue of mobile terminals' spatial uniqueness, thus increasing the system capacity. The scheme is commonly referred as SDMA. However, the conventional beamforming based SDMA scheme fails in a multipath rich environment. We propose a technique to circumvent the problem. The technique employs the constrained optimization method to form nulls to undesired users' channel responses.

- From the review we have learned that the development trend of smart antennas technique is to realize high data rate transmissions for 3G multimedia communications. Diversity antennas are installed at both transmitters and receivers to construct a MIMO system. Transmitting STCs through a MIMO system dramatically increases the data rate over the wireless channel. The beamforming techniques have long been accepted as an efficient way for interference suppression. To bring the benefits of the beamforming techniques to the MIMO systems, we concatenate beamforming with the Alamouti STBC receivers. The results show significant improvement in BER performance and the achieved channel capacity.

In the following chapters, the proposed techniques will be presented. Firstly, a fast adaptive beamforming algorithm for reactively controlled directive array antennas are presented. After a theoretical study of the beamforming ability and the convergence of the algorithm, testbed experiments are carried out in an anechoic chamber to verify the algorithm's performance. We also develop a high-resolution direction finding technique for single RF-port smart antennas. Again, the algorithm is justified with experiment results. Not solely confined to mobile terminal smart antennas, we also develop algorithms for cellular BS implementations. An SDMA technique for implementations in multipath rich environments are proposed. Finally, we combine beamforming techniques with STC transmissions at receivers to achieve better BER performances and

higher channel capacities.

2.8 Conclusions

The thorough literature survey provides the fundamental knowledge of the PhD study. Guided with the development trend of the smart antenna technologies, we propose fast beamforming and direction finding techniques for reactively controlled directive array antennas. To realize SDMA scheme, we develop a technique for the implementation in multipath rich environment. The latest trend is to implement diversity antennas at both transmitters and receivers to construct MIMO communication systems. In response to that tendency, we combine beamforming with MIMO STBC for higher data rate transmissions.

Chapter 3

Adaptive Beamforming with Parasitic Array Antennas

Abstract: *In this chapter, we describe a low power consumption, small size reactively controlled directive array. The beamforming of this antenna is achieved by tuning the load reactances at parasitic elements surrounding the active central element. Currently, the beam forming algorithms for the reactively controlled parasitic array antennas require long training time. In comparison with these algorithms, a faster beamforming algorithm based on simultaneous perturbation stochastic approximation (SPSA) theory with a MCCC criterion is proposed. The simulation and experimental results validate the algorithm. In an environment where the SIR is 0 dB, the algorithm converges within 50 iterations and achieves an output SINR of 10 dB. With the fast beamforming ability and its low power consumption attribute, the antenna makes the mass deployment of smart antenna technologies practical. To give a comparison of the beamforming algorithm with the one of the standard beamforming algorithm for a DBF antenna array, we compare the proposed algorithm with the LMS beamforming algorithm. Since the parasitic antenna is in nature an analog antenna, it cannot suppresses correlated inter-*

ference. Here, we assume that the interferences are uncorrelated.

3.1 Introduction

The evolution of wireless communications systems requires new technologies to support better quality communications, new services and applications. Smart antennas have become a hot topic of research. The directive beam patterns can be steered toward the desired signal and deep nulls can be formed toward the interference, thus spatial filtering is realized. This brings the benefits such as lower power transmission, higher spectrum efficiency, better link quality and higher system capacity [6, 7, 15–25].

Various beamforming and DOA estimation algorithms have been designed [5, 7, 8, 18, 24, 27, 31, 35–38]. The simulation and experiments carried out by many researchers have shown the abilities of these algorithms [18, 39–43]. Most of these algorithms are designed based on the DBF antenna arrays (see Fig. 2.11). Signals received by individual antenna elements are down-converted into baseband signals. These signals are digitized and fed into DSP chip where the algorithms reside in. However, RF circuit branches connected to the array elements, ADCs and the baseband DSP chip consume a considerable amount of dc power. Furthermore, each channel connected to the array sensor has the same structure, so the cost of fabrication increases with the number of array elements [18, 30, 32]. Thanks to the recent development of GaAs MMIC technologies, the beamformer could be integrated into a single chip at the RF front end such as MBF [44], instead of the baseband. The advantages are the reduced quantization errors and the increased dynamic range. However, their costs of fabrication still limit the range of implementations. All these factors make DBF and MBF antennas unsuitable for low power consumption and low cost systems and hinder the mass applications

of the smart antenna technologies. For example, it could be too costly to equip DBF antenna arrays at battery powered lap-tops or mobile computing terminals within a wireless network [144].

As the wireless communications become more ubiquitous, the demands for smart antennas expand. Smart antennas are now not only required to be installed at BSs but also strongly desired to be installed at MSs such as vehicles, laptops, and even commercial mobile phones. The challenges emerge. Due to the size and power limit of the MSs, the smart antennas should be low power consumption, small in size and low cost so that it is affordable to mobile users. With this goal in mind, researchers are now investigating the parasitic array antennas [30, 32–34, 45, 108–120]. The antenna normally has one RF port, therefore, the size and power consumption can be significantly reduced [30, 32].

In [108–113, 115–118, 120], various forms of switched parasitic array antennas have been documented. Beam steering of these antennas are achieved by switching ON and OFF parasitic elements, or switching the position of the active element. However, beam patterns can only be steered to a predefined set of directions in a way similar to that for switched beam antennas. This limits the performance, especially for the applications scenarios where desired signals impinging from different directions due to multipath reflection.

Another category of parasitic array antennas, reactively controlled directive array antenna, circumvents the problem with switched parasitic array antennas. The first work was presented in [45]. The antenna has one central element connected to the sole RF port and a number of surrounding parasitic elements form the array. Beam steering is achieved by tuning the load reactances at parasitic elements surrounding the central active element (see Fig. 2.13). Loaded with tunable reactance, each parasitic element

has more than just two states, ON or OFF. Therefore, a smoother beam pattern steering is achieved.

Recently, the adaptive beamforming with reactively controlled directive array antennas has been studied [30, 34, 47–52, 119, 124, 145–148]. The antenna is named as “electronically steerable parasitic array radiator (ESPAR)”. In [145], a random search method is applied to optimize the reactance values. A Hamiltonian approach is presented in [146]. It projects an optimization problem on the motion of a particle in a M -dimensional space, by making the objective function and the unknown variables correspond to the potential energy and the coordinates, respectively. The optimum result is only achieved through a large number of calculation cycles. The Steepest Gradient Algorithm (SGA) is applied in [47, 147]. The algorithm maximizes the cross correlation coefficient (CCC) between the received signal and the reference signal, using steepest gradient method based on sequential perturbation [149]. To cope with the noise effect, maladjustment and uncertainty in estimating the objective function, another beamforming algorithm is proposed in [48, 148]. The algorithm is based on the stochastic approximation theory. A blind adaptive algorithm maximizes the higher-order moment using steepest gradient method is presented in [50]. The algorithm does not require the training signals.

However, the above algorithms are not efficient enough for real time implementations in the wireless communications systems, where a fast beam tuning ability is required. In this chapter a fast beamforming algorithm based on the simultaneous perturbation stochastic approximation (SPSA) theory [150] is proposed. After the theoretical study the performance is examined with simulations. The experiment [34] is presented in the next chapter. Our study proves the adaptive beamforming ability of the ESPAR antennas and demonstrates its suitability for mass commercial implementations of smart antenna technologies to wireless networks.

3.2 Beamforming Architectures and Theory

3.2.1 Antenna Configuration

Fig. 3.1 shows the configuration of a 7-element ESPAR antenna. One active central element (monopole) is surrounded by 6 parasitic elements uniformly placed on a circular grid of radius R on a circular grounded baseplate. The length of each monopole L and the radius R of the circular grid are one-quarter wavelength ($\lambda/4$) of the transmitting RF signal at 2.484 GHz. According to Harrington [45], large spacing among elements (for example $\lambda/2$) leads to reduced gain and significant back lobes. Moreover, further reduction of the inter-element spacing to $\lambda/8$ leads to the super-gain array. Therefore, a realistic design for spacing = $\lambda/4$ is chosen. The baseplate transforms monopoles with their images to dipoles with a length of $2L$. The central monopole is connected to a RF receiver and each parasitic monopole is loaded with a tunable varactor.

Figure 3.1: Configuration of a 7-element ESPAR antenna. The length of each monopole L and the radius R of the circular grid are one-quarter wavelength of the transmitting RF signal at 2.484 GHz.

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The working principle of ESPAR antennas is different from that of DBF array antennas. We first assume that the antenna is working in the transmitting mode. The antenna generates a directional beam based on tuning load reactances $(x_1, x_2, \dots, x_6)$ on the parasitic monopoles. Signals received or transmitted from the central RF port excite the parasitic monopoles with substantial induced mutual currents on them. Vectors $\mathbf{i}$ and $\mathbf{v}$ represent the currents and the voltages on the monopoles, respectively.

$$\mathbf{i} = [i_0 \ i_1 \ i_2 \ i_3 \ i_4 \ i_5 \ i_6]^T \quad (3.1)$$

$$\mathbf{v} = [v_0 \ v_1 \ v_2 \ v_3 \ v_4 \ v_5 \ v_6]^T \quad (3.2)$$

The superscript T represents the transpose. i_0 and v_0 represent the current and the voltage on the active central element ($\#0$), respectively. Mutual admittances are represented by the matrix $\mathbf{Y}_C$ with each entity y_{ij} denoting the mutual admittance between the i th element and the j th element. The induced currents are represented with mutual admittances:

$$\mathbf{i} = \mathbf{Y}_C \mathbf{v} = \begin{bmatrix} y_{00} & y_{01} & y_{02} & y_{03} & y_{04} & y_{05} & y_{06} \\ y_{10} & y_{11} & y_{12} & y_{13} & y_{14} & y_{15} & y_{16} \\ y_{20} & y_{21} & y_{22} & y_{23} & y_{24} & y_{25} & y_{26} \\ y_{30} & y_{31} & y_{32} & y_{33} & y_{34} & y_{35} & y_{36} \\ y_{40} & y_{41} & y_{42} & y_{43} & y_{44} & y_{45} & y_{46} \\ y_{50} & y_{51} & y_{52} & y_{53} & y_{54} & y_{55} & y_{56} \\ y_{60} & y_{61} & y_{62} & y_{63} & y_{64} & y_{65} & y_{66} \end{bmatrix} \times \begin{bmatrix} v_0 \\ v_1 \\ v_2 \\ v_3 \\ v_4 \\ v_5 \\ v_6 \end{bmatrix} \quad (3.3)$$

Because of the symmetric structure of the ESPAR antenna, the admittances are:

$$y_{11} = y_{22} = y_{33} = y_{44} = y_{55} = y_{66}$$

$$y_{01} = y_{02} = y_{03} = y_{04} = y_{05} = y_{06}$$

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$$y_{12} = y_{23} = y_{34} = y_{45} = y_{56} = y_{61}$$

$$y_{13} = y_{24} = y_{35} = y_{46} = y_{51} = y_{62}$$

$$y_{14} = y_{25} = y_{36}$$

Therefore, the admittance matrix $\mathbf{Y}_C$ is described with only 6 independent parameters. These values are obtained with the Numerical Electromagnetic Code (NEC) [151]. The voltages on the active central monopole and the m th parasitic monopole are represented by (3.4) and (3.5) respectively:

$$v_0 = v_s - z_0 i_0, \quad (3.4)$$

$$v_m = -jx_m i_m. \quad (3.5)$$

$m = 1, \dots, 6$. z_0 is the characteristic impedance of 50Ω at the RF port. v_s represents the transmitted voltage signal source with the amplitude and the phase from the driving RF port at the central element. Equations (3.4) and (3.5) are written into a matrix form as:

$$\mathbf{v} = \begin{bmatrix} v_s \\ 0 \\ 0 \\ 0 \\ 0 \\ 0 \\ 0 \end{bmatrix} - \mathbf{X}_L \mathbf{i} = v_s \mathbf{U}_1 - \mathbf{X}_L \mathbf{i}, \quad (3.6)$$

and define

$$\mathbf{X}_L = \text{diag} \left[z_0, jx_1, \dots, jx_6 \right], \quad (3.7)$$

where

$$\mathbf{U}_1 = \begin{bmatrix} 1 & 0 & 0 & 0 & 0 & 0 & 0 \end{bmatrix}^T \quad (3.8)$$

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Combining (3.3) and (3.6), we get:

$$\mathbf{i} = \mathbf{Y}_C \mathbf{v} = \mathbf{Y}_C (v_s U_1 - \mathbf{X}_L \mathbf{i}). \quad (3.9)$$

After a simple mathematical manipulation, we can get:

$$\mathbf{i} = v_s (\mathbf{Y}_C^{-1} + \mathbf{X}_L)^{-1} U_1 = v_s \mathbf{w}_{ESP}, \quad (3.10)$$

where $(\mathbf{Y}_C^{-1} + \mathbf{X}_L)^{-1} U_1$ is represented as $\mathbf{w}_{ESP}$, a 7-by-1 dimensional vector, called as “ESPAR equivalent weight vector”. The far-field radiation pattern is the superposition of all monopoles’ radiation patterns [121]. Therefore, the far-field current signal in the azimuthal direction ϕ with its amplitude and the phase is represented as:

$$y_{far}(\phi) = \mathbf{i}^T \boldsymbol{\alpha}(\phi) = \mathbf{w}_{ESP}^T \boldsymbol{\alpha}(\phi) v_s. \quad (3.11)$$

The steering vector $\boldsymbol{\alpha}(\phi)$ is defined based on the array geometry (see Fig. 3.1).

$$\boldsymbol{\alpha}(\phi) = \begin{bmatrix} 1 \\ e^{j\frac{\pi}{2}\cos(\phi)} \\ e^{j\frac{\pi}{2}\cos(\phi-\frac{\pi}{3})} \\ e^{j\frac{\pi}{2}\cos(\phi-\frac{2\pi}{3})} \\ e^{j\frac{\pi}{2}\cos(\phi-\pi)} \\ e^{j\frac{\pi}{2}\cos(\phi-\frac{4\pi}{3})} \\ e^{j\frac{\pi}{2}\cos(\phi-\frac{5\pi}{3})} \end{bmatrix}. \quad (3.12)$$

According to the reciprocity theory for radiation patterns [121], if the antenna is working in the receiving mode, the voltage signals $y(t)$ at the RF port are:

$$y(t) = \mathbf{w}_{ESP}^T \boldsymbol{\alpha}(\phi) s(t). \quad (3.13)$$

where, $s(t)$ represents the far-field incident current waves with the amplitude and the phase in the azimuthal direction ϕ .

3.2.2 Calculation of Mutual Admittance Matrix

In the previous section, we have shown that beamforming of an ESPAR antenna is achieved based on strong mutual coupling among neighboring elements (one active center monopoles and 6 parasitic monopoles loaded with tunable reactance).

The calculation of the mutual impedance matrix $\mathbf{Y}_C$ is achieved with Numerical Electromagnetic Code, i.e. NEC simulator. NEC is the moment method code for analyzing the interaction of electromagnetic waves with arbitrary structures. The NEC simulator we used is from Poynting Software (Pty) Ltd [152]. It is an object-oriented software which uses an electric-field integral equation (EFIE) to model the electromagnetic response of general structures. Surface and elements are modelled with wire segments. In Fig. 3.2, each monopole is modelled with 21 wire segments. All the monopoles are loaded with 50Ω impedance at the center segment. By exciting one monopole at the center segment with a unitary voltage, we can get the currents at the center segments of all monopoles. Since the exciting voltage is unitary, the currents on the neighboring monopoles are the mutual admittance and the current on the excited monopole is the self admittance. The values are listed below.

$$y_{00} = 0.00023344 - j0.0066918$$

$$y_{10} = -0.0001816 + j0.00242469$$

$$y_{11} = 0.00300676 - j0.0044176$$

$$y_{21} = 0.00097427 + j0.00299986$$

$$y_{31} = -0.0003066 - j0.0003067$$

$$y_{41} = -0.0000973 - j0.0001353$$

Figure 3.2: Structure of a seven-element ESPAR antenna on an infinite ground plane in NEC simulator. The ground plane transforms the monopoles with their image into $\lambda/2$ dipoles. The element spacing is $\lambda/4$ at 2.484 GHz

3.2.3 ESPAR Antenna Beam Pattern

In the previous two subsections, we have described the theoretical foundation of ESPAR antennas, and how to calculate the mutual impedance matrix of an ESPAR antenna using NEC simulator. In this subsection, we investigate the beam pattern generated by an ESPAR antenna.

Firstly, we compared the beam pattern generated using MATLAB[®] code with equation (3.13) with the pattern generated using NEC simulator. The reactance values x_1, x_2, x_3, x_4, x_5 and x_6 are set to $-90 \Omega, 0 \Omega, 0 \Omega, 0 \Omega, 0 \Omega$ and 0Ω , respectively. This produces a sector beam pattern. The three dimensional (3D) current distributions gen-

Figure 3.3: The current distributions on the monopoles of a 7-element ESPAR antenna generated using NEC simulator. The ground plane transforms the monopoles with their image into $\lambda/2$ dipoles. The inter-element spacing is $\lambda/4$. The reactance values x_1, x_2, x_3, x_4, x_5 and x_6 are set to $-90 \, \Omega, 0 \, \Omega, 0 \, \Omega, 0 \, \Omega, 0 \, \Omega$ and $0 \, \Omega$, respectively

Figure 3.4: The radiation patterns of a 7-element ESPAR antenna generated using NEC simulator and MATLAB® code. The inter-element spacing is $\lambda/4$. The reactance values x_1, x_2, x_3, x_4, x_5 and x_6 are set to $-90 \, \Omega, 0 \, \Omega, 0 \, \Omega, 0 \, \Omega, 0 \, \Omega$ and $0 \, \Omega$, respectively

Figure 3.5: The radiation patterns of a 7-element ESPAR antenna generated using NEC simulator correspond to different inter-element spacings. The reactance values x_1, x_2, x_3, x_4, x_5 and x_6 are set to $-90 \, \Omega, 0 \, \Omega, 0 \, \Omega, 0 \, \Omega, 0 \, \Omega$ and $0 \, \Omega$, respectively

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erated using the NEC simulator is shown in Fig. 3.3. The color bar indicates the current amplitude. The 2D radiation pattern in the azimuthal plane ($\theta = 90^\circ$) is compared the radiation pattern calculated from MATLAB[®] code using (3.13). As shown in Fig. 3.4, the close match between the two patterns justifies the validity of (3.13).

Figure 3.6: The radiation pattern gains of a 7-element ESPAR antenna generated using NEC simulator correspond to different inter-element spacings. The reactance values x_1, x_2, x_3, x_4, x_5 and x_6 are set to $-90 \Omega, 0 \Omega, 0 \Omega, 0 \Omega, 0 \Omega$ and 0Ω , respectively

So far, we have simulated beam patterns of a 7-element ESPAR antenna with quarter wavelength inter-element spacing. Here, we keep the length of the antenna elements and examine the influence of inter-element spacing to the beam forming performance. From the previously published study [45], it is understood that a larger inter-element spacing (e.g. $\lambda/2$) leads to less beam pattern controllability by loaded reactances. From Fig. 3.5, it can be observed that pattern 1 has a larger back lobe than that of pattern 2 for which the inter-element spacing is $\lambda/4$. Further decreasing of the inter-element

spacing leads to more directive patterns (see pattern 3). However, this tends to be a supergain excitation of the array [121], which means the optimum design is very sensitive to frequency and design variations. This is illustrated in Fig. 3.6. As the inter-element spacing decreases, the pattern gain becomes more sensitive to the frequency.

3.2.4 Adaptive Beamforming Theory

Examining $\mathbf{w}_{ESP}$ in (3.13), we found that it is dependent on the reactance at each parasitic monopole. Desired beam patterns could be formed by tuning the reactances $(x_1, x_2, \dots, x_6)$. This is achieved by changing the control voltages $(V_{cc1}, V_{cc2}, \dots, V_{cc6})$ of the diodes, which are connected to parasitic elements.

Note that (3.13) has the same form as the beamforming equation in array processing literatures [35, 128]. Signals at the antenna receiver are the sum of the *weighted* signal samples at individual antenna sensor elements.

For beamforming based on conventional DBF array antennas, signal samples are obtained at the baseband, and the desired output is the linear combination of these samples. If the objective function is the MSE between the training signals and the combined output, the beamforming is a Wiener filtering problem. The optimum solution of the combining weight is obtained by solving the Wiener-Hopf equation [35, 134]. To reduce the computational complexity, the LMS algorithm is employed to obtain the optimum solution through iteration using the gradient of the MSE objective function which is in a linear quadratic form [134]. On the contrary, the ESPAR antenna has only one RF port. Signals impinging on the antenna elements are combined together due to the strong mutual coupling. The baseband receiver receives the combined signals from the sole RF port. This combination is a nonlinear function of the loading reactance as

shown in the above equations.

Furthermore, for the DBF array antennas, the optimum *weight* obtained is multiplied with the signal sample vector digitally at the baseband. While for ESPAR antennas, the *weight* is tuned by changing the reactance values at antenna elements based on a nonlinear relationship between the control voltages and the beam pattern. Therefore, $\mathbf{w}_{ESP}$ defined in (3.10) is called “equivalent weight vector”. This unique property of the ESPAR antenna initiates our design of the adaptive beamforming algorithm based on a numerical iterative manner as for a nonlinear filtering problem.

3.3 Adaptive Algorithm

As aforementioned, the optimum *weight* has to be obtained numerically. The objective function of the iteration process is the normalized mean square error (NMSE) between the received signal and the desired signal (reference signal). It is simply represented by $1 - |\rho_{yr}|^2$, where ρ_{yr} is the cross correlation coefficient (CCC) between the desired and the received signals [48]. ρ_{yr} is expressed as:

$$\rho_{yr} = E[y(t)r(t)^*] / \sqrt{E[y(t)y(t)^*]E[r(t)r(t)^*]} \quad (3.14)$$

where $r(t)$ is the reference signal. Since the objective function is based on the CCC, and the goal of the iteration is to maximize the CCC to minimize the NMSE, we also call this algorithm a maximum CCC (MCCC) algorithm. The output SINR is approximately represented by $|\rho_{yr}|^2 / (1 - |\rho_{yr}|^2)$. In this optimization problem, the system parameters are the reactance values, which are controlled by the reversely biased (RB) voltages ($V_{cc1}, V_{cc2}, \dots, V_{cc6}$) at tunable varactors. The desired beam pattern is

obtained by changing the reactance values through iterations.

The reported algorithm [47] is based on the SGA [149], a classical deterministic optimization method. It assumes that one knows exactly the loss function (and its derivatives if relevant) and that information is used to calculate the gradient of the optimization search route in a deterministic manner at each step of the iteration. Perturb each parameter of the system; measure the loss function; and determine the gradient with respect to each individual parameter. This is not computationally efficient if the system has a considerable amount of parameters. As for the 7-element ESPAR antenna, there are 6 parameters (load reactances at six parasitic elements). This algorithm hinders us to implement more parasitic elements to improve the performance and impedes the reduction of the iteration process. Furthermore, we do not have the perfect information about the loss function, because the measurement of the loss function is quite *noisy*. This *noise* denoted by ϵ in following equations is not the additive noise, rather it is the randomness and the inevitable error in observing the loss function [153]. For example, we cannot compute the *mean* square error. In practice, it is a squared error over a finite length of observations. In our algorithm the observation length of the loss function is 50 data bits. Therefore, the optimization process of the ESPAR antenna should be modelled more properly as the stochastic optimization problem, especially in the Kiefer-Wolfowitz setting [153], in which one uses only *noisy* measurements of the objective function to estimate the gradient of the objective function.

3.3.1 Signal Model and Assumption

To model the propagation environment and simulate the adaptive beamforming of an ESPAR antenna, we assume a line-of-sight (LOS) propagation environment, and the multipath components are not considered. These assumptions give a direct insight into

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the beamforming ability of the ESPAR antenna.

Signals are assumed to come in the azimuth plane. When there are total K signals impinging on the antenna, the received RF signal $y(t)$ is the sum of all the RF signals from individual signal sources:

$$y(t) = \sum_{k=1}^K \mathbf{w}_{ESP}^T \boldsymbol{\alpha}(\phi_k) s_k(t) + n(t) \quad (3.15)$$

ϕ_k is the azimuthal angle of the k th impinging signal. $n(t)$ is the added white Gaussian noise (AWGN) component with power σ^2 . $s_k(t)$ denotes the k -th signal wave impinging on the antenna array. Here we define a scalar $\tilde{\alpha}(\phi_k)$ as $\mathbf{w}_{ESP}^T \boldsymbol{\alpha}(\phi_k)$.

3.3.2 Optimization Procedure

The objective function is the NMSE. It is represented as:

$$L(X(n)) = 1 - |\rho_{yr}|^2 \quad (3.16)$$

where $X(n)$ is the reactance values $(x_1, x_2, \dots, x_6)$ at the n -th iteration stage. ρ_{yr} is the CCC between the received signal and the reference signal. The estimated gradient of the optimization problem is:

$$\widehat{\mathbf{g}}_n = (\widehat{L}_n^{(+)} - \widehat{L}_n^{(-)}) / 2c(n)\Delta_n \quad (3.17)$$

where $\widehat{}$ denotes estimated values. $\widehat{L}_n^{(\pm)}$ is the estimated value of the objective function with the parameters (6 reactance values) simultaneously perturbed to the positive and

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negative directions, respectively, at the n -th iteration.

$$\widehat{L}_n^{(\pm)} = L(X(n) \pm c(n)\Delta_n) \pm \epsilon^{(\pm)} \quad (3.18)$$

$$\Delta_n = \begin{bmatrix} \Delta_n^{(1)} & \Delta_n^{(2)} & \Delta_n^{(3)} & \Delta_n^{(4)} & \Delta_n^{(5)} & \Delta_n^{(6)} \end{bmatrix} \quad (3.19)$$

Each entity is selected based on the Bernoulli distribution. Their values are either 1 or -1 with a probability of $1/2$ [150]. $c(n)$ is the perturbation size at the n -th iteration:

$$c(n) = C/(n+1)^\gamma \quad (3.20)$$

C is a constant. ϵ is the *noise* component as aforementioned. Please note that, although the observation of the loss function is a scalar, $\mathbf{g}_n$ is a vector with 6 entities. New reactance values are updated as:

$$X(n+1) = X(n) - a(n)\widehat{\mathbf{g}}_n \quad (3.21)$$

where, $a(n)$ is the updating step size, a and A are the constants. α and γ are non-negative coefficients.

$$a(n) = a/(A+n+1)^\alpha \quad (3.22)$$

Only two measurements of the loss function are needed to get the gradient g_n . This is obviously because all the parameters are always perturbed together instead of one by one as in the steepest gradient optimization process [47, 149]. The step size goes to zero as $n \rightarrow \infty$. This characteristic allows an average of the *noise* in the long run [153]. $c(n)$ also follows $\sum_{n=0}^{\infty} c(n) = \infty$ to satisfy the convergence condition.

3.3.3 Selection of Parameters

In this chapter, we select the values of α , γ , C , and a from [150]. It is recommended to choose α and γ as 0.602 and 0.101, respectively (practically effective and theoretically valid). C is set at the level approximately equal to the standard deviation of the measured noise in the objective function. a is selected together with A while satisfying that the magnitude of the gradient times $a/(A+1)^\alpha$ approximately equals to the smaller perturbation size of the parameters in early iterations. In our design, the parameters are the control voltages ($V_{cc1}, \dots, V_{cc6}$) at the tunable reactances. We selected 50 as the maximum number of iterations in the algorithm, because it is not too long, but is sufficient for the algorithm to converge.

Although the selection of parameters is based on the guideline in [150], in the experiment we cannot directly use the parameters that are optimal to simulation environment as differences between two systems exist. The practical parameters have to be selected through trials and errors. In our system, we set $a = 8002000$, $C = 280$, $A = 100$,

3.4 Simulation Results

3.4.1 Speed of Convergence

In the simulation the DOA of the desired signal is 0° . The length of each data sample block P is 50 bits. The total number of iterations N is 50. For the proposed algorithm based on SPSA, the total length of training signals required is $P \times 2 \times N$ [150]. Based on the varactor catalog data [154], we model the relationship between the digital control

voltage V_{cc_i} and the reactance x_i ($i=1, \dots, 6$) with a linear equation:

$$x_i = -0.0217V_{cc_i} - 49.21. \quad (3.23)$$

First, we study the convergency of the algorithm. The interference is impinging on the antenna from 90° . We set the SNR 30 dB and vary the SIR from -5 dB to 5 dB. Secondly, we vary the SNR keeping the SIR at 0 dB. The learning curves of CCCs for different settings are plotted in Fig. 3.7.

When we fix the SNR at 30 dB and vary the SIR from 5 dB to -5 dB, the algorithm converges to a point where the output SINR is more than 15 dB. At the beginning of the iteration, the value of ρ_{yd} is influenced by the input SIR. When the SIR is kept at 0 dB and lower the SNR to 20 dB, ρ_{yd} still reaches more than 0.95 (solid line with plus signs), which corresponds to an output SINR of 10 dB. However, the algorithm could not converge when the SNR is ≤ 10 dB. In a high noise level environment (SNR ≤ 10 dB), it is difficult to estimate correctly the CCC between the training signals and the received signals with only 50 bit samples. To ensure a correct calculation of the CCC, the sample length P needs to be increased. In Fig. 3.7 the CCCs with 100-bit and 1,000-bit block samples are plotted, respectively.

We also note that the performance of the algorithm varies at each time. This is because of the randomness of the CCC between the shot training signal and the signals transmitted by the interference. Their correlation is varying randomly rather than being zero. This phenomenon is also observed in other experiments and field tests [155–157]. Averaged results for different environment settings are listed in table 3.1. Simulation shows that 50 iterations are more than sufficient for the algorithm to converge.

Figure 3.7: Cross correlation coefficient between the received signal and the training signal in different environment settings for the proposed ESPAR SPSA beamforming algorithm. The azimuthal DOAs of one desired signal and one interference signal are 0° and 90°, respectively.

Table 3.1: Averaged Simulation Results (Averaged over 10 Times of Executing the Program) For Different Environment Settings. $P = 50$. $N = 50$. DOAs of One Desired Signal and One Interference are 0° and 90° in the Azimuth Plane, Respectively. (Results in the Shaded Region is Obtained When $P = 1,000$)

SIR (dB)	SNR = 30 dB		SNR (dB)	SIR = 0 dB	
	Output SINR (dB)	Formed Null (dB)		Output SINR (dB)	Formed Null (dB)
-5	15	24	10	2	16
0	16	21	20	10	17
5	18	17	30	16	21

Figure 3.8: Cross Correlation coefficient between the received signal and the training signal for LMS adaptive beamforming algorithm. The azimuthal DOAs of one desired signal and one interference signal are 0° and 90° , respectively. The length of each data sample block is 50 bits. The step size is 0.07.

3.4.2 Comparison with Standard LMS Beamforming

To give a comparison on the convergence of the proposed beamforming algorithm for reactively controlled directive array with the standard LMS beamforming algorithm based on DBF array antennas of the same array geometry of the 7-element ESPAR antenna, we plot the learning curve of the LMS algorithm in Fig. 3.8. The mutual coupling of the array antenna is also considered. The length of each data sample block is 50 bits. The step size is 0.07. From the figure we can see that the LMS beamforming algorithm requires comparatively less iterations to achieve the same CCC (the same output SINR). However, the penalties are the storing of seven times the data as for the ESPAR beamforming algorithm and the increase of the computational

complexity. Needless to say the complexity of the circuit and the power consumption as aforementioned.

3.4.3 Null Forming Ability

To justify the null forming ability of the proposed algorithm, the formed beam patterns are also plotted in Fig. 3.9. We observe that the algorithm still forms deep nulls even when the interference level increases. This is because the CCC between the received signals and the training signals is mostly influenced by the interference instead of the AWGN. While the MCCC based algorithm is searching to maximize the CCC, the interference is getting suppressed.

Figure 3.9: Averaged ($P = 1,000$) simulated beam patterns (normalized gain in dB) in the azimuth plane for different SIRs. SNR = 30 dB.

We also set the azimuthal DOAs of the interferer at 180° , 225° and 315° , respectively,

Table 3.2: Simulation Results (Averaged over 10 Times of Executing the Program) for Different DOAs in the Azimuth Plane of the Interference Signal. DOAs of the Signal of Interest is 0°.

Interferer's direction in the azimuth plane		90°	180°	225°	315°
SNR 20 dB	Output SINR (dB)	10	8	9	6
	Formed Null (dB)	17	15	16	11
SNR 30 dB	Output SINR (dB)	18	11	14	11
	Formed Null (dB)	22	14	18	13

Figure 3.10: Spatial correlation coefficients at the ESPAR antenna between the desired signal ($\phi_1 = 0^\circ$) and the interference signal with respect to different azimuthal angles (ϕ_2 is from 0° to 360°).

while keeping the DOA of the desired signal at 0° . The averaged results are produced in Table 3.2. Among four cases, the performances are the best when the interferer is at 90° or 225° . This could be explained by examining the spatial correlation between the ESPAR steering vector for the desired signal's direction ϕ_1 and the one for the interferer's direction ϕ_2 . The spatial correlation coefficient at the ESPAR antenna is plotted in Fig. 3.10, where the DOA of the desired signal is fixed at 0° , the DOA of the interferer is changed from 0° to 360° . The performance degrades when the angular separation between the desired signal and the interference is getting small. The received signals at the RF port could be represented as:

$$y(t) = \mathbf{w}_{ESP}^T(s_1(t)\boldsymbol{\alpha}(\phi_1) + s_2(t)\boldsymbol{\alpha}(\phi_2)) + n(t) \tag{3.24}$$

Analogous to the duplicate channel multi-user division in the SDMA scheme [78, 80], the less spatial correlation between the desired signal and the interference signal is, the better the null forming is to the interference.

Table 3.3: Simulation Results (Averaged over 10 Times of Executing the Program) Where There are Two Interferers. Case I: Interferers' Directions are 90° and 270° . Case II: Interferers' Directions Are 225° and 270° . DOA of the Desired Signal is 0° .

	SNR (dB)	Formed null at interferer 1's direction (dB)	Formed null at interferer 2's direction (dB)	Output SINR (dB)
Case I	20	15	13	9
	30	16	16	15
Case II	20	14	6.5	8
	30	17	10	12

Finally we increase the number of interferers to two. The powers of the interference

Figure 3.11: Simulated beam patterns (normalized gain in dB, $P = 1,000$) in the azimuth plane when there are two interferers. SIR = 0 dB. SNR = 30 dB. $N = 50$. Case I: Interferers' directions are 90° and 270°. Case II: Interferers' directions are 225° and 270°.

signals are equally set to be half of the desired signal's power. Therefore, the SIR is kept 0 dB. The SNR is set to be 30 dB. Directions of two interferers are 90° and 270° in the first case, and 225° and 270° in the second case. Averaged results over 10 times of executing the algorithm are listed in Table 3.3. Although the performances with respect to the output SINR and the formed null degrade around 2 dB comparing to the same condition with one interferer, about 10 dB nulls are still formed at the interferers' directions. Beam patterns for the two cases when the SNR is 30 dB are plotted in Fig. 3.11. As it can be seen in the figure for case II, more than 10 dB null is formed at interference 1 which is at 225°, and about 9 dB null is formed at interference 2 which is at 270°.

3.5 Conclusions

Based on the knowledge gained from the thorough literature review, we notice that parasitic array antennas are suitable for power-consumption limited mobile terminal applications. Two categories of parasitic array antennas are examined. In comparison with switched parasitic array antennas, reactively controlled directive array can realize a smoother beam pattern steering, thus achieving a better interference suppression performance. Therefore, we develop a fast beamforming for a reactively controlled directive array antenna named “ESPAR”. The beamforming technique, maximizing the CCC between desired signals and reference signals based on the SPSA theory, adaptively adjust the tunable reactance at each parasitic element to form a beam pattern that gives efficient interference suppression. Simulation studies are carried out to examine the beamforming ability, convergence, and interference suppression of the proposed technique. Based on the simulation studies in this chapter, the following conclusions can be made.

1. The reactively controlled directive array antenna is an analog adaptive antenna. Adaptive beamforming is achieved by adaptively adjusting the load reactance of parasitic elements. The problem of beamforming cannot be modelled as a linear filtering problem as for DBF array antennas. Instead of solving the linear equation to obtain the Wiener solution, iterative approach is employed to achieve the optimum solution.
2. In DBF array antennas, mutual coupling influences the beam steering ability and degrades the beamforming and interference suppression performances. On the contrary, the reactive controlled directive array antennas take the advantage of strong mutual coupling between neighboring elements to achieve beamforming.

3. We also compare the convergence speeds between the adaptive algorithm for a 7-element ESPAR antenna and the standard LMS beamforming algorithm using a 7-element circular array antenna with the same geometry as that of the ESPAR antenna. Results reveal that ESPAR antennas converge slower than DBF array antennas. However, the ESPAR antenna store much less signal datum and has much lower power consumption and system complexity. The sufficient beamforming ability and low power consumption feature make the ESPAR antenna a strong candidate for mobile terminal applications.
4. The beamforming algorithm suppresses the interference by maximizing the CCC between the desired signal and the reference signal. It is more efficient to combat against the interference than to the noise. Therefore, the algorithm is suitable and efficient for the interference dominant environment, for example in ad hoc network applications where the link quality and throughput at a node are mainly limited by the interferences from the neighboring nodes (mobile terminals). In a moderate condition (SNR is around 20 to 30 dB), the algorithm could achieve an output SINR of around 10 dB within a few iterations. Thus, a sound link ($\text{BER} \leq 10^{-3}$) could be established quickly. When the SNR is lower than 10 dB, the method breaks down.
5. When the DOA of the impinging interference signal is close to that of the desired signal, the null forming ability degrades. This is due to the high spatial correlation between the desired signal and the interference signal.

After presenting the simulation study, in the next chapter, we will present experiment testbed and results.

Chapter 4

Experimental Evaluation of the Adaptive Beamforming Algorithm

***Abstract:** To verify the proposed adaptive beamforming algorithm of the reactively controlled directive array, the testbed was set up in a radio anechoic chamber where the SNR is 20 dB and the SIR is 0 dB. Using an ESPAR antenna working at 2.484 GHz as the receive adaptive antenna, the algorithm converges within 5,000 training signal symbols, and achieves an output SINR of 10 dB. Therefore, a sound communication link ($BER \leq 10^{-3}$) is established. The excellent agreement between the simulation and the experimental results reflects the effectiveness of the proposed beamforming algorithm.*

4.1 Introduction

As we have discussed in the previous chapters, the evolution of wireless communications systems calls for new services and better link quality. The demand has excited the applications of smart antennas. Steering directive beam patterns towards desired

signals and forming null patterns in the direction of interference perform spatial filtering. Therefore the spatial domain is exploited to increase the spectrum efficiency and system capacity.

The implementation of DBF array antennas is costly, especially at mobile terminals, where power consumption and cost are the main limitation factors. Parasitic array antennas are potential candidates as mobile terminal smart antennas. Two categories of parasitic array antennas have been studied. Reactively controlled directive array antennas can achieve a smoother beam steering, in comparison with their counterpart—switched parasitic array antennas, thus improving the performance.

An adaptive beamforming algorithm of a reactively controlled directive array antennas has been presented in Chapter 3. Simulations have been carried out to examine the performance of the algorithm. To investigate the feasibility of implementation of the antenna and the algorithm for real application, testbed evaluations are presented in this chapter. The adaptive antenna is used for receive beamforming. After execution of the adaptive beamforming algorithm that we have proposed, the formed beam patterns are measured and the suppression to interference is examined.

4.2 Testbed

The functional block diagram of the testbed is shown in Fig. 4.1 [34]. It consists of a transmitter block, a receiver block, a baseband digital block, and the antenna motor control block (not shown in the figure). Followings are the detailed descriptions of different functional blocks of the testbed.

Figure 4.1: ESPAR adaptive beamforming algorithm experiment testbed functional block diagram.

4.2.1 Transmitter and Receiver Block Setup

The data, a 13-order m -sequence, is generated at the field programmable gate arrays (FPGA) m -sequence generator-1 and generator-2 for the transmitter and for generating reference signals at the receiver, respectively. The inner signal reference certifies that two generators produce the same sequence. At the transmitter part, with this m -sequence as modulating signals the vector signal generator produces a binary phase-shift keying (BPSK) RF signal at 2.484 GHz, the channel-14 in IEEE 802.11b WLAN standard. Signals are transmitted with a 17 dBi horn antenna (Scientific Atlanta Standard Horn antenna. Model: 12-17.0) pointing to the ESPAR antenna. It is about 5 meters above the ground floor. The SNR is adjusted by tuning the power of the transmitter's signal.

The ESPAR antenna, as the receive antenna, is mounted on a test positioner. (Fig. 4.2 shows the prototypes of ESPAR for beam pattern tests.) The received signals of the ESPAR antenna pass through a low noise amplifier (LNA). The amplified signals are directly converted down to 70 MHz intermediate frequency (IF) signals at the mixer with a 2.414 GHz local oscillator signal. After passing through the 2nd amplifier and the bandpass filter (BPF), the signals are splitted into in-phase (I) and quadrature (Q) components. Two ADCs (500 KHz sampling rate) simultaneously convert the I and Q components into 12-bit digital signals.

In our experiment, the interference is randomly generated by the signal generator and is transmitted with a 14 dBi Yagi-Uda antenna (WLE-HG-DYG Yagi). The SIR is tuned by adjusting the power of the interference signals. The interference signal's DOA is set by placing the Yagi-Uda antenna at different places in the radio anechoic chamber.

Figure 4.2: Fabricated ESPAR antenna prototype at ATR for beam pattern tests.

4.2.2 Baseband Digital Block

Sampled digital signals from the A/Ds of the I/Q channel are sent to the TMS320C6701 DSP chip, where the algorithm is executed. With the reference signals, the CCC between the received signals and the reference signals ρ_{yr} is calculated and displayed on the computer control panel. The DSP chip outputs the updated reactance 12-bit digital control voltages ranging from 2,047 to $-2,048$ (decimal value of the 12-bit binary digit). 6 digital-to-analog (D/A) converters generate analog control voltages (V_{cc1} , V_{cc2} , ..., V_{cc6}) from 20 V to -0.5 V correspondingly.

The reference signals and the received signals from the I/Q channels are displayed on

Figure 4.3: Setting of the guard time at DSP chip. For each iteration, 50×2 bits are required to calculate the gradient.

the oscilloscope for a real time monitoring. The CCC, the received signals, the reference signals and the control voltages of the load reactances are saved for post-processing. For calculating the gradient at each iteration stage, the training signals should be the same when perturbing the parameters to the positive and negative directions. This is achieved by setting a *guard time* on the FPGA generated 13-order *m*-sequence. Since we use 50-bit signal samples to calculate the CCC ρ_{yr} , our guard time is $2^{13} - 1 - 50$ bits. The DSP chip waits during the guard time until it gets the same 50-bit prefix of the 13-order *m*-sequence. Fig. 4.3 illustrates this implementation.

Due to the hardware limitation, we set a relatively long time for DSP processing and D/A conversion. This eliminates the influence from the hardware processing delay and assures that we get exactly the same training signals for the gradient calculation during the experimental stage. For real implementations with high speed D/A converters (DACs) and DSP chips, the *guard time* could be greatly reduced to a few bit duration.

4.2.3 Implementation of the Algorithm

When implementing the algorithm, the hardware limitations need to be considered. For our fabricated ESPAR antenna, two diodes (1SV287, V_R : -20 V to 0.5 V , C : 0.7 pF to 9.0 pF , $C_{\text{max}}/C_{\text{min}}$ is 7.6 at 1 MHz) are positioned in parallel and terminate at each parasitic monopole. The analogue control voltages V_{cc} (from 20 V to -0.5 V) impress on reverse polarities of the diodes (see Fig. 4.4) to make them reverse biased. The two parallel diodes are tuned to function as one tunable reactance that gives a reactance from $-45.8j\ \Omega$ to $-3.6j\ \Omega$ [34]. The reactance value cannot be tuned over a wide range, because of the physical limitation of the diode and the constraint imposed on the input impedance of the RF port.

Figure 4.4: Vertical cross section view of the ESPAR antenna and a reactance circuit of one of the six parasitic elements. The length of each monopole is one-quarter wavelength ($\lambda/4$) of the transmitting RF signal with the frequency of 2.484 GHz .

The beam steering of the ESPAR antenna is based on the strong coupling among the active center element and 6 varactor loaded parasitic elements. The impedance and the resonant frequency of the antenna array elements vary as the beam pattern is tuned.

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This may lead to an unacceptable level of input return loss (S_{11}) for practical applications. A constraint is imposed on the practical range that is from $-45.8j\ \Omega$ to $-3.6j\ \Omega$. The corresponding digital control voltage (Decimal value of 12-bit binary representation of analog control voltage from 20 V to -0.5 V) is from 2,047 to $-2,048$. The measured return loss is shown in Fig. 4.5. For 0° sector pattern $V_{cc1} = 2047$, $V_{cc2} = 2047$, $V_{cc3} = -2048$, $V_{cc4} = -1506$, $V_{cc5} = -2048$ and $V_{cc6} = 2047$. For 30° sector $V_{cc1} = 2047$, $V_{cc2} = 2047$, $V_{cc3} = -2048$, $V_{cc4} = -1939$, $V_{cc5} = -1967$ and $V_{cc6} = -2048$. Within the range the antenna has -10 dB return loss (S_{11}) bandwidth of 30 MHz centered at the desired resonant frequency as the beam pattern is tuned over 360° in the azimuth plane [158]. This proves that the antenna does not suffer significant scan blindness.

Figure 4.5: The measured return loss (S_{11} parameter) over frequencies.

The constraint on the range of the tunable reactances at the parasitic elements is realized by adding two limits on the control voltages in the algorithm implemented at the DSP board.

One constraint is just before the measurement of the loss function with the perturbation. When the program generates the random Δ_n and perturbs the reactances to the positive direction, the program checks the resulted reactance values. In case any reactance is beyond the range, that reactance is resumed to the initial value before the perturbation. The same check procedure is carried out for the perturbation to the negative direction. The maximum perturbation size is always smaller than the range of the reactance. In another word, each reactance value is within the range at least for one perturbation. The gradient with respect to $X(n)$ is obtained with one measurement of the loss function with the perturbed parameters and one without the perturbation as in (3.17) [150]. This is also a valid estimation of the gradient [150]. Another constraint is imposed on the updating of the reactance values $X(n)$. If the updating of the reactances results in a value exceeding the range, the parameter is assigned the boundary value.

4.3 Experimental Results

In this section, we discuss the results of the experiment carried out in the radio anechoic chamber. The placements of the test equipments are shown in Fig. 4.1. The transmitter horn antenna, the receiver ESPAR antenna and the interference generating Yagi-Uda antenna are placed in the same height level. Signals are assumed to be in the same azimuth plane. The number of iterations is 50 ($N = 50$). The length of the data sample block is 50 bits ($P = 50$). Therefore, the length of total training signals is $P \times 2 \times N = 5,000$ bits. The SNR at the chamber is set at 20 dB, whereas the SIR is kept 0 dB. The DOA of the desired signal is always 0° .

In the first test, the interference is coming from 90° in the azimuth plane. The output

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SINRs for 10 tests are shown in Fig. 4.6. In most cases, the output SINR exceeds 10 dB. The average output SINR is around 15 dB. Results exhibit a standard deviation of 5.4 dB. This is due to the short data sample block used to calculate the correlation coefficient as we have mentioned in the previous section. One learning curve of the test is plotted in Fig. 4.7. In this case, the CCC ρ_{yd} reaches 0.96. The corresponding output SINR is 11 dB.

Figure 4.6: Output SINRs for 10 tests. The average output SINR is 15 dB. The standard deviation is 5.4 dB.

We also compare the proposed algorithm with an ESPAR antenna beamforming algorithm [47] which employs the well-known steepest gradient algorithm (SGA) [149]. The CCC curves for both algorithms are plotted in Fig. 4.7. For the proposed algorithm, we set $P = 50$. For the beamforming algorithm with SGA, we set $P = 100$. These values are the experimentally tested minimum length of data sample block for each algorithm to be stable. The reason why the SPSA beamforming algorithm could use much shorter observation sample block is that the effect of the *noise* which is the

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Figure 4.7: Comparison between the SGA beamforming algorithm and the SPSSA beamforming algorithm over number of training bits. For the SPSSA beamforming algorithm, the CCC converges to 0.96 within $P \times 2 \times N = 50 \times 2 \times 50 = 5,000$ bits. For the SGA beamforming algorithm, the number of the training signals is $P \times 7 \times N = 100 \times 7 \times 50 = 35,000$ bits.

randomness of the observation of the objective function is eliminated by allowing the parameter updating step size (see (3.22)) to go to zero as $n \rightarrow \infty$ [153]. While for the classical deterministic SGA, it is kept constant [47, 149].

With the same number of iteration ($N = 50$), the proposed algorithm uses $P \times 2 \times N = 50 \times 2 \times 50 = 5,000$ bits, while the SGA beamforming algorithm uses $P \times 7 \times N = 100 \times 7 \times 50 = 35,000$ bits. If the number of training signals for SGA beamforming ($P \times 7 \times N$) is limited to around five thousand bits, N has to be set 8 with $P = 100$. As shown in the figure, within roughly the same length of training bits, it could not converge. This is because 8 observations of the objective function are required to calculate the gradient at one iteration [47]. With a limited number of training bits, the algorithm cannot converge within a few iterations.

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However, the beamforming algorithm with SGA converges to the optimal point within 50 iterations if a long training signal is available. Because the *noise* effect is eliminated by using longer data sample block, for example 1,000 bits. In Fig. 4.8, the performances of the SGA beamforming algorithm and the proposed SPSA beamforming algorithm are compared using the same number of iterations. From the figure, we could observe that SGA converges to the optimal point. But the total number of training signal bits is $P \times 7 \times N = 1,000 \times 7 \times 50 = 350,000$. This could be impractical for real applications.

Figure 4.8: Comparison between the SGA beamforming algorithm and the SPSA beamforming algorithm over the number of iterations. $N = 50$ for both algorithms. For the SPSA beamforming algorithm, $P = 50$; The total number of training bits is $P \times 2 \times N = 5,000$. For the SGA beamforming algorithm, $P = 1,000$; The total number of training bits is $P \times 7 \times N = 350,000$.

The digital control voltages ($V_{cc1}, \dots, V_{cc6}$) for one test of the proposed algorithm at different iteration stages are plotted in Fig. 4.9. The forming of the null with iterations is clearly shown in Fig. 4.10. From Fig. 4.9 we can see that the control digital voltages V_{cc1} and V_{cc6} converge to positive values. The reactance values of x_1 and x_6 are tuned to high capacitance (see (3.23)). Others are tuned to be negative which result in low

Figure 4.9: Learning curves of the control voltages (V_{cc1} , ..., V_{cc6}) (Decimal value (2,047 to $-2,048$) of 12-bit binary representation of analog control voltage (20 V to -0.5 V)). The cross correlation coefficient reaches 0.98. The output SINR is 13 dB.

capacitances. Among them, the capacitances x_2 and x_3 are the smallest. This could be clearly explained with a look at the geometry (Fig. 3.1) of the ESPAR antenna. Because the desired signal is at 0° in the azimuth plane and the interference is at 90° . The ESPAR antenna, by tuning the reactance values adaptively to the signal environment, exhibits more directive beam in 0° direction while nulls in 90° (see Fig. 4.10). As the load capacitance values are varied, the parasitic monopoles' resonant frequencies change, thus the electrical lengths of the elements vary. The parasitic elements surrounding the active central element act like the directors or reflectors respectively as in a Yagi-Uda antenna to maximally receive the desired signals. Therefore the first element (#1 in Fig. 3.1) is tuned to be a director while the second (#2) and third (#3) elements are adjusted to be reflectors to reject the interference. For DBF array antennas, beamforming is achieved by summing the (phase and amplitude) weighted signals. MBF is based on the same mechanism by move the weighting up to the RF front end.

The constellations of the received base band BPSK signals before and after the itera-

Figure 4.10: Formed successive beam patterns at different iteration stages (Normalized antenna gain in dB). The formed null is around 19 dB. The CCC reaches 0.98. The output SINR is 13 dB. $P = 50$, $N = 50$. (#– positions of 7 elements in an ESPAR antenna.)

Figure 4.11: BPSK signal constellations at the beginning (a) and the end (b) of the iterations. The cross correlation coefficient reaches 0.98. The output SINR is 13 dB.

Figure 4.12: The measured azimuthal beam pattern ($P = 1,000$). DOAs of the desired signal and one interference signal are 0° and 225° , respectively.

tions are shown in Figs. 4.11 (a) and (b), respectively. At the beginning of the iteration, the BPSK signal constellation is distorted as shown in Fig. 4.11 (a). After the convergence of the algorithm, the correct constellation is shown in Fig. 4.11 (b). The CCC reaches 0.98. The output SINR is 13 dB.

Next, the DOA of the interference signal is changed to 225° . The averaged output SINR for ten tests is 14.5 dB with a 5.4 dB standard deviation. The averaged beam pattern, obtained with 1,000-bit data samples ($P = 1,000$) is plotted in Fig. 4.12. A 24 dB null is formed towards the interference direction. However, when we move the interferer to 315° , which is only 45° away from the desired signal, the performance degrades. The beam pattern is plotted in Fig. 4.13. The averaged output SINR is 6 dB with 1.4 dB standard deviation. This is because the interference is close to the desired

Figure 4.13: The measured azimuthal beam pattern ($P = 1,000$). DOAs of the desired signal and one interference signal are 0° and 315° , respectively.

signal. Their spatial correlation coefficient reaches 0.7.

We also have tested the beamforming ability when there are two interferences. The SIR is kept at 0 dB. The SNR is still 20 dB. In the previous chapter, the simulation results with the same condition have been shown in Fig. 3.9 and output SINR data is produced in Table 3.3. The averaged output SINRs over 10 experiments with $P = 50$ are listed in Table 4.1. For $P = 1,000$, measured beam patterns for two cases (case I – desired signal is at 0° and interferers are at 90° and 270° ; case II – desired signal is at 0° and interferers are at 225° and 270°) are shown in Figs. 4.14 and 4.15, respectively.

The experimental output SINRs and the null levels (see Table 4.1) agree with the simulation results (see Tables 3.2 and 3.3). Nulls could still be formed even there are two interferences; but when the interferer is close to the desired signal (see Fig. 4.13),

Table 4.1: Averaged Digital Control Voltages. (Decimal Value of the 12-bit Binary Digital Value) after Convergence and the Averaged Performance with $P = 1,000$ (bits). (The Digital Voltage, Ranging from 2,047 to $-2,048$, is Associated to the Analog Bias Voltage from 20 V to -0.5 V.)

Interference(s)' azimuthal direction	Vcc ₁	Vcc ₂	Vcc ₃	Vcc ₄	Vcc ₅	Vcc ₆	Output SINR (dB)	Standard Deviation (dB)	Formed Null (dB)
90°	-48	-1322	-1530	-760	362	1506	15	5.4	19
225°	609	1105	-295	-1353	-1665	-681	14.5	5.6	24
315°	-685	929	-1503	1007	-1419	-623	6	1.4	7
90° & 270°	740	188	-1378	-322	-1344	736	3.7	1.1	7 & 3
225° & 270°	633	1663	-551	-807	-1641	-543	6.6	1.7	15 & 10

or the two interferers are separated with a large angle (see Fig. 4.14), the null forming performance degrades that leads to a lower output SINR. The performance is better when one interferer is at 90° or at 225° as shown in Figs. 4.10 and 4.12; or when the two interferers are close as in Fig. 4.15.

4.4 Performance Analysis

There are a few reasons that limit the beamforming performance of the ESPAR antenna. The first one is the range of the reactance values. As we have mentioned previously, when the desired reactance values exceed the practical reactance range, the value has to be assigned the maximum or the minimum values. It is foreseeable that broadening the reactance ranges brings an improvement in the performance, but there is a practical limitation in this case.

Figure 4.14: The measured azimuthal beam pattern ($P = 1,000$). DOAs of the desired signal and two interference signals are 0° , 90° and 270° , respectively.

The second reason is that the ESPAR has only 7 elements (6 parasitic elements). It is reasonable to increase the number of the parasitic elements. As more parasitic elements could function as directors (as in a Yagi-Uda antenna), sharper beam patterns could be formed.

Finally, the ESPAR antenna does not form sharp beams/nulls when there are more than one interferer. This is because the algorithm does not exploit the spatial information or the channel information. The beamforming is done only based on the maximization of the CCC through the iteration. The performance could be further enhanced by exploiting more channel information, for example, the DOA information by the modified MUSIC algorithm [33,159].

Figure 4.15: The measured azimuthal beam pattern ($P = 1,000$). DOAs of the desired signal and two interference signals are 0° , 225° and 270° , respectively.

The beauty of the algorithm is that when we increase the number of the antenna array elements, the total training signals remains unchanged. Because only two measurements of the objective function are required to calculate the gradient at each iteration stage, which is not dependent on the number of parameters (number of parasitic elements). Therefore, the total number of training bits ($P \times 2 \times N$) remains unchanged. This is the most attractive feature of the proposed algorithm. With more parameters to be optimized, the advantage of SPSA beamforming algorithm over other algorithms becomes more obvious.

4.5 Conclusions

In these two chapters (3 and 4), an economical solution to the deployment of smart antennas at WLAN mobile terminals has been presented. We first described the design of the ESPAR antenna. A fast beamforming algorithm based on SPSA was proposed. The simulation and experiments were carried out to justify the algorithm. In a moderate noisy environment ($\text{SNR} = 20$ to 30 dB, $\text{SIR} = 0$ dB), the algorithm converges within 5,000 data bits, achieving an output SINR of around 10 dB. Thus, a sound communication link ($\text{BER} \leq 0.001$) could be established quickly.

The following conclusion can be drawn from the experimental study

1. Because the algorithm maximizes the CCC between the received signals and the reference signals by iteratively adjusting the digital control voltages of the tunable reactance of the parasitic elements, the implementation of the adaptive beamforming algorithm for reactively controlled directive array antenna does not prerequisite calibration. Strong mutual coupling effect is exploited to realize beamforming.
2. The adaptive beamforming performance is limited by the physical limitation of the tunable reactance and the limited range of the antenna impedance matching and the resonant frequency and the active scan reflection coefficient.
3. The results showed that the proposed method is good at interference suppression. Nulls can be formed at the interferers' directions within a few iterations. This proves that ESPAR antennas are suitable for ad hoc networks or WLAN applications, where the performance is mainly influenced by the interferences from the neighboring mobile terminals (laptops, PDA *etc.*).
4. In case the SNR drops to a low level ($\text{SNR} \leq 10$ dB), the performance with respect

to the output SINR deteriorates. This indicates that MCCC based algorithm with training signals is not sufficient for noise suppression. Unlike the minimum variance beamforming algorithm [128], the algorithm does not limit the total output power, while it only maximizes the CCC between the reference signals and the training signals. To keep a sound communication, power control algorithms should be employed at the transmitter to keep the SNR at a certain level.

Compared with the DBF antenna arrays and MBF antennas, the ESPAR antennas feature low power consumption, low cost and ease of fabrication. The smart antenna is efficient because the beamforming is realized at the analog front end instead of at the digital baseband. The proposed fast beamforming algorithm makes the ESPAR antenna a suitable candidate for the mass implementation of smart antennas, especially at the battery operated wireless mobile terminals.

Chapter 5

Direction Finding with Parasitic Array Antennas

Abstract: *A novel DOA estimation technique that uses the conventional MUSIC algorithm with periodic signals is applied to a single port smart antenna for incoherent waves. Results show that the proposed method gives a high-resolution (1 degree) DOA estimation in an uncorrelated signal environment. The novelty lies in that the MUSIC algorithm is applied to a simplified antenna configuration. Only one ADC is used in this antenna, which features low power consumption, low cost, and ease of fabrication. Modifications of the conventional MUSIC algorithm do not bring much additional complexity. The proposed technique is also free from the negative influence by the mutual coupling among antenna elements. Therefore, it offers an economical way to extensively implement smart antennas into the existing wireless mobile communications systems, especially at power consumption limited mobile terminals such as laptops in wireless networks*

5.1 Introduction

As we have presented in the second chapter, DOA estimation has extensive applications in wireless communications systems, such as PL services and ad hoc networks. Various DOA estimation techniques have been designed based on DBF antennas [26, 37, 137, 138]. However, RF circuit branches connected to the array elements, ADCs and the baseband DSP chip consume a considerable amount of dc power. Furthermore, each channel connected to the array elements has the same structure, so the cost of fabrication increases with the number of array elements [119]. All these factors make the DBF antenna arrays unsuitable for low power consumption and low cost systems and hinder the mass implementation of the smart antenna technologies. Installing DBF array antennas at battery powered laptops or mobile computing terminals within a wireless network are costly.

Recently, direction finding techniques using parasitic array antennas have been proposed in [33, 108, 109, 114, 118, 120, 160–162]. In [108, 109], the direction finding with parasitic array antennas is realized with rotating the beam pattern by switching ON and OFF different set of parasitic elements. In this way the DOA of the impinging waves is estimated by examining the received power from beams pointing to different directions. In [123], a directive beam pattern is rotated twelve times to cover 360° by changing the reactance values at the parasitic monopoles in a reactively controlled directive array antenna. The pattern which gives the maximum received signal power indicates the direction of the impinging waves. The above methods are simple in implementation but gives low resolution direction finding. Preston *et al.* [115] presented another way of direction finding with switched parasitic array antennas. The antenna pattern is rotated four times each with 90° to cover 360° in the azimuth plane. Under the assumption that the beam pattern is known precisely and the rotation does not

influence the pattern, a linear equation between the received signals at four rotations and beam patterns are formed. The signal direction and strength are obtained by solving a set of linear equations. However, the assumption that the beam pattern does not change at each rotation will be invalid if the antenna element's gain, phase, and position error are taken into account. The MUSIC algorithm is applied to a switched parasitic array antenna in [118]. However, the problem has not been intensively studied.

In this chapter, we propose a smart antenna structure which has only one RF port [33, 160, 162]. It is a parasitic array antenna [32]. One central element connected to the sole RF port and a number of surrounding switched and reactively loaded parasitic elements form the array (See Fig. 5.1). We call the antenna "switched and reactively loaded parasitic array (SARLPA)". The system has only one RF port connected to the active central element and one subsequent down conversion circuit. The load reactances are realized by the RB varactor diodes where there is only a small RB leakage current. Therefore, it obviously consumes much less power than a DBF antenna array.

A high-resolution direction finding using the MUSIC algorithm with periodic signals is employed on this single port smart antenna. Activating different beam patterns by switching ON and OFF different parasitic elements, while the signal is periodically transmitted, the received scalar signal from the RF-port can be stacked into a vector over one period. The beam pattern is shifted with a predefined angle. So, for the application of the MUSIC algorithm, it is sufficient to know these angular shifts, without knowing exactly the response to one subset. This is similar to a uniform circular or linear antenna array in which we know the phase shift between consecutive antenna responses, without having to know those responses exactly. The MUSIC algorithm is carried out with the constructed signal vector over one period and the modified steering vector based on the beam pattern angle shift. We called the algorithm "angular-domain

MUSIC". Simulation justifies the proposal. Our heuristic study demonstrates the feasibility of employing the eigenstructure-based direction finding algorithm with a parasitic array that has only one RF port. Another algorithm with the simplified antenna structure is also proposed. Coincidentally, a similar algorithm is independently developed at ATR. The algorithm is called "reactance-domain MUSIC" [161, 163, 164]. After presenting the design of the antenna, we will describe both algorithms in this chapter.

5.2 Design of SARLPA Antenna

Fig. 5.1 depicts the functional block diagram of a SARLPA antenna. Fig. 5.2 depicts the top view of the antenna. One active central element (monopole) is surrounded by 36 parasitic elements on a circle of radius R on the circular grounded baseplate. The length of each monopole (L) and the radius (R) are one quarter wavelength ($\lambda/4$) of the transmitting RF signal. The SARLPA antenna is half wavelength ($\lambda/2$) in diameter. It may not be suitable to be integrated into a mobile phone, but could be easily mounted on a mobile terminal in wireless networks. For an example, the diameter of the array operating at 2.4 GHz is 62.25 mm, which is suitable for mounting on laptops. The baseplate transforms the monopoles with their images to dipoles with a length of $2L$. The central monopole is connected to a RF receiver and each parasitic monopole is loaded with a tunable reactor and is controlled by a switch. The circuitry could be simply realized with a single port (1-36 way) radial switch as in [165]. Since, both the switches and the tunable reactances are implemented at individual parasitic elements, the antenna provides great flexibility in operation. It can be reconfigured into a switched parasitic array [32] or a reactively controlled directive array antenna [30, 45].

Figure 5.1: Functional block diagram of the switched and reactively loaded parasitic array (SARLPA).

Figure 5.2: Top view of a 37-element SARLPA.

As shown in Fig. 5.2, those 36 parasitic monopoles are grouped into 6 subsets. Each subset consists of 6 parasitic elements equispaced on the circle of radius R . The central monopole is designated as $\#0$, whereas the parasitic element $\#m - n$ ($m = 1, \dots, 6$; $n = 1, \dots, 6$) refers to the n th element of the subset $\#m$. The antenna is operated in such a way that at any time only one parasitic monopole subset is switched on and the rest of the subsets are switched off. The switched-off elements do not contribute to beam forming and can be viewed as “transparent” to other switched-on elements. This is realized either by switching the parasitic monopoles to open-circuit as in [115, 117], or by tuning the load reactance value to make the element invisible electrically [166]. The central element is always activated as signals pass through the sole RF port. Therefore, at any time only 6 parasitic monopoles from the selected subset and one active central monopole are in operation. In another word, each subset of the antenna has an equivalent structure of an ESPAR antenna that comprises one active central monopole and 6 parasitic monopoles. The working principle of an ESPAR antenna is presented in Chapter 3.

During operation the first subset (parasitic monopoles $\#1 - 1$, $\#1 - 2$, $\#1 - 3$, $\#1 - 4$, $\#1 - 5$, and $\#1 - 6$) is selected. After a certain period, the subset $\#2$ (parasitic monopoles $\#2 - 1$, $\#2 - 2$, $\#2 - 3$, $\#2 - 4$, $\#2 - 5$, and $\#2 - 6$) is selected. As this process continues, the antenna array is virtually rotating anti-clockwise on its vertical axis. The loads at parasitic monopoles are tuned in the same manner for all subsets. We have selected 36 parasitic and 1 active monopoles to satisfy the modified MUSIC algorithm as applied to the proposed SARLPA configuration. For practical applications, the number of the monopoles could be chosen with the trade-off between the complexity and the performance (number of waves to be estimated, beam pattern directivity *etc.*). Please also be noted that in our proposed DOA estimation technique, the antenna is not used for adaptive beamforming, though it still has the ability to do so. Desired beam patterns

are formed to explore the spatial information of the signal sources.

5.3 MUSIC Algorithm for Single Port Smart Antennas

It is obvious that we can't apply the conventional MUSIC algorithm to the SARLPA antenna, because it has only one RF port. Forming the signal sample vector is impossible at one time. This is the distinction between the proposed modified MUSIC algorithm based on a single port smart antenna and the conventional MUSIC algorithm based on a DBF array antenna. However, we can see from the explanation in the following part that the received signals from the RF port contain signal sources' direction information. And, the signal correlation matrix can still be formed.

In operation, six subsets of parasitic elements are selected sequentially as indicated in Fig. 5.2. The beam pattern rotates by 10° six times consecutively. Therefore, the antenna is virtually rotated. The sample signal vector $\mathbf{y} = [y_1 \ \dots \ y_m \ \dots \ y_6]^T$ is constructed with signal samples from the RF output. Each entity y_m , $m = 1, 2, \dots, 6$ is obtained in the sampling period when the subset $\#m$ is in operation. The maximum number of sampling periods M equals to the number of subsets. For the SARLPA antenna with six subsets, after rotating 6+1 times the beam pattern returns to the beginning position. The load reactance values of parasitic monopoles are the same for all the sampling periods. The load reactance values of parasitic monopoles are the same for all the sampling periods. Therefore, $\mathbf{w}_{ESP}$ defined in (3.10) is the same for the six sampling periods with the antenna rotation. However, the signal directions with respect to the antenna X-Y coordinate (cf. Fig. 5.2 and Fig. 2.13) change as the antenna is virtually rotated. According to (3.15), the received signal sample at each sampling

5.3. MUSIC Algorithm for Single Port Smart Antennas

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period is represented as:

$$y_m = \mathbf{w}_{ESP}^T \sum_{k=1}^K \alpha \left(\phi_k - \frac{2\pi}{36}(m-1) \right) s_k + n_m = \sum_{k=1}^K \tilde{\alpha} \left(\phi_k - \frac{2\pi}{36}(m-1) \right) s_k + n_m \quad (5.1)$$

where the variables without a time argument represent digital samples of the signal.

For simplicity we define

$$\tilde{\alpha}_m(\phi_k) = \tilde{\alpha} \left(\phi_k - \frac{2\pi}{36}(m-1) \right). \quad (5.2)$$

We can see that $\tilde{\alpha}_m(\phi_k)$ contains the k th signal source's direction information ϕ_k . n_m is the noise component at the m th sampling period. Suppose each signal source transmits periodic signals and the length of one signal period is the same as the sampling period. Then received signal samples are the same for all sampling periods. Therefore, the received signal vector $\mathbf{y}$ after six sampling periods can be written into the following matrix multiplication form:

$$\begin{aligned} \mathbf{y} &= \begin{bmatrix} y_1 \\ y_2 \\ y_3 \\ y_4 \\ y_5 \\ y_6 \end{bmatrix} = \begin{bmatrix} \tilde{\alpha}_1(\phi_1) & \tilde{\alpha}_1(\phi_2) & \dots & \tilde{\alpha}_1(\phi_K) \\ \tilde{\alpha}_2(\phi_1) & \tilde{\alpha}_2(\phi_2) & \dots & \tilde{\alpha}_2(\phi_K) \\ \tilde{\alpha}_3(\phi_1) & \tilde{\alpha}_3(\phi_2) & \dots & \tilde{\alpha}_3(\phi_K) \\ \tilde{\alpha}_4(\phi_1) & \tilde{\alpha}_4(\phi_2) & \dots & \tilde{\alpha}_4(\phi_K) \\ \tilde{\alpha}_5(\phi_1) & \tilde{\alpha}_5(\phi_2) & \dots & \tilde{\alpha}_5(\phi_K) \\ \tilde{\alpha}_6(\phi_1) & \tilde{\alpha}_6(\phi_2) & \dots & \tilde{\alpha}_6(\phi_K) \end{bmatrix} \begin{bmatrix} s_1 \\ s_2 \\ \vdots \\ s_K \end{bmatrix} + \begin{bmatrix} n_1 \\ n_2 \\ n_3 \\ n_4 \\ n_5 \\ n_6 \end{bmatrix} \\ &= \begin{bmatrix} \bar{\alpha}(\phi_1) & \bar{\alpha}(\phi_2) & \dots & \bar{\alpha}(\phi_K) \end{bmatrix} \begin{bmatrix} s_1 & s_2 & \dots & s_K \end{bmatrix}^T + \mathbf{n} \end{aligned} \quad (5.3)$$

In the first matrix on the right-hand side of the equation, we define each column as steering vector $\bar{\alpha}(\phi_k)$. It corresponds to the steering vector for a 6-element array antenna as reported in array signal processing literature [128]. But each entity is not

a phase delay of a source signal induced when the plane wave passes the array with an impinging angle relative to the antenna. We call the algorithm “angular-domain MUSIC” and $\bar{\alpha}(\phi_k)$ “SARLPA-direction Vector” to differentiate it from the *steering vector* [128] for conventional antenna arrays. It corresponds to the k th signal source impinging at the azimuthal AOA ϕ_k . Equation (5.3) has the same form as the output from an array antenna with six elements [cf. (2.5)]. The signal correlation matrix at the antenna output is $\mathbf{R}_{yy}$:

$$\mathbf{R}_{yy} = E[\mathbf{y}\mathbf{y}^H] = \frac{1}{P}\mathbf{y}\mathbf{y}^H \quad (5.4)$$

where P is the length of the data sample block. Under the assumption that noise components at each individual periods are modelled as statistically independent and identically distributed (i.i.d) Gaussian variables. Then, we have

$$\mathbf{R}_{yy} = \begin{bmatrix} \bar{\alpha}(\phi_1) & \bar{\alpha}(\phi_2) & \dots & \bar{\alpha}(\phi_K) \end{bmatrix} \mathbf{R}_{ss} \begin{bmatrix} \bar{\alpha}(\phi_1) & \bar{\alpha}(\phi_2) & \dots & \bar{\alpha}(\phi_K) \end{bmatrix}^H + \sigma^2 \mathbf{I}_6. \quad (5.5)$$

$\mathbf{I}_d$ represents a $d \times d$ identity matrix. The signal sources' correlation matrix $\mathbf{R}_{ss}$ is

$$\mathbf{R}_{ss} = E \left[\begin{bmatrix} s_1 & s_2 & \dots & s_K \end{bmatrix}^T \begin{bmatrix} s_1 & s_2 & \dots & s_K \end{bmatrix}^* \right]. \quad (5.6)$$

Based on (5.4), the MUSIC algorithm is executed in a similar procedure as the conventional MUSIC algorithm. Fig. 5.3 illustrates the flowchart. The SARLPA-Direction Vector $\bar{\alpha}(\phi)$ is used to obtain the MUSIC spectrum. The algorithm requires the signals to be periodic. That means the proposed technique cannot be applied when mobile terminals are transmitting message signals, because message signals are quite indeterministic. It could, however, be implemented in a pilot or a command channel or by

assigning periodic codes at message channels. Therefore, we call it *MUSIC algorithm with periodic signals*. Its procedure is described below:

1. Construct the signal sample vector given by $\mathbf{y} = [y_1 \ y_2 \ \dots \ y_6]^T$. y_m , $m = 1, \dots, 6$, is the signal sample when the antenna is at sampling period $\#m$. Given $\mathbf{y}$, the signal correlation matrix $\mathbf{R}_{yy}$ is formed.
2. Eigendecompose the signal correlation matrix $\mathbf{R}_{yy}$ and form the noise subspace $\mathbf{V}_N$ using eigenvectors corresponding to the small eigenvalues.
3. Plot the MUSIC spectrum by calculating:

$$P_{MU}(\phi) = \frac{1}{\bar{\alpha}(\phi)^H \mathbf{V}_N \mathbf{V}_N^H \bar{\alpha}(\phi)} \quad (5.7)$$

4. Finally, extract the impinging waves' direction information.

Besides the additional procedure for forming the signal sample vector, the proposed technique follows the same procedure as that for the conventional MUSIC algorithm. An extra amount of time is required for storing the data to form the signal sample vector.

5.4 DOA Estimation Performance of SARLPA

In our simulation, three BPSK signals with equal power levels are randomly generated. The signal impinging directions are set arbitrarily to 120° , 150° and 300° in the azimuth plane, respectively. During each of the six antenna sampling periods, the length of signal samples is 16 bits. This indicates that P is 16 in (5.4).

Figure 5.3: Flowchart of the conventional and the proposed MUSIC algorithm and the proposed MUSIC algorithm based on a single port smart antenna.

Table 5.1: Setting of six parasitic loads of each subset at the sampling period $\#m$ ($m = 1, \dots, 6$). Once the setting is decided, the values do not change during the process of the DOA estimation.

Case	x_1 Ω	x_2 Ω	x_3 Ω	x_4 Ω	x_5 Ω	x_6 Ω	DOA information
1	30	30	30	30	30	30	Not shown in the spectrum
2	0	30	0	30	0	30	Not shown in the spectrum
3	0	30	0	30	30	30	Shown in the spectrum
4	0	10	0	-30	70	30	Shown in the spectrum

The monopoles on the ground plane of the SARLPA antenna are oriented vertically. We have simulated the polarization pattern of a 37-element SARLPA using MoM NEC simulator. We found that in our simulation, the pattern is linearly polarized parallel to alignment of monopole elements. Therefore, in our study, we only consider impinging waves co-polarized with the antenna. The impinging signals are assumed co-polarized with the antenna. The cross-polarization coupling typically produced by scatters and reflectors in multipath propagation environment [167] is not considered.

5.4.1 Influence of Parasitic Load Reactance

We study four different reactance load settings to investigate their influence on the performance of the DOA estimation. Table 5.1 summarizes the reactance values for these four cases. MUSIC spectra with different reactive load settings in uncorrelated signal environments where the SNR is 40 dB for each individual signal source are presented and explained as follows.

Figure 5.4: MUSIC Spectrum for Case 1: $x_l = 30 \Omega$ ($l = 1, \dots, 6$), SNR = 40 dB and the signal directions are 120° , 150° and 300° in the azimuth plane.

Case 1. Equal Load Reactance of 30Ω

In case 1, we set all the parasitic load reactances to 30Ω . The antenna exhibits a nearly omnidirectional beam pattern. It could not be virtually rotated in a way as described previously. The MUSIC spectrum is plotted in Fig. 5.4. From the figure we can observe that the signal direction could not be extracted.

Case 2. Ordered But Unequal Load Reactances of 0Ω and 30Ω

In case 2, we set the loads to $x_1 = 0 \Omega$, $x_2 = 30 \Omega$, $x_3 = 0 \Omega$, $x_4 = 30 \Omega$, $x_5 = 0 \Omega$, $x_6 = 30 \Omega$. The beam pattern for rotation is shown in Fig. 5.5. The MUSIC spectrum is plotted in Fig. 5.6. It also cannot give the direction information. From the simulation studies in cases 1 and 2, we conclude that we should not set the loads “ordered”, which will result in insufficient beam patterns for rotation.

Figure 5.5: Beam pattern in the azimuth plane for the antenna sampling periods #1 to #6 in Case 2. Load setting: $x_1 = 0 \, \Omega$, $x_2 = 30 \, \Omega$, $x_3 = 0 \, \Omega$, $x_4 = 30 \, \Omega$, $x_5 = 0 \, \Omega$, $x_6 = 30 \, \Omega$.

Figure 5.6: MUSIC spectrum for case 2: $x_1 = 0 \, \Omega$, $x_2 = 30 \, \Omega$, $x_3 = 0 \, \Omega$, $x_4 = 30 \, \Omega$, $x_5 = 0 \, \Omega$, $x_6 = 30 \, \Omega$. SNR = 40 dB. Signal directions: 120° , 150° and 300° in the azimuth plane.

Case 3. Random Load Reactances of 0Ω and 30Ω

In case 3, we arbitrarily set $x_1 = 0 \Omega$, $x_2 = 30 \Omega$, $x_3 = 0 \Omega$, $x_4 = 30 \Omega$, $x_5 = 30 \Omega$, $x_6 = 30 \Omega$. The beam pattern is shown in Fig. 5.7. The signals DOA information is clearly resolved from the MUSIC spectrum as shown in Fig. 5.8.

Figure 5.7: Beam pattern in the azimuth plane for the antenna sampling periods #1 to #6 in case 3. Load setting: $x_1 = 0 \Omega$, $x_2 = 30 \Omega$, $x_3 = 0 \Omega$, $x_4 = 30 \Omega$, $x_5 = 30 \Omega$, $x_6 = 30 \Omega$.

Case 4. Random Load Reactances

Finally, we randomly set the reactance values to $x_1 = 0 \Omega$, $x_2 = 10 \Omega$, $x_3 = 0 \Omega$, $x_4 = -30 \Omega$, $x_5 = 70 \Omega$, $x_6 = 30 \Omega$. The MUSIC spectrum is plotted in Fig. 5.8. From the figure we can see that if the reactance is not set ordered, they only slightly influence the MUSIC spectrum but not the performance. All the spectra reach the peak values at the respective signal directions. The beam pattern for this case is shown in Fig. 5.9. Note

Figure 5.8: MUSIC spectra for cases 3 and 4. SNR = 40 dB. The load settings are listed in Table 5.1. Signal directions: 120°, 150° and 300° in the azimuth plane.

that the proposed technique is used for DOA estimation and not for beam forming. So the beam pattern does not necessarily correspond to the signals' directions.

From the above investigations, we conclude that the manner of load settings greatly influences the algorithm's DOAs estimation performances. When setting the reactive loads, we should avoid setting them ordered, which produces symmetric azimuthal beam patterns. With a symmetric pattern, signals impinging from different directions may produce the same output at the RF port. It is analogous to image peaks problem for DOA estimation using a linear array antenna. Impinging waves from two sides of the array with the same AOA cause the identical time delays on array elements. Rather, we can set the reactance values arbitrarily to precisely estimate the DOA information.

Forming directive beam patterns is not important to the DOA estimation performance. The technique is dependent of the angular shifting of the beam pattern. If only the pattern is not symmetric, the DOA can be estimated correctly. However, the load set-

Figure 5.9: Beam pattern in the azimuth plane for the antenna sampling periods #1 to #6 in case 4. Load setting: $x_1 = 0 \Omega$, $x_2 = 10 \Omega$, $x_3 = 0 \Omega$, $x_4 = -30 \Omega$, $x_5 = 70 \Omega$, $x_6 = 30 \Omega$.

ting that gives deep nulls in radiation patterns in certain directions should be avoided for the DOA estimation. Because, signals coming from that regions could be received with low power levels, or even be nulled out. Consequently, the DOA information of signals in this null region could not be extracted.

5.4.2 Performance in Different SNRs Environments

We also simulate the MUSIC spectra in signal environments with different SNRs. We suppose that all signal sources are totally uncorrelated and have the same power. The SNRs for all signal sources are equal. The MUSIC spectra with three different SNRs from 0 dB to 40 dB are shown in Fig. 5.10. The performance of our proposed MUSIC algorithm degrades as SNR decreases, but all three MUSIC spectra reach nearly the

Figure 5.10: MUSIC spectra in different SNRs signal environments. The load setting is the same as in case 4. Signal directions: 120°, 150° and 300° in the azimuth plane.

same peak values at the signal directions.

5.4.3 Performance in a Multipath Propagation Environment

Our study assumes a LOS propagation environment (Ricean channel). In a macroscopic area, where PL services such as intelligent transportation services are employed, base station antennas are mounted well above the roof of the buildings to provide LOS propagations between base stations and mobile terminals [64]. Ideally, signals arrive at the antenna without phase delays. The plotted MUSIC spectrum is shown in Fig. 5.11 with a “x” marked solid line. In a practical wireless communication environment, multipath propagation delays of the signals are expected. Received signals contain the LOS component and the multipath components from each far field signal source. We assume that LOS components are uncorrelated with the multipath components [168] and the

Figure 5.11: MUSIC spectra in multipath environment. The load setting is the same as in case 4. Signal directions: 120°, 150° and 300° in the azimuth plane.

LOS components have much higher signal power level, we could still reasonably represent the received signals at the antennas as a summation of only the LOS components from different directions with phase differences induced by the propagation delays. The angular distribution in this situation can be modelled as Gaussian distributed. Recent measurement supports Laplacian distribution being a better fit than the Gaussian distribution [66].

$$f(\phi) = \frac{c}{\sqrt{2}\sigma} e^{-\frac{\sqrt{2}|\phi|}{\sigma}}, -\pi \leq \phi \leq \pi \quad (5.8)$$

where

$$c = \left(1 - e^{-\frac{\sqrt{2}\pi}{\sigma}}\right)^{-1} \quad (5.9)$$

is the normalizing constant and σ controls the angular spread as shown in Fig. 5.12.

In a microscopic area or indoor applications, the base station antennas are mounted

Figure 5.12: Laplacian angular distribution with different angular spread.

below the rooftop level to confine the signal coverage into a small local area [169]. In this case, the base stations are located in a rich scattering environment. There are not only LOS component but also multipath components. In this situation, the angular distributions at both the base station and mobile station are normally assumed uniformly distributed over the azimuth plane. Fig. 5.13 shows the impinging angles for different propagation delays in a typical indoor environment.

In our study, we assume the uniform angular distribution. Signals received at the base station antenna are a collection of multipath components that are correlated. For one realization of the simulation, we arbitrarily set the two coherent multipath directions as 120° and 150° , respectively, to test the performance of the proposed algorithm. Fig. 5.11 shows the resulting MUSIC spectrum with a “ ∇ ” marked solid line. Two multipath components from 1 far-field signal source arriving at the antenna from 120° and 150° in the azimuth plane are correlated. The MUSIC algorithm could not resolve

Figure 5.13: Impinging angles for different propagation delays in a typical indoor environment.

correlated signals. To circumvent the rich multipath problem, the mobile terminal's spatial signature [170] can be used to locate the mobile terminal's location.

5.4.4 Resolution and Limitation in Number of Resolvable Signals

The proposed technique is capable to estimate the DOAs of two signal sources with 1° angular separation. As can be seen in Fig. 5.14 signals impinging from 149° and 150° in the azimuth plane could be clearly resolved from the spectrum. The SNR is 30 dB. The maximum number of uncorrelated waves to be estimated is $M - 1$. M is the number of subsets, the number of sampling periods and is also the dimension of the formed signal correlation matrix $\mathbf{R}_{yy}$ in (5.4). The scenario is similar to the

Figure 5.14: MUSIC spectrum. The load setting is the same as in case 4. Two signal directions are 149° and 150° in the azimuth plane, respectively. $\text{SNR} = 40 \text{ dB}$. The noise is ideally white.

Figure 5.15: MUSIC spectrum. Load setting is the same as for case 3. SNR=30 dB. Signal directions: 50°, 120°, 150°, 200°, and 300° in the azimuth plane.

conventional MUSIC algorithm, in which $\mathbf{R}_{yy}$ with a dimension of M could produce up to $M - 1$ signals [137, 138]. In our design, there are six parasitic monopoles, thus we can estimate up to five DOAs as shown in Fig. 5.15. Signal directions are set to be: 50°, 120°, 150°, 200°, and 300° in azimuth plane.

In the above simulation, we assume that noise samples of antenna's six sampling periods are uncorrelated. This is an ideal case where $E[\mathbf{nn}^H]$ is an identity matrix. In a practical environment, the correlation among noise samples reduces the DOA estimation resolution. This is illustrated in Fig. 5.16, where the two signal directions are 140° and 150° in the azimuth plane, respectively.

Figure 5.16: MUSIC spectrum in a practical environment. The load setting is the same as in case 4. Two signal directions are 140° , 150° in the azimuth plane, respectively.

5.4.5 Sampling Period and Periodicity of Transmitted Signals

Finally, we discuss the requirement of the periodicity of the transmitted signals for employing the proposed DOA estimation technique. In the derivation of $\mathbf{R}_{yy}$, we require that each individual signal source transmits periodic signals. In a practical wireless communication environment, time offsets among periodic signals and the antenna sampling periods, induced by different propagation path delays are expected.

However, the signal period does not need to be aligned with the sampling period (cf. Fig. 5.17). This is true because the signal source transmits signals longer than the length of total six sampling periods; and each signal period is the same as the antenna sampling period T . The received signals at the antenna are still periodic, therefore, (5.3) is still valid. The requirement means that when doing DOA estimation, mobile

Figure 5.17: Periodic signal time slots, and antenna virtual sampling periods in the proposed DOA estimation technique. Each signal contains six slots. The antenna rotates six sampling periods. Time offsets are expected in a multipath environment

terminals will transmit periodic signals instead of message signals. With the priori knowledge of the location of wireless base stations, where the antennas are mounted, and with the database of the map of the local area, PL service could be provided at the expense of a short message period. This is applicable in cases when mobile subscribers are in dangerous situations and need to make a rescue call, such as the E999 call in Singapore and the E911 call in America. For applications in ad hoc networks where the DOA information is required for data package routing the periodical signal can be assigned during the network setup period.

5.4.6 Speed of DOA Estimation

In our design, sixteen signal bits are stored at each sampling period. A minimum length of $16 \times 6 = 96$ bits of the periodic signals for six antenna sampling periods is required. For TDMA IS-136 System, one frame comprises six time slots with the length of 6.67 ms each. In each time slot, 260 data bits are transmitted. This indicates that the DOA estimation is possible to be carried out within one TDMA time slot. The tuning of the reactance could be done in nanoseconds [171]. And, the off-the-shelf diodes can realize

state switching in a few nanoseconds.

5.5 Influence of the Switched-off Parasitic Elements

In the previous study, we assume that the switched-off elements do not contribute to the beamforming and are therefore viewed as “transparent”. Therefore, in operation the SARLPA is equivalent to a 7-element ESPAR antenna. In this section, we investigate the influence of the switched-off elements by looking at the current distributions on the switched-off monopoles and the formed beam patterns using the 37-element SARLPA.

Firstly, Fig. 5.18 describes the structure of the 37-element SARLPA in the NEC simulator. According to the previously published study [166], the parasitic elements can be made transparent by loading large reactance instead of large impedance. So in our simulation, we loaded all the switched-off parasitic elements with reactances of $-j450\Omega$. The antenna is working at subset #1. The reactance values are $x_{1-1} = -90\Omega$, $x_{1-2} = 0\Omega$, $x_{1-3} = 0\Omega$, $x_{1-4} = 0\Omega$, $x_{1-5} = 0\Omega$, and $x_{1-6} = 0\Omega$ (cf. Fig. 5.2). The center active monopole is excited at the central segment with unit voltage. Fig. 5.19 gives a 3D view of the current distributions on the array elements (monopoles). It is obvious from the figure that the switched-off elements have little current excitations while most currents are distributed on elements #1 – 1, #1 – 2, and #1 – 6. Fig. 5.20 shows the current distributions on the neighboring elements. Beside the center active element #0, the element #1 – 1 in operation (loaded with -90Ω reactance) has the highest current distribution. The switched-off elements (#2 – 1, #2 – 2, and #2 – 3) have very small current distributions. Therefore, they are transparent in this operating frequency.

Finally, we examine the beam patterns produced by the 37-element SARLPA and

Figure 5.18: The structure of a 37-element SARLPA in the NEC simulator. The ground plane transforms the monopoles with their image into $\lambda/2$ dipoles. The radius of the array is $\lambda/4$.

Figure 5.19: 3D view of current distributions on all parasitic elements. The ground plane transforms the monopoles with their image into $\lambda/2$ dipoles. The color bar indicates the amplitude of the current. The antenna is working at subset #1. The switched-off parasitic elements are loaded with reactances of $-j450\Omega$. The reactance values are $x_{1-1} = -90\Omega$, $x_{1-2} = 0\Omega$, $x_{1-3} = 0\Omega$, $x_{1-4} = 0\Omega$, $x_{1-5} = 0\Omega$, and $x_{1-6} = 0\Omega$ (cf. Fig. 5.2).

Figure 5.20: Current distributions on neighboring elements. The color bar indicates the amplitude of the current. The antenna is working at subset #1. The switched-off parasitic elements are loaded with reactances of $-j450\Omega$. The reactance values are $x_{1-1} = -90\Omega$, $x_{1-2} = 0\Omega$, $x_{1-3} = 0\Omega$, $x_{1-4} = 0\Omega$, $x_{1-5} = 0\Omega$, and $x_{1-6} = 0\Omega$ (cf. Fig. 5.2).

7-element ESPAR. The SARLPA is working at subset #1. The reactance values are $x_{1-1} = -90\Omega$, $x_{1-2} = 0\Omega$, $x_{1-3} = 0\Omega$, $x_{1-4} = 0\Omega$, $x_{1-5} = 0\Omega$, and $x_{1-6} = 0\Omega$ (cf. Fig. 5.2). the switched-off parasitic elements are loaded with reactances of $-j450\Omega$. In this situation, the SARLPA is equivalent to a 7-element ESPAR antenna with the reactance setting $x_1 = -90\Omega$, $x_2 = 0\Omega$, $x_3 = 0\Omega$, $x_4 = 0\Omega$, $x_5 = 0\Omega$, and $x_6 = 0\Omega$ (cf. Fig. 3.1). The 2D beam pattern in the azimuthal plane of the ESPAR antenna and that of the SARLPA antenna working at subset #1 are compared in Fig. 5.21. This close match justifies that the SARLPA working in each subset is equivalent to that of a 7-element ESPAR antenna and the pattern generated by the SARLPA can be simply presented by that of the 7-element ESPAR.

Figure 5.21: Beam patterns in the azimuthal plane ($\theta = 90^\circ$) generated by SARLPA operation in subset #1 and the 7-element ESPAR, respectively. For the SARLPA, the switched-off parasitic elements are loaded with reactances of $-j450\Omega$. The reactance values are $x_{1-1} = -90\Omega$, $x_{1-2} = 0\Omega$, $x_{1-3} = 0\Omega$, $x_{1-4} = 0\Omega$, $x_{1-5} = 0\Omega$, and $x_{1-6} = 0\Omega$ (cf. Fig. 5.2). For the ESPAR, the reactance values are $x_1 = -90\Omega$, $x_2 = 0\Omega$, $x_3 = 0\Omega$, $x_4 = 0\Omega$, $x_5 = 0\Omega$, and $x_6 = 0\Omega$ (cf. Fig. 3.1).

5.6 The Algorithm with a Simplified Antenna Structure

The proposed DOA estimation technique using a single port smart antenna relies on rotating beam patterns to construct a signal sample vector which is analogous to that linear array antennas built signal sample vector with data samples at each array element. In considering the complexity of implementing 36 parasitic monopoles, we simplified the proposed method. We found that using a reactively controlled directive array antenna, the same procedure can be carried out by adjusting the reactance values. (Beamforming by adjusting reactance values is described in Chapter 3.) Similar algorithms were

developed independently at ATR [161]. We call it “reactance-domain MUSIC”.

Slightly different from (5.3), the sampling data vector is formed with seven different ESPAR equivalent weight vectors $\mathbf{w}_{ESP}$.

$$\mathbf{y} = \mathbf{W}_{ESP} \begin{bmatrix} \boldsymbol{\alpha}(\phi_1) & \boldsymbol{\alpha}(\phi_2) & \dots & \boldsymbol{\alpha}(\phi_K) \end{bmatrix} \begin{bmatrix} s_1 \\ s_2 \\ \vdots \\ s_K \end{bmatrix} + \mathbf{n} \quad (5.10)$$

where

$$\mathbf{W}_{ESP} = \begin{bmatrix} \mathbf{w}_{ESP}^{(1)} & \mathbf{w}_{ESP}^{(2)} & \mathbf{w}_{ESP}^{(3)} & \mathbf{w}_{ESP}^{(4)} & \mathbf{w}_{ESP}^{(5)} & \mathbf{w}_{ESP}^{(6)} & \mathbf{w}_{ESP}^{(7)} \end{bmatrix}^T \quad (5.11)$$

In the equation, $\mathbf{W}_{ESP}$, a 7×7 dimensional matrix, is called “ESPAR equivalent weight matrix”. It comprises seven *ESPAR equivalent weight vectors* that produce 7 different beam patterns including one omnidirectional and six directive beam patterns. The MUSIC spectrum is plotted by calculating:

$$P_{MU} = \frac{1}{\boldsymbol{\alpha}(\phi)^H \mathbf{W}_{ESP}^H \mathbf{V}_N \mathbf{V}_N^H \mathbf{W}_{ESP} \boldsymbol{\alpha}(\phi)} = \frac{1}{\underline{\boldsymbol{\alpha}}(\phi)^H \mathbf{V}_N \mathbf{V}_N^H \underline{\boldsymbol{\alpha}}(\phi)}. \quad (5.12)$$

In the above equation, we define $\underline{\boldsymbol{\alpha}}(\phi)$ as the steering vector for reactance domain MUSIC.

5.7 Calibration Method and Experiment

From (5.3) and (5.12) we notice that $\mathbf{w}_{ESP}$ is involved, which contains mutual coupling information and reactance values at parasitic elements. Precise knowledge of this

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information is critical to obtaining the steering vectors. However, there exist many uncertainties due to array sensors' location errors, element gain, phase errors and frequency errors. These factors are neglected in calculating mutual coupling using NEC simulator. Therefore, application of the algorithm prerequisites array calibration. This section describes the calibration method that carried out in the radio anechoic chamber and the experimental results of the proposed reactance-domain MUSIC algorithm.

We applied a signal-subspace method [107, 125, 163] to calibrate the antenna. Let us assume at this moment that there is one calibration signal source s_k from the azimuthal angle ϕ_k and define $\mathbf{R}_{yy}^{(k)}$ as the corresponding correlation matrix of the constructed signal sample vector $\mathbf{y}$ in (5.10), we have

$$\mathbf{R}_{yy}^{(k)} = \mathbf{W}_{ESP} \boldsymbol{\alpha}(\phi_k) E[s_k s_k^H] \boldsymbol{\alpha}(\phi_k)^H \mathbf{W}_{ESP}^H + \sigma^2 \mathbf{I}_7 = \sum_{i=1}^7 \lambda_i^{(k)} \mathbf{e}_i^{(k)} \{\mathbf{e}_i^{(k)}\}^H \quad (5.13)$$

where $\lambda_i^{(k)}$ ($\lambda_1^{(k)} \geq \lambda_2^{(k)} \geq \dots \geq \lambda_7^{(k)} \geq 0$) and $\mathbf{e}_i^{(k)}$ are the eigenvalue and the corresponding eigenvector, respectively. Since there is only one signal source, the noise subspace is spanned by the six eigenvectors corresponding to the six smallest eigenvalues.

$$\mathbf{V}_N = \text{span} \left[\begin{array}{cccccc} \mathbf{e}_2^{(k)} & \mathbf{e}_3^{(k)} & \mathbf{e}_4^{(k)} & \mathbf{e}_5^{(k)} & \mathbf{e}_6^{(k)} & \mathbf{e}_7^{(k)} \end{array} \right]. \quad (5.14)$$

The equation indicates that

$$\mathbf{e}_1^{(k)} \propto \boldsymbol{\alpha}(\phi_k). \quad (5.15)$$

The goal of the calibration is to get $\mathbf{W}_{ESP}$ such that given any $0^\circ \leq \phi_k \leq 360^\circ$,

$$\mathbf{e}_1^{(k)} = \mathbf{W}_{ESP} \boldsymbol{\alpha}(\phi_k). \quad (5.16)$$

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For calibrating an ESPAR antenna with seven elements, at least seven independent signal sources with known directions are required. In this work we increased number of calibration waves to twelve to increase the calibration accuracy. These 12 waves are uniformly placed over the azimuthal plane with 30° separation. If the above procedure is carried out consecutively for each individual signal source, it is equivalent to placing a calibration wave at a fixed direction and rotating the ESPAR antenna 30° each time. After that, we can get a set of linear equations of the twelve array factors $\alpha(\phi_k)$ for $k = 1, 2, \dots, 12$ and the corresponding twelve eigenvectors $\mathbf{e}_1^{(k)}$. The equation can be written in a matrix multiplication form as:

$$\begin{bmatrix} \mathbf{e}_1^{(1)} & \mathbf{e}_1^{(2)} & \dots & \mathbf{e}_1^{(12)} \end{bmatrix} = \mathbf{W}_{ESP} \begin{bmatrix} \alpha(\phi_1) & \alpha(\phi_2) & \dots & \alpha(\phi_{12}) \end{bmatrix}. \quad (5.17)$$

$\mathbf{W}_{ESP}$ is obtained by solving the above equation. We now solve the equation using the least square method. Let

$$\mathbf{E} = \begin{bmatrix} \mathbf{e}_1^{(1)} & \mathbf{e}_1^{(2)} & \dots & \mathbf{e}_1^{(12)} \end{bmatrix} \quad (5.18)$$

and

$$\mathbf{A} = \begin{bmatrix} \alpha(\phi_1) & \alpha(\phi_2) & \dots & \alpha(\phi_{12}) \end{bmatrix}. \quad (5.19)$$

$\mathbf{W}_{ESP}$ is solved by minimizing the function

$$\min_{\mathbf{W}_{ESP}} \|\mathbf{E} - \mathbf{W}_{ESP} \mathbf{A}\|. \quad (5.20)$$

The least square solution to (5.20) is given by

$$\mathbf{W}_{ESP} = \mathbf{E} \mathbf{A}^H \{ \mathbf{A} \mathbf{A}^H \}^{-1}. \quad (5.21)$$

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The procedure for calibrating the ESPAR antenna is summarized below. The testbed is shown in Fig. 4.1.

1. Position the ESPAR on the motor as the receive antenna. The transmit antenna is a horn antenna pointing to the ESPAR antenna from 0° degree.
2. Transmit calibration signals periodically. At the same time, consecutively apply seven consecutive equivalent vectors $\mathbf{w}_{ESP}^{(i)}$, $i = 1, 2, \dots, 7$ by adjusting the bias voltages on the six tunable reactances.
3. With the received signal samples, construct the received signal vector $\mathbf{y}$ [cf. (5.10)].
4. Calculate the signal sample correlation matrix $\mathbf{R}_{yy}^{(k)}$. k is the index of the calibration wave.
5. Eigendecompose the correlation matrix and find the eigenvector $\mathbf{e}_1^{(k)}$ correspond to the maximum eigenvalue. Since the direction of the calibration wave is known, the array factor $\alpha(\phi_k)$ is obtained based on the array geometry.
6. Control the motor to rotate the ESPAR antenna in the azimuthal plane by 30° .
7. Repeat step 2 until the ESPAR antenna goes back to the original position.
8. Obtain $\mathbf{W}_{ESP}$ by solving equation (5.20).

With the calibrated $\mathbf{W}_{ESP}$, the reactance domain DOA estimation is carried out in the same radio anechoic chamber using (5.12). The SNR is 20 dB. Fig. 5.22 substantiates the importance of calibration to the DOA estimation. Without the calibration, we obtain $\mathbf{W}_{ESP}$ by first calculating mutual impedance using NEC simulator and calculating $\mathbf{W}_{ESP}$ using (3.10) and (5.11). It is clear from the figure that without the calculation we cannot achieve the precise value of steering vector. Therefore, MUSIC spectrum does not show peaks, thus leading to large estimation error.

Figure 5.22: Experimental spectra of the reactance-domain MUSIC technique with and without calibration. SNR = 20 dB. DOA of the impinging wave is 120°

5.8 Conclusions

In this chapter a novel approach for executing the high-resolution MUSIC algorithms (angular-domain MUSIC and reactance-domain MUSIC) based on a single RF port smart antenna has been proposed. After presenting the configuration and the working principle of the antenna, the performance of the proposed technique for various aspects have been studied by simulations. The results justified that the technique is capable for a high-resolution DOA estimation of 1 degree, which is comparable with a conventional MUSIC algorithm based on DBF array antennas. We also modified the technique called reactance-domain MUSIC to suit it to a simplified parasitic array antenna structure. Array calibration and DOA estimation using the reactance-domain technique were also carried out in anechoic chamber to verify the proposed technique.

The proposed technique for DOA estimation based on a single RF port parasitic antenna has many advantages over that for a DBF. They are as follows.

1. In operation only six parasitic elements are switched on. The switches consume a few milli-watt of power and the leakage current of a reverse biased diode is negligible. The parasitic array antenna features low power consumption and low complexity. It is suitable for commercial implementations at mobile terminals such as laptops in ad hoc wireless networks [30, 165] to provide direction information. It can also be easily mounted on vehicles to realize position-based services.
2. Contrary to the DOA estimation based on a DBF, the mutual coupling is utilized to steer the beam [45]. Thus, the modified MUSIC algorithm is free from the negative influences of the mutual coupling among antenna elements. In a DBF this influence has to be compensated for the enhancement of system performance in a high-resolution DOA estimation as studied in [172, 173].
3. The technique gives high quality performance in low SNR signal environments. The performance in a $\text{SNR} = 0$ dB environment is shown in Fig. 5.10.
4. The proposed SARLPA antenna is very flexible in operation. The parasitic elements are reactively loaded and controlled with switches. The antenna can be configured into a reactively controlled directive array with a certain amount of reactance loaded parasitic elements. Adaptive beamforming algorithm could be employed at the baseband to tune the beam direction [34] as an adaptive directional antenna helping to reduce the channel interferences and improve the channel capacity. If the directions of the impinging waves are known, the antenna could be reconfigured into a switched parasitic antenna. Beam patterns could be controlled by simply switching on and off different parasitic elements to function as a switched beam antenna, or sector antenna [62].

5. Parasitic elements are arranged to form a circular array antenna. When used for DOA estimation, it could give the signal direction information covering 360° in the azimuth plane.

Simulation studies have also pointed out that further improvements are needed to enhance system performances to make it more applicable to practical wireless communication systems. The areas of improvements are as follows:

1. The proposed technique works in an uncorrelated signal environment. New methods should explore signal sources' spatial signatures in a correlated signal environment for PL services.
2. The proposed antenna will not suffer from the negative influence from the mutual coupling between antenna elements. However, the calibration of the antenna aperture over DOA, frequency and temperature, weather environment, and fabrication error are still unavoidable and they influence the antenna performances.
3. The Doppler effect of the fast moving mobile terminals needs to be investigated.
4. The antenna having 36 parasitic monopoles equispaced on the circle surrounding the active central element has been used to theoretically demonstrate the idea of eigenstructure-based direction finding algorithm with a single RF port smart antenna. For practical implementations, using less parasitic antenna elements could reduce the complexity. Furthermore, the parasitic array antennas with elements in the form of microstrip patch also render a much compacter design.
5. The correlation among noise samples compromises the DOA estimation performance. Longer sample blocks are required to reduce the correlation effect.

As we have presented in Chapter 2, the single RF-port parasitic array antennas feature low power consumption and low fabrication cost. The architecture provides an economical and commercial way of implementing smart antennas at mobile terminals especially at those power-consumption limited applications. In chapters 3 and 4, we have proposed a fast beamforming algorithm for a parasitic array antenna. In this chapter, a DOA estimation technique is proposed. Embedded the proposed beamforming and DOA estimation algorithms, the parasitic array antennas provide great potential for application in WLAN and ad hoc networks. The beamforming ability of the parasitic array antennas provide an improved wireless link quality; The DOA estimation ability supports location-aware services and avails the packet routing in a wireless network.

For those smart antenna applications where the main concern is the performance rather than the cost and power consumption (e.g. cellular BSs) DBF array is a suitable implementation architecture. The SDMA scheme significantly increases the capacity of frequency resource limited wireless system by virtue of the multibeam forming ability of BS DBF array antennas. In the next chapter, we will present a technique developed for BS DBF array antennas.

Chapter 6

Space Division Multiplexing With DBF Antenna Arrays

Abstract: *With powerful base band DSP, beamforming and interference suppression ability of DBF array antennas bring about a significant capacity improvement to the frequency resource limited wireless communication systems. We study the SDMA scheme in a multipath-rich environment in this chapter. The channel responses correlation coefficient (CRCC) is proposed as a duplicate channel allocation criterion for a null-forming SDMA. Instead of forming null patterns at DOAs of interference signals, the array antenna's weight vector for a desired user is formed within the null space spanned by the interferers' channel responses vectors. Multipath, nonexistence of LOS path and angular spread, which are conventionally viewed as undesirable features to a beamforming SDMA scheme, are exploited to realize the null-forming SDMA. The performance is studied with respect to the BER. The performance degradation in the presence of mutual coupling between antenna array elements is also studied. The study reveals the trade-off between the given target BER performance, the channel duplication threshold, and the coverage capacity. In the presence of mutual coupling the BER performance*

degrades, the SNR at the MS transmitters needs to be increased to maintain a certain BER performance. In another word, the presence of mutual coupling brings an SNR loss at the MS transmitters. Results show when average BER is 10^{-2} , the SNR loss at MS transmitters due to the presence of mutual coupling at the BS array antenna is almost 10 dB.

6.1 Introduction

In the previous chapters, we have focused on analog array antennas and their algorithms. The parasitic array antennas are a strong candidate for mass implementations at mobile terminals due to the low power consumption feature. Their counterpart, DBF array antennas, are more suitable for installations at BSs where power consumption and operation cost is not the main limitation factor. Conventionally, the antenna arrays are deployed at BSs. With beamforming algorithms, spot beams are directed towards desired MSs. The MSs served under different spot beam areas could be allocated within the same physical channel (frequency and time slots). This technique is called channel duplication. The direction information of MSs is required at the BS for pointing beams to right directions of the users; and angular separation of distinct MSs in the same physical channels is the criterion for the duplicate channel allocation [17, 31, 42, 79]. This scheme is referred to beamforming SDMA. Trial test is carried out for Enhanced Digital Codeless Telephone (EDCT) radio standard. Using a DBF array antenna at the BS, the DOAs of the signals from MSs are estimated with MUSIC algorithm. The Kalman filter technique is used to track the positions of the MSs. However, the minimum separation of two MSs accommodated in the same frequency channel is limited to 3dB beamwidth of the DBF array antenna. This is mainly attributed to the difficulties associated with the synthesis of sharp nulls near the main lobe of the beam.

Furthermore, the direction based beamforming SDMA scheme is not effective in indoor and outdoor blocked scenarios. Walls and buildings obstruct LOS propagation that causes erroneous DOA estimation. The inaccuracy results in null patterns pointing to the desired MSs [42, 156]. In addition, large angular spread also precludes the scheme [170]. From the antenna side, overlapping side lobes of antenna beam patterns induce the MAI.

To alleviate all the constraints of beamforming SDMA, null steering techniques have been proposed to isolate co-channel users in SDMA scheme [77, 78]. In [78] spatial correlation coefficient (SCC) has been proposed as the duplicate channel allocation criterion. Using the spatial response to the interferer at the BS antenna array, an array null pattern is formed toward the interference direction. However, only the LOS component is considered in that model. The accurate DOA information of the interference signal is required to form a null pattern. The criterion for channel duplication is still dependent on the angular separation. This technique also does not consider the implementation issues in multipath rich environment.

To solve these problems in [78], we examine the performance of the null-forming SDMA with respect to BER in multipath rich environment in this chapter. Being different from the conventional null-forming SDMA in [78], DOAs of interferers are not required. The BS estimates the channel responses of the MSs. With the constrained optimization method, the antenna weight vector of the desired user is formed within the null-space spanned by interferers' channel responses. The CRCC is proposed as the criterion for channel duplicating. The threshold for duplicate channel with $\text{BER} \leq 10^{-2}$ is given. The multipath wireless channel is modelled using a two-dimensional ray-tracing model. Furthermore, we also study the influence of mutual coupling between BS array antennas on the SDMA channel allocation. With the ray tracing model and the inclusion of mu-

tual coupling, the system is much closer to the practical implementation environment.

6.2 Theory of Channel Duplication in a Multipath Environment

6.2.1 Ray-tracing Channel Model

We use two dimensional (2D) shooting and bouncing rays to simulate a wireless multipath environment as shown in Fig. 6.1. The coordinate is scalable. In this model, the LOS component and the first-order reflected rays are considered. The BS uniform linear array antenna has four elements with half wavelength ($\lambda/2$) inter-element separations. Received signals at the BS array antenna elements are represented with an $R_X \times 1$ dimensional vector $\mathbf{u}(t)$, where R_X is the number of receive antennas at the BS.

$$\mathbf{u}(t) = \mathbf{h}_1 s_1(t) + \mathbf{h}_2 s_2(t) + \mathbf{n}(t). \quad (6.1)$$

$\mathbf{h}_i$ ($i=1, 2$) are the channel responses at the BS linear array antenna to the MS- k , where $s_k(t)$ is the signal transmitted from MS- k with power p_{sig} .

$$E(s_k(t)s_k(t)^*) = p_{sig}. \quad (6.2)$$

$\mathbf{n}(t)$ denotes the zero-mean AWGN vector at the array antenna elements with the power of σ^2 .

$$E(\mathbf{n}(t)\mathbf{n}(t)^H) = \sigma^2 \mathbf{I}_{R_X}. \quad (6.3)$$

It is assumed that at each MS- k 's instant location, the channel status is constant over a short period; but the time period is long enough so that the infinite time horizon

Figure 6.1: 2-D ray tracing model for the null-forming SDMA scheme. The coordinate is scalable. The MSs are located in the far-field of the BS array antenna.

information theoretic description of communications portrays a meaningful idealization. The channel response at the BS antenna array for three impinging waves (one LOS and two reflected components in Fig. 6.1) from MS- k is expressed as [128]:

$$\begin{aligned}
 \mathbf{h}_k = & \begin{bmatrix} 1 \\ \exp(-j\pi \cos(\phi_k^{(0)})) \\ \exp(-2j\pi \cos(\phi_k^{(0)})) \\ \exp(-3j\pi \cos(\phi_k^{(0)})) \end{bmatrix} + \exp(-j\beta_k^{(1)}) \begin{bmatrix} 1 \\ \exp(-j\pi \cos(\phi_k^{(1)})) \\ \exp(-2j\pi \cos(\phi_k^{(1)})) \\ \exp(-3j\pi \cos(\phi_k^{(1)})) \end{bmatrix} \\
 & + \exp(-j\beta_k^{(2)}) \begin{bmatrix} 1 \\ \exp(-j\pi \cos(\phi_k^{(2)})) \\ \exp(-2j\pi \cos(\phi_k^{(2)})) \\ \exp(-3j\pi \cos(\phi_k^{(2)})) \end{bmatrix} \quad (6.4)
 \end{aligned}$$

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where $\phi_k^{(m)}$ represents azimuthal DOAs of one LOS wave ($m = 0$) and two reflected waves ($m = 1, 2$) from the MS- k , impinging on the BS uniform linear antenna array. For MS- k , $\beta_k^{(m)}$ is the relative propagation phase delay of the m th reflected wave with respect to the LOS component.

Here, we define CRCC as:

$$\rho_{kl} = \frac{|\mathbf{h}_k^H \mathbf{h}_l|}{\sqrt{(\mathbf{h}_k^H \mathbf{h}_k)(\mathbf{h}_l^H \mathbf{h}_l)}}. \quad (6.5)$$

Figure 6.2: Channel responses correlation coefficient (CRCC) of 2-D shooting and bouncing ray tracing model for the SDMA scheme. The color bar indicates the ρ_{12} . MS- k is fixed at $X = 5$, $Y = 5$.

With (6.5), we calculate the CRCC between two MSs, while the MS-1's position is fixed at $X = 5$, $Y = 5$ and the MS-2 is moving in the area $0 \leq X \leq 10$, $0 \leq Y \leq 10$ as shown in Fig. 6.1. The CRCC is plotted in Fig. 6.2 with respect to the X-axis and Y-axis of

the moving MS-2's position. From the figure we can see that even two MSs are located very close to each other, the probability of having a high correlation coefficient is still rare. This means that the channel duplication for two closely located MSs is possible.

6.2.2 Mutual Coupling

Mutual coupling exists among closely spaced antennas. The current on each element contains not only its own current from the feeding port, but also induced currents from the neighboring elements [121].

$$i_1 = y_{11}v_1 + y_{12}v_2 + y_{13}v_3 + y_{14}v_4. \quad (6.6)$$

y_{ij} ($i \neq j$) is the mutual admittance between the i th and the j th element. y_{ii} ($i = j$) is the self-admittance. v_i and i_i are the voltage and current, respectively, at the i th element induced by impinging plane waves on the antenna array or the open-circuit voltages as in [27].

The mutual admittance among 4 elements is represented by a mutual admittance matrix.

$$\mathbf{Y}_C = \begin{bmatrix} y_{11} & y_{12} & y_{13} & y_{14} \\ y_{21} & y_{22} & y_{23} & y_{24} \\ y_{31} & y_{32} & y_{33} & y_{34} \\ y_{41} & y_{42} & y_{43} & y_{44} \end{bmatrix}. \quad (6.7)$$

The currents and voltages are related by (6.8).

$$\mathbf{i} = \mathbf{Y}_C * \mathbf{v}, \quad (6.8)$$

$$\mathbf{i} = \begin{bmatrix} i_1 & i_2 & i_3 & i_4 \end{bmatrix}^T, \quad (6.9)$$

$$\mathbf{v} = \begin{bmatrix} v_1 & v_2 & v_3 & v_4 \end{bmatrix}^T. \quad (6.10)$$

Since, the array configuration is symmetric, $y_{12} = y_{43}$, $y_{13} = y_{42}$, $y_{14} = y_{41}$, $y_{21} = y_{34}$, $y_{23} = y_{32}$, $y_{24} = y_{31}$. We calculate the mutual admittance with Numerical Electromagnetic Code (NEC) (See Fig. 6.3).

For analysis of influence of the mutual coupling to the received signals, both open circuit model [27] and antenna transmission model [34] can be applied. Two models are in essence the same with reciprocity theory [121]. Using the open circuit model, the voltages at individual RF ports, $\hat{\mathbf{v}} = [\hat{v}_1 \ \hat{v}_2 \ \hat{v}_3 \ \hat{v}_4]$, connected to four dipoles are represented as [27, 34]:

$$\hat{\mathbf{v}} = \mathbf{X}_L(\mathbf{Z}_C + \mathbf{X}_L)^{-1}\mathbf{v} = \mathbf{C}\mathbf{v}, \quad (6.11)$$

where

$$\mathbf{X}_L = \text{diag} \begin{pmatrix} z_1 & z_2 & z_3 & z_4 \end{pmatrix}, \quad (6.12)$$

$z_i, (i = 1, \dots, 4)$ are the input impedances at the RF-ports. $\mathbf{Z}_C$ is the mutual impedance matrix:

$$\mathbf{Z}_C = \mathbf{Y}_C^{-1}. \quad (6.13)$$

Equation. (6.11) builds a relationship between the plane wave voltages $\mathbf{v}$ on the antenna array and the voltages at the RF ports that are fed to the baseband digital signal processor (DSP). In the presence of array antenna mutual coupling, signal samples at the RF ports is represented as:

$$\mathbf{u}(t) = \mathbf{C}\mathbf{h}_1 s_1(t) + \mathbf{C}\mathbf{h}_2 s_2(t) + \mathbf{n}(t), \quad (6.14)$$

Calculating Y_C with NEC Simulator

The mutual admittances are calculated using NEC [151], in which each monopole is modelled with 11 wire segments. All the monopoles are loaded with $50\ \Omega$ impedances at center segments. By exciting one monopole at the center segment with a unitary voltage, we get the currents at the center segments of neighboring monopoles. Since the exciting voltage is unity, the currents at the central segment of the monopoles are the mutual admittance. The antenna structure in NEC simulator is shown in Fig. 6.3. Using the described method, we calculate the mutual admittance for the array antennas with different numbers of array elements. The values are listed in Table 6.1.

Figure 6.3: Structure of a 4-dipole uniform linear array with $\lambda/2$ inter-element spacing on a infinite ground plane in the NEC simulator. The unit of the X-Y-Z coordinate is wavelength (λ).

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Table 6.1: Mutual admittance for different numbers of array elements calculated using Moment Method NEC simulator. The array elements are half wavelength dipoles with half wavelength inter-element spacing.

	Two Antennas	Three Antennas	Four Antennas	Five Antennas
V_{11}	0.009188-j0.003784	0.0093-j0.003764	0.009334-j0.003759	0.00935-j0.003757
V_{12}	0.003723+j0.0005995	0.003395+j0.0004478	0.003342+j0.0004454	0.003322+j0.0004456
V_{13}		-0.001071+j0.0001418	-0.0008911+j0.000222	-0.0008562+j0.0002228
V_{21}		0.003395+j0.0004477	0.003342+j0.0004455	0.003322+j0.0004456
V_{22}		0.01007-j0.003049	0.001015-j0.003053	0.01018-j0.003057
V_{23}			0.00319+j0.0003528	0.003067+j0.0003575
V_{24}			-0.0008911+j0.000222	-0.0007364+j0.0002657
V_{14}			0.0005725-j0.00008386	0.0004728-j0.0001404
V_{15}				-0.0004029+j0.00006211
V_{25}				0.0004728-j0.0001404
V_{31}				-0.0008561+j0.0002228
V_{32}				0.003067+j0.0003575
V_{33}				0.01023-j0.003059

6.2.3 Generating Antenna Weight for Null-forming SDMA

In a conventional beamforming SDMA scheme, directional beams are formed towards MSs' locations. An acceptable signal quality is maintained while the BS individual receivers receive only the weighted signals in their own desired beam directions. As aforementioned, the side lobes induce interference from the duplicated channels. Furthermore, minimum angular separations among the MSs are required for them to be allocated to the same physical channel. In case of multipath environments, large angular spread results in DOA estimation errors. This leads to nulls and beams being placed towards wrong directions [170]. In our scheme the antenna weight is formed in the null space spanned by the interferers channel response vectors. The functional block diagram of the null-forming SDMA receiver is shown in Fig. 6.4. Note that the null forming is referred to forming an antenna weight vector in the null space, not forming null patterns towards the directions of interfering waves.

The BS estimates the channel response of the MSs and examines the CRCC. Based on

Figure 6.4: Functional block diagram of the null-forming SDMA receiver for one physical (frequency, time) channel. To MS-1 other MSs are multiple access interference (MAI). $R_X = 4$.

the channel correlation threshold, the links between the MSs and the BS are allocated to different physical channels. MSs which exhibit low CRCC for channels responses are assigned to the same physical channel. For MS-1, the scalar signal received after null-forming at the first individual receiver is:

$$y^{(1)}(t) = \mathbf{w}_1^H \mathbf{u}(t) = \mathbf{w}_1^H \mathbf{C} \mathbf{h}_1 s_1(t) + \mathbf{w}_1^H \mathbf{C} \mathbf{h}_2 s_2(t) + \mathbf{w}_1^H \mathbf{n}(t) \quad (6.15)$$

where $\mathbf{w}_1^H$ is the desired weight vector for the communication link with MS-1. The moving MS-2 is viewed as an interferer to MS-1. Let us consider the situation that K MS communicate with the BS simultaneously in the same physical channel and MS-1 is the desired user. Applying the constrained optimization method in [128], [149] and including the mutual coupling, $\mathbf{w}_1^H$ should be obtained by minimizing $\mathbf{w}_1^H \mathbf{R}_{uu} \mathbf{w}_1$ while satisfying:

$$\mathbf{w}_1^H \mathbf{C} \mathbf{h}_1 = 1. \quad (6.16)$$

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$$\mathbf{w}_1^H \mathbf{C} [\mathbf{h}_2, \dots, \mathbf{h}_M] = \mathbf{0}, \quad (6.17)$$

where the signal sample correlation matrix is defined as:

$$\mathbf{R}_{\mathbf{u}\mathbf{u}} = \mathbf{E}(\mathbf{u}(t)\mathbf{u}(t)^H). \quad (6.18)$$

In the above equations, $\mathbf{0}$ is a $1 \times (K - 1)$ dimensional zero vector. It is clear that the equation is solvable only if [135]

$$\text{rank}([\mathbf{h}_2 \dots \mathbf{h}_K]) < R_X. \quad (6.19)$$

This proves that the maximum number of users (K) can be accommodated in the same channel is equal to the number of antennas (R_X) at BS. Adding more antennas increases the system capacity. From (6.16) and (6.17), it is obvious that low CRCC (see (6.5)) leads to optimum solution. This is also demonstrated with simulation in the following section.

After solving this constrained optimization problem using Lagrange multiplier method [128], [149], $\mathbf{w}_1$ is:

$$\mathbf{w}_1 = \mathbf{R}_{\mathbf{u}\mathbf{u}}^{-1} \mathbf{c}^H (\mathbf{c} \mathbf{R}_{\mathbf{u}\mathbf{u}}^{-1} \mathbf{c}^H)^{-1} \mathbf{d}, \quad (6.20)$$

with

$$\mathbf{c} = \begin{bmatrix} \{\mathbf{C} \mathbf{h}_1\}^H \\ \{\mathbf{C} \mathbf{h}_2\}^H \end{bmatrix} \quad (6.21)$$

and

$$\mathbf{d} = \begin{bmatrix} 1 \\ 0 \end{bmatrix}. \quad (6.22)$$

Please note that generating antenna weight in this way does not aim at forming nulls towards interferences' directions. Rather, the weight is a vector within the null of the linear space spanned by the interferer channel vectors. Therefore, beam patterns are not a key issue to the performance.

6.2.4 Criterion for Channel Duplication

Knowing that CRCC is critical to forming weight vector at the null space interferers' channel responses vectors, we examine the BER performance of each link to determine the threshold below which two channels can be allocated to the same physical channel. Simulation results are shown in the next section. The BER of the MS-1's link while the closely located MS-2 is moving along the Y-axis is represented by (6.23) [66, 78], while the closely located MS-2 is moving along the Y-axis. For MS-1, signals from MS-2 are the interference, and vice versa.

$$\text{BER}(SNR_1, INR_1) = \frac{1}{4\sqrt{INR_1}} \int_{-\sqrt{INR_1}}^{\sqrt{INR_1}} \text{erfc}\left(x + \sqrt{\frac{SNR_1}{2}}\right) dx \quad (6.23)$$

SNR_1 and INR_1 are SNR and interference-to-noise-ratio (INR) at the MS-1. Suppose that the noise components at all the elements are independently and identically distributed (i.i.d.). From (6.15)

$$SNR_1 = \frac{\mathbf{w}_1^H \mathbf{C} \mathbf{h}_1 s_1(t) \{\mathbf{w}_1^H \mathbf{C} \mathbf{h}_1 s_1(t)\}^H}{\mathbf{w}_1^H \mathbf{n}(t) \{\mathbf{w}_1^H \mathbf{n}(t)\}^H} = \frac{p_{sig} \mathbf{w}_1^H \mathbf{C} \mathbf{h}_1 \mathbf{h}_1^H \mathbf{C}^H \mathbf{w}_1}{\sigma^2 \mathbf{w}_1^H \mathbf{w}_1} \quad (6.24)$$

$$INR_1 = \frac{\mathbf{w}_1^H \mathbf{C} \mathbf{h}_2 s_2(t) \{\mathbf{w}_1^H \mathbf{C} \mathbf{h}_2 s_2(t)\}^H}{\mathbf{w}_1^H \mathbf{n}(t) \{\mathbf{w}_1^H \mathbf{n}(t)\}^H} = \frac{p_{sig} \mathbf{w}_1^H \mathbf{C} \mathbf{h}_2 \mathbf{h}_2^H \mathbf{C}^H \mathbf{w}_1}{\sigma^2 \mathbf{w}_1^H \mathbf{w}_1} \quad (6.25)$$

6.3 Simulation Studies

6.3.1 BER Performances at Different SNRs

In the simulations, each MS transmits quadrature phase-shift keying (QPSK) signals with transmit powers (SIR is 0 dB for the two user case.). MS-1 (desired user) is fixed at location $X = 5$, $Y = 5$. MS-2 is moving from $X = 5$, $Y = 4$ to $X = 5$, $Y = 6$. The ray-tracing model considers up to sixth-order reflection multipath component, propagation loss and reflection loss. The CRCC is shown in Fig. 6.5. The BER performance of MS-1 versus the position of the MS-2 with and without the antenna array mutual coupling are plotted in Figs. 6.6, 6.7 and 6.6. From these figures, we observe that:

1. The BER exhibits peak values at some discrete positions where CRCC shows high correlation.
2. At the same position that has a certain CRCC value, a lower SNR results in a higher BER. This hints that the CRCC threshold, at which acceptable the communication link is maintained, is environment dependent.
3. Mutual coupling degrades the BER performance at all positions.

In Fig. 6.9, BER performances versus CRCC with the presence of mutual coupling at different SNRs environments are plotted. It can be seen that given an SNR level, to keep a certian BER performance, the CRCC between MS-1 and M-2 should be keep below a certain level. This imply that we can use CRCC values as a threshold for channel duplication. At any position of MS-2 where the CRCC is lower than the channel duplication threshold Γ , MS-1 and MS-2 can be allocated to the same physical channel. Here, we define the coverage capacity for a given BER performance C_r as the

Figure 6.5: Channel correlation coefficient at different between the two users while MS-2 (interferer) is moving along the Y-axis ($X = 5$). MS-1 is fixed at position $X = 5$, $Y = 5$.

percentage that the CRCC is lower than a given channel duplication threshold Γ while MS-2 is moving within the given area.

$$C_{\Gamma} \doteq P(X, Y, |CRCC < \Gamma) \quad (6.26)$$

6.3.2 BER Performance and Coverage Capacity at Different CRCCs

In this subsection, we examine the inter-relationship between the BER performance and the coverage capacity. Figs. 6.10 and 6.11 reveal the BER performance and coverage capacity over different channel duplication threshold without and with the mutual

Figure 6.6: BER at MS-1 in SNR 30 dB environment, while MS-2 (interferer) is moving along the Y-axis ($X = 5$). To MS-1, MS-2 is interferer.

Figure 6.7: BER at MS-1 in SNR 20 dB environment, while MS-2 (interferer) is moving along the Y-axis ($X = 5$). To MS-1, MS-2 is interferer.

Figure 6.8: BER at MS-1 in SNR 10 dB environment, while MS-2 (interferer) is moving along the Y-axis ($X = 5$). To MS-1, MS-2 is interferer.

Figure 6.9: BER at the MS-1 for different SNRs environment in the presence of mutual coupling. To MS-1, MS-2 is interferer. SIR = 0 dB.

Figure 6.10: Trade-off between capacity and BER performance (the probability that $\text{BER} < 10^{-2}$) for given SNR levels without mutual coupling at the BS array antenna. BS has 4 array elements.

coupling effect. As from the figures the coverage capacity grows with the channel duplication threshold. This can be explained by looking at Fig. 6.2. When MS-2 is moving within the given area, the correlation between the uplink channels of the two MSs changes. When the channel duplication threshold increases, there are positions at which CRCC is lower than Γ . Therefore, the coverage capacity defined in (6.3.1) increases. However, there exists a trade off between the BER performance and the coverage channel capacity. As Γ increases, the probability that the BER is lower than a target level (10^{-2}) drops. This indicates a degradation of uplink quality of MS-1. The uplink quality can be sustained at a given level either by decreasing the channel duplication threshold, which means less coverage capacity; or by increasing the SNR which means an increase of transmitted signal power at the MSs.

The comparison between Fig. 6.10 and Fig. 6.11 shows the influence of mutual coupling to the performance. We notice that in the presence of mutual coupling, the SNR

Figure 6.11: Trade-off between capacity and BER performance (the probability that $BER < 10^{-2}$) for given SNR levels in the presence of mutual coupling at the BS array antenna. BS has 4 array elements.

has to be increased to achieve a comparable performance of that without mutual coupling effect. Given a channel duplication threshold $\Gamma = 0.8$, the coverage capacity C_Γ is 90%. To achieve an uplink quality that the $P(BER < 10^{-2}) = 0.95$, from Fig. 6.10 we know that the requested SNR is 2 dB. However, as shown in Fig. 6.11, the requested SNR is 12 dB in the presence of mutual coupling. This means that the presence of mutual coupling at BS antenna array inquires an increase of the SNR. In another word, the mutual coupling incurs an SNR loss at the MS transmitters.

6.3.3 SNR Loss Due to Mutual Coupling

In this subsection, the mutual coupling influence to the requested SNR at the MS transmitter is investigated. Fig. 6.12 shows the requested SNRs at the transmitter versus different channel duplication threshold levels for the MS-1's uplink quality that $P(BER < 10^{-2}) = 1$. Given the channel duplication threshold $\Gamma = 0.6$, the requested

Figure 6.12: SNR loss versus the BER for a given channel duplications CRCC threshold (at a given channel percentage capacity).

minimum SNRs to guarantee the BER performance are 15 dB and 2 dB, respectively, with and without the mutual coupling effect at the BS antenna array. The mutual coupling brings an SNR loss of around 13 dB in this situation. In the above simulations, we assume the number of antenna array elements is four at the BS. Here, we examine the SNR Loss due to the mutual coupling effect for different numbers (from 2 to 5) of BS antenna array elements. Their mutual admittance values are listed in Table 6.1. Fig. 6.13 plots the requested SNRs for the average BER of 10^{-2} at different CRCC levels. The requested SNR decreases as the number of BS antenna elements increases. With more array elements, the BS array antenna has more degree of freedom to adjust for interference suppression. However, by comparing the two groups of curves we observe that regardless of the numbers of array elements, the mutual coupling incurs almost 10 dB SNR loss at different CRCC levels.

Figure 6.13: The requested SNRs at MS transmitters for a given BER performance versus CRCC between $\mathbf{h}_1$ and $\mathbf{h}_2$. Two groups of curves indicate the situations with and without mutual coupling at the BS antenna array, respectively.

6.4 Scopes and Limitations

6.4.1 Practical Communication Environment

In this work, we simulate the multipath channel with a 2D shooting and bouncing ray tracing model. One LOS component and two first order reflected rays are considered in this model. Each path is assumed to be a Gaussian channel. In a realistic wireless communication environment, there exist more reflected rays and each signal path has an angular spread, and the existence of LOS components is not guaranteed, especially in urban or indoor scenarios. Conventional SDMA scheme fails because of the difficulty

in estimating the DOAs of the MSs. Thus, beams could not be steered towards the MSs to realize the channel duplicating. On the contrary, these factors benefit null-forming SDMA. With a more complicated environment, the probability that two channels are highly correlated is low. With the proposed CRCC as the criterion for channel allocation, the application of SDMA in urban or indoor environment is possible.

6.4.2 Mitigating Negative Influence of Mutual Coupling

The mutual coupling decreases the BER performance. To maintain certain BER performance, the CRCC threshold has to be tuned down. With a lower threshold, there are less areas in which CRCC are under the threshold. Therefore, the mutual coupling decreases the potential channel coverage capacity C_T . In order to keep a sound communication for the duplicate channel, compensation techniques [107, 174, 175] need to be employed to reduce the negative effect of the mutual coupling. Multiplying the inverse of the coupling matrix $\mathbf{C}$ with the received signal vectors at the antenna array restores the voltages at the antenna elements and suppresses the mutual coupling effect [107].

6.4.3 Polarization Loss

The dipoles at the BS array antenna are oriented vertically. Impinging signals are assumed co-polarized with the dipole array antenna. Cross-polarization coupling typically produced by scatters and reflectors in the multipath propagation environment [167] is not considered in this work. Exploiting the polarization diversity in multipath environment realizes MIMO channel to increase transmission data rate [68] as we will discuss in the next chapter.

6.5 Conclusions

In this chapter, we first presented a typical multipath wireless communication environment with a 2-D shooting and bouncing ray tracing model. The mutual coupling effect was included in the performance study. CRCCs between one stationary MS and one moving MS were plotted. Since, the possibility of having high correlation coefficient is low in a multipath rich environment, we proposed the CRCC as the criterion for a duplicate channel allocation. With the constrained optimization null-forming algorithm, the performance with respect to the BER was studied. The benefits of the scheme are:

1. Enable the deployment of SDM scheme at multipath environment (indoor or urban areas) where conventional beamforming based SDM fails to function due to the difficulty in DOA estimation and larger angular spread.
2. Since the spatial separation is based on channel responses, not on angular separation, two users located close to each other can be allocated to the same physical channel, provided their CRCC is lower than the threshold.
3. Since the antenna weight vector is formed using the channel responses, DOAs of interferers are not required to form sharp null patterns, this relaxes the pattern synthesis process. Calibration of mutual coupling will improve the system performance.
4. High resolution, high accuracy DOA estimation, if possible in a multipath environment, is not required. Now, we need to estimate the uplink channel and CRCC is the criterion for channel allocation.
5. Capacity increase is achieved by simply installing more antennas at the BS to provide more freedom for interference suppression. For beamforming SDMA, in-

creasing the system capacity requires forming sharper beams to resolve two closely located MSs.

The contribution is that we exploited the capability of null-forming SDMA scheme, and proposed a new duplicate channel allocation criterion that was previously dependent on or related to the angular separation of the MSs. Simulations included both LOS component and reflected waves. Mutual coupling was also included. These made the simulation closer to the practical environment. Results show that even MSs are located in the same angular regions; they could still be allocated to the same physical channel, provided that the CRCC is lower than the threshold for $\text{BER} \leq 10^{-2}$. With the proposed method, we could further improve the cell capacity by adding more antenna elements at BS. Cell capacity is increased without the expense of the extra frequency resources. With R_X antennas at BS, the system capacity is R_X -folded.

Chapter 7

Beamforming in MIMO Communication Scenarios

Abstract: *Installing multiple antennas at both transmitters and receivers constructs a MIMO channel which promises a higher channel capacity. In this chapter, two receiver structures that concatenate optimum receive beamforming and STBC decoder (both ML and MMSE decoders) are proposed. Applying the Alamouti STBC at receiver, the receive beamforming process improves the BER performance and the achieved MIMO channel capacity in the presence of interference. The beamformer maximizes the SINR while retaining the orthogonality of the MIMO channel. Furthermore, the STBC decoding process is modified to adopt beamforming. We also investigate the implementation of beamforming with dual-polarization diversity antennas. Finally, the BER performances and the achieved MIMO channel capacities are compared for the two receivers.*

7.1 Introduction

Along with the evolution of wireless communications systems, there is always a desire to reach a higher channel capacity which allows higher data rate transmissions over wireless channels. It was concluded from Adivor *et al.*'s study that "capacity enhancing technologies are key in securing the financial validity of the operator's business" [92]. It is estimated that a rough doubling of the capacity every three years is desirable. Measured by the data rate per second per Herz, the channel capacity can be increased through aggressive frequency reuse and higher-order modulation [12, 176]. However, the frequency reuse increases CCI and thus lowers the modulation order that can be sustained. Recently, technologies on smart antennas are showing promises. As Dr. A. Viterbi once put "Spatial processing remains the most promising if not the last frontier in the evolution of multiple access systems." Smart antennas and MIMO systems are the right candidates to provide such capacity upgrade.

Over the last decade, research on smart antennas has been extensively carried out. Initially, the goal of implementing a smart antenna system is to coherently combine signals from multiple ports of the systems including analog adaptive arrays such as reactively controlled directive arrays (see Chapters 3, 4) or DBF array antennas (see Chapter 5). Directive antenna beam patterns are steered towards the desired signals while null patterns are formed in the directions of the interference signals. The SINR is improved by the system, which leads to a higher spectrum efficiency and a higher system capacity.

The propagation study shows that the increase of spectrum efficiency by forming directive antenna beam patterns is limited by the scattering typical to an urban wireless communications scenario. Recently, a new view of the multipath propagation is that

when there are significant scattering, the wireless transmission data rate can be increased significantly by exploiting the decorrelation of signals from multiple antennas at both transmitters and receivers [56, 57, 91, 93–96, 177, 178]. Theoretically, if the scattering exists to an extent that the channel links between different transmit antennas and receive antennas could be modelled as statistically i.i.d. variables, the system capacity increases proportionally with the number of antennas [93]. Transmitting STCs over MIMO wireless channels takes the advantage of spatial and temporal diversities. It provides an efficient way to increasing the transmission data rate [55, 58, 97–99, 179]. This high data rate transmission provided by the MIMO system is fundamental to the high speed wireless applications such as the fixed broad band wireless access networks [91, 176] and mobile wireless internet access.

Beamforming techniques have long been recognized as an effective means for suppressing interference or interference cancellation (IC) that improves the spectrum efficiency. Furthermore, with successful IC, a smaller frequency reuse factor can be adopted. Thus, it also increases the system capacity [10, 142]. In Chapters 3 and 4, we have presented adaptive beamforming technique using a reactively controlled directive antenna. In this chapter, we investigate the beamforming techniques for IC in MIMO STC transmissions. Since the output signals from multiple antennas are assumed uncorrelated in MIMO communications scenarios, the meaning of “beamforming” in this context is not limited to forming any specific beam pattern physically. The term is generalized to include combining independent signals or transmit and receive optimization as proposed by many researchers. In [140], beamforming is referred to pre-multiplying the signals with a weighting matrix before transmitting and receiving, respectively, to diagonalize the MIMO channel matrix. The capacity of the MIMO system becomes the sum of the capacities of multiple independent signal-input single-output (SISO) channels. This technique is known as eigen-mode transmission or eigen-beamforming [140]. In [180],

the signals are weighted at the transmitter with a beamforming matrix to maximize the delivered power to the receiver according to the feedback channel state information (CSI).

Interference exists in various wireless communications scenarios. In a multiuser communications scenario, the wireless link of each individual user is susceptible to MAI. In this situation, we assume that there is cooperation among all the MSs and BSs, so that the CSI of all the transmitters are known at the receivers. The standard multiuser detection techniques include: the ML detection [143], which is optimal but the complexity grows exponentially with the number of multiple users, and linear receiving techniques such as MMSE and zero forcing (ZF) methods [142]. Applied to Alamouti STBC MIMO communications, a MMSE linear receiving technique exploits the algebraic structure of the Alamouti code words to suppress interference [58, 142, 143, 181, 182]. Beside these techniques, beamforming has been employed in the MIMO multiuser communications recently. “Beamforming” in this context refers to suppressing MAI to realize joint detection. In [183], signals are weighted at the transmitter with a beamforming weight matrix that is located in the null space of other users’ channel matrices. The technique presented in [184] generates a transmit beamforming weight vector that is the linear combination of zero eigenvectors (eigenvectors corresponding to zero eigenvalues) of MAI channel matrices. The linear combination is so formed that the transmit power to the desired user is maximized and that to the interferer is minimized. These techniques requires CSI of all users at transmitters or receivers or both. Besides, matrix inversion and eigen-decomposition are involved.

Wireless links in multi-cell environment suffer from CCI which is attributed to the nature of frequency reuse. In WLAN or ad hoc networks, mobile terminals are susceptible to interference from neighboring terminals. In contrast to the multiuser communication

scenario, there is no cooperation among multiple transmitters and receivers, furthermore the CSI of interfering MS is usually not known or only the covariance matrix of unwanted signals is known. Therefore, the aforementioned IC techniques (ML receiver, MMSE and ZF linear receivers, beamforming) that require CSI of MAI fail to function. Therefore, designing IC techniques that do not require interferers' CSI is critical to the wireless link quality of each user. For the MIMO STC transmission in the presence of CCI, a receiver beamforming matrix that minimizes the MMSE before STC decoder is realized using LMS method with reference signals [185]. However, the technique requires both channel estimation and adaptation with reference signals. Transmission of long reference signals reduces the spectrum efficiency.

This chapter presents two receiver structures for Alamouti STBC transmission in the scenario where each MS transmitter has two transmit antennas. Furthermore, we assume that the receivers only know the CSI of the desired MS and the correlation matrix of unwanted signals. Both beamformers maximize the SINRs before STBC decoders at the receivers. The CSI is not required at the transmitter. The STBC decoding processes are modified to adopt the beamforming. The property of the Alamouti STBC is preserved after optimum beamforming to support higher data rate transmissions.

Section 7.2 presents the receiver structure that combines optimum beamforming with the conventional Alamouti STBC decoder. We found that the conventional Alamouti STBC decoder is in essence a ML-single user (ML-SU) receiver. The BER performances and the achieved MIMO capacities in various application scenarios are presented. In Section 7.3, we discuss the implementation of receiver structure with dual-polarized antennas (DPAs). We found that the beamforming is channel-oriented in that it forms directive beam patterns in a high Ricean- K environment. Section 7.4 presents the receiver structure that combines beamforming with the Alamouti MMSE decoder. Sim-

ulation results show significant BER performance improvement and a MIMO channel capacity gain over the conventional scheme. Studies in sections 7.3 and 7.4 show that both receiver structures improve the performance of the conventional decoders. In Section 7.5, we compare the two receiver structures in terms of their BER performances and the achieved MIMO channel capacities. Finally, Section 7.6 concludes the chapter.

7.2 Combining Beamforming with Alamouti STBC ML Decoder

7.2.1 System Model and Assumptions

Throughout the discussion we assume that the desired signals and the interference signals have the synchronized time slots, same frequency (single carrier modulation) and modulation scheme. It is also assumed that each transmitter transmits Alamouti STBC with two transmit antennas ($T_X = 2$). The MIMO channel is assumed to be stationary flat Rayleigh fading. The channel state remains unchanged over the period of a few consecutive symbols. The receiver employs $R_X = M_r$ number of receive antennas as shown in Fig. 7.1. Therefore the CSI from the k th MS transmitter to the receiver is an $R_X \times T_X$ dimensional matrix $\mathbf{H}^{(k)}$ with each entity modelled as an i.i.d. complex Gaussian variable with zero mean and unit variance.

$$\mathbf{H}^{(k)} = \begin{bmatrix} h_{1,1}^{(k)} & h_{1,2}^{(k)} \\ h_{2,1}^{(k)} & h_{2,2}^{(k)} \\ \vdots & \vdots \\ h_{R_X,1}^{(k)} & h_{R_X,2}^{(k)} \end{bmatrix} \quad (7.1)$$

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where $h_{i,j}^{(k)} \sim \tilde{\mathcal{N}}(0, 1)$ for $i = 1, 2, \dots, R_X$ and $j = 1, 2, \dots, T_X$.

Figure 7.1: Functional block diagram of the proposed receiver that combines beamforming with the Alamouti decoder.

Over two consecutive symbol epochs, the received signal samples at the array elements before the beamformer are simply the sum of received signals from all MSs and the AWGN components as:

$$\mathbf{U} = \sum_{k=1}^K \mathbf{H}^{(k)} \mathbf{S}^{(k)} + \mathbf{N}, \quad (7.2)$$

where $\mathbf{U}$ is a $R_X \times 2$ dimensional matrix. K is the number of total MSs in the same network, or the same frequency spectrum. Therefore, to any one MS, there are $K - 1$ co-channel interferers. $\mathbf{S}^{(k)}$ is Alamouti code word transmitted over MS- k 's two transmit antennas [139] during two symbol epochs. It is a 2×2 dimensional matrix. The symbol energy is defined as E_s . We assume that signals at each MS's two antennas have the same power. Therefore, for $T_X = 2$, the transmit power at each MS is $P_T = 2E_s$.

$$\mathbf{S}^{(k)} = \begin{bmatrix} \mathbf{s}^{(k)}(NT_s) & \mathbf{s}^{(k)}((N+1)T_s) \end{bmatrix} = \begin{bmatrix} s_0^{(k)} & -s_1^{(k)*} \\ s_1^{(k)} & s_0^{(k)*} \end{bmatrix}, \quad (7.3)$$

where T_s denotes the symbol duration.

$$\mathbb{E} \left[\mathbf{S}^{(k)H} \mathbf{S}^{(k)} \right] = P_T \mathbf{I}_2 \quad \text{for } k = 1, 2, \dots, K. \quad (7.4)$$

$\mathbf{N}$ is the $R_X \times 2$ dimensional AWGN matrix at the BS antenna array elements. Each entity is modelled as an i.i.d. zero mean Gaussian variable with variance σ^2 . We assume that all the STBC signal vectors are uncorrelated, and the AWGN vector is uncorrelated with all STBC signal vectors. The SNR defined in this context is P_T/σ^2 .

7.2.2 Optimum Beamforming for Maximum SINR

As shown in Fig. 7.1, beamforming is carried out at the receiver before the STBC decoder. After two symbol epoches, the received signals $\mathbf{U}$ from the array elements are combined at the beamformer to produce two signals for STBC decoding. Let us assume that the signals from MS-1 are the desired signal. Signals from other $K - 1$ MSs and noise are unwanted signals. Two signals produced at the beamformer are presented as:

$$\mathbf{y} = \begin{bmatrix} y_1 & y_2 \end{bmatrix} = \mathbf{w}^H \mathbf{H}^{(1)} \mathbf{S}^{(1)} + \mathbf{w}^H \left\{ \sum_{k=2}^K \mathbf{H}^{(k)} \mathbf{S}^{(k)} + \mathbf{N} \right\}, \quad (7.5)$$

where $\mathbf{w}^H$ is the $1 \times R_X$ dimensional optimum receiver beamforming weight vector for MS-1. Since it is assumed that the receiver communicates with MS-1 and signals from other MSs are CCI, we define the desired signal covariance matrix of MS-1 at the receiver as:

$$\mathbf{R}_s = \mathbb{E} \left[\mathbf{H}^{(1)} \mathbf{S}^{(1)} \{ \mathbf{H}^{(1)} \mathbf{S}^{(1)} \}^H \right]. \quad (7.6)$$

The desired signals contain the data streams transmitted over two transmit antennas of MS-1. Though they present interferences to each other, the two data streams are well separated at the Alamouti STBC decoder (See Appendix A for more detail). $\mathbf{R}_{ni}$

is defined as the covariance matrix of those unwanted signals (interference plus noise).

$$\mathbf{R}_{ni} = \mathbb{E} \left[\left\{ \sum_{k=2}^K \mathbf{H}^{(k)} \mathbf{S}^{(k)} + \mathbf{N} \right\} \left\{ \sum_{k=2}^K \mathbf{H}^{(k)} \mathbf{S}^{(k)} + \mathbf{N} \right\}^H \right]. \quad (7.7)$$

At the receiver beamformer, the output SINR is expressed as:

$$\gamma_b = \frac{\mathbf{w}^H \mathbf{R}_s \mathbf{w}}{\mathbf{w}^H \mathbf{R}_{ni} \mathbf{w}}. \quad (7.8)$$

It is a generalized Rayleigh quotient and its value is bounded by the maximum eigenvalue λ_{max} and the minimum eigenvalue λ_{min} of $\mathbf{R}_{ni}^{-1} \mathbf{R}_s$ [135]. Maximizing γ_b , the optimum beamforming weight vector $\mathbf{w}$ is the eigenvector corresponding to the maximum eigenvalue λ_{max} of $\mathbf{R}_{ni}^{-1} \mathbf{R}_s$. Therefore, λ_{max} is the output SINR at the beamformer.

After receiving two consecutive data outputs y_1 and y_2 from the beamformer, the virtual Alamouti MIMO channel $\mathbf{A}^{(1)}$ is constructed at the STBC decoder by exploiting the algebraic structure of the Alamouti STBC as

$$\mathbf{A}^{(1)} = \begin{bmatrix} a_0 & a_1 \\ a_1^* & -a_0^* \end{bmatrix}, \quad (7.9)$$

where

$$\begin{bmatrix} a_0 & a_1 \end{bmatrix} = \mathbf{w}^H \mathbf{H}^{(1)}. \quad (7.10)$$

At the STBC decoder, two signals for the symbol detection of $s_0^{(1)}$ and $s_1^{(1)}$ are produced as:

$$\begin{bmatrix} \widehat{s_0^{(1)}} \\ \widehat{s_1^{(1)}} \end{bmatrix} = \{\mathbf{A}^{(1)}\}^H \begin{bmatrix} y_1 \\ y_2^* \end{bmatrix}. \quad (7.11)$$

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In the presence of beamforming, the SINR γ_d at the STBC decoder is (See Appendix A for more detail.)

$$\gamma_d = \frac{E_s (|a_0|^2 + |a_1|^2)}{\frac{1}{2} \mathbf{w}^H \mathbf{R}_m \mathbf{w}} = \gamma_b. \quad (7.12)$$

Following the approach in [63], it can be proven that for the Alamouti STBC decoder in [139] the output SINRs after the STBC decoder γ_d is equal to that at the beamformer (See Appendix A for more detail). Therefore, we can use γ_b to evaluate the performance.

7.2.3 BER Performance

In this section, we will discuss the performances of the proposed technique in terms of the BER and the achieved MIMO channel capacity. Firstly, we look at the BER performance. We know that it is dependent on the probability density function (pdf) of the output SINR at the STBC decoder, which equals the output SINR at the beamformer. Therefore, the key step is to find the pdf of γ_d . In the following performance discussions, we will analyze the performance theoretically. Monte-Carlo simulations are also executed for system BER-SNR performances. BPSK is employed as the modulation scheme. After each individual STBC decoder, the receiver performs ML symbol detection. With the pdf $f(\gamma_d)$ of γ_d the average BER for BPSK modulated signals is calculated as [63]:

$$P_e^{(k)} = \frac{1}{2} \int_{-\infty}^{\infty} \text{erfc}(\sqrt{\nu}) f(\gamma_d)(\nu) d\nu, \quad (7.13)$$

where $\text{erfc}()$ is the complementary error function [149]. $\mathbf{H}^{(1)} \mathbf{H}^{(1)H} \sim \mathcal{CW}_{R_X}(T_X, \mathbf{I}_{R_X})$, and $\sum_{k=2}^K \mathbf{H}^{(k)} \{\mathbf{H}^{(k)}\}^H \sim \mathcal{CW}_{R_X}(T_X(K-1), \mathbf{I}_{R_X})$ [186] are complex Wishart distributed having the same dimension of R_X , but with different ranks of T_X and $T_X(K-1)$, respectively.

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The Wishart distribution of a complex covariance matrix $\mathbf{R} \sim \mathcal{CW}_{L'}(K', \Sigma)$, $\Sigma = \mathbb{E}[\mathbf{R}]$ ($K' \geq L'$) is given as [186, 187]:

$$f(\mathbf{R}) = \frac{e^{-\text{Tr}(\Sigma^{-1}\mathbf{R})} \text{Det}(\mathbf{R})^{K'-L'}}{\pi^{\frac{L'(L'-1)}{2}} \left\{ \prod_{i=1}^{L'} (K' - i)! \right\} \text{Det}(\Sigma)^{K'}} \quad (7.14)$$

$\text{Tr}(\bullet)$ is the trace of a matrix. In the case $K' \leq L'$, $\mathbf{R}$ follows the pseudo-Wishart distribution as:

$$f(\mathbf{R}) = \frac{e^{-\text{Tr}(\Sigma^{-1}\mathbf{R})} \text{Det}(\mathbf{R})^{K'-L'}}{\pi^{\frac{K'(2L'-K'-1)}{2}} \left\{ \prod_{i=1}^{L'} (K' - i)! \right\} \text{Det}(\Sigma)^{K'}} \quad (7.15)$$

For Wishart distributed $\mathbf{R}$, the joint distribution of ordered eigenvalues $\lambda_1, \lambda_2, \dots, \lambda_{L'}$ ($\lambda_1 > \lambda_2 > \dots > \lambda_{L'}$) is given as:

$$f(\lambda_1, \dots, \lambda_{L'}) = \text{Det}(\Sigma)^{-K'} {}_0F_0(-\Sigma^{-1}, \mathbf{R}) \frac{\pi^{m(m-1)}}{\tilde{\Gamma}_{L'}(K') \tilde{\Gamma}_{K'}(L')} \text{Det}(\mathbf{R})^{K'-L'} \prod_{i < j}^{L'} (\alpha_i - \alpha_j)^2 \quad (7.16)$$

For the pseudo-Wishart distributed $\mathbf{R}$, the joint distribution of ordered K' eigenvalues has the same form as (7.16) with K' and L' exchanged [188]. The complex multivariate gamma function is given as [186]:

$$\tilde{\Gamma}_{L'}(a) = \pi^{\frac{1}{2}L'(L'-1)} \prod_{i=1}^{L'} \Gamma(a - i + 1) \quad (7.17)$$

${}_0F_0(-\Sigma^{-1}, \mathbf{R})$ follows the hypergeometric function given in [186]. The probability density function of the maximum eigenvalue λ_1 is:

$$f(\lambda_1) = \int_0^{+\infty} \dots \int_0^{+\infty} f(\lambda_1, \dots, \lambda_{L'}) d\lambda_2 \dots d\lambda_{L'} \quad (7.18)$$

In the absence of CCI

In this case we discuss the performance when only one MS is serviced in the channel. It provides a benchmark for comparison with multiuser cases. When only one MS that has one transmit antenna communicate with the BS has two receive antennas, there is no interferences, therefore the system is a MRC problem [189]. When the MS has two transmit antennas and the BS has two receive antennas, the channel is a single user 2×2 MIMO channel without interferences. The output SINR for MS-1 at the receiver ($K = 1, T_X = 2, R_X = 2$) is given as:

$$\gamma_b = \frac{\mathbf{w}^H \mathbf{R}_s \mathbf{w}}{\mathbf{w}^H \mathbf{R}_{ni} \mathbf{w}} = \frac{P_T \mathbf{w}^H \mathbf{H}^{(1)} \{\mathbf{H}^{(1)}\}^H \mathbf{w}}{2\sigma^2 \mathbf{w}^H \mathbf{w}} = \frac{P_T \lambda_1}{2\sigma^2} \quad (7.19)$$

where, λ_1 is the maximum eigenvalue of the 2×2 covariance matrix $\mathbf{H}^{(1)} \mathbf{H}^{(1)H}$. According to (7.16)–(7.18), its pdf is given as:

$$f_{\lambda_1}(\alpha_1) = (1 - e^{-\lambda}) e^{-\lambda_1} \lambda_1^2 - 2 \left(1 - e^{-\lambda} \sum_{k=0}^1 \frac{\lambda_1^k}{k!} \right) e^{-\lambda_1} \lambda_1 + 2 \left(1 - e^{-\lambda} \sum_{k=0}^2 \frac{\lambda_1^k}{k!} \right) e^{-\lambda_1} \quad (7.20)$$

Given the pdf of λ_1 the output SINR at the beamformer is plotted in Fig. 7.2. The numerical outcome matches with the theoretical result.

In the presence of CCI

In this section, we discuss the performance when there exist CCI. We increase the number of receiving antennas to suppress CCI. When the receiver has more than 2 antennas ($R_X > 2$), the desired signal covariance matrix and the unwanted signal

Figure 7.2: Probability density function of the output SINR at the beamformer. SNR = 5 dB. $T_X = 2$, $R_X = 2$, $K = 1$.

covariance matrix follow the pseudo-Wishart matrix distribution. As reported in [188], it is impossible to obtain the closed form for the joint eigenvalue distribution since it involves zonal polynomial that is notoriously difficult to compute. Therefore, we resort to Monte Carlo simulation to obtain the beamformer output SINR. The pdf of the output SINR at the decoder is obtained with Monte Carlo simulation. BER is calculated numerically. In Monte Carlo simulations each data is obtained with 1,000,000 samples.

When the SNR is 5 dB, the simulated probability densities of the output SINRs at the beamformer where there are K users are shown in the Fig. 7.3. The cumulative distributions of the output SINRs are shown in Fig. 7.4. The SINR improvement due to the array gain can be obviously seen from Fig. 7.4.

Figure 7.3: Probability densities of output SINRs (unitless) at the beamformer for different number ($K - 1$) of interferers. $T_X = 2$, $R_X = 4$. SNR = 5 dB. All MS transmit signals with the same power. Therefore, all interferers have the same transmit power as the desired MS.

Figure 7.4: Cumulative distribution of the output SINR [unitless] at the beamformer.

In the absence of beamforming

We examine the CCI influence to the conventional Alamouti scheme without beamforming. This justifies the necessity of concatenating beamforming with the Alamouti STBC decoder in the presence of CCI. BER performances are shown in Fig. 7.5. The line ($K = 1, T_X = 2$ and $R_X = 1$) gives the performance of the conventional Alamouti scheme with one receive antenna. The line ($K = 1, T_X = 2, R_X = 2$) gives the BER performance of single user MIMO communication where both the transmitter and the receiver have two antennas, respectively. From the figure, we observe that in the presence of CCI, increasing the number of receive antennas at the receiver brings a better BER performance. However, the two curves ($K = 2, T_X = 2, R_X = 2$ and $K = 2, T_X = 2$ and $R_X = 4$) exhibit floor as the SNR increases. The slopes of the curves indicates that the in the presence of CCI, no diversity advantage is obtained from transmitting Alamouti STBC. The Alamouti STBC decoder process is a space MRC in nature [142], which ignores the CCI and coherently combine the signal samples according to the CSI of the desired user. It has implicit interference suppression ability but the performance is dependent on the instant CSI of both desired users and interfering users.

BER with beamforming

Fig. 7.6, shows the performances when beamforming is combined via an increased number of receive antennas at the BS. Performances for the cases of $K = 1, T_X = 2, R_X = 2$ and $K = 2, T_X = 2, R_X = 4$ are very similar, and the curves have the same slope. These prove that the added diversity freedom at the receiver (the added antennas) is scarified for suppressing CCI. In the situation that $K \geq 3, T_X = 2, R_X = 4$, the curves exhibit a floor. The receive antenna array has not enough freedom to cope with the

Figure 7.5: BERs without beamforming for two conventional Alamouti schemes.

Figure 7.6: Theoretical BERs with numerical integration and BERs with computer simulation. Beamforming is combined with Alamouti STBC for cases when $R_X \geq 2$.

interferences which become the main reason of errors as SNR increases. Secondly, we fix $K = 2$ and increase the R_X from 4 to 6. Results show that in the presence of unknown CCI suppressing one CCI requires two receive antennas. Adding more receive antennas brings a better performance and the receiver has higher CCI tolerance. What is worth to mention here is that the beamforming process brings a performance which needs two receive antennas to suppress one co-channel user with two transmit antennas, which fall into a classical interference suppression technique [141]. For the Alamouti STBC, there is a MMSE receiving technique that suppresses one co-channel user with only one extra antenna. From this point of view, the advantages of proposed concatenation structure seems trivial. However, we will show in the following sections that concatenating the receiver beamforming is also beneficial to the MMSE receiving technique. More importantly, there are some circumstances that the receiver combines beamforming with Alamouti decoder outperforms with the one that concatenates beamformer with MMSE decoder.

7.2.4 Maximum Channel Capacity with Optimum Beamforming

As presented by (7.10), in the presence of beamforming, an equivalent 2×1 vector channel is constructed at the receiver for MS-1 with channel gains a_0 and a_1 , respectively. Furthermore it can be seen from (7.9) that the Alamouti virtual MIMO channel $\mathbf{A}^{(1)}$ retains the orthogonality in that $\mathbf{A}^{(1)} \{\mathbf{A}^{(1)}\}^H = (|a_0|^2 + |b_0|^2) \mathbf{I}_2$. Therefore, the joint ML decoding of $s_0^{(1)}$ and $s_1^{(1)}$ is decoupled into two independent processes, each realizes a dual branch MRC and gives the SINR γ_d [139]. Under the assumption that each CCI signal waveform is Gaussian, the classical Shannon capacity is obtained for each symbol [179, 190]. For the full rate Alamouti scheme (two symbols are transmitted

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over two epochs), the achieved channel capacity for each MS is given by:

$$C_M = \log_2(1 + \gamma_d) \text{ bits/s/Hz.} \quad (7.21)$$

From (7.21), we can directly observe that the beamforming process maximizes the achievable capacity.

Figure 7.7: Channel capacities achieved by MIMO communication links with and without beamforming at the Alamouti conventional decoder.

MIMO channel capacities for various cases studied in the previous subsection are shown in Fig. 7.7. Theoretical capacities for different number of transmit and receive antennas without beamforming [56, Eqn (4)] are presented with dashed lines as the yardstick. When $K = 1, R_X = 1$, it is the conventional Alamouti transmission scheme. When $R_X = 2$ the beamforming is combined with the STBC decoder. The virtual 2×2 channel is constructed in the presence of the beamforming (see Eqn. (7.9)). When $K = 2, R_X = 4$, the capacity is close to that for $K = 1, R_X = 2$, so are their

Figure 7.8: Ergodic capacities with beamforming over different numbers (R_X) of receive antennas.

BER performances in Fig. 7.6. With more antennas higher capacity is achieved (e.g. $K = 2, R_X = 6$). When $K \geq 3, R_X = 4$ the capacity drops, the BER curves exhibit a floor (see Fig. 7.6). This capacity drip is because of the lack of the diversity freedom for CCI suppression.

The ergodic capacities (mean information rate) over the numbers of receive antennas for different numbers of co-channel MSs are shown in Fig 7.8. The beamforming process increases the ergodic capacity of a conventional 2×1 Alamouti scheme ($K = 1, T_X = 2, R_X = 1$) from 3.2 bits/s/Hz to 4 bits/s/Hz by adding one antenna. Furthermore, the archived ergodic capacity is also increased in the presence of CCI. Comparing the three curves, it is also clear that suppressing one additional CCI while achieving the same capacity requires two extra receive antennas in the presence of CCI with unknown CSI.

In the above discussion, we assume that the reciter only knows the CSI of the transmitter (the desired user) and the covariance matrix of the unwanted signals. The additional beamforming processing brings a higher interference tolerance and thus improves the performance of conventional single-user Alamouti decoder is reported in [139] in the presence of CCI. In the next section, we discuss the implementation of the proposed structure with polarization diversity antenna arrays.

7.3 MIMO Communication with Polarization Diversity

Because of the space limitation, polarization diversity has been considered as the suitable and compact realization of MIMO system at size-limited mobile terminals. For example, with a single low-profile microstrip patch antenna, dual polarization with high isolation can be achieved [191]. It also has the advantage that the diversity order is preserved in the line of sight propagation environment. However, it has been demonstrated that high cross polarization discrimination and high Ricean- K factor lead to performance degradation for the Alamouti STBC scheme [71, 127]. In this section, we study the feasibility of implementing optimum receive beamforming with polarization diversity antennas in different Ricean- K factors environments.

7.3.1 System Model

Let us assume that the BS has R_X number of receive DPAs, and each MS has one transmit DPA ($T_X = 1$). The antenna elements can be realized with microstrip patches that have $+45^\circ$ and -45° slanted linear polarizations [191]. As shown in Fig. 7.9 the

BS has 4 dual-polarized antennas ($R_X = 4$). To any one MS, the other $K - 1$ MSs are MAIs. The Alamouti STBC transmission scheme is employed. The uplink channel for the k th user is represented as:

$$\begin{bmatrix} \mathbf{h}_{A,a}^{(k)} & \mathbf{h}_{A,b}^{(k)} \\ \mathbf{h}_{B,a}^{(k)} & \mathbf{h}_{B,b}^{(k)} \end{bmatrix} = \begin{bmatrix} \mathbf{H}_A^{(k)} \\ \mathbf{H}_B^{(k)} \end{bmatrix} = \begin{bmatrix} h_{1,a}^{(k)} & h_{3,a}^{(k)} & h_{5,a}^{(k)} & h_{7,a}^{(k)} & h_{2,a}^{(k)} & h_{4,a}^{(k)} & h_{6,a}^{(k)} & h_{8,a}^{(k)} \\ h_{1,b}^{(k)} & h_{3,b}^{(k)} & h_{5,b}^{(k)} & h_{7,b}^{(k)} & h_{2,b}^{(k)} & h_{4,b}^{(k)} & h_{6,b}^{(k)} & h_{8,b}^{(k)} \end{bmatrix}^T \quad (7.22)$$

In the equation, those $+45^\circ$ slanted linear polarization ports at the MS and BS antennas are noted with indexes a and A , respectively. In the same manner indexes b and B denote the -45° slanted linear polarization ports. $\mathbf{h}_{A,a}^{(k)}$ and $\mathbf{h}_{B,b}^{(k)}$ are the co-polarized channels for MS- k . $\mathbf{h}_{A,b}^{(k)}$ and $\mathbf{h}_{B,a}^{(k)}$ are the cross-polarized channels. In order to include cross-polarization discrimination (XPD) (Γ , $0 \leq \Gamma \leq 1$) and Ricean- K factor ($\mathcal{K}$), the channel matrix $\mathbf{H}^{(k)}$ is presented as:

$$\mathbf{H}^{(k)} = \sqrt{\frac{\mathcal{K}}{1 + \mathcal{K}}} \begin{bmatrix} \overline{\mathbf{h}_{A,a}^{(k)}} & \overline{\Gamma \mathbf{h}_{A,b}^{(k)}} \\ \overline{\Gamma \mathbf{h}_{B,a}^{(k)}} & \overline{\mathbf{h}_{B,b}^{(k)}} \end{bmatrix} + \sqrt{\frac{1}{1 + \mathcal{K}}} \begin{bmatrix} \widetilde{\mathbf{h}_{A,a}^{(k)}} & \widetilde{\Gamma \mathbf{h}_{A,b}^{(k)}} \\ \widetilde{\Gamma \mathbf{h}_{B,a}^{(k)}} & \widetilde{\mathbf{h}_{B,b}^{(k)}} \end{bmatrix} \quad (7.23)$$

Entries with $\widetilde{(\quad)}$ denote diffuse components [66]. Those with $\overline{(\quad)}$ are fixed components [71]. The fixed components are the array responses to the impinging waves. As shown in Fig. 7.9, the BS has four DPAs with half-wavelength ($\lambda/2$) inter-element spacing. Let the DOA of the impinging wave be ϕ , we have the fixed component channel as:

$$\overline{\mathbf{h}_{A,a}^{(k)}} = \overline{\mathbf{h}_{A,b}^{(k)}} = \overline{\mathbf{h}_{B,a}^{(k)}} = \overline{\mathbf{h}_{B,b}^{(k)}} = \begin{bmatrix} 1 & e^{-j\pi \cos(\phi)} & e^{-j2\pi \cos(\phi)} & e^{-j3\pi \cos(\phi)} \end{bmatrix}^T. \quad (7.24)$$

The diffuse components are modelled as i.i.d. complex Gaussian variables with zero mean and unit variance.

Figure 7.9: MIMO communication using polarization diversity antenna array with the Alamouti transmission scheme. Looking at MS–1 other MSs are MAIs. The MSs are assumed located in the far-field of the BS array antenna.

7.3.2 Beamforming with Polarization Diversity

Two optimum receiver beamformers are employed at the each polarization channel in parallel. The beamforming weights of MS–1 at two polarization channels are $\mathbf{w}_A^{(1)}$ and $\mathbf{w}_B^{(1)}$, respectively. They are $R_X \times 1$ dimensional vectors. Looking at one polarization channel (channel-A), the received signals after beamforming over two symbol epoches

are:

$$\mathbf{y}_A^{(1)} = \mathbf{w}_A^{(1)H} \left\{ \mathbf{H}_A^{(1)} \mathbf{S}^{(1)} + \sum_{k=2}^K \mathbf{H}_A^{(k)} \mathbf{S}^{(k)} + \mathbf{N} \right\} \quad (7.25)$$

To maximize the output SINR, beamforming weight is obtained as the eigenvector corresponding the maximum eigenvalue of $\mathbf{R}_{ni}^{-1} \mathbf{R}_d$ as we have presented in the previous section. At the STBC decoder, output signals from two beamformers (for both polarization channels) are summed. The virtual Alamouti channel is constructed as [139]:

$$\mathbf{A} = \begin{bmatrix} a_A^{(1)} & b_A^{(1)} & a_B^{(1)} & b_B^{(1)} \\ (b_A^{(1)})^* & -(a_A^{(1)})^* & (b_B^{(1)})^* & -(a_B^{(1)})^* \end{bmatrix} \quad (7.26)$$

where,

$$\begin{bmatrix} a_A^{(1)} & b_A^{(1)} \end{bmatrix} = \left\{ \mathbf{w}_A^{(1)} \right\}^H \begin{bmatrix} \mathbf{h}_{A,a}^{(1)} & \mathbf{h}_{A,b}^{(1)} \end{bmatrix} \quad (7.27)$$

$$\begin{bmatrix} a_B^{(1)} & b_B^{(1)} \end{bmatrix} = \left\{ \mathbf{w}_B^{(1)} \right\}^H \begin{bmatrix} \mathbf{h}_{B,a}^{(1)} & \mathbf{h}_{B,b}^{(1)} \end{bmatrix} \quad (7.28)$$

Notice that beamforming weights are involved in the virtual channel. Therefore, different from the conventional decoding process as in [139], the decoder must know the beamforming weights for two polarization channels.

7.3.3 Numerical Results

In simulation, we assume that there are only one fixed component from each MS. The DOA of MS-1 (desired MS) is set 130° , whereas the DOA of the impinging interference signal is 50° . BPSK is the modulation scheme. The values of Ricean- K and XPD are chosen from the experimental results presented in [192]. All MSs have the same transmit power P_T . In the presence of beamforming, an equivalent Alamouti scheme with two transmit antennas and two receive antennas is constructed with DPAs. After receiving $\mathbf{y}_A^{(1)}$ and $\mathbf{y}_B^{(1)}$, the conventional Alamouti decoding process for two transmit

Figure 7.10: BER performances for different Ricean- K factors and XPDs. DOA of LOS fixed component is 130° . When $K = 2$, the DOA of the interference MS is 50° .

antennas and two receive antennas [139] is carried out. The BER performance is shown in Fig. 7.10. The reason why the curves cross is that in a high Ricean- K factor environment, the beamforming process forms a directional pattern pointing to the desired user, and a higher SNR gives a better beamforming performance. Thus, the performance is improved with the array directivity gain. The output SINR at the STBC decoder is shown in Fig. 7.11. With higher a Ricean- K factor, the SINR is more centered. This is due to the strong fixed component.

The aim of the section is to examine the array patterns formed in different Ricean- K factors and XPDs environment. The output beam patterns at two polarization chan-

Figure 7.11: Probability density of output SINRs (unitless) at the STBC decoder after beamforming. $\text{SNR} = 5 \text{ dB}$. $K = 2, T_X = 1, R_X = 4$. DOAs of the desired user and interference user are 130° and 50° , respectively. $\text{XPD} = 0 \text{ dB}$.

nels for different Ricean- K factors are shown in Figs. 7.12 and 7.13. Each figure shows three snapshots of the formed beam patterns. We can observe that the beam patterns are environment oriented. In a low Ricean- K factor environment, the diffuse components dominate. The patterns are not directive but they are randomly formed. The beam pattern varies randomly at each snap, depending on the channel status. At this time, the system suppresses interferences by combining independent signals. When the Ricean- K factor increases, the fixed components dominate. The optimum beamformer forms the directive pattern towards the desired impinging wave. And, the pattern varies only a little.

Figure 7.12: Three instant beam patterns at $+45^\circ$ polarization channel-A and -45° polarization channel-B. Ricean- $K=0$. XPD=0 dB. $T_X = 1$, $R_X = 4$, $K = 2$. DOAs of the desired user and interference user are 130° and 50° , respectively. SNR = 5 dB.

The obvious advantage of this environment-oriented beamforming is that in a high Ricean- K factor environment, directional beam patterns are formed. Therefore, the interference to other links are reduced with directive beam patterns. For example, in an ad hoc network, when the environment has a high Ricean- K factor, the directional patterns are formed at each MS, this reduces the interference to other MS. The overall throughput of the network is increased.

Figure 7.13: Three instant beam patterns at $+45^\circ$ polarization channel-A and -45° polarization channel-B. Riccan- $K=10$. XPD=6 dB. $T_X = 1$, $R_X = 4$, $K = 2$. DOAs of the desired user and interference user are 130° and 50° , respectively. SNR = 5 dB.

7.4 Combining Beamforming with Alamouti STBC MMSE Decoder

In the previous sections, we have concatenated the optimum receive beamforming with the conventional Alamouti STBC decoder. Without CSI of interferers at the receiver, the beamforming process suppresses interference and improves the performance of Alamouti STBC decoder in the presence of interference. In this section, we examine the concatenation of the optimum receive beamforming with the MMSE Alamouti decoder.

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The Alamouti STBC [139] has been widely accepted as a transmit diversity technique owing to its implementation simplicity [89]. Furthermore, to increase the system capacity without extra frequency resources, signal processing algorithms have been investigated to allow for synchronous co-channel multiuser communications with the Alamouti STBC [58, 143, 181, 182]. The ML-multiuser (ML-MU) joint detection technique [143] can decouple multiple users into a single Alamouti STBC user. However, the computation is burdensome.

7.4.1 System Model

Aiming at a lower receiver complexity, suboptimal IC techniques have been proposed. In [181], the MMSE method exploits the algebraic structure of the Alamouti STBC to fulfill IC in a scenario where two co-channel MSs each with two transmit antennas ($K = 2$, $T_X = 2$) communicate with a BS that has two receive antennas ($R_X = 2$). One CCI is suppressed at the expense of one receive antennas. To further enhance the performance, we increase the number of receive antennas at the BS to $2M_r$ ($R_X = 2 \times M_r$, $M_r \geq 2$) and concatenate optimum receiver beamforming with the MMSE IC technique as shown in Fig. 7.14. These $2 \times M_r$ number of receive antennas are grouped into two subsets, A and B , each having M_r antennas. The beamformer at each subset maximizes the SINR independently while preserving orthogonality of the Alamouti virtual MIMO channel. The BER performance is improved considerably. Here, for simplicity, we assume $T_X = 2$ and $K = 2$. Combining the MMSE IC technique presented in [182], our proposal can be easily extended to the situations of $K > 2$. Under the assumption that $T_X = 2$ and $R_X = 2M_r$ as shown in the figure, the Rayleigh block fading uplink channel for MS- k

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Figure 7.14: Functional block diagram for receiver beamforming concatenated with MMSE IC.

is modelled as a $2M_r \times 2$ dimensional matrix:

$$\mathbf{H}^{(k)} = \begin{bmatrix} h_{1,1}^{(k)} & \dots & h_{M_r,1}^{(k)} & h_{M_r+1,1}^{(k)} & \dots & h_{2M_r,1}^{(k)} \\ h_{1,2}^{(k)} & \dots & h_{M_r,2}^{(k)} & h_{M_r+1,2}^{(k)} & \dots & h_{2M_r,2}^{(k)} \end{bmatrix}^T = \begin{bmatrix} \mathbf{H}_A^{(k)} \\ \mathbf{H}_B^{(k)} \end{bmatrix} \quad (7.29)$$

where $\mathbf{H}_A^{(k)}$ and $\mathbf{H}_B^{(k)}$ are $M_r \times 2$ dimensional channel matrices for MS- k at BS receive antenna subsets A and B, respectively. Since there are two MSs, the received signals over two symbol epoches at BS antenna subsets A and B are represented as:

$$\mathbf{U}_A = \mathbf{H}_A^{(1)} \mathbf{S}^{(1)} + \mathbf{H}_A^{(2)} \mathbf{S}^{(2)} + \mathbf{N}_A, \quad (7.30)$$

and

$$\mathbf{U}_B = \mathbf{H}_B^{(1)} \mathbf{S}^{(1)} + \mathbf{H}_B^{(2)} \mathbf{S}^{(2)} + \mathbf{N}_B, \quad (7.31)$$

where $\mathbf{N}_A$ and $\mathbf{N}_B$ are the noise components at BS receive antenna subset-A and B over two symbol epoches. We assume that MS-1 is the desired user, the optimum

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beamforming process is carried out at each antenna subsets. $\mathbf{w}_A$ and $\mathbf{w}_B$ are beamforming weighter at two subsets. The out signals at the two beamformers over two symbol epoches are:

$$\begin{bmatrix} y_A(NT_s) & y_A((N+1)T_s) \end{bmatrix} = \mathbf{w}_A^H \mathbf{U}_A, \quad (7.32)$$

and

$$\begin{bmatrix} y_B(NT_s) & y_B((N+1)T_s) \end{bmatrix} = \mathbf{w}_B^H \mathbf{U}_B. \quad (7.33)$$

Here we define a noise vector after beamforming as:

$$\eta = \begin{bmatrix} \mathbf{w}_A^H \mathbf{N}_A(:, 1) & \{\mathbf{w}_A^H \mathbf{N}_A(:, 2)\}^* & \mathbf{w}_B^H \mathbf{N}_B(:, 1) & \{\mathbf{w}_B^H \mathbf{N}_B(:, 2)\}^* \end{bmatrix} \quad (7.34)$$

where $\mathbf{N}_A(:, 1)$ denotes the first column of $\mathbf{N}_A$. In the presence of beamforming, an equivalent 2×1 vector channel is constructed for MS-1 at subset-A as $\begin{bmatrix} a_{A,1}^{(1)} & a_{A,2}^{(1)} \end{bmatrix} = \mathbf{w}_A^H \mathbf{H}_A^{(1)}$. Employing the algebraic structure of the Alamouti STBC, the virtual Alamouti MIMO channel matrix for two users are expressed as:

$$\mathbf{A}_A^{(1)} = \begin{bmatrix} a_{A,1}^{(1)} & a_{A,2}^{(1)} \\ \{a_{A,2}^{(1)}\}^* & -\{a_{A,1}^{(1)}\}^* \end{bmatrix}. \quad (7.35)$$

For the co-channel user MS-2, the 2×1 vector channel is constructed as $\begin{bmatrix} a_{A,1}^{(2)} & a_{A,2}^{(2)} \end{bmatrix} = \mathbf{w}_A^H \mathbf{H}_A^{(2)}$. The Alamouti virtual channel is expressed as:

$$\mathbf{A}_A^{(2)} = \begin{bmatrix} a_{A,1}^{(2)} & a_{A,2}^{(2)} \\ \{a_{A,2}^{(2)}\}^* & -\{a_{A,1}^{(2)}\}^* \end{bmatrix}. \quad (7.36)$$

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Including the virtual MIMO channels from BS receive antennas subset-B, the overall MIMO channel is formed as:

$$\mathbf{H}' = \begin{bmatrix} \mathbf{A}_A^{(1)} & \mathbf{A}_A^{(2)} \\ \mathbf{A}_B^{(1)} & \mathbf{A}_B^{(2)} \end{bmatrix}. \quad (7.37)$$

The MMSE IC technique as presented in [181] is executed with signals from two beamformers and the overall MIMO channel matrix $\mathbf{H}'$. The procedure is described below.

- Construct $\mathbf{y} = \begin{bmatrix} y_A(NT_s) & \{y_A((N+1)T_s)\}^* & y_B(NT_s) & \{y_B((N+1)T_s)\}^* \end{bmatrix}^T$.
- Calculate $\mathbf{M} = \frac{1}{E_s} \mathbb{E} [\mathbf{y}\mathbf{y}^H]$.
- Obtain $\boldsymbol{\alpha}_1 = \mathbf{M}^{-1}\mathbf{H}'(:, 1)$, where $\mathbf{H}'(:, 1)$ is the first column of $\mathbf{H}'$.
- Obtain $\boldsymbol{\alpha}_2 = \mathbf{M}^{-1}\mathbf{H}'(:, 2)$, where $\mathbf{H}'(:, 2)$ is the second column of $\mathbf{H}'$.
- The signals for the detection of $s_0^{(1)}$ and $s_1^{(1)}$ are: $\widehat{s_0^{(1)}} = \boldsymbol{\alpha}_1^H \mathbf{y}$ and $\widehat{s_1^{(1)}} = \boldsymbol{\alpha}_2^H \mathbf{y}$, respectively.

7.4.2 BER Performance

Figure 2 shows the estimated BERs obtained from Monte-Carlo simulations over different SNRs. BPSK is the modulation scheme. The case $K = 1$, $T_X = 2$, $R_X = 1$ represents a conventional single user Alamouti scheme with two transmit and one receive antennas [139]. The curve “ $K = 2$, $T_X = 2$, $R_X = 2$, + MMSE IC” corresponds to the MMSE IC method for two users as reported in [182]. This suboptimal approach exploits the algebraic structure of the Alamouti STBC and gives a diversity order of

7.4. Combining Beamforming with Alamouti STBC MMSE Decoder 198

two for each MS. When each antenna array subset at the BS has two receive antennas ($K = 2, T_X = 2, R_X = 2 \times 2$), the BER performance for each MS is improved by the SINR gain from the beamformers. The curve is parallel to those for the previous two cases. The reason is that the extra diversity freedom (two additional receive antennas) at the BS is consumed for co-channel interference suppression to obtain a higher SINR. In this situation, beamforming does not provide extra diversity advantage. Adding more receive antennas at the BS ($R_X = 2 \times 4$) provides the SINR gain and offers extra diversity freedom. Therefore, the BER performance is further improved and the curve is parallel to that for the conventional single user Alamouti scheme with two transmit and two receive antennas ($K = 1, T_X = 2, R_X = 2$), which has a diversity order of four.

Figure 7.15: BER performance comparison of various techniques.

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7.4.3 Maximum Channel Capacity

As for the MMSE decoder, the output SINR for $s_0^{(1)}$ is:

$$\gamma_d = \frac{E_s \alpha_1^H \mathbf{H}'(:, 1) \{\mathbf{H}'(:, 1)\}^H \alpha_1}{E_s \alpha_1^H \mathbf{H}'(:, 2-4) \{\mathbf{H}'(:, 2-4)\}^H \alpha_1 + \alpha_1^H \eta^H \eta \alpha_1} \quad (7.38)$$

where $\mathbf{H}'(:, 2-4)$ is the second to fourth column of $\mathbf{H}'$. Therefore, for the full rate Alamouti STBC transmission, the channel capacity can be obtained with (7.21). For various scenarios, the capacities are plotted in Fig. 7.16. We can observe from the equation that the beamforming process maximizes the output SINR in the presence of CCI and therefore maximizes the achieved channel capacity.

Figure 7.16: Achieved MIMO channel capacity at the STBC receiver that combines beamforming with MMSE decoding.

7.5 Discussion

7.5.1 Comparison between Two Receiver Structures

The conventional Alamouti STBC decoder presented in [139] is actually a ML decoder which is simplified to a linear processing process by virtue of the orthogonality of the virtual MIMO channel (See Appendix B.). Therefore, we will call Alamouti STBC decoder “ML-SU decoder” to differentiate it with the ML-MU decoder in [143]. We will compare the BER performance and the ergodic channel capacity of two receivers. One concatenates beamformer with the ML-SU decoder and the other with the MMSE decoder. In Fig. 7.17 and Fig. 7.18 the achieved BERs and ergodic capacities for ML-SU decoder (cf. Fig. 7.1) and MMSE decoder (cf. Fig. 7.14) are compared, respectively. From Fig. 7.17, it can be seen that when $R_X = 2$, the performance of ML-SU decoder exhibits a floor. The BS receiver has insufficient receive antennas to suppress interference. While, the MMSE decoder suppresses one interference with only one receive antenna and achieves a diversity order of two. When the number of receive antennas is increased, the beamforming process fully suppresses CCI, the ML-SU decoder outperforms the MMSE decoder. From Fig. 7.18, we can observe that the two lines for ergodic capacities cross. When the number of receive antennas is smaller than three, the performance of the structure that concatenates beamforming with MMSE decoder outperforms the one that concatenates beamforming with the ML-SU decoder. The reason can be explained as follow: When there are two MSs and the number of receive antennas is two, the MMSE technique exploit the algebraic structure of the Alamouti STBC, and suppresses one interference with one diversity freedom (one antenna) and archives a diversity order of two (cf. Fig. 7.17). However, for the structure that concatenates beamforming with the ML-SU decoder, the decoder perform joint ML-SU decoding of the transmitted two data symbols not matter the existence of interference. The beam-

Figure 7.17: BER for ML-SU decoder and MMSE decoder over different numbers of receive antennas. $K = 2$. SNR = 10 dB.

forming stage needs two receive antennas suppressing one interferer. In another word, the two receive antennas are sacrificed for interference suppression. Therefore, the lack of the diversity freedom leads to the performance degradation. When the number of receive antennas at each subset is greater than two, the structure with ML-SU decoder performances better in that two out of three receive antennas at each subset are used to fully suppress the interference. With sufficient IC conventional Alamouti decoder approaches the joint ML detection which gives the optimal solution, while MMSE is known to be the suboptimal solution.

Figure 7.18: Achieved ergodic capacities for ML-SU decoder and MMSE decoder over different numbers of receive antennas. $K = 2$. SNR = 10 dB.

7.5.2 Multiuser Communications

In this chapter, we concatenate beamforming with Alamouti STBC decoders. The extra beamforming processes suppress interferences and bring a higher SINR to the decoders. At the beamformer, only the unwanted signal covariance matrix is required. The technique is useful for MIMO communication in the presence of unknown interference. However, the technique proposed in this chapter is also suitable for multiuser detections. Beamforming processing which improves the SINR will also enhance the performance of multiuser communications. In that scenario, the focus is to discuss the multiuser rate region. This is left for the future work.

7.5.3 Time Dispersive Channel

In this chapter, we assume that the orthogonality of the Alamouti STBC is maintained during transmission. The performance of the proposed scheme is studied in the flat fading channel. For practical high data rate (wideband) communications in a multipath rich environment, multipath propagation delay and high data rate transmission cause the dispersiveness of the channel. And this reduces the orthogonality of the Alamouti data streams from two transmitting antennas. The performance degradation is expected. To mitigate the negative influence of the dispersive channel, various techniques including equalization [193, 194], orthogonal frequency division multiplexing (OFDM) [58] have been exploited. With the help of the techniques orthogonality might be restored but with the expense of the complexity.

7.5.4 Mutual Coupling Effect

In the previous chapters, we have noticed the effects of array mutual coupling, both positive and negative. For the reactively controlled directive array antennas, mutual coupling is fundamental to the beamforming. In Chapter 6, we conclude that mutual coupling reduces the threshold of CRCC, thus the system capacity is decreased. Aiming a fully understanding of ideal beamforming for MIMO STC communication, In the context of MIMO communications discussed in this chapter, we do not include the mutual coupling effect in the simulation. In the work of Janaswamy [195] and Wallace [196], the mutual coupling effect to the MIMO channel capacity has been analyzed. Including mutual coupling, the simulation model is closer to a practical environment. Further investigation on its effect is left for our future work.

7.6 Conclusions

We have combined the beamforming with STBC MIMO communication schemes. The STBC decoding process was modified to adopted beamforming. The data rate and the BER improvement provided by the Alamouti STBC are preserved to support higher data rate transmissions. Furthermore, beamforming maximizes the SINRs to improve the STBC communication performance in the presence of interference. From the above discussion, we can draw the following conclusions:

1. Beamforming does not separate signal streams transmitted from the desired MS's two transmit antennas. Though these two stream are co-antenna interference to each other, they are separated at the decoding stage. Whereas, the beamforming process fulfills the CCI suppression.
2. The diversity gain provided by the STBC is preserved after the beamforming to reduce the BER. The improved BER performance allows higher mode modulation schemes, thus increases the link data rate.
3. The beamforming maximizes the output SINR before the STBC decoding process. At the decoding, the virtual MIMO channel is built with the increased system SINR. For the ML-SU decoding receiver, the achieved MIMO channel capacity is a function of the output SINR, therefore, beamforming maximizes the instantaneous achieved MIMO channel capacity.
4. Also, adding more antennas at BS further increases the capacity. With more diversity freedom, more interferences can be suppressed.
5. Since the CSI of interferers are unknown. The optimum beamforming techniques that require joint optimizations at both transmitters and receivers fail to function in this scenario. The proposed technique is a receiver-processing only algorithm.

Concatenated with optimum receiver beamforming, a higher interference tolerance is achieved for the conventional Alamouti STBC decoder and the MMSE Alamouti STBC decoder.

6. By exploiting the algebraic structure of the Alamouti STBC, the MMSE decoder suppresses one CCI (or accommodates one co-channel user) at the expense of one diversity freedom (cf. Fig. 7.15). However, the comparison between two receiver structures shows that, when there is sufficient diversity freedom to suppress the interference, the structure that concatenates beamformer with the conventional Alamouti decoder brings a better performance in that it is in essence a ML-SU decoder.
7. Combining beamforming improves the performance of both conventional Alamouti STBC decoder and MMSE STBC decoder. With the extra degree of freedom (more antennas at the receiver), the beamforming process brings a higher CCI tolerance. The structure is also applicable for multiuser communications scenarios, where the CSI of all interference is known.
8. Since array antennas are installed at the BS, with DOA information of the impinging waves with respect to the broadside of the array it is possible to realize direction finding and forming directive beam pattern to further increase the channel capacity with antenna directivity gain, especially in a high Ricean- K factor environment where the MIMO capacity is doomed to decrease [72].

The contribution of this chapter is that we propose two receiver structures that combine beamforming with STBC decoders. Applied to the Alamouti STBC, beamforming is combined with both the conventional Alamouti decoder and the MMSE decoder. The beamforming process does not require CSI of CCI. It maximizes the output SINR before the STBC decoding process. At the decoder, the virtual MIMO channel constructed

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in the presence of beamforming exhibits a higher SINR. Therefore, the performances of both decoders are significantly improved. The techniques are suitable for implementations where CCI with unknown CSI exist, such as in ad hoc networks where the CCI are signals from other mobile terminals in the same network.

Chapter 8

Conclusions and Recommendations

8.1 Conclusions

The explosive growth of wireless communications industries is creating a huge market opportunity. Wireless operators are currently searching for new technologies that can be implemented into the existing wireless communication infrastructures to provide a broader bandwidth per user channel, better quality, and new value-added services. Smart antennas or adaptive arrays have gained great interest among researchers. With the beamforming ability, they bring a significant link quality improvement by suppressing interferences. The capacity of frequency resource limited systems also increases with an efficient frequency reuse scheme (e.g. SDMA). New value-added services, such as the PL services for emergency calls, the fraud detection, intelligent transportation systems *etc.* are also coming into reality with the implementation of the direction finding techniques of the smart antennas. Recently, antenna arrays that provide uncorrelated signal outputs are installed at both transmitters and receivers to build MIMO wireless communication channels. The multipath phenomenon achieves a channel capacity that

increases with the number of array antenna elements. Transmitting STCs through the MIMO system dramatically increases the data rate over wireless channels.

In response to the recent demand of exploring new technologies in smart antennas for mobile wireless communications, the main goals of the thesis are to design algorithms and investigate implementation issues for both analog adaptive array antennas and digital array antennas. To fulfill the goals, we have designed: 1) a fast adaptive beamforming algorithm for reactively controlled directive array antennas based on simultaneous perturbation stochastic approximation theory; 2) A high-resolution direction finding technique was proposed for parasitic array antennas; 3) The implementation of SDMA in a multipath rich environment was also studied and the channel responses correlation coefficient was suggested as the channel duplication criterion; 4) Finally we proposed two receiver structures that concatenate optimum receive beamforming with the Alamouti ML decoder and the MMSE decoder for MIMO communications in the presence of interferences. In the following section, we describe how the goals are fulfilled.

8.1.1 Fulfilling the Goal of the Thesis

Firstly, we carried out a thorough literature survey on the state-of-art smart antenna technologies, which has brought a new classification of smart antenna hardware and algorithms, formulated the motivation and set the goals of the thesis. We concluded that smart antennas can be classified as: switched beam antennas, phased array antennas, DBF array antennas, parasitic array antennas and diversity array antennas. They are playing important roles in the migration to the future wireless communications systems. The parasitic array antennas are strong candidates for realizing commercial implementations at power consumption limited mobile terminals. Whereas DBF array

antennas are suitable for applications at cellular BSs where the main limiting factor is the performance rather than the cost and the power consumption. Diversity array antennas have been implemented for many years to combat multipath fading. They are now utilized at both transmitter and receivers to build up MIMO communication systems which promise a higher spectrum efficiency, thus higher data rate transmissions are supported. The outcome of the literature review is a new classification of smart antennas based on their hardware, algorithms and various applications. This also has created the motivation of the research and formulated the purpose and goals of the PhD thesis project. Based on the knowledge gained from the thorough literature review, we began our study on parasitic array antennas.

In the last decades, various beamforming and direction finding algorithms have been designed for DBF array antennas. However, due to the high power consumption, fabrication and operation costs, the implementation of the DBF array antennas has been restricted to BS and military scenarios. Recently, the development of wireless networks calls for implementations of smart antennas at the mobile terminals that consume low power. In this circumstance, the parasitic array antennas, which feature lower power consumption and lower fabrication cost, are strong candidates for mass commercial implementation at those battery powered mobile terminals. The adaptive beamforming of parasitic array antennas can be realized by switching ON and OFF different parasitic elements in a manner analogous to the operation of switched beam antennas, or by changing the reactance loads at these parasitic elements. The latter realization, referred to “reactive controlled directive array” achieves a smoother beam pattern tuning, thus leading to a better interference suppression performance. However, the reported beamforming algorithms are not sufficient in terms of convergence speed for practical implementations. Therefore, we designed an adaptive beamforming algorithm for the reactively controlled directive array antennas. The algorithm adaptively adjusts the

beam pattern by tuning the reactance values to maximize the correlation between the received signals and the reference signals. Because the algorithm maximizes the correlation by iteratively adjusting the digital control voltages of the tunable reactance at the parasitic elements, the implementation of the adaptive beamforming algorithm for reactively controlled directive array antenna does not prerequisite array calibration. The strong mutual coupling between array elements is exploited to realize beamforming. Trials and experiments have been carried out during the internship at ATR, Japan. Results justify the algorithm. The algorithm converges within 5000 training bits. Nulls can be formed at the interferers' directions within a few iterations. In comparison with the DBF antenna arrays and MBF antennas, the reactive controlled directive array antennas feature low power consumption, low cost and ease of fabrication. The antenna is efficient because the beamforming is realized at the analog front end instead of at the digital baseband. The proposed fast beamforming algorithm makes the antenna a suitable candidate for the mass implementation of smart antennas, especially at the battery operated wireless mobile terminals such as laptops. Furthermore, the sufficient null forming and interference suppression abilities of the proposed algorithm make the antenna suitable for applications in ad hoc networks or WLAN, where the performance is mainly influenced by the interferences from the neighboring mobile terminals such as laptops, PDA *etc.*

New services such as PL and ad hoc networks require the direction finding ability of the smart antennas. In order to realize direction finding with low power consumption smart antennas, various direction finding techniques have been designed for parasitic array antennas. We proposed a novel parasitic array antenna structure. The antenna has one central element connected to the sole RF port and a number of switched and reactively controlled parasitic elements surrounding the active element in a circle. We call the antenna "switched and reactively loaded parasitic array (SARLPA)". Since the

system has only one RF port connected to the active central element and one subsequent down-conversion circuit, it obviously consumes much less power than a DBF array antenna does. A high-resolution direction finding using the MUSIC algorithm with periodic signals was employed on this single port smart antenna. Rotating the antenna beam patterns, while the signal is periodically transmitted, the received scalar signals from the RF port can be stacked into a vector. The beam pattern is shifted with a predefined angle. So, for the application of the MUSIC algorithm, it is sufficient to know these angular shifts, without knowing exactly the response to one subset. This is similar to a uniform circular or linear antenna array in which we know the phase shift between consecutive antenna responses, without having to know those responses exactly. The MUSIC algorithm is carried out with the constructed signal vector over one period and the modified steering vector based on the beam pattern angle shift. Furthermore, the algorithm was also modified for implementation with a reactively controlled directive array (e.g. ESPAR antenna). The PhD study demonstrated the feasibility of employing the eigenstructure-based MUSIC algorithm with a parasitic array antenna that has only one RF port.

As coding and modulation techniques have exhausted their ability to increase the system capacity, exploiting spatial domain provides a new frontier for research. Due to their multibeam forming and DOA estimation abilities, DBF array antennas are installed at cellular BSs to realize the SDMA scheme. Currently, the SDMA scheme is based on forming directional beam patterns to separate multiple users. The angular separation of the users is the criterion of channel duplication. However, the performance of this scheme is limited when there are rich scatterers. We proposed the channel responses correlation coefficient as channel duplication criterion for a null-forming based SDMA scheme. Different users are distinguished based on their spatial uniqueness (channel responses) at the base station antenna array. The multipath wireless channel

was modelled using a two dimensional ray-tracing method. Applying the constrained optimization method, the beam weight vector is formed within the null space spanned by the channel responses of interfering users. Including the mutual coupling effect, the BER performances evaluation is closer to practical environments. Results show that by exploiting the uniqueness of MSs' channel responses, MSs that locate in the same angular region, can be allocated in the same physical channel. Since the antenna weight is formed using the channel responses, DOAs of interferers are not required to form sharp null patterns, this relaxes the pattern synthesis process, hence the performance improvement. The study proves the increase of cell capacity with SDM in multipath environment where the conventional DOA based beamforming SDMA scheme fails to function.

Originally, diversity array antennas are utilized to combat multipath fading. By combining signals from independent branches, the probability of signal fade is greatly reduced. Recently, diversity array antennas are installed at both transmitters and receivers to construct a MIMO system. Transmission STCs through the MIMO system provides an economical way to increase the data rate over wireless channels. The meaning of beamforming in the context of MIMO communications has been extended to include combining independent signals. We proposed two receiver structures that concatenate optimum beamforming with STBC decoders. Applied to the Alamouti STBC, the beamforming is combined with the conventional Alamouti decoder and the MMSE decoder. The receiver beamformer maximizes the SINR before the STBC decoders. The decoding processes were modified to adopt beamforming. The properties of Alamouti STBC are preserved after the optimum beamforming to support higher data rate transmissions. Beamforming maximizes the SINR before STBC decoding. Thus, the CCIs are suppressed. The co-antenna interferences typical to the STBC transmission are suppressed at the STBC decoders. Results show that both structures improve the

performances in terms of BER and the achieved MIMO channel capacity of the original STBC decoders. Furthermore, the performances of combining beamforming with the conventional Alamouti decoder and the MMSE decoder were compared. The MMSE decoder exploits the algebraic structure of the Alamouti STBC to suppress one CCI (or accommodates one co-channel user) at the expense of one diversity freedom. However, the comparison between two receiver structures shows that, when there is sufficient diversity freedom to suppress the interference, the structure that concatenates beamformer with the conventional Alamouti decoder brings a better performance because it is in essence a ML-SU decoder. We also investigated the beamforming with polarization diversity antennas and found that the beamforming is environment-oriented in that a directive pattern is formed in high Ricean- K factor environment corresponding to the direction of the impinging wave.

8.2 Recommendations for Further Research

1. *Adaptive beamforming with parasitic array antennas.* Previous studies have shown the beamforming ability of a parasitic array antenna. Further investigation can be done into algorithms with less computational complexity and faster convergence to achieve commercial implementations of adaptive beamforming with parasitic array antennas. Performance evaluations for implementation of the parasitic array antennas embedded with the the proposed beamforming and direction finding algorithms in wireless networks will also be carried out to substantiate the benefits of installing the adaptive antennas at mobile terminals.
2. *Direction finding with parasitic array antennas and its potentials in realizing new value-added services.* Eigenstructure-based high-resolution direction finding with parasitic antennas has been studied. As shown in the simulation study, the per-

formance of direction finding degrades in a multipath environment. Extra investigations will be carried out to improve the estimation accuracy. New techniques should be exploited to cope with the problems of direction finding in multipath propagation. Furthermore, it has been shown that the calibration accuracy directly influences the direction finding ability. However, there are always differences between the anechoic chamber, where the calibration is carried out, and the application environment. This calibration discrepancy shall be examined when the antennas are implemented for both indoor and outdoor practical applications.

3. *Beamforming for STBC MIMO communications to improve channel capacity.* As an extension of smart antenna technologies, MIMO systems represent an economic way to increase the channel capacity and realize high data-rate broadband wireless transmissions. We have combined optimum receive beamforming with STBC for MIMO communications in the presence of interference. As an extension, beamforming multiuser communications shall be investigated, especially, on how beamforming influences multiuser rate region for broadcast and access channels.
4. *Influence of mutual coupling to DBF array antennas.* Mutual coupling exists among closely spaced antennas. The accuracy of high-resolution direction finding algorithm and the beam pattern tuning ability degrade in the presence of mutual coupling. Furthermore, mutual coupling substantially reduces the MIMO channel capacity. Characterization and compensation methods are to be investigated to optimize the system performance.
5. *Investigating implementations of MIMO communications.* To realize the compact installation of the MIMO system at the mobile terminals, the polarization diversity has been proposed. It has been shown that the rank of a multi-polarized MIMO channel matrix does not degrade when the channel has a high Ricean- K

factor where the LOS fixed component dominates. The receivers that combine beamforming with STBC decoders have been proposed in the thesis. Transmit beamforming algorithms are to be investigated. Trials and experiments need to be carried out to verify the proposal. Furthermore, compact realizations of antennas (polarization diversity antennas) for MIMO systems are to be investigated as this will benefit the commercial implementation of MIMO system for future generation wireless communication systems.

Finally, smart antenna technologies for mobile wireless communications is a vast research area. New algorithms and hardware implementation issues will evolve with the demands of emerging technologies and applications.

Appendix A

Proof of (7.12)

From (7.5), the output signal sample y_1 from the beamformer at the first symbol epoch (NT_s) is represented with a scalar.

$$y_1 = \underbrace{\mathbf{w}^H \mathbf{H}^{(1)} \mathbf{s}^{(1)}(NT_s)}_{\text{desired signal}} + \underbrace{\mathbf{w}^H \left\{ \sum_{k=2}^K \mathbf{H}^{(k)} \mathbf{s}^{(k)}(NT_s) + \mathbf{n}(NT_s) \right\}}_{\text{unwanted signal}}. \quad (\text{A-1})$$

We represent the unwanted signals in the equation with

$$z(NT_s) = \mathbf{w}^H \left\{ \sum_{k=2}^K \mathbf{H}^{(k)} \mathbf{s}^{(k)}(NT_s) + \mathbf{n}(NT_s) \right\}. \quad (\text{A-2})$$

At the first symbol epoch NT_s , the transmitted Alamouti STBC over the two transmit antennas at MS-1 is $\mathbf{s}^{(1)}(NT_s) = [s_0^{(1)} s_1^{(1)}]^T$. With (7.9), we rewrite (A-1) in the following form:

$$y_1 = a_0 s_0^{(1)} + a_1 s_1^{(1)} + z(NT_s) \quad (\text{A-3})$$

During the subsequent symbol epoch, $\mathbf{s}^{(1)}((N+1)T_s) = [-s_1^{(1)*} \ s_0^{(1)*}]^T$ is transmitted. The received signal after beamforming y_2 is:

$$y_2 = -a_0 s_1^{(1)*} + a_1 s_0^{(1)*} + z((N+1)T_s), \quad (\text{A-4})$$

where

$$z((N+1)T_s) = \mathbf{w}^H \left\{ \sum_{k=2}^K \mathbf{H}^{(k)} \mathbf{s}^{(k)}((N+1)T_s) + \mathbf{n}((N+1)T_s) \right\}. \quad (\text{A-5})$$

Because of the orthogonality of the virtual channel matrix $\mathbf{A}$ the ML decoder is simplified to a linear combiner that performs a temporal domain diversity combining and produces two decoded symbols for the detection of $s_0^{(1)}$ and $s_1^{(1)}$:

$$\begin{aligned} \begin{bmatrix} \widehat{s_0^{(1)}} \\ \widehat{s_1^{(1)}} \end{bmatrix} &= \begin{bmatrix} a_0^* & a_1 \\ a_1^* & -a_0 \end{bmatrix} \begin{bmatrix} y_1 \\ y_2^* \end{bmatrix} \\ &= \begin{bmatrix} (|a_0|^2 + |a_1|^2)s_0^{(1)} + a_0^* z(NT_s) + a_1 z((N+1)T_s)^* \\ (|a_0|^2 + |a_1|^2)s_1^{(1)} + a_1^* z(NT_s) - a_0 z^{(k)}((N+1)T_s)^* \end{bmatrix} \end{aligned} \quad (\text{A-6})$$

$a_0^* z(NT_s) + a_1 z((N+1)T_s)^*$ and $a_1^* z(NT_s) - a_0 z^{(k)}((N+1)T_s)^*$ are interferences plus noise components for symbols $s_0^{(1)}$ and $s_1^{(1)}$, respectively. For $s_0^{(1)}$ the SINR is represented as:

$$\gamma_d = \frac{(|a_0|^2 + |a_1|^2)s_0^{(1)}s_0^{(1)*}(|a_0|^2 + |a_1|^2)}{(a_0^* z(NT_s) + a_1 z((N+1)T_s)^*)(a_0^* z(NT_s) + a_1 z((N+1)T_s)^*)^*} \quad (\text{A-7})$$

Since two transmitted STBC signals at NT_s and $(N+1)T_s$ instants are uncorrelated and with equal powers, $z(NT_s)$ is uncorrelated with $z((N+1)T_s)$ and have the same powers. The expression for γ_d is simplified to:

$$\gamma_d = \frac{(|a_0|^2 + |a_1|^2)E_s}{|z(NT_s)|^2} = \frac{(|a_0|^2 + |a_1|^2)E_s}{|z((N+1)T_s)|^2}. \quad (\text{A-8})$$

From (7.7) and (A-1) we know that

$$|z(NT_s)|^2 + |z((N+1)T_s)|^2 = \mathbf{w}^H \left\{ \sum_{k=2}^K \mathbf{H}^{(k)} \mathbf{S}^{(k)} + \mathbf{N} \right\} \left\{ \sum_{k=2}^K \mathbf{H}^{(k)} \mathbf{S}^{(k)} + \mathbf{N} \right\}^H \mathbf{w}. \quad (\text{A-9})$$

Therefore, we have

$$\gamma_d = \frac{2E_s \mathbf{w}^H \mathbf{H}^{(1)} \mathbf{H}^{(1)H} \mathbf{w}}{\mathbf{w}^H \mathbf{R}_{n_i} \mathbf{w}} = \frac{2(|a_0|^2 + |a_1|^2)E_s}{2|z(NT_s)|^2} = \gamma_b. \quad (\text{A-10})$$

Now we have proved that the output SINRs at the beamformer and at the STBC decoder are the same.

Appendix B

In this appendix we will prove that the conventional Alamouti STBC decoder is in essence a ML decoder. The orthogonality of the Alamouti virtual MIMO channel simplifies the ML decoding process into a linear decoder.

Let us assume that an Alamouti STBC code word is transmitted over two transmit antennas and is received over one receive antenna. See (2.22) and (2.23), over two symbol epochs, the signals at the receive antenna is represented as:

$$\begin{bmatrix} u(NT_s) \\ u((N+1)T_s)^* \end{bmatrix} = \begin{bmatrix} h_{11} & h_{12} \\ h_{12}^* & -h_{11}^* \end{bmatrix} \begin{bmatrix} c_1 \\ c_2 \end{bmatrix} + \begin{bmatrix} n(NT_s) \\ n((N+1)T_s)^* \end{bmatrix}. \quad (\text{B-11})$$

Let us denote

$$\mathbf{c} = \begin{bmatrix} c_1 & c_2 \end{bmatrix}^T, \quad (\text{B-12})$$

and

$$\mathbf{u} = \begin{bmatrix} u(NT_s) & u((N+1)T_s)^* \end{bmatrix}^T, \quad (\text{B-13})$$

and

$$\mathbf{n} = \begin{bmatrix} n(NT_s) & n((N+1)T_s)^* \end{bmatrix}^T. \quad (\text{B-14})$$

It is a general data model [197]:

$$\mathbf{u} = \mathbf{H}\mathbf{c} + \mathbf{n} \quad (\text{B-15})$$

where $\mathbf{c}$ is the parameter to be estimated. $\mathbf{H}$ is the known matrix and $\mathbf{n}$ is a AWGN noise vector, $\mathbf{n} \sim \mathcal{N}(0, \mathbf{R})$. Under these conditions, the pdf is:

$$f(\mathbf{u}, \mathbf{c}) = \frac{1}{2\pi \det(\mathbf{R})^{\frac{1}{2}}} \exp \left[-\frac{1}{2} (\mathbf{u} - \mathbf{H}\mathbf{c})^H \mathbf{R}^{-1} (\mathbf{u} - \mathbf{H}\mathbf{c}) \right]. \quad (\text{B-16})$$

The ML estimation of $\mathbf{c}$ is obtained by minimizing:

$$L(\mathbf{c}) = (\mathbf{u} - \mathbf{H}\mathbf{c})^H \mathbf{R}^{-1} (\mathbf{u} - \mathbf{H}\mathbf{c}). \quad (\text{B-17})$$

Solving for $\hat{\mathbf{c}}$ produces the ML estimation:

$$\hat{\mathbf{c}} = (\mathbf{H}^H \mathbf{R}^{-1} \mathbf{H})^{-1} \mathbf{H}^H \mathbf{R}^{-1} \mathbf{u}. \quad (\text{B-18})$$

If the noise components at two symbol epoches $n(NT_s)$ and $n((N+1)T_s)$ are uncorrelated, then we have $\mathbf{R} = \sigma^2 \mathbf{I}_2$. Because of the orthogonality of $\mathbf{H}$ that

$$\mathbf{H}\mathbf{H}^H = (|h_{11}|^2 + |h_{12}|^2) \mathbf{I}_2, \quad (\text{B-19})$$

we have

$$\hat{\mathbf{c}} = \frac{1}{(|h_{11}|^2 + |h_{12}|^2)} \mathbf{H}^H \mathbf{u}, \quad (\text{B-20})$$

which is exactly the conventional Alamouti STBC decoder as reported in [139].

Author's Publications

Book Chapter

N. C. Karmakar and **C. Sun**, "Adaptive Array Antennas", *Encyclopedia of RF and Microwave Engineering*, John Wiley and Sons, Kai Chang (Editor), 2005.

Journal Papers

1. **C. Sun** and N. C. Karmakar, "Direction of arrival estimation based on a single port smart antenna using MUSIC algorithm with periodic signals," *International Journal of Signal Processing*, ISBN 1304-4494, in press 2005.
2. **C. Sun**, A. Hirata, T. Ohira, and N. C. Karmakar, "Fast Beamforming of Electronically Steerable Parasitic Array Radiator Antennas: Theory and Experiment," *IEEE Trans. Antennas Propagat.* vol. 52, no. 7, July 2004.
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1. **C. Sun** and N. C. Karmakar, "A DOA estimation technique based on a single-port smart antenna for position location services," in *Proc. Asia-Pacific Microwave Conf. 2002*, Kyoto, Japan, Nov. 2002, vol. 1, pp. 196-199.
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