Compensation for the mutual coupling effect in smart antenna arrays

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Compensation For Mutual Coupling Effect In Smart Antenna Arrays

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School of Electrical and Electronic Engineering

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

2007

Statement of Originality

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

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Date

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Zhang TongTong

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Summary

Smart antennas techniques are considered to be the best means to mitigate the effects of severe and dynamic interference on the performance of modern communication systems. But in a practical array system, the receiving elements of the smart array must be of some physical size. The elements receive and re-radiate the incident signals, which results in the mutual coupling among them. The mutual coupling distorts the linear phase front of the incoming signal which results in an inaccurate DOA estimation and significant degradation of the system performance. Therefore, any practical implementation of smart antenna systems requires compensation for mutual coupling effect in smart antennas.

Many methods have been proposed to identify or compensate for the mutual coupling effect in antenna arrays. Among of them, the Method-Of-Moment and Open-circuit-voltage method were two popular methods to tackle the problem. However, in these two conventional methods, some unrealistic assumptions were used and the accuracy was not enough to further improve the system performance. For examples, in MOM method, the knowledge of the entire incident field and current distribution on the antenna elements is assumed known. In practice, this information is not available.

In open-circuit voltage method, it assumes that open-circuit voltages at each terminal load of antenna elements are free of mutual coupling effect among them. This assumption is only valid in a limited sense. Moreover, this method fails to account for the effect of the compensation process on the noise. A new method was proposed for the compensation of the mutual coupling effect by using a new mutual impedance which requires only an estimated current distribution. Except that, there is no assumption required for the new method.

In present work, four novel ways that can significantly reduce the mutual coupling in Uniform Linear Arrays (ULAs) and Uniform Circular Arrays (UCAs) are proposed. DOA estimation algorithms and adaptive signal processing algorithms are studied together with the proposed compensation methods to improve the overall system performance. Comparing with the conventional method, the proposed method does not involve any unrealistic assumptions but relies on the calculation of new mutual impedance. Computer simulations show that the proposed method make a significantly improvement of the performance and produce more accurate detection results than the conventional open-circuit-voltage method and MOM. The self-designed EM solver based on the MOM is used to carry out the simulation.

A structured least square method for simultaneous estimation of the mutual coupling matrix and DOAs of signal sources in ULA and UCA is proposed. This method is very useful in the calibration of ULA and UCA when the DOAs of calibration sources are not known. Simulations show that the proposed method obtain a very accurate estimation of mutual coupling matrix and DOAs at same time.

This research work deals with the mutual coupling compensation in the area of smart array antennas and some array signal processing issues. A number of problems within this diverse area are analyzed using theory and methods not only from signal processing but also from antennas, computation of electromagnetic field and mathematic and statistic theory. In fact, this interdisciplinary character is a major theme and contribution of the thesis since most of the literatures on the analyzed topics did not make this connection.

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List of Abbreviations and Symbols

Abbreviations

1D	One dimensional
2D	Two dimensional
3D	Three dimensional
2G	Second-generation mobile systems
2.5G	Evolution of Second-generation systems
3G	Third-generation mobile systems
4G	Fourth-generation mobile systems
3GPP	Third Generation Partnership Project
AA	Adaptive antennas
ABC	Always Best Connected Concept
ADC	Analog to Digital Converter
AFP	Adjacent-fitness-pairing
AMPS	Advanced mobile phone services
AR	Auto-regressive
BS	Base station

BER	Bit error rate
BF	Beamforming
BlueTooth	Enabling devices to communicate over short distances
BPSK	Binary Phase Shift Keying
BSC	base station controller
BTS	base transceiver stations
BWA	broadband wireless access
CCI	Co-channel interference
CCIs	Co-channel interferences
CDMA	code-division multiple access
CRB	Cramer-Rao bound
CEPT	Conference of Postal and Telecommunications Administrations
CMA	Constant Modulus Algorithm
C/N	Carrier-to Noise ration
CT2	Second generation cordless Telephone
DBF	Digital beamforming
DBS	Direct Broadcasting System
D/C	Down Converter
DCS	Digital communication System
DCS-1900	Digital communication System–1900
DECT	Digital European Cordless Telecommunication
DFT	Discrete Fourier Transform
DPE	Deterministic Parametric Estimation

DOA	Direction of Arrival
DSL	Digital subscriber Line
DSP	Digital Signal Processing
DCS	Digital communication system
DF	Direction Finding
EDGE	Enhanced Data Rates for GSM Evolution
EIRP	Effective Isotropically Radiated Power
EIR	Equipment Identity Register
ES	Element space
ESA	European Space Agency
ESPRIT	Estimation of signal parameters via rotational invariance technique
ESTEC	European Space Research and Technology Center
ETACS	European total access communication system
EFTA	European Free Trade Association
EU	European Union
EV-DO	Evolution-Data Only
FCC	Federal Communications Commission
FDD	Frequency division duplex
FDMA	Frequency division multiple access
FOV	Field of view
FM	Frequency modulation
FPLMTS	Future Public Land-Mobile Telecommunications Systems
GA	Genetic algorithm

GEOS	Geostationary satellite
GGSN	Gateway GPRS Support Node
GMSK	Gaussian minimum shift keying
GPRS	General Packet-switched radio service
GSM	Global System for Mobile communications
G/T	Gain-to-Noise temperature ration
HEOS	Highly elliptical orbit satellite
HLR	Home location register
HSDPA	High Speed downlink packet access
HS-DSCH	High Speed Downlink Shared Channel
IF	Intermediate frequency
ISI	Inter-symbol interference
IS-95	EIA interim Standard for U.S.Code Division Multiple Access
IS-96A	EIA interim Standard for 8.6 kbps vocoder
IS-136	EIA interim Standard 136 for USDC with Digital Control Channels
ISDN	Integrated Services Digital Network
JTACS	Janpanese Total access communications system
LEOS	Low earth orbit satellite
LP	Linear prediction
LMDS	Local multipoint distribution service
LMS	Least Mean Square
LS	Least Square
MAI	Multiple-Access Interference

MBF	Microwave Beamforming
MCM	Mutual Coupling Matrix
MANETs	Mobile ad-hoc networks
MEOS	Medium earth orbit satellite
MMDS	Multichannel multipoint distribution service
MMSE	Minimum Mean Squared Error
MMIC	Monolithic microwave integrated circuit
MIMO	Multiple-Input-Multiple-Output
MISO	Multiple-Input-Single-Output
ML	Maximum Likelihood
ML-GA	ML algorithm optimized by GA
MLM	Maximum Likelihood Method
MN	Minimum-norm
MSC	Mobile switching center
MSE	Mean Squared Error
MUSIC	Multiple signal classification
MVDR	Minimum variance distortional response
NMT	Nordic Mobile Telephone
NN	Neural network
NTT	Nippon Telephone and Telegraph
OCV	Open circuit voltage
PAN	Personal area network
PACS	Personal Access Communication System

PDC	Personal digital cellular
PHS	Pocket hand-phone service
PL	Position location
PTTs	Telegraph and Telephone organizations
PN	Pseudo-random sequence
PSBE	Parametric Subspace Based Estimation
PSK	Phase shift keying
PSO	Particle swarm optimization
PSTN	Public Switched Telephone Network
QoS	Quality-of-service
QPSK	Quadrature Phase shift keying
RF	Radio Frequency
RLS	Recursive Least Squares
RSS	Received signal strength
RMS	Root mean square
RMSE	Root-mean-squared error
RNCs	Radio network controllers
SA	Smart antennas
SCARP	Smart Communication Antenna Research Program
SDMA	Space division multiple access
SGSN	Serving GPRS Support Node
SER	Symbol error rate
SLL	Side-lobe level

SFIR	spatial filtering for interference reduction
SIMO	Single-Input Multiple-Output
SIR	Signal-to-Interference ration
SMI	Sample matrix Inversion
SNIR	Signal-to-noise-and-interference ratio
SNR	Signal-to-Noise ratio
STD	Standard Deviation of Error
STCs	Space-time codes
TACS	Total access communications system
TDD	Time Division Duplex
TDM	Time division multiplex
TDMA	Time division multiple access
TLS	Total least square
UE	User equipment
UCA	Uniform Circular Array
ULA	Uniform Linear Array
UHF	Ultrahigh Frequency
UMTS	Universal Mobile Telecommunications systems
UM	Uniform mutation
UTRAN	Universal Terrestrial Radio Access Network
VHF	Very High Frequency
VLR	Visiting locationregister
WCDMA	Wide-band CDMA

WLANs Wireless Local Area Networks

Symbols

a	Scalar
$a_i(\theta_l)$	i -th antenna element complex response
$\mathbf{a}$	Vector in time domain
$\mathbf{A}$	Matrix in time domain
A_k^2	Average power in the k th signal
A_z	The z -component of the magnetic vector potential
$\mathbf{a}(\theta_l)$	Array steering vector for signals from direction θ_l
α	Radius of antenna element
c	Propagation speed of light
d	Inter-element distance of ULA
e_n	corresponding eigenvectors
$E\{*\}$	Expectation operator
$\mathbf{E}_n$	Noise subspace
$\mathbf{E}_s$	Signal subspace
E^i	Incident field
E^s	Scattered field
$\vec{E}_{kl}$	Incident field at the k th antenna due to the l th antenna
$\vec{E}_m^l$	Field at the m th mode due to the current on the l th antenna element
f_c	Carrier frequency
$\mathbf{I}$	Identity matrix
i_j	Terminal current of j -th array element
$[I]$	Current vector

j	Imaginary unit
$\vec{J}_m$	m th test function
Jdv'	A Z -directed volume current element
ℓ	Length of antenna element
L	Length of the array
M	Number of signal sources
$n(t)$	Random noise component
$\mathbf{n}(t)$	Noise vector
N	Number of antennas in the antenna array
n	Number of antennas in the antenna array
$P_{\mathbf{s}}$	Source covariance matrix
P_k	Power of the signal
p_k	Polarization of the k th signal
P_{MUSIC}	MUSIC spectrum
P_{noise}	Projection on the noise subspace
q_i^2	Average power of the i -th source
r	Distance from the source to the array
$r(t)$	Received signal
$\mathbf{R}$	Correlation matrix
R	Distance between the observation point and the source point
$\text{Re}\{\cdot\}$	Take the real part
$R_{\mathbf{y}}$	Covariance matrix of $\mathbf{y}(t)$
$\tilde{s}_k(t)$	k -th signal source

$\mathbf{s}(t)$	Signal vector
$\mathbf{s}(\phi_k, \theta_k)$	Array steering vector response to a signal from ϕ_k, θ_k
S_{21}	Scattering parameter
N_e	Number of segments in an antenna
$u_k(t)$	Time varying function of amplitude of k -th signal source
$\mathbf{U}$	Unitary matrix
V_g	Open circuit voltage of array element
V_t	Terminal voltage vector for antenna array
v_i	Output voltage of i -th array element
V_c	Corrected voltage matrix
V_{kl}^T	OCV at the k th antenna terminal due to the current on the l th antenna
w	Complex weights
$\mathbf{x}$	Complex valued signal amplitude vector
$y_i(t)$	The output of i -th array element at time t
$\mathbf{y}(t)$	Array output vector
$[Y]$	Admittance matrix
z_{mn}	Mutual impedance between the m -th and n -th segments of the antenna
Z_g	Internal impedance of array element
Z_L	Terminal load
$[Z]$	Impedance matrix
Z^T	Terminal impedance matrix
Z_0	Normalized impedance matrix
λ	Operating wavelength

λ_l	l -th eigenvalue of the matrix
Λ	Diagonal matrix
θ	Elevation angle of signal of interest
ϕ	Azimuth angle of signal of interest
Φ	Scalar potential
$\tau_i(\theta_l)$	Propagation delay between the reference point and i -th antenna
σ^2	Noise power
$\nu_k(t)$	Time varying function of phase of k -th signal source
Δ	Length of each segment
σ_n	Standard deviation of random additive noise
σ	Conductivity of wire antenna
ϕ_k	Phase shift of the signal
ψ_k	Carrier phase of the k th signal at the coordinate origin
ω_c	Carrier frequency
ρ_{kj}	k th signal phase at the j th array element
$\psi(z, z')$	The free space Green's function
$\vec{I}_p$	Current on the p -th antenna element
ℓ_p	Length of the p -th antenna element
$\overline{\overline{G}}_{e0}(\vec{R}, \vec{R}')$	Free-space dyadic Green's function
ψ_k	Carrier phase of the k -th signal
$(\cdot)^H$	Complex conjugate transpose of a vector or matrix
$(\cdot)^T$	Transpose of a vector or matrix
$\ \cdot\ $	Matrix or vector norm (default is the Euclidean norm)

Chapter 1

Introduction

1.1 Motivation

Wireless communications are growing at an explosive rate and playing an increasing role in today's society. There are many technical areas that are covered in the field of wireless communications. As a result, a diverse range of products and services are currently on the market.

Smart antennas (SA) with adaptive BF and direction finding are the emerging technologies which have many advantages such as higher network capacity and better transmission quality. The research on SA therefore attracted much attentions in recent mobile communications [1].

1.1.1 The Evolution of Mobile Communication Systems

To gain a historic perspective, it is worthy noting that land mobile radio systems were introduced as early as the 1920s to provide two-way communications to automobiles. These radio systems evolved into a number of specialized services and offered paging, dispatch, two-way voice and data communications to mobile users. The evolution of wireless communication is illustrated in table 1.1. In the early 1980s, the first generation analog FM systems such as the Advanced Mobile Phone System (AMPS) in the United States, the Total Access Communication System (TACS) in Europe and the Japanese TACS system (JTACS) in Japan, were deployed, bringing untethered wireless voice access to the Public Switched Telephone Network (PSTN).

Table 1.1: Evolution of wireless communications.

	First Generation systems	2G systems	3G systems
Time Frame	1984-1996	1996-2000	2000-2010
Services	Analogy Mobile Telephony Voice Band Data	Digital voice messaging	High speed data Broadband video Multimedia
Architecture	Macrocellular	Microcellular picocellular Wireless Local Loop	Circuit switched packet switched
Radio Technology	Digital modulation Analog FM FDD-FDMA	CDMA, possibly CDMA TDMA using FDD and TDD	combined with TDMA ,with TDD and FDD vairants
Frequency Band	800 MHz	800+1900 MHz	2 GHz+
Examples	AMPS TACS ETACS NMT450/900 NTT JTACS /NTACS	cdmaone(IS-95) GSM/DCS-1900 US TDMA IS-136 PACS PHS	WCDMA cdma2000 EDGE

The second generation (2G) mobile systems were introduced in 1990s to shift communication systems from analog to digital. Short messages, low-rate data services and wide area roaming for the user were added in the 2G mobile system. The 2G systems became universal due to lightweight, low-power, portable subscriber handsets with the clarity of digital voice and enhanced features. There are Global System for Mobile communications (GSM), Time Division Multiple Access (TDMA), Personal Digital Cellular (PDC), and Code Division Multiple Access cdmaOne systems in 2G communications society worldwide. Adding the technique of packet switched radio connections over the air, 2G systems were evolved to 2.5G mobile systems. For GSM systems, General Packet-Switched Radio Service (GPRS) is an indication of 2.5G systems. For the users, GPRS offers a higher transfer data rate, more accurate billing information and always online service.

Continuing development of wireless communication networks shifted to the Third-Generation (3G) systems, and the worldwide introduction of Wideband CDMA (WCDMA) took place in 2001 and 2002. As shown in Figure 1.1, WCDMA, Enhanced Data Rates for GSM Evolution (EDGE), cdma2000 systems are the main standards in the world communications market [3]. While the 3G wireless communication systems started to take their role in current market, more advanced technologies for improvement or augment of 3G services have entered the next generation wireless communication era [3].

Beyond 3G evolution are 4G mobile communication systems which are built on 3G and evolved 3G systems. At the moment, many possible applications of 4G systems, such as augmented reality application, ad hoc networking and multihop networks have

been discussed and developed. The 4G systems are to a large extent in the research stage and are expected to be launched sometime after 2010 [4]. The worldwide research activities in this growth industry indicate an anticipation that communication with a mobile device anywhere on the globe at all times will be available in the near future [5]. In the path of evolution of wireless mobile communication systems toward future generations, the wireless communication technologies advanced with the consideration of aiming at higher channel capacity, better system performance and smaller size of portable communication devices. Under these performance objectives and design constraints, as a result, the technique of SA is introduced and developed. The technology of smart antenna has been developed in 2G, 3G and 4G communication systems.

There are many efforts on the design of smart antenna arrays and the associated techniques and algorithms. A smart antenna consists of several identical antennas, whose operation and timing are usually controlled by an adaptive signal processor which processes signals adaptively in order to exploit the spatial dimension of the mobile radio channel [7]. Exploitation of the spatial dimension can enhance the system capacity, improve link quality by mitigating multi-path fading and co-channel interference, increase the transmission data rate by the simultaneous transmission of multiple streams through different antennas.

From the above, it is known that the principal reason for introducing SA is to increase the capacity and enhance the quality. The origins of smart antenna systems can be traced to the 1960's. During World War II, the Ultrahigh Frequency (UHF) and mi-

Figure 1.1: The path to future generation mobile systems [3].

microwave array antennas were introduced in radar systems. Today, arrays at microwave frequencies and above are used extensively in satellite communication systems [8]. In these applications, SA has a great capability to perform direction finding tasks and to null out unwanted interferers. Moreover, SA has been used for military application for consideration time. However, only in recent years have they gained acceptance in the commercial wireless industry.

Several projects and field trials have emerged to reinforce the benefits of smart antenna systems. Many applications of antenna arrays have also been introduced in recent years for mobile communication systems. In the context of commercial mobile communication systems, the array outputs can be modified to enhance the desired signal reception and meanwhile suppress the undesired ones. The smart antenna receivers provide acceptable of accepting error performance and hence maximize SINR for each user in the communication systems. Smart antennas are proven to be robust against channel fading and interference, hence to achieve higher data rates and are being used widely in communication systems [9].

1.1.2 Demands for Compensation for the Mutual Coupling Effect in Smart Antennas

In mobile communications, SA can be integrated into the communication systems to improve the system performance. Physical limitation in some of these systems and devices forces multiple antennas to be spaced closely which results in significant amount of mutual coupling effect among the antenna elements, and severely degrades the per-

formance of the smart antenna array.

Current worldwide research efforts in the SA area are focusing on the following critical issues [4]:

- The design and development of advanced smart antenna processing algorithms that should adapt to varying propagation and network conditions and robustness against network impairments.
- The design and development of innovative smart antenna strategies for optimization of performance at the system level.
- Realistic performance evaluation of the proposed algorithms and strategies and the introduction of suitable performance metrics and simulation methodologies.
- Analysis of implementation, complexity and cost efficiency issues involved in realization of the proposed smart antenna techniques for future generation wireless systems.

Although SA techniques are considered to be the best means available to mitigate the effects of severe, dynamic interferences on the performance of modern communication systems, but in practical array systems, the receiving elements of the smart array must be of some physical size. Therefore, the elements of receiving antennas not only receive but also re-radiate the incident signals. The re-radiated signals cause mutual coupling effect among the elements of receiving antennas. The mutual coupling effect distorts the linear phase front of the incoming signal, which causes inaccurate DOA estimates and significantly degrades the system performance. Any practical implemen-

tation of DOA estimation therefore requires compensation for mutual coupling effect. Moreover, signals from a target can undergo reflection which creates delayed multiple returns and amplitude-weighted replicas of the direct signal to the array. In addition, intentional jamming signals generate coherent interference [22], [23]. For example, in radar, multi-path returns give rise to secondary signals that are completely coherent with the original signal.

Since electromagnetic coupling signals received by array elements will no longer be independent, and become dependent on each other. Mutual coupling effect influences the cross-correlation between the received signals which depends on the array geometry, antenna element type, frequency of signals, near-field scatterers and DOAs of the incoming waves.

The impedance and radiation pattern of an antenna element change when the element is radiating in the surrounding area of other antennas, causing the maximum and nulls of the radiation pattern to shift. Furthermore, these detrimental effects intensify as the inter-element spacing is reduced [21], [67]. Therefore, when these mutual coupling effects are not taken into account by the adaptive array algorithms, the overall system performance will degrade severely. These can not be neglected in communication systems. As a result, compensation for mutual coupling effects in SA will be an important strategy for optimization of performance at the system level. Thus, the systems and communication devices which are using SA have been posed some challenges. Methods are required to accurately and effectively mitigate the effect of the mutual coupling of antenna arrays in wireless communication systems and devices.

In most of mobile communication systems, capacity and performance are also limited by another two major impairments, multipath and co-channel interference. Multipath is a condition which arises when a transmitted signal undergoes reflection from various obstacles in the propagation environment. This gives rise to multiple signals arriving from different directions. Since the multipath signals follow different paths, they have different phases when arriving at the receiver. The degradation in signal quality is therefore caused when they are combined at the receiver due to the phase mismatch. Co-channel interference is the interference between two signals that operate at the same frequency. In cellular communication the interference is usually caused by a signal from a different cell occupying the same frequency band.

In most studies, the co-channel interference, the effects of mutual coupling among array elements and scattering effect due to multi-path propagation are not considered, thus leading to potentially less optimum system performance. In this thesis, some improved methods where the effects of both mutual coupling and scattering are considered to enhance the system performance will be presented in later chapters.

There are several reasons that drive us to find the techniques of compensation of mutual coupling effect in smart antenna systems and can be summarized as following [10]:

- Mutual coupling changes the induced current phase and current distribution on each antenna element. This effect results in the unwanted changing of antenna gain, beam-width, radiation pattern and input impedance.
- Mutual coupling effect is rather drastic as the inter-element spacing drops below

half wavelength.

- The failure to recognize the presence of mutual coupling degrades the system performance.
- With strong mutual coupling among the array elements, the output SINR of SA drops significantly compared with the mutual coupling compensated case.
- The presence of mutual coupling among the antenna elements reduce the response speed of smart antenna systems.

From the above, it is understood that although tremendous researches and development works have been done in advancing the technology of SA to improve the performance of wireless communication systems, there are still plenty of room for improvement in these area. Some accurate and simple models of mutual coupling effect are essential for better system analysis. The thesis addresses these problems which motivate my research for robust techniques to enable the benefits of SA to be enhanced and augmented. This is the motivation of my Ph.D thesis.

1.2 Objectives

The objectives of this research are to study various conventional compensation techniques and proposed new mutual coupling compensation techniques for SA system. DOA estimation and adaptive signal processing algorithms are studied together with these compensation techniques to improve the system performance.

In practical active sensing situations such as radar and sonar, a known waveform of finite duration is generated which in turn propagates through a medium and is reflected by some targets back to the point of origin. Both amplitude and phase of the transmitted signal are usually modified by the target characteristics, which by themselves might be changing with time and its position in space. These disturbances give rise to a random return signal. The problem is that these signal processing algorithms generally ignore the electromagnetic behavior of the receiving antennas. The receiver is assumed to be an ideal, equally spaced linear or planar array of isotropic point sensors. In these cases, it was assumed that the array only receives and not re-radiates the incident signals. Most signal processing techniques rely heavily on this assumption. Therefore, some efficient compensation techniques and high-resolution DOA estimator need to be proposed for smart antenna systems on more practical assumptions. There are five focused areas:

- Firstly, this thesis presents an insight into a general overview of modern wireless communication systems. The evolutionary path of the wireless communication systems and the quest for more capacity are illustrated. The basic problems of mobile communications and how the SA can be exploited to mitigate those problems are introduced. The principal theories of a wide range of SA are presented. The related knowledge and techniques covering a wide range of array signal processing, characteristic of Electromagnetic (EM) field of antenna arrays and applications is presented. The theory of underlying physical principles of linear and planar phased array antennas is analyzed.

- Secondly, the mutual coupling in antenna array is a well-known phenomenon. The thesis analyzes the impact of effect of mutual coupling on adaptive array antenna signal processing.
- Thirdly, the conventional definition of self impedance and mutual impedance matrixes are developed via the method of moments. The proposed new definition of mutual impedance and practical measurement procedures for the new mutual impedance are described.
- Fourthly, the performance for array antennas in terms of probability of detection of DOAs of signal interests combined with three compensation techniques is demonstrated .
- Finally, the monopole and dipole antennas are designed, simulated, and fabricated. These antenna elements are used in my research work in the measurement of self and mutual impedances.

The outcomes of the Ph.D research have originated a few fundamental contributions and developments over the existing methods which are outlined in next section.

1.3 Major Contributions of the Thesis

In responding to the aforementioned problems, we focus our research efforts in the direction in which robust methods should be developed to realize the mitigation of those detrimental effects in smart antenna arrays. The thesis deals with techniques for mutual coupling compensation in the area of adaptive array antennas and some array

signal processing issues. A number of problems within this diverse area are analyzed using theory and methods not only from signal processing but also from antennas, computation of electromagnetic field, and mathematic, statistic theory. In fact, this interdisciplinary character is a major theme and contribution of the thesis since most of the literatures on the analyzed topics did not make this connection. The following tasks have been envisaged in my research work:

The thesis proposed four novel ways that can reduce or at least control the mutual coupling in ULAs and UCAs.

- 1.) In first method, an effective method was introduced to compensate the mutual coupling effect in the Estimation of Signal Parameters via Rotational Invariance Techniques (ESPRIT). Study of two closely spaced arriving signals showed that the high-resolution capability of ESPRIT can be achieved only when the mutual coupling effect is compensated by using the new compensation method. Critical situations such as a large signal level difference, an increased mutual coupling effect resulting from a more compact-size antenna array, and signals coming from a non-horizontal elevation angle were also studied by using the ESPRIT algorithm. Results showed that the new compensation method, when applied to ESPRIT, is more accurate and more robust than the previous methods.
- 2.) The second technique is to solve the problem of two dimensional (2-D) angle estimation (azimuth and elevation) of multiple plane waves incident on an UCA. In this method, a previously proposed model for mutual coupling in method 1 was modified to be applicable to an UCA. The effect of near-field scatterers with

the mutual coupling effect were specifically investigated to validate the model. Self-designed EM simulators are utilized for this purpose. Once the effect of near-field scatterers is characterized, correlation properties between antenna elements in an circular antenna array can be obtained more accurately. The correlation estimation algorithms can be developed accordingly. This method circumvents the difficulty of dealing with coherent signals in the estimation of 2-D DOAs. The new 2-D DOA algorithm Maximum Likelihood-Genetic Algorithm (ML-GA) which using ML technique and being optimized by the GA is introduced. The ML-GA is shown to be less computationally demanding and statistically more efficient than the Maximum Likelihood Method (MLM). The difference in the structure of the mutual impedance matrixes of UCA and ULA under the current compensation method is fully derived. Results of simulations which validate the theoretical performance expressions are presented.

- 3.) The third method in the thesis is to apply newly proposed technique of compensating for the mutual coupling effect in a linear adaptive dipole array employed in adaptive nulling of interference signals. The new method does not require the knowledge of the elevation angles of the Signal of Interest (SOI) and the interferences and still works if the elevation angles of the SOI and the interferences do not deviate too much from the horizontal direction. Computer simulations for a number of rather extreme signal environments have been carried out to testify the robustness and the capability of the new method.
- 4.) The fourth technique in my thesis, which is totally based on signal processing perspective, is to propose a structured Least Square (LS) method for simultaneous

estimation of the Mutual Coupling Matrix (MCM) and DOA of signal sources in ULA and UCA. An analytical model for the mutual coupling is presented. The effect of mutual coupling are modelled in the form of a complex Toeplitz and Circulant matrix. The DOAs and MCM can be simultaneously estimated by the proposed method when the observations of at least two different DOAs are available. This method is especially useful for the calibration of ULA and UCA. It significantly simplifies the procedure of the calibration of the mutual coupling. Simulations also show the strong dependency of the convergence behavior on the signal to noise ratio. We find that only at very low signal-to-noise ratios, more iteration rounds are necessary, e.g. at a SNR value of 0dB only 50 iterations are needed. Simulation results confirm the efficiency of the proposed method.

- 5.) Finally, the self-designed EM solver which is based on the methods of moment is used to carry out the simulation.

1.4 Organization of the Thesis

The overall organization of the thesis is as follows. After some introductory remarks on SA about its historical background and the state of the art in Chapter 1, the impact of their introduction into current and future mobile communication systems is discussed. The major emphasis in the Chapter 1 is on the “Demands for compensation of mutual coupling in SA”, it shows how the effect of mutual coupling decreases the main aspects of the performance of smart antenna systems. The chapter is concluded with the motivations , objectives and the main contributions of my Ph.D research study.

Chapter 2 gives a comprehensive description of the principle concepts, system architecture and analytic models of adaptive arrays. Simple examples are given that state how an adaptive array accomplishes its major purpose: tracking desired signals and nulling interferences. It reviews analytic techniques from different viewpoints for adaptive antenna system, such as analytic approaches of Computational Electromagnetic (CEM), beam-forming, DOA estimation, Interferences nulling, Multiple-Input-Multiple-Output (MIMO) system. It also discuss the practical effects that actually control the performance of AA. Mutual coupling effect is analyzed in a general way to show how arbitrary desired signal and interference may be handled in an adaptive array analysis. It reviews the different strategies for compensation of mutual coupling effect in adaptive array antennas, and compares the various schemes in the terms of DOA estimation and interferences nulling. It also presents the discussion of the DOA estimation. Covariance matrix eigenvalue behavior is described. It shows how eigenvalues depend on signal arrival angles, number of elements, element spacing, and element patterns. This chapter summarizes many practical DOA estimation techniques for adaptive arrays, and compares their performances, sensitivities and limitations.

Chapter 3, presents the newly proposed method of compensation for the mutual coupling effect and gives the detailed description of it. The new mutual impedance measuring procedure are presented, and experimental results which is compared with simulation results obtained by using new method are analyzed.

Chapter 4, studying of two closely spaced arriving signals and shows that the high-resolution capability of ESPRIT can be achieved only when the mutual coupling effect

is compensated for by using the newly proposed compensation method.

Chapter 5, the problem of 2-D angle estimation of multiple plane waves incident on a UCA is considered. Results of simulations validate the theoretical performance expressions.

In Chapter 6, a more practical structured LS method for simultaneous estimation of the MCM and DOA of signal source is proposed. The DOAs and MCM can be simultaneously estimated by the proposed method when the observations of at least two different DOAs are available. Simulation results also confirm the efficiency of the proposed method.

Chapter 7 continues the discussion of the proposed compensation method, with focus on a linear adaptive dipole array employed in adaptive nulling of interference signals. Computer simulations for a number of rather extreme signal environments are carried out to testify the robustness and the capability of the new method.

I conclude the thesis in Chapter 8 by addressing the impact of the new compensation technique on the performance of DOA estimation and adaptive nulling of interferences for both ULA and UCA. The compensation technique is attractive for communication applications because it can improve the performance of automatically tracking the DOA of desired incoming signals. Finally, some concluding remarks and recommendations for future work are provided.

Chapter 2

Overview of Smart Antennas Technology

In this chapter, some introductory remarks regarding the evolutionary path are presented. Some of the basic problems of mobile communications and the possibility of using the SA to mitigate these problems are described as well. An overview of the SA technologies for mobile communication systems is introduced. The overview is focused on describing the influence that SA have on wireless communication systems. Finally, a comprehensive review of algorithms for DOA estimation is studied.

2.1 Smart Antennas

2.1.1 Introduction

Over the last few years the demand for service provision via wireless communication bearer has risen beyond all expectations. The radio frequency spectrum is a finite and valuable resource. For a fixed bandwidth of spectrum, there is a fundamental limit on the number of radio channels that is realized by a mobile communication system operating over this bandwidth. In order to increase the capacity, a considerable amount of work has been done on the use of time, frequency and coding techniques. Some of these efforts have resulted in multiple-access standards, such as frequency-division multiple access (FDMA), TDMA, and CDMA. To meet the ambitious requirements introduced for future wireless systems, new ‘intelligent’ or ‘self-configured’ and highly efficient systems will most certainly be required. In the pursuit of schemes that will solve current wireless communication problems, attention has recently turned to spatial filtering methods using advanced antenna techniques: adaptive or SA. Filtering in the space domain can separate spectrally and temporally overlapped signals from multiple mobile units and thus the spatial dimension can be exploited as a hybrid multiple access technique complementing FDMA, TDMA and CDMA [24]. Many efforts have been made on the design of “smart” antenna arrays and the associated technologies.

Smart antenna systems, which employ array antennas coupled with adaptive signal processing techniques at the Base Station (BS), are used to enhance system capacity, range coverage, and tracking efficiency. They have been considered as core technologies

for current and future wireless communication systems. Smart antenna arrays can also help reduce multi-path fading. Note that several different definitions for SA are used in the literatures. One useful and consistent difference between a smart/adaptive antenna and a fixed antenna is their respective adaptive and fixed lobe-patterns properties. We

Figure 2.1: The basic block diagram of smart antenna system.

define a smart antenna as an array of antennas which can be arranged in space and interconnected to produce a directional radiation pattern as shown in the Figure 2.1. Such a configuration of multiple radiating elements is referred to as a smart antenna array whose signals are processed adaptively in order to exploit the spatial domain of the mobile radio channel. Usually, in the narrow-band antenna cases, the signals received at the different antenna elements are multiplied with complex weights (w) and then summed up. The weights are chosen by the smart antenna system adaptively.

Smart antenna arrays may be used in various configurations for mobile communications, some of which are discussed here. The geometry of the antenna locations can

Figure 2.2: A rectangular array antennas configuration [13].

vary widely. But the most common configurations are to place the antennas along a straight line (linear array) or around an arcade (circular array). When array element centers are located in a plane, it is said to be a planar array. Common examples of planar arrays are circular and rectangular arrays. For rectangular array, the element

centers are contained within a rectangular area, as shown in Figure 2.2. The array geometries that are widely surveyed in literatures and practical applications are ULAs and UCAs. The ULAs have relatively simple structures. Normally, this type of arrays consists of uniformly spaced antenna elements which may be dipoles, monopoles or helix antennas. All these elements are mounted on a ground plane.

ULAs have front to back ambiguity. It cannot distinguish the signal coming to the front or from the back of the array. For example, if the second signal is exactly 180 degrees from the first signal, ULA will not be able to tell the difference between the first and second signal. ULAs do not provide an appropriate solution to scenarios wherein 360° field of view is required.

Planar array provides a large aperture and may be used for directional beam control by varying the relative phase of each element. It can also be used with a reflecting screen behind the active plane. In planar arrays, the UCAs are more versatile comparing to others as they provide more symmetrical patterns with lower side lobes and much higher directivity (narrow main beam). They can be used to scan the main beam toward any point in spaces provide 360° azimuthal coverage and information on source elevation angles.

2.1.2 Types of Smart Antennas

In the category of smart antenna systems, the systems can be classified into five categories according to their structures and mechanism: phased array antennas, parasitic array antennas, switched beam antennas, digital beam-forming array antennas, and

diversity array antennas [27]- [36].

- 1.) The phased array antennas cover a wide spectrum of frequency, from 1 MHz up to and including millimeter wavelengths. They have wide range of communication and radar applications including cellular telephone BS, tracking satellite and GPS antennas, commercial broadcasting antennas, linear radar arrays and planar-rectangular, planar-circular radar arrays. Phased array antenna based communications links are anticipated to deliver high data rates without the risk of the single-point failure of the gimbaled motors and transmitters used in reflector-based systems. Phased-array antennas contain a multitude of radiating elements, typically arranged in a rectangular or triangular tessellation. Beams are formed by electrically adjusting the relative phase of the radiating elements using ferrite or semiconductor devices.
- 2.) Parasitic array antennas and switched beam antennas have played important roles in satellite communication and military scenarios over the past few years [27]. Parasitic array antenna mitigate the digital beam-forming (DBF) and microwave beam-forming (MBF) deployment problems, in which one central element is connected to the sole radio frequency (RF) port and a number of surrounding parasitic elements form the array. Beam steering is achieved by adjusting the load reactance at parasitic elements surrounding the central active element or by switching ON and OFF parasitics elements. In terms of power consumption and cost of fabrication, it is obvious that parasitic array antennas are suitable for mass implementation of SA, especially for battery powered portable mobile

terminals [28], [29].

- 3.) For radiation patterns the switched beam antennas are extension of the current micro-cellular or cellular sectorization method by splitting a typical cell. The approach further subdivides macro-sectors into several micro-sectors as a means of improving range and capacity [27]. The switched beam antenna system uses a number of fixed beams at an antenna site. Ever so often, the system scans the outputs of each beam and selects the beam with the largest output power. If the signal/user moves from one beam to another, the antenna switches to the new beam. The switched-beam antennas are normally used only for reception of signals since there can be ambiguity in the system's perception in the location of the received signal. This system selects one of several predetermined fixed-beam patterns with the greatest output power in the remote user's channel. These choices are driven by RF or baseband DSP hardware and software. The systems may not offer the degree of performance improvement by adaptive systems. But they are often much less complex and easily compatible with existing wireless technologies [24], [31].

Figure 2.3 shows the basic block diagram of the switched beam SA. At the RF front end, an array of N antennas is combined using a Butler Matrix which sums up the signals at the RF stage and generates an output for M different beams. This method uses a separate receiver or a group of receivers to monitor the signal power at each antenna port for each subscriber and determines the port with the highest received signal strength (RSS) for a specified user. The separate switch unit then generates a path from the antenna output port (switch matrix

input) to the desired radio channel (switch matrix output). As shown, the switch matrix outputs are cabled to the BS receivers with each operating at different frequencies. The control receiver unit monitors all beams and radio frequencies in the cell site (or the sector). For each frequency channel (i.e. receiver), the optimum beam is chosen and a command is sent to the switch matrix to generate the RF path from the selected beam to the receiver [32].

- 4.) The DBF adaptive antenna systems deal with communication between a user and BS in a different way, in effect adding a dimension of space. By adjusting to an RF environment, the adaptive antenna technology can dynamically alter the signal patterns to near infinity in order to optimize the performance of the wireless system. Adaptive arrays utilize sophisticated signal-processing algorithms to continuously distinguish among desired signals, multi-path, and interfering sig-

Figure 2.3: Block diagram of switched beam systems [32].

nals and calculate their directions of arrival. This approach continuously updates its transmit strategy based on changes in both the desired and interfering signal locations. The ability to track users smoothly with main lobes and interferers with nulls ensures that the link budget is constantly maximized because there are neither micro-sectors nor predefined patterns [32]- [33].

Figure 2.4: Block diagram of adaptive array systems [32].

Figure 2.4 shows the block diagram of a digitally adaptive smart antenna. The RF signals from the antennas are coherently down-converted to an IF frequency, low enough for quality digitization of the signals. The beam-former then processes the digital outputs for each channel. The process includes amplitude and phase adjustment, which results in beam and null steering. The adaptive antenna system can be viewed as a spatial filter in which the pass and stop band are created along the direction of the signal and interferers respectively [32].

5.) Diversity antennas merely switch operation from one working element to another.

Spatial diversity exploits multiple antennas either separated in space or polarized differently. Although this approach mitigates severe multi-path fading, its use of one element at a time offers no uplink gain improvement over any other single-element approach. Diversity antennas provide uncorrelated signal outputs installed at both receiver and transmitter side to build the MIMO wireless communication channels [34]- [35]. Transmitting space-time codes (STCs) over a MIMO system dramatically increase the data transmitting rate over wireless communication channels [36].

From the above introductory remarks, we know that SA can be deployed in different ways. In the following sections, a thorough review and survey on the application of the smart antenna system is presented.

2.1.3 Applications

Along with the development of smart antenna technologies, applications of arrays including multi-path mitigation, BF, cancellation of CCIs, diversity, and direction finding, etc, are shown in Figure 2.5. In many research works, a number of possibilities of using SA are discussed [5] [26], [37]- [40].

Let's first look at the indoor systems. Multi-path fading becomes a severe problem in indoor-mobile communication since there are a large number of scattering and reflections inside a building. In dense urban areas there are many buildings and obstacles that give rise to multi-path propagation which occurs due to multiple copies of the signals arriving at the receiver with different amplitudes and time delays. It's

Figure 2.5: Arrays in mobile communication systems.

clear that the antenna has to process a number of copies of the desired signal, each of which is corrupted by undesirable interference. For a conventional antenna, all interfering noise sources and desired signals are mixed up at the receiver with random phase delays, leading to fading and thus an increased error probability. The sensitivity of an adaptive array to interfering noise sources can be reduced by suitable processing of the outputs of the individual array elements. The combination of array and processing acts as a filter in both space and frequency. The results shown in [44] indicate that replacing the omni-directional base-station antenna with an adaptive array which capable of forming multiple dynamically allocated beams steered toward the mobiles in a building complex, can substantially reduces the delay spread [5].

Research on smart antenna arrays for satellite-mobile systems and satellite-to-satellite communications has intensified in the recent years. By using an array of antennas, satellite communications form beams which point in the direction of the desired satellite and have nulls in the direction of others operating in the same band. In these systems, multiple antenna elements may be utilized on a mobile as well as on a satellite. The capability of the system to generate multiple spot beams with independent power control and frequency had been evaluated in the European geostationary satellite system [5] [45]- [46]. The array was used to track the satellites, cancel interferences arising from the transmission of other satellites and provide protection against unfriendly jamming.

In addition to the aforementioned applications in Figure 2.5, new applications will be incorporated into future wireless communication systems. Future generations of

wireless communications require adaptive array techniques capable of operating in a wide range of environment with respect to traffic, interference, propagation, antenna configuration, and radio access technology. Extensive investigation and development have been carried out to realize these new applications, such as routing selection, easy beam pattern control and reduced power consumption in ad-hoc networks, as well as the high data rate transmission over MIMO systems.

2.1.4 Benefits and Drawbacks for Smart Antennas

In wireless communication environments, a smart antenna system can be used to achieve a number of major benefits [30]:

- Higher system capacity: Smart antennas can be used to allow the subscriber and BS to operate at lower power and at the same range as a conventional system. This may allow FDMA and TDMA systems to be re-channelized to reuse frequency channels more often than those systems using conventional fixed antennas due to the fact that the carrier-to-interference ratio is much greater when SA are used.
- Increased coverage: The coverage, or coverage area, is simply the area in which communication between a mobile and the BS is possible. Smart antennas provide enhanced coverage through range extension, hole filling, and better building penetration.

- Better transmission quality: Smart antennas help to mitigate the impact of multi-path or even exploit the diversity inherent in multi-path. The systems can be used to reduce the delay spread, improve user position estimation and obtain spatial diversity gain to effectively improve link quality.
- Cost reduction: Smart antenna system can be adaptively optimized to steer to the desired users, which results in lower power consumption and cost reduction.
- Increased spectral efficiency: Interference cancellations, reuse of time, frequency and codes resources can increase data transmission rate and spectral efficiency will be therefore increased.
- Potential to introduce new services.

Two projects stated below have emerged to reinforce the benefits of smart antenna system. Ericsson-Mannesmann employed DBF at the uplink and down-link. By which the capacity was increased by 100 - 200 percent [47]. NTT company applied SA for a 3G UMTS W-CDMA network and a substantial improvement in average bit error rate (BER) was obtained from test results [20] .

Although there are many benefits of using SA, some important drawbacks are also existed. Below are some point which must be taken into account in designing an adaptive systems [48], [49]:

- The smart antenna transceiver is much more complicated than a traditional BS transceiver.

- The adaptive beam-forming is a computationally intensive process. So the smart antenna BS must include very powerful numeric processors and control systems. The requirements for hardware and software are increased.
- The computational time and complexity of array signal processing algorithms can be very significant.
- The practical antenna arrays maybe adversely affected by calibration errors, noises and interferences from environment, so accurate real-time calibration for each of the array antenna elements is required.
- More complex radio resource management.

From aforementioned applications, SA is very attractive to mobile communications system designers. As a result, many efforts have been made on the design of smart antenna arrays and the associated techniques. Below we start to review these important techniques and algorithms that are related to SA. In the following section, we reviewed the DOA estimation algorithms as well as their performance and limitations since they are very important to SA.

2.2 DOA Estimation Algorithms

Researchers from different specialities tend to view SA from quite different perspectives. For instance, antenna engineers are inclined to ask about antenna patterns, side-lobe levels, null depths, the effects of element pattern and polarization. Research on antenna design has focused on the selection of attractive radiating elements and antenna

architectures that can accommodate the physical and electrical requirements of the system [50], [51]. Communication engineers tend to regard the SA as a time-varying processor. They want to know how signal modulation and bandwidth, signal arrival angles, signal and noise powers, and the number of signals affect output signal-to-noise ratios, bit error probabilities, and the like. Research on the signal processing aspects of smart antenna systems has concentrated on the development of efficient algorithm for DOA estimation and adaptive BF [52]- [55]. Control engineers usually ask about stability and rate of convergence of the adaptive weight control loops [56].

The demand for service provision via wireless communication networks is growing at an explosive rate. Users may anticipate more multimedia services that are ubiquitous and customized to individual needs. The rising expectation requires mammoth channel capacity. The demand for increased capacity has motivated recent researches that exploit space selectivity. As a result, attention has been tuned to advanced adaptive spatial filtering SDMA by using smart antenna techniques. With advanced SDMA, multiple users can be allocated and tracked simultaneously with separated beams, and each beam has the ability to reject the signals from all the interferers. Hence, virtual channels in an angle domain are provided, and the co-channel interference is significantly reduced. The intended user and interferers are located and tracked by using the direction of arrival estimation algorithms. The DOAs information is feed into the adaptive beamforming algorithms to adjust the weights.

As shown in Figure 2.6, direction of arrival estimation algorithms analyze signal samples obtained from the antennas and give direction information of the impinging

waves.

Figure 2.6: DOA estimation based adaptive antennas configuration.

High-resolution DOA estimation at antenna arrays is a critical electronic supported activity in both electronic warfare systems and mobile communication systems. The problem has been an active research topic for more than 30 years, and numerous DOA estimation schemes have been proposed.

DOA estimation algorithms are classified by [59] into three categories: spectral estimation, parametric subspace based estimation (PSBE), and deterministic parametric estimation (DPE). As shown in Figure 2.7, the most frequently used algorithm of the first category is the multiple signal classification algorithm (MUSIC) [57]. ESPRIT [60] and all its variants such as least square, total least square method [58]

and unitary ESPRIT [61] belong to the parametric subspace based estimation techniques. The deterministic parametric estimation techniques include: maximum entropy (ME) [62] maximum likelihood (ML) [82], space-alternating generalized expectation-maximization (SAGE) [83], and weighted subspace fitting (WSF) [84] methods. A brief review on DOA estimation algorithms is presented in the following sections.

Figure 2.7: DOA estimation algorithm classification.

2.2.1 Data Models

Let's consider an array with arbitrarily spaced antenna elements. M narrow band signal sources impinge at the array from distinct directions $\theta_1, \dots, \theta_M$. Assume that

the incident sources are located sufficiently far from the array such that in homogenous isotropic transmission media the wavefronts are planar. A source is considered to be in the far-field if $r > 2L^2/\lambda$, where r is the distance from the source to the array, L is the length of the array and λ is the wavelength of the arriving wave.

For simplicity, ignoring the mutual coupling among the array elements for a moment and let us firstly formulate the data signal model for an ideal array which has no mutual coupling need to be considered.

The radiation impinging on the array is in the form of a sum of plane waves. The signals are assumed to be narrow band processes. The assumption of narrow band which implies that the effect of a time delay on the received waveforms is simply a phase shift (i.e, $s(t - \tau) \approx s(t)e^{-j\omega_c\tau}$). The signals are assumed to have the same known center frequency ω_c . Additive noise is present at all N antenna elements. The k -th signal can be written as,

$$\tilde{s}_k(t) = u_k(t) \cos[\omega_c t + \nu_k(t)], k = 1, \dots, M \quad (2.1)$$

where $u_k(t)$ and $\nu_k(t)$ are varying functions of time that define the amplitude and phase of $\tilde{s}_k(t)$, respectively. A complex envelope representation is usually used, $\tilde{s}_k(t) = \text{Re}\{s_k(t)\}$. In the function, $\text{Re}\{ \}$ denotes the real part of the complex variable, and $s_k(t) = u_k(t)e^{j(\omega_c t + \nu_k(t))}$.

For i -th element at time t , the output can be written as

$$\begin{aligned} y_i(t) &= \sum_{l=1}^M a_i(\theta_l) s_l(t - \tau_i(\theta_l)) \\ &= \sum_{l=1}^M a_i(\theta_l) s_l(t) e^{-j\omega_c \tau_i(\theta_l)} \quad (i = 1, \dots, N) \end{aligned} \quad (2.2)$$

where $\tau_i(\theta_l)$ is the propagation delay between a reference point and the i -th sensor for the l -th wavefront impinging on the array from direction θ_l , $a_i(\theta_l)$ is the corresponding antenna element complex response (gain and phase) at frequency ω_c . Employing vector notation for the outputs of the array antennas, the data model becomes,

$$\mathbf{y}(t) = \sum_{l=1}^M \mathbf{a}(\theta_l) s_l(t) \quad (2.3)$$

The output vector of the array is

$$\mathbf{y}(t) = [y_1(t), y_2(t), \dots, y_N(t)] \quad (2.4)$$

where

$$\mathbf{a}(\theta_l) = [a_1(\theta_l) e^{-j\omega_c \tau_1(\theta_l)}, a_2(\theta_l) e^{-j\omega_c \tau_2(\theta_l)}, \dots, a_N(\theta_l) e^{-j\omega_c \tau_N(\theta_l)}]^T \quad (2.5)$$

$\mathbf{a}(\theta_l)$ is often termed as array steering vector for signals from direction θ_l which is determined by the array geometry and antenna element characteristics. We can also

write the array response and signal sources in the form of vectors ,

$$\mathbf{A}(\theta) = [\mathbf{a}(\theta_1), \mathbf{a}(\theta_2), \dots, \mathbf{a}(\theta_M)] \quad (2.6)$$

$$\mathbf{s}(t) = [s(t_1), s(t_2), \dots, s(t_M)]^T \quad (2.7)$$

Additive noise $\mathbf{n}(t)$ which is assumed to be uncorrelated with $\mathbf{s}(t)$ and modelled as temporally white and zero-mean complex gaussian process, satisfies,

$$E[\mathbf{n}(t)] = 0, \quad E[\mathbf{n}(t)\mathbf{n}^H(t)] = \sigma^2 \mathbf{I} \quad (2.8)$$

where $E\{\bullet\}$ and $\{\bullet\}^H$ denote statistical expectation operator and conjugate transpose, respectively. σ^2 stands for the noise power and $\mathbf{I}$ is a N -dimensional identity matrix.

Therefore, for narrow band signal processing, the output of the receiving antenna array is described as below:

$$\mathbf{y}(t) = \mathbf{A}(\theta)\mathbf{s}(t) + \mathbf{n}(t) \quad (2.9)$$

which can be expanded in matrix form

$$\begin{bmatrix} \mathbf{y}_1(t) \\ \mathbf{y}_2(t) \\ \vdots \\ \mathbf{y}_N(t) \end{bmatrix} = \begin{bmatrix} a_1(\theta_1)e^{-j\omega_c\tau_1(\theta_1)} & \dots & a_1(\theta_M)e^{-j\omega_c\tau_1(\theta_M)} \\ a_2(\theta_1)e^{-j\omega_c\tau_2(\theta_1)} & \dots & a_1(\theta_M)e^{-j\omega_c\tau_1(\theta_M)} \\ \vdots & \ddots & \vdots \\ a_N(\theta_1)e^{-j\omega_c\tau_N(\theta_1)} & \dots & a_N(\theta_M)e^{-j\omega_c\tau_N(\theta_M)} \end{bmatrix} \begin{bmatrix} \mathbf{s}_1(t) \\ \mathbf{s}_2(t) \\ \vdots \\ \mathbf{s}_M(t) \end{bmatrix} + \begin{bmatrix} \mathbf{n}_1(t) \\ \mathbf{n}_2(t) \\ \vdots \\ \mathbf{n}_N(t) \end{bmatrix} \quad (2.10)$$

where $a_n(\theta_m)$ is the n -th array element response for signal sources $s(t)$ from direction θ_m and taking the first element in the array as the phase reference. Thus the covariance matrix of $\mathbf{y}(t)$ is

$$R_{\mathbf{y}} = E[\mathbf{y}\mathbf{y}^H] = \mathbf{A}P_s\mathbf{A}^H + \sigma^2\mathbf{I} \quad (2.11)$$

where P_s is the $M \times M$ sources cross-covariance matrix and $P_s = E[\mathbf{s}(t)\mathbf{s}^H(t)]$. $E[\mathbf{n}(t)\mathbf{n}(t)^H] = \sigma^2\mathbf{I}$ is the noise covariance matrix. The noise covariance structure is a reflection of the noise having a common variance σ^2 at all sensors and being uncorrelated among all sensors. The source covariance matrix, P_s , is often assumed to be nonsingular or near-singular for highly correlated signals. In the later development, the spectral factorization of $R_{\mathbf{y}}$ will be of central importance and its positivity guarantees the following representation,

$$R_{\mathbf{y}} = E[\mathbf{y}\mathbf{y}^H] = \mathbf{A}P_s\mathbf{A}^H + \sigma^2\mathbf{I} = \mathbf{U}\mathbf{\Lambda}\mathbf{U}^H \quad (2.12)$$

with $\mathbf{U}$ is unitary matrix and $\mathbf{\Lambda} = \text{diag}\{\lambda_1, \lambda_2, \dots, \lambda_N\}$. The diagonal matrix of real eigenvalues are ordered such that $\lambda_1 \geq \lambda_2 \geq \dots, \geq \lambda_N > 0$. Observe that any vector

orthogonal to $\mathbf{A}$ is an eigenvector of R_y with the eigenvalue σ^2 . There are $N - M$ such linearly independent vectors.

R_y can be partitioned in terms of its eigenvalues, λ_n , and their corresponding eigenvectors e_n ($n = 1, 2, \dots, N$) into eigenvalue pairs and noise pairs. The first M eigenvalues will correspond to the directional sources and their values are larger than σ^2 . The remaining $(N - M)$ eigenvalues are equal to σ^2 . The eigenvectors corresponding to the signal eigenvalues can be used to describe the signal subspace:

$$E_S = [e_1 \ \dots \ e_M] \quad (2.13)$$

E_N contains the remaining $N - M$ noise eigenvectors that describes the noise subspace, which is the orthogonal complement to the signal subspace:

$$E_N = [e_{M+1} \ \dots \ e_N] \quad (2.14)$$

2.2.2 MUSIC Algorithm

Most of the spectral estimation methods estimate the DOAs by computing the spatial spectrum and then determining the local maximums [27]. These techniques have their roots in time-series analysis. A brief overview and comparison of some of these methods could be found in [103]. MUSIC is a relatively simple and efficient spectral DOA estimation method. It has many variations and is perhaps the most studied method in its class. In its standard form, also known as spectral MUSIC, the method

estimates the noise subspace from the available samples. This can be done by either eigenvalue decomposition of the estimated array correlation matrix or singular value decomposition of the data matrix, with its columns being the snapshots or the array signal vectors.

Schmidt [57] was the first one to correctly exploit the measurement model in the case of sensor arrays of arbitrary form. Schmidt, in particular, accomplished this by first deriving a complete geometric solution in the case of absence of noise, then cleverly extending the geometric concepts to obtain a reasonable approximate solution in the presence of noise. The resulting algorithm was called MUSIC and has been widely studied. Although the performance advantages of MUSIC are substantial, they are achieved at a considerable cost in computation(searching over parameter space) and storage (of array calibration data). A crucial problem in the MUSIC method is in estimating the eigensystem of the covariance matrix of observed signals from the sensor array. In this method, a sample signal covariance matrix is formed by the snapshots of the signals from antenna array elements. Then eigen-decomposition is performed to obtain the noise subspace and signal subspace. Therefore the MUSIC spectrum is formed by using the steering vector which is dependent on the antenna array geometry and the corresponding signal direction orthogonal to the noise subspace. Signal direction information can thus be obtained from the MUSIC spectrum.

Using the standard data model (2.9), the ideal covariance matrix of the data received by the array, which is estimated by using an infinite number of data snapshots, is given

by (2.12), where P_s is the $M \times M$ sources cross-covariance matrix,

$$P_s = E[\mathbf{s}(t)\mathbf{s}^H(t)] \quad (2.15)$$

Assuming that the M signal sources are uncorrelated and P_s is a diagonal matrix. Then, by applying eigen-decomposition to P_s , the eigenvalues under ideal conditions will satisfy

$$\lambda_1 \geq \lambda_2 \geq \cdots, \geq \lambda_M > \sigma^2 = \lambda_{M+1} = \lambda_{M+2} = \cdots, = \lambda_N > 0 \quad (2.16)$$

We can define λ_i , $i = 1, \dots, M$ as the principal eigenvalues which are associated by eigenvector e_i , ($i = 1, \dots, M$). The e_i , ($i = M+1, \dots, N$) are noise eigenvectors. The representation of P_s in terms of its eigenvalues and its corresponding eigenvectors is as follows:

$$P_s = \sum_{i=1}^M (\lambda_i - \sigma^2) e_i e_i^H + \sigma^2 I \quad (2.17)$$

Since the signals sources are uncorrelated, P_s can be expressed as

$$P_s = \sum_{i=1}^M |q_i|^2 \mathbf{a}(\theta_i) \mathbf{a}(\theta_i)^H + \sigma^2 I \quad (2.18)$$

where q_i^2 is the average power of the i -th source. Therefore,

$$\mathbf{A} P_s \mathbf{A} = \sum_{i=1}^M |q_i|^2 \mathbf{a}(\theta_i) \mathbf{a}(\theta_i)^H = \sum_{i=1}^M (\lambda_i - \sigma^2) e_i e_i^H \quad (2.19)$$

The noise subspace can be formed and given explicitly by

$$P_{noise} = \sum_{j=M+1}^N e_j e_j^H \quad (2.20)$$

The signal subspace is spanned by E_S . The key idea of MUSIC is: Since signal subspace is self-adjoint, the noise subspace is orthogonal to the signal subspace. Therefore, a vector f is in the E_S , if and only if its projection onto the noise subspace is zero, i.e.,

$$\|P_{noise}f\| = 0 \quad (2.21)$$

and equivalently,

$$\frac{1}{\|P_{noise}f\|} = \infty \quad (2.22)$$

We note that for an operator that is not self-adjoint, MUSIC can be used with the singular value decomposition instead of the eigenvalue decomposition. The procedure of using MUSIC algorithm in estimation of DOA is described as follow:

1. Get the signal samples vector from the array elements.
2. Form the signal correlation matrix R_y as in (2.11).
3. Eigen-decompose R_y .
4. Get the noise subspace E_n corresponding to the small value eigenvalues.
5. Search for looking directions orthogonal to the noise subspace.

From the orthogonality of the signal and noise subspaces, we can find the peaks in the estimator function as below,

$$P_{Music} = \frac{1}{[\mathbf{a}(\theta)]^H E_{\mathbf{n}} E_{\mathbf{n}}^H [\mathbf{a}(\theta)]} \quad (2.23)$$

(2.23) is the standard form. MUSIC has many variations, e.g. root-MUSIC, beam-space MUSIC [59], and etc. The root-MUSIC algorithm [103] is applicable to the uniformly spaced linear array, in which the search for DOA can be made by finding the roots of a polynomial. Root-MUSIC has better performance than spectral MUSIC. The beam-space MUSIC has a beam-forming processor. The array data are passed through this processor before applying the MUSIC estimation algorithm. The beam-space MUSIC estimation has a number of advantages such as a reduced computation, improved solution compared to spectral MUSIC method.

2.2.3 ESPRIT Algorithm

ESPRIT, which dramatically reduces computation and storage costs is very attractive in many practical signal-processing problems. In the context of DOA estimation, the reduction in computation and storage cost is achieved by requiring that the sensor array possess displacement invariance, i.e., sensors occur in matched pairs with identical displacement vectors. Fortunately, there are many practical problems in which these conditions are or can be satisfied. ESPRIT is also manifestly more robust (i.e. less sensitive) with respect to array imperfections than previous techniques including MUSIC.

ESPRIT is a computationally efficient and robust method for estimating DOA, which was developed in order to overcome the disadvantages of MUSIC. Although the performance advantages of MUSIC are substantial, they are achieved at a considerable cost in computation and storage. To simplify the description of the basic ideas behind ESPRIT, discussions will deal only with single dimensional parameter spaces, e.g., azimuth-only direction finding of far field point sources, since the basic concepts are most easily visualized in such spaces. Narrow-band signals of known center frequency will be assumed. A planar array with arbitrary geometry composed of N antenna

Figure 2.8: General doublet antenna structure.

element doublets is considered. The elements in each doublet have identical sensitivity patterns and are translational separated by a known constant displacement vector Δ , which sets the reference direction, and all angles are measured with reference to this vector. Assume that there are M , $M \leq N$ narrow band sources centered at frequency ω_c , and that the sources are located sufficiently far from the array such

that the waterfronts impinging on the array are planar. As before, the sources may be assumed to be stationary zero-mean random processes or deterministic signals. Additive noise is present at all $2N$ sensors and is assumed to be a stationary zero-mean random process with a variance σ^2 . As shown in Figure 2.8, it is convenient to describe the array as being comprised of two sub arrays Z_X and Z_Y , where Z_X and Z_Y are composed of elements of $\{x_1, x_2, x_3, \dots, x_N\}$ and $\{y_1, y_2, y_3, \dots, y_N\}$, respectively. The signals received at the i -th doublet can then be expressed as

$$\begin{aligned} x_i(t) &= \sum_{k=1}^M s_k(t) a_i(\theta_k) + n_{xi}(t) \\ y_i(t) &= \sum_{k=1}^M s_k(t) e^{j\omega_c \Delta \sin \theta_k / c} a_i(\theta_k) + n_{yi}(t) \end{aligned} \quad (2.24)$$

Where θ_k is the direction-of-arrival of the k -th source relative to the direction of the translation displacement vector Δ and c is the speed of light. Since ESPRIT does not require any knowledge of the sensitivities, the sub array displacement vector Δ sets not only the scale for the problem, but the reference direction as well. The obtained DOA estimates are angles-of-arrival with respect to the direction of the vector Δ .

Combining the outputs of each of the sensors in the two sub-arrays, the received data vectors can be written as follows:

$$\begin{aligned} \mathbf{x}(t) &= A\mathbf{s}(t) + \mathbf{n}_x(t) \\ \mathbf{y}(t) &= A\Phi\mathbf{s}(t) + \mathbf{n}_y(t) \end{aligned} \quad (2.25)$$

Where $\mathbf{x}(t)$ and $\mathbf{y}(t)$ denote the output of each sub array. The $\mathbf{n}_x(t)$ and $\mathbf{n}_y(t)$ denote the noise induced on the two sub arrays, respectively. Vector $\mathbf{s}(t)$ is the $M \times 1$ vector of

impinging signals as observed at the reference sensor of sub array Z_X . Φ is a diagonal $M \times M$ matrix of the phase delay between the doublets sensors for the M wavefronts:

$$\Phi = \text{diag}\{ e^{j\omega_c \Delta \sin \theta_1 / c} \quad \dots \quad e^{j\omega_c \Delta \sin \theta_M / c} \} \quad (2.26)$$

The diagonal matrix Φ mutually relates the outputs of the two sub arrays and it is used to estimate the DOA. This is simply carried out by separating signal space E_S into two matrixes spanning the same M dimensional signal space. This over-determined set of equations is then used to find the estimation of the diagonal matrix, Φ , which leads to the DOA estimates.

Defining the total array output vector as $\mathbf{Z}(t)$, the sub-array outputs can be obtained by yielding,

$$\mathbf{Z}(t) = \begin{bmatrix} \mathbf{x}(t) \\ \mathbf{y}(t) \end{bmatrix} = \bar{\mathbf{A}}\mathbf{s}(t) + \mathbf{n}_z(t) \quad (2.27)$$

$$\bar{\mathbf{A}} = \begin{bmatrix} \mathbf{A} \\ \mathbf{A}\Phi \end{bmatrix}, \quad \mathbf{n}_z(t) = \begin{bmatrix} \mathbf{n}_x(t) \\ \mathbf{n}_y(t) \end{bmatrix} \quad (2.28)$$

It is the structure of $\bar{\mathbf{A}}$ that is exploited to obtain estimates of the diagonal elements of Φ without having to know $\mathbf{A}$. The basic idea behind ESPRIT is the relevant signal subspace that contains the outputs from the two sub-arrays described above, Z_X and Z_Y . Simultaneous sampling of the output of the arrays leads to two sets of vectors, E_x

and E_y , that span the same signal subspace.

Analogously to the other subspace-based algorithms, ESPRIT relies on properties of the eigen-decomposition of the array covariance matrix.

The signal subspace can also be obtained from the covariance matrix of $\mathbf{Z}(t)$,

$$R_Z = E[\mathbf{Z}(t)\mathbf{Z}^H(t)] = \bar{A}P_s\bar{A}^H + \sigma^2I \quad (2.29)$$

When applying eigen-decomposition to (2.29), there must exist a unique nonsingular matrix T . The signal subspace E_S can be expressed as blow:

$$E_S = \begin{bmatrix} E_X \\ E_Y \end{bmatrix} = \begin{bmatrix} AT \\ A\Phi T \end{bmatrix} \quad (2.30)$$

It yields:

$$\begin{aligned} E_X &= T^{-1}\Phi T \\ E_Y &= A\Phi T \end{aligned} \quad (2.31)$$

which, by defining $\Psi = T^{-1}\Phi T$, becomes

$$E_Y = E_X\Psi \quad (2.32)$$

Note that Ψ and Φ are related by a similarity transformation, and hence have the same eigenvalues. Therefore, the eigenvalue of Ψ must be equal to the diagonal elements of Φ , and DOA estimates can be obtained according to (2.26). By solving the approximate

relation (2.32) in either a least-square sense or a total-least-square sense. This may lead to many versions of ESPRIT, including TLS-ESPRIT, multiple invariance ESPRIT, etc.

2.2.4 ML Algorithm

The maximum likelihood (ML) DOA estimation is a more complex technique comparing with MUSIC, minimum variance and minimum norm. The ML algorithm gives a superior performance compared to other methods, particularly when the SNR is low, the number of snapshots is small, or the sources are highly correlated due to multipath propagation. However, the computational cost associated with optimization of the complex, multidimensional and highly nonlinear likelihood function is expensive, which prevents the method from practical use, unless in the scenarios where a slow post-processing procedure is allowed. The method is to apply the ML principle to the statistics of the observed raw data. The ML method gives an almost optimal performance compared to other methods, particularly when the SNR is small, the number of samples are small, or the sources are correlated, and thus is of practical interest [94]. The estimates of ML method are nearly equal to their true values. That is, it may be used as a standard to compare the performance of other methods. Normally it needs to make an assumption first. One assumes the number of signal sources are known. This method is computationally intensive, it requires iterative schemes for solutions. However, several researchers have proposed various schemes to optimize the log-likelihood function to increase the probability of global convergence and com-

putation efficiency [95]- [98]. Unfortunately, most of them can not guarantee global convergence in general cases [98].

Consider an array composed of N sensors located on an UCA with array radius of r . Assume that M narrow band sources, centered around a known wavelength, say λ_0 , impinge on the array from M distinct directions $\theta_1, \dots, \theta_M$ asynchronously, with the reference point set at the center of the array. For the k -th signal source, ϕ_k is azimuth angle which is measured with respect to the line connecting the reference point with the first array element. θ_k is the elevation angle measured with respect to the Z axis. The N -dimensional array steering vector $\mathbf{s}(\phi_k, \theta_k)$, ($k = 1, \dots, M$), is an array response to a narrow-band signal of wavelength λ_0 arriving from angle (ϕ_k, θ_k) , with $\theta_k \in [-\pi, \pi]$, and $\phi_k \in [0, 2\pi]$. Then, $\mathbf{s}(\phi_k, \theta_k)$ is given by

$$\mathbf{s}(\phi_k, \theta_k) = G_n(\phi_k, \theta_k) \exp[j \frac{2\pi r}{\lambda_0} \sin(\theta_k) \cos(\phi_k - \frac{2\pi(n-1)}{N})] \quad (2.33)$$

where $G_n(\phi_k, \theta_k)$ is the complex gain pattern of the k -th element. Suppose the array elements are all identical and isotropic, then $G_n(\theta_k, \varphi_k) = 1$, ($k = 1, \dots, N$). The array manifold vector $\mathbf{s}(\phi_k, \theta_k)$ for a UCA can be written as

$$\mathbf{s}(\phi_k, \theta_k) = \begin{bmatrix} e^{-j2\pi \frac{r}{\lambda_0} \sin(\theta_k) \cos(\phi_k)} \\ e^{-j2\pi \frac{r}{\lambda_0} \sin(\theta_k) \cos(\phi_k - \frac{2\pi}{N})} \\ \vdots \\ e^{-j2\pi \frac{r}{\lambda_0} \sin(\theta_k) \cos(\phi_k - \frac{2\pi(N-1)}{N})} \end{bmatrix} \quad (2.34)$$

In matrix notation, the $N \times 1$ voltage vector $\mathbf{y}$ received by the array can be expressed

by:

$$\mathbf{y} = \mathbf{S}\mathbf{x} + \mathbf{n} \quad (2.35)$$

where $\mathbf{n}$ denotes the $N \times 1$ additive noise vector of the array, $\mathbf{x} = [x_1, x_2, \dots, x_M]$ is a complex valued signal amplitude vector and S is the steering matrix of the array. $\mathbf{S} = [\mathbf{s}(\phi_1, \theta_1), \mathbf{s}(\phi_2, \theta_2), \dots, \mathbf{s}(\phi_M, \theta_M)]$. Let R_Y is the obtained covariance matrix of measurable observations,

$$\begin{aligned} R_{\mathbf{y}} &= E[\mathbf{y}\mathbf{y}^H] \\ &= E[(\mathbf{S}\mathbf{x} + \mathbf{n})(\mathbf{S}\mathbf{x} + \mathbf{n})^H] \\ &= E[\mathbf{S}\mathbf{x}\mathbf{x}^H\mathbf{S}^H] + E[\mathbf{n}\mathbf{n}^H] \\ &= \mathbf{S}\mathbf{A}\mathbf{S}^H + \sigma^2\mathbf{I} \end{aligned} \quad (2.36)$$

$\mathbf{A} = E[\mathbf{x}\mathbf{x}^H]$ is a $M \times M$ matrix that represents the covariance matrix of the source signals, where A is a positive definite. σ^2 is the noise variance at each sensor. The ML problem can be considered as solving the following maximization function [101] [102]:

$$F_{ML}(\phi, \theta) = \text{tr}[\mathbf{I} - \mathbf{S}(\mathbf{S}^H\mathbf{S})^{-1}\mathbf{S}^H]\hat{R}_Y \quad (2.37)$$

where,

$$\hat{R}_{\mathbf{y}} = \frac{1}{N} \sum_{t=1}^N \mathbf{y}(t)\mathbf{y}(t)^H \quad (2.38)$$

For one snapshot of observations, $\hat{R}_{\mathbf{y}} = R_{\mathbf{y}}$. The problem in maximizing (2.37) is

severe. It is a multidimensional maximization problem, and searching function may not converge, resulting in an intense computation load.

In this section a set of general methods are presented for DOA estimation using a set of phaser voltages measured at the feed points of a uniform spaced antenna elements. In next chapter, we will review the techniques of compensation for the mutual coupling effect in the antenna arrays.

2.3 Concluding Remarks

In this chapter, introductory remarks about the evolutionary path and the basic problems of mobile communications are presented. Some of the these problems of mobile communications and the possibility of using the SA to mitigate these problems are described as well. An overview of the SA technologies for mobile communication systems is introduced. The review on the algorithms of DOA estimation is presented. In next chapter, the detailed description on our new method and other conventional methods of compensation for mutual coupling effect in array antennas will be presented and discussed.

Chapter 3

The Compensation Techniques for the Mutual Coupling Effect in Antennas

3.1 Introduction

In the previous chapter, we have surveyed smart antenna systems in regard with their architectures, performance, and related algorithms on direction finding application. In this chapter, we will focus our interest in the electromagnetic characteristics of smart antenna array. The radiation pattern of an array is determined by the type of individual elements used, their orientations, their positions in space, and the amplitude and phase of the currents feeding them [8]. When antenna are in close proximity they interact in a complicated manner. This interaction is called mutual coupling. The resulting effect

is to change the current on each antenna element. The phase change of the current is most noticeable although the amplitude may be different as well. The current of a given element depends not only on the voltage source at its own terminal, but also on the current in all other nearby antennas. Thus, the radiation characteristics of an individual antenna element is altered considerably in the presence of other identical elements in an array due to mutual coupling effect. It has been shown in [65] and [66], that the performance of an adaptive antenna array is strongly affected by the electromagnetic characteristics of the antenna array. The effect of mutual coupling is a serious problem in the application of adaptive antenna arrays.

Mutual coupling effect has been studied by many researchers [63]- [77]. It has been very difficult in the past to calculate mutual coupling effects and thus analysis and design of arrays were initially based on the assumption of constant current generators at the terminals of the array elements. The current numerical procedures made it possible to take mutual coupling effect into account more or less routinely. In this section, we will make a discussion of mutual coupling effects and give an overview of past and current techniques on compensation for mutual coupling effects.

Many methods have been suggested to identify or to compensate for the mutual coupling effect in antenna arrays. In the research works [63] [64], mutual coupling between the antenna elements were ignored and the antenna elements were assumed to be isolated from each other. In [65] the method of moments (MoM) was used to quantify the effect of mutual coupling on DOA estimation algorithms. The author solved the entire MOM problem which requires knowledge of the entire incident fields.

In practice, this information is not available. It shall be shown in [66], the effects of mutual coupling have been reduced but not eliminated. In this work, the author has shown a way to correct the actual voltage matrix using the terminal impedance matrix which is derived from the MOM impedance matrix. In [67], the authors analyzed the performance of an adaptive linear array and compensated for the effects of mutual coupling. The analysis was restricted to a linear array of thin half-wavelength dipoles. They developed an expression of the linear array by considering the N element array as $N + 1$ terminal linear, bilateral network responding to an outside source. The authors obtained a relationship between the open circuit voltages (voltages at the ports of the array if all were open circuited) and the voltages measured at the ports. The expression was derived to study the effect of mutual coupling on the performance of adaptive array. The authors used the concept of mutual impedance to derive the open-circuit voltages from the terminal voltages and showed that mutual coupling had a significant effect on the performance of adaptive arrays. Using the defined mutual impedance of two such dipoles to demonstrate that even for large inter-element spacing, mutual coupling degrades the ability of statistical algorithms to suppress interference. In that research work, the stated assumption is that open-circuit voltages are free of mutual coupling. This assumption is only valid in a limited sense. Moreover, the authors failed to account for the effect of the compensation process on the noise.

This method was later used in [68] and [69] to compensate for the mutual coupling effect of a dipole array for direction finding. The author studied the performance of the MUSIC and ESPRIT algorithm respectively to determine the DOA using dipole arrays and including the effects of mutual coupling. It is obvious that open-circuit voltages

so derived fail to account for the scattering effect of the antenna elements even when they are in the open circuit state. Subsequent studies in [66] and [70] revealed that this open-circuit voltage method failed to take into account the scattering effect due to the presence of other antenna elements in the array. In [66], the authors tried to remedy this problem by assuming that the current distributions on the antenna elements can be exactly known. But in real situations these current distributions are, of course, not available. The similar problem of an unrealistic assumption can also be found with the method in [70] which requires the exact knowledge of the elevation angles of the incoming signals. This requirement is also not easy to meet in real situations. Recently, it has been shown by our work [73]- [77] that improvement can be obtained by taking this open-circuit scattering effect into account by redefining mutual impedance. In our papers, a new method is proposed to tackle this problem in a different way. This method considers the coupling effect (due to the scattered fields) from other antenna elements on a particular antenna element as excitation sources and seeks to quantify their strengths. The calculation of the mutual impedance is based on an estimated current distribution which is caused by a plane wave impinging on a dipole antenna. In this way, the mutual impedances will be provided with a reference to the incoming signal. Unlike the open-circuit-voltage method which seeks to find the open-circuit voltages on the antenna terminals, this proposed method recovers the signal voltages on the antenna terminal loads which are free from the mutual coupling effect. This method does not involve any unrealistic assumption but relies on the calculation of new mutual impedance. The new proposed method does not require the known current distributions on the antenna elements as in [66] or a pre-determined incoming elevation

angle of the signal as in [70] but it uses only a single estimated current distribution for all the antenna elements. Computer simulations show that the proposed method has a significantly better performance and can produce more accurate detection results over the conventional open-circuit-voltage method and MOM method.

In antenna analysis and design, there are various ways to classify the analysis approaches in computation of electromagnetic field (CEM). Generally, we divide CEM into two categories: numerical methods and high-frequency methods. For the most part, numerical techniques are used in the region where the size of the antenna is on the order of the wavelength to a few tens of wavelengths. On the other hand, high-frequency methods, which are best suited to objects that are many wavelengths in extent. Two numerical methods in CEM stand out: MoM and finite difference time domain (FDTD). MOM is a technique that is integral equation based and in the frequency domain. FDTD is a technique that is differential-equation-based and in the time domain [10].

In this thesis, the principle of conventional compensation methods for the mutual coupling effect including method of moments (MOM) and Open-circuit-voltage method are introduced and studied in the following sections. These compensation techniques have their own advantages and disadvantages. They are applicable to different kinds of array imperfections and application scenarios. In addition to the introduction on general compensation techniques for the mutual coupling effect, this chapter gives out a detailed description on a newly proposed compensation method as well.

3.2 Method of Moments

To begin a discussion of mutual coupling effects, consider a wire antenna along the Z-axis. The wire length L is divided into N_e subintervals of length $\Delta = L/N_e$. The MOM is a solution procedure for approximating an integral equation of the form $\int I(z')K(z, z')dz' = -E^i(z)$ to a system of simultaneous linear algebraic equations in terms of the unknown current $I(z')$. The kernel $K(z, z')$ depends on specific integral equation formulation used. As we know, once the current is known, it is fairly straightforward to determine the radiation pattern and impedance of the antenna. It is relatively straightforward to obtain the near- and far-zone fields created by the current.

In 1897, Pocklington [107] derived a commonly used integral equation and showed that the current distribution on thin wires is approximately sinusoidal and propagates with nearly the speed of light.

Consider a wire antenna with conductivity σ surrounded by free space (μ_0, ϵ_0) . Assume the conductivity of the wire is high such that the current is largely confined to the surface of the wire. The equivalent model for the wire where current on the material wire is replaced by an equivalent surface current in free space. This step is necessary so that the vector potential, which employs the free space Green's function, can be used.

When the wire radius is much less than the wavelength, we may assume only Z-directed currents are present. From the Lorentz gauge condition, one have the equation

as below:

$$\frac{\partial A_z}{\partial z} = -j\omega\epsilon_0\Phi \quad (3.1)$$

where, Φ is the scalar potential and A_z is the z -component of the magnetic vector potential.

The vector electric field arising from potentials is

$$\vec{E} = -j\omega\vec{A} - \nabla\Phi \quad (3.2)$$

which can be reduced to the scalar equation,

$$E_z = -j\omega A_z - \frac{\partial\Phi}{\partial z} \quad (3.3)$$

in this situation.

If we consider a Z -directed volume current element Jdv' , taking the derivative of (3.1) and substituting it into (3.3) gives:

$$dE_z = \frac{1}{j\omega\epsilon_0} \left(\frac{\partial^2\psi(z, z')}{\partial z^2} + \omega^2\psi(z, z') \right) Jdv' \quad (3.4)$$

where $\psi(z, z')$ is the free space Green's function given as below:

$$\psi(z, z') = \frac{\exp -j\omega^2 R}{4\pi R} \quad (3.5)$$

and R is the distance between the observation point (x, y, z) and the source point (x', y', z') or:

$$R = \sqrt{(x - x')^2 + (y - y')^2 + (z - z')^2} \quad (3.6)$$

The total contribution to the electric field is the integral over the wire volume:

$$E_Z = \frac{1}{j\omega\epsilon_0} \iiint \left[\frac{\partial^2 \psi(z, z')}{\partial z^2} + \omega^2 \psi(z, z') \right] J dv' \quad (3.7)$$

For radius $a \ll \lambda$, the current distribution is nearly uniform with respect to ϕ' , Thus:

$$E_Z = \frac{1}{j\omega\epsilon_0} \int_{-L/2}^{L/2} \left[\frac{\partial^2 \psi(z, z')}{\partial z^2} + \omega^2 \psi(z, z') \right] I(z') dz' \quad (3.8)$$

In accordance with the surface equivalence principle, we can denote the quantity E_z as the scattered field E_z^s , that is E_z^s is the field radiated in free space by the equivalent current $I(z')$. The other field present is the incident or impressed field E_z^i . At the surface of the perfectly conducting wire and also interior to the wire, the sum of tangential components of the scattered field and the incident field must be zero. Hence, $-E_z^s = +E_z^i$, and using (3.8), we write

$$\frac{-1}{j\omega\epsilon_0} \int_{-L/2}^{L/2} \left[\frac{\partial^2 \psi(z, z')}{\partial z^2} + \omega^2 \psi(z, z') \right] I(z') dz' = E_z^i(z) \quad (3.9)$$

We solve this equation using the MOM. This solution is based on a Galerkin formulation with both the expansion and testing functions being piecewise sinusoids functions, as

describe in [106]. This reduces the integral Equation (3.9) to the matrix equation,

$$[V] = [Z][I] \Rightarrow [I] = [Y][V] \quad (3.10)$$

where $[I]$ is the current vector with coefficients of the expansion of the current in the sinusoidal basis. $[V]$ is voltage vector representing the inner product of the weighting functions and the incident field. $[Z]$ and $[Y]$ are the impedance and admittance matrix respectively.

The elements of the impedance and voltage matrix are given [104] [105], respectively by:

$$z_{mn} = \int (\vec{J}_m \cdot \vec{E}_n^s) dv' \quad (3.11)$$

$$v_m = - \int (\vec{J}_m \cdot \vec{E}_i) dv' \quad (3.12)$$

where $\vec{J}_m$ is the m th test function, $\vec{E}_n^s$ is the field radiated by n th mode of unite amplitude, and $\vec{E}_i$ is the incident field. The size of the impedance matrix is determined by the number of segments into which the antenna is divided. The mutual impedance z_{mn} gives the mutual coupling between the various segments of the antenna. From (3.12), we know the voltage element, v_m , is dependent on the excitation, $\vec{E}_i$.

Once (3.10) is solved to obtain the current matrix $[I]$, the terminal voltage in the

n th antenna is given by:

$$v_n^T = I_n^T \cdot Z_n \quad (3.13)$$

where I_n^T and Z_n are the modal current and load impedance at the n th antenna terminal respectively. The voltage matrix:

$$V_{actual} = [v_1^T, v_2^T, \dots, v_M^T] \quad (3.14)$$

is the actual voltage across the antenna terminal with all the mutual coupling effects included. The current through the load impedance at the antenna terminals is not proportional to just the incident field at the antenna terminals. It also depends upon the coupling from other elements in the array. Thus, we derive a procedure to obtain the corrected voltage which were no coupling between elements of the array. These corrected voltage at the terminals of the antennas are proportional to the incident field at the location of the antennas.

In practice, only the terminal currents are measured. Let us assume an array with M antenna elements. Each antenna has N_e segments. Therefore, the size of the matrix is $N \times N$, where $N = M \times N_e$. Here, we want to reduce this matrix to $M \times M$ which represents the reaction between the antenna terminals. The mutual impedance between the terminal k and l is defined as:

$$Z_{kl}^T = \frac{V_{kl}^T}{I_l^T} \quad (3.15)$$

where V_{kl}^T is the open circuit voltage at the k th antenna terminal due to the current on the l th antenna. The open circuit voltage can be expressed as:

$$V_{kl}^T = -\frac{1}{V_k^T} \int_{kth} \vec{E}_{kl} \cdot \vec{I}^k dl \quad (3.16)$$

where $\vec{E}_{kl}$ is the incident field at the k th antenna due to the l th antenna.

The current on the antenna was expanded as a finite series of piecewise sinusoidal functions, the current on the l th and k antenna are given by [72]:

$$I^l = \sum_{p=1+(l-1)N_e}^{lN_e} I_P^l F_P(z) \quad (3.17)$$

$$I^k = \sum_{m=1+(k-1)N_e}^{kN_e} I_m^k F_m(z) \quad (3.18)$$

where I_P^l and I_m^k are the mode currents on the l th and k th antenna elements, respectively. $F_P(z)$ and $F_m(z)$ are the piecewise sinusoidal expansion functions [72], shown below as:

$$F_m(z) = \begin{cases} \frac{\sin[k(z-z_{m-1})]}{\sin[k\Delta]} & z_{m-1} \leq z \leq z_m \\ \frac{\sin[k(z_{m+1}-z)]}{\sin[k\Delta]} & z_m \leq z \leq z_{m+1} \\ 0 & elsewhere \end{cases} \quad (3.19)$$

The electric field at the k th antenna element due to l th antenna element may be written

as

$$\vec{E}_{kl} = \sum_{m=1+(k-1)N_e}^{kN_e} \vec{E}_m^l G(z - z_m) \quad (3.20)$$

where $\vec{E}_m^l$ is the field at the m th mode due to the current I^l on the l th antenna element and $G(z - z_m)$ is unity over the m th mode and zero elsewhere. $\vec{E}_m^l$ is the sum of the total field from all the modes on the l th antenna.

$$\vec{E}_m = \sum_{p=1+(l-1)N_e}^{lN_e} \vec{E}_{mp} \quad (3.21)$$

therefore, substituting (3.21) into (3.20),

$$\vec{E}_{kl} = \sum_{m=1+(k-1)N_e}^{kN_e} \sum_{p=1+(l-1)N_e}^{lN_e} \vec{E}_{mp} G(z - z_m) \quad (3.22)$$

substituting (3.22) in (3.16), we have,

$$\begin{aligned} V_{kl}^T &= -\frac{1}{I_k^T} \int_{kth} [\sum_{m=1+(k-1)N_e}^{kN_e} \sum_{p=1+(l-1)N_e}^{lN_e} \vec{E}_{mp} G(z - z_m)] \cdot \vec{I}^k dl \\ &= -\frac{1}{I_k^T} \sum_{m=1+(k-1)N_e}^{kN_e} \sum_{p=1+(l-1)N_e}^{lN_e} \int_{kth} \vec{E}_{mp} \cdot \vec{I}^m dl \end{aligned} \quad (3.23)$$

The mutual impedance between modes m and p is given by,

$$Z_{mp} = \frac{V_{mp}}{I_p} = -\frac{1}{I_p I_m} \int_{mth} \vec{E}_{mp} \cdot \vec{I}^m dl \quad (3.24)$$

Therefore,

$$Z_{kl}^T = \frac{V_{kl}^T}{I_{kl}^T} = \frac{1}{I_k^T I_l^T} \sum_{m=1+(k-1)N_e}^{kN_e} \sum_{p=1+(l-1)N_e}^{lN_e} Z_{mp} \vec{I}_m \vec{I}_p \quad (3.25)$$

Once the terminal impedance matrix Z^T is determined, the corrected voltage matrix V_c is readily obtained

$$V_c = Z^T I^T \quad (3.26)$$

where I^T the actual terminal current are obtained from (3.13).

3.3 Open-Circuit-Voltage Method

In 1983, the authors of [67] studied the effect of mutual coupling between array elements on the performance of adaptive arrays. The study includes both steady state and transient performance. It has been shown in their previous work that the performance of an adaptive antenna array is strongly affected by the electromagnetic characteristics of the antenna array, and an important electromagnetic characteristic of an antenna array is the mutual coupling between its elements. The effect of mutual coupling affects the gain, beam-width, etc., of the array. An analytic expression for the steady state output SINR of adaptive arrays is derived when taking the mutual coupling between the array elements into account. The expression is used to study the effect of mutual coupling on the performance of adaptive arrays. It has been shown that the output SINR of the array depends upon the mutual coupling between the its elements.

The output SINR of an adaptive array is the most commonly accepted measure of its steady performance. We will, therefore, first develop an expression for the element output voltages when the mutual coupling is taken into account. The signals received by antenna elements is multiplied by a complex weight, and then these signals are summed to produce the array output.

The required expression can be obtained by considering the N element array as an $N+1$ terminal linear, bilateral network responding to an outside source as shown in Fig 3.1. The array has a driving source as a generator with open circuit voltage V_g and internal impedance Z_g . The each port of the N element array is shown terminated in a known load impedance Z_L . Using standard notation, the expression for array element output voltage was derived in [67] based on $N + 1$ terminal linear bilateral network model,

$$\begin{aligned}
 v_1 &= i_1 Z_{11} + \cdots + i_j Z_{1j} + \cdots i_N Z_{1N} + i_s Z_{1s} \\
 &\vdots \\
 v_j &= i_1 Z_{j1} + \cdots + i_j Z_{jj} + \cdots i_N Z_{jN} + i_s Z_{js} \\
 &\vdots \\
 v_N &= i_1 Z_{N1} + \cdots + i_j Z_{Nj} + \cdots i_N Z_{NN} + i_s Z_{Ns}
 \end{aligned} \tag{3.27}$$

where Z_{ij} is mutual impedance between the ports (array elements) i and j . Furthermore, making use the relationship between terminal current and load impedance, we can obtain an equation as following:

$$i_j = -\frac{v_j}{Z_L}, j = 1, 2, \cdots, N \tag{3.28}$$

Now, we assume that all the elements in the array are in an open circuit condition, that means:

$$i_j = 0, j = 1, 2, \dots, N \quad (3.29)$$

From (3.27), we obtain,

$$v_j = v_{0j} = Z_{js} i_s \quad (3.30)$$

Figure 3.1: Antenna array as an N+1 terminal network.

then substituting v_j and (3.28) into (3.27), we have:

$$\begin{bmatrix} 1 + \frac{Z_{11}}{Z_L} & \frac{Z_{12}}{Z_L} & \cdots & \frac{Z_{1N}}{Z_L} \\ \frac{Z_{21}}{Z_L} & 1 + \frac{Z_{22}}{Z_L} & \ddots & \frac{Z_{2N}}{Z_L} \\ \vdots & \ddots & \ddots & a_2 \\ \frac{Z_{N1}}{Z_L} & \cdots & \frac{Z_{NN-1}}{Z_L} & 1 + \frac{Z_{NN}}{Z_L} \end{bmatrix} \begin{bmatrix} v_1 \\ v_2 \\ \cdots \\ v_N \end{bmatrix} = \begin{bmatrix} v_{01} \\ v_{02} \\ \cdots \\ v_{0N} \end{bmatrix} \quad (3.31)$$

more compactly,

$$Z_0 V = V_0 \quad (3.32)$$

In (3.32), Z_0 is the normalized impedance matrix and V_0 is the open circuit voltages at the antenna terminals. As Z_0 is nonsingular, we can obtain the element output voltages from the open circuit voltages. The element output voltages will be given as:

$$V = Z_0^{-1} V_0 \quad (3.33)$$

From (3.31), we known that matrix Z_0 is an impedance matrix which is normalized to the load impedance. It acts like a transformation matrix, transforming the open circuit element voltages to the terminal voltages. Normally, in the analyzing adaptive antenna systems, it is assumed that the element spacing is large enough so that the mutual coupling between the elements is small and consequently the open circuit voltage matrix Z_0 becomes diagonal. In addition, it is further assumed that the self-impedances Z_{ii} , ($i = 1, 2, \cdots, N$) are all equal, the input signal vector will be equal to the open circuit voltage vector multiplied by $1 + \frac{Z_{ii}}{Z_L}$. Thus, the array performance will

be same as the calculated one using the open circuit voltages as the input signals to an adaptive processor.

Let M signals of the same frequency be incident on the array. The open circuit voltages at the antenna terminals are given by

$$V_0 = \sum_{k=1}^M \mathbf{s}_k \quad (3.34)$$

where

$$\mathbf{s}_k = A_k e^{j(\omega_c t + \psi_k)} U_k \quad (3.35)$$

In (3.35), A_k^2 is the average power in the k th signal, ω_c is the carrier frequency, ψ_k is the carrier phase of the k th signal at the coordinate origin and U_k is the k th signal vector defined as follows:

$$U_k = \begin{bmatrix} f_1(\theta_k, \phi_k, p_k) e^{j\rho_{k1}} \\ f_2(\theta_k, \phi_k, p_k) e^{j\rho_{k2}} \\ \vdots \\ f_N(\theta_k, \phi_k, p_k) e^{j\rho_{kN}} \end{bmatrix} \quad (3.36)$$

where p_k is the polarization of the k th signal, $f_j(\theta_k, \phi_k, p_k)$ is the pattern response of the j th array element to a signal incident from direction (θ_k, ϕ_k) with the polarization p_k , and ρ_{kj} is the k th signal phase at the j th array element, measured with respect to the coordinate origin.

Using (3.33) and (3.34), the input signal to the adaptive processor will be

$$\begin{aligned} V &= Z_0^{-1} \left(\sum_{k=1}^M \mathbf{s}_k \right) \\ &= Z_0^{-1} \left(\sum_{k=1}^M A_k e^{i(\omega_c t + \psi_k)} U_k \right) \end{aligned} \quad (3.37)$$

In Z_0 matrix, the element Z_{ii} is the self-impedance of the i -th element when all other elements are open circuited. Mutual impedance, Z_{ij} ($= Z_{ji}$ by reciprocity), is the open circuit voltage produced at the first terminal pair divided by the current supplied to the second when all other terminals are open circuited, that is,

$$Z_{ij} = \left. \frac{v_i}{i_j} \right|_{I_n=0} \quad \text{for all } n \text{ except } j = n \quad (3.38)$$

Using this approach, several researchers studied the performance of MUSIC and ESPRIT algorithms to determine DOA using antenna array with different array geometries and including the effects of mutual coupling. These modifications involve either modifying the search vector using the inverse of the impedance matrix or reconstructing the signal subspace by solving a generalized eigenvalue problem.

3.4 New method

The idea of the new compensation method [73]- [74] is different than the open-circuit voltage method [67]. The most critical step in many mutual coupling compensation methods is to identify and quantify the amount of mutual coupling effect in an antenna array.

In new method, the first step is to re-define the mutual impedance. The re-defined mutual impedance is then used to quantify the amount of the mutual coupling in the array signals. Once the amount of the mutual coupling is known, mutual coupling among the antenna elements can be removed from the array signals.

The remaining question is how to approximately know the amount of mutual coupling through the measured or accessible quantities. In an antenna array, the measured or accessible quantity is the terminal voltage or current on an antenna terminal load, which is a single value of the whole voltage or current distribution function along the antenna. The inadequacies of the conventional method, the open-circuit voltage method, to estimate the amount of mutual coupling have been explained before. For the sake of completeness, we summarize the salient points of the new compensation method here, especially the calculation of the new mutual impedance. Unfortunately, in real practical antenna arrays, the amount of mutual coupling is very difficult to be exactly known due to the unavailability of the knowledge of the current distribution on each antenna element. The accuracy of this method depends on the accuracy in the calculation of the mutual impedances which in turn depends on how well we know the current distributions on the antenna elements. We define the mutual impedances differently from the conventional methods. The calculation of the new mutual impedance takes into account the presence of other antenna elements and uses a more accurate current distribution. Unlike the open circuit voltage method which seeks to find the open circuit voltages on the antenna terminals, the new method seeks to recover the signal voltages across the antenna terminal loads which are free from the mutual coupling effect. The new method does not require the known current distributions on the

antenna elements as in [66] or a pre-determined incoming elevation angle of the signal and the interferences as in [70]. Instead, it uses a single estimated current distribution for all the antenna elements.

3.4.1 Problem Formulation

Let's consider a linear antenna array with n antennas. The antenna elements are of equal dimensions and made of thin wires. It is shown in Fig.3.2, the array is placed along the x -axis with the feed point of the first element is coincident with the coordinate origin. An equal spacing d between the antenna elements is assumed and the antennas are oriented with their axes parallel to the z -axis. Every antenna is connected to an equal input load. The signal of interest (SOI) and all the interferences are coming from the upper half space. By using the standard moment method analysis [112], an

Figure 3.2: A linear array with n antenna elements.

electric field integral equation for the antenna array can be formulated by enforcing

the boundary condition of the tangential component of the total electric field on the surfaces of the antenna wires. That is,

$$\vec{E}_s + \vec{E}^{inc} = 0 \quad (3.39)$$

where $\vec{E}_s$ is the scattered electric field due to the induced electric currents along the antenna wires of the array and $\vec{E}^{inc}$ is the incident field due to the SOI and the interferences. The expression of $\vec{E}_s$ can be cast by using the free-space dyadic Green's function $\overline{\overline{G}}_{e0}(\vec{R}, \vec{R}')$ [115] in the most general form as:

$$\vec{E}_s = -j\omega\mu \sum_{p=1}^n \int_{\ell_p} \overline{\overline{G}}_{e0}(\vec{R}, \vec{R}') \cdot \vec{I}_p(\vec{R}') d\ell' \quad (3.40)$$

where ℓ_p and $\vec{I}_p$ denote, respectively, the length and the current on the p th antenna element. $\vec{R}$ is the point at which the field is to be calculated and $\vec{R}'$ is the point where the current source is located. All fields and sources are assumed to vary harmonically with the time-dependent factor $e^{j\omega t}$ being suppressed throughout unless stated otherwise. By discretizing each wire antenna into m segments and matching (3.39) using the Galerkin method [112], a system of $m \times n$ linear equations can be obtained in the following matrix form:

$$\begin{bmatrix} [Z^{11}] & [Z^{12}] & \dots & [Z^{1n}] \\ [Z^{21}] & [Z^{22}] & \dots & [Z^{2n}] \\ \vdots & \vdots & \ddots & \vdots \\ [Z^{n1}] & [Z^{n2}] & \dots & [Z^{nn}] \end{bmatrix} \begin{bmatrix} [I^1] \\ [I^2] \\ \vdots \\ [I^n] \end{bmatrix} = \begin{bmatrix} [V^1] \\ [V^2] \\ \vdots \\ [V^n] \end{bmatrix} \quad (3.41)$$

where the matrix blocks are respectively

$$[Z^{\alpha\beta}] = \begin{bmatrix} z_{11}^{\alpha\beta} - Z_L & z_{12}^{\alpha\beta} & \cdots & z_{1m}^{\alpha\beta} \\ z_{21}^{\alpha\beta} & z_{22}^{\alpha\beta} & \cdots & z_{2m}^{\alpha\beta} \\ \vdots & \vdots & \ddots & \vdots \\ z_{m1}^{\alpha\beta} & z_{m2}^{\alpha\beta} & \cdots & z_{mm}^{\alpha\beta} \end{bmatrix} \quad (3.42)$$

$$[I^\alpha] = \begin{bmatrix} I_1^\alpha & I_2^\alpha & \cdots & I_m^\alpha \end{bmatrix}^T \quad (3.43)$$

$$[V^\alpha] = \begin{bmatrix} V_1^\alpha & V_2^\alpha & \cdots & V_m^\alpha \end{bmatrix}^T \quad (3.44)$$

The elements in (3.42) $z_{\mu\nu}^{\alpha\beta}$ (with $\alpha, \beta = 1, 2, \dots, n$ and $\mu, \nu = 1, 2, \dots, m$) have the following general expression:

$$z_{\mu\nu}^{\alpha\beta} = -j\omega\mu \iint \hat{\ell}_\mu^\alpha(\vec{R}) \cdot \overline{\vec{G}_{e0}}(\vec{R}, \vec{R}') \cdot \hat{n}_\nu^\beta(\vec{R}') d\ell' d\ell \quad (3.45)$$

where ℓ and ℓ' are, respectively, the coordinates along the μ th segment of the α th antenna element and the ν th segment of the β th antenna element, $\hat{\ell}_\mu^\alpha(z)$ and $\hat{\ell}_\nu^\beta(z')$ are the tangential unit vectors at the coordinates ℓ and ℓ' , respectively. In (3.43) and (3.44), I_β^α (respectively V_β^α) denotes the current (respectively the voltage due to the incident field) on the β th segment of the α th antenna element. The superscript “T” in (3.43) and (3.44) denotes the transpose operation on the respective row vector. The

elements of the voltage vector in (3.44) is calculated by the following expression,

$$V_\mu^\alpha = -j\omega\mu \int \hat{\ell}_\mu^\alpha(\vec{R}) \cdot \vec{E}^{inc}(\vec{R}) d\ell \quad (3.46)$$

From (3.41), we see that the existence of the mutual coupling effect is due to the off-diagonal matrix blocks, i.e., $[Z^{\alpha\beta}]$ with $\alpha \neq \beta$. If the current distributions on the antenna elements, $[I^\alpha]$ $\alpha = 1, 2, \dots, n$, are exactly known, the mutual coupling effect can be exactly quantified and removed. For example, if we want to eliminate the mutual coupling effect due to all other antenna elements on the p -th antenna element, then we can rewrite the p -th row of (3.41) as

$$\begin{aligned} & [Z^{pp}] [I^p] \\ &= - [Z^{p1}] [I^1] - \dots - [Z^{p(p-1)}] [I^{p-1}] - [Z^{p(p+1)}] [I^{p+1}] - \dots \\ & \quad \dots - [Z^{pn}] [I^n] + [V^p] \end{aligned} \quad (3.47)$$

We see that the left-hand side of (3.47) is exactly the system of equations resulted from the moment-method analysis for the p -th antenna element as if it were standing alone without the presence of all other antenna elements. However, on the right-hand side, we have two excitation sources now: the voltage due to the incident field on the p -th antenna element $[V^p]$ which we want to find, and the voltage due to the current distributions on the other antenna elements $- [Z^{p1}] [I^1] - \dots - [Z^{p(p-1)}] [I^{p-1}] - [Z^{p(p+1)}] [I^{p+1}] - \dots - [Z^{pn}] [I^n]$ which we want to eliminate. By the principle of superposition, the effects on the p -th antenna element due to these two excitation sources are independent, meaning that the field generated by any one of these two excitation sources satisfies the boundary conditions alone and hence can exist independent

of the field generated by the other source. Thus if the currents on the antenna elements are completely and exactly known, we can find $[Z^{pp}] [I^p]$ and $-[Z^{p1}] [I^1] - \dots - [Z^{p(p-1)}] [I^{p-1}] - [Z^{p(p+1)}] [I^{p+1}] - \dots - [Z^{pn}] [I^n]$ and then from (3.47) we obtain $[V^p]$ which is completely free of any mutual coupling effect.

The problem now is that we can only measure the voltages or currents on the antenna terminal loads only. We do not have the knowledge of the distributions of the currents or voltages on every point of the antenna elements. That is, at most we can only measure one point (the first one) of the current or voltage distribution in (3.43) or (3.44). To solve this problem, we make a “single-mode” approximation. That is, we treat the current distribution on each antenna element in the array as consisting of one segment only. Furthermore, we use the terminal currents and voltages on the antenna elements to describe the antenna array as an n-port network. By doing so, (3.41) is reduced to

$$\begin{bmatrix} Z_t^{11} - Z_L & Z_t^{12} & \dots & Z_t^{1n} \\ Z_t^{21} & Z_t^{22} - Z_L & \dots & Z_t^{2n} \\ \vdots & \vdots & \ddots & \vdots \\ Z_t^{n1} & Z_t^{n2} & \dots & Z_t^{nn} - Z_L \end{bmatrix} \begin{bmatrix} I_t^1 \\ I_t^2 \\ \vdots \\ I_t^n \end{bmatrix} = \begin{bmatrix} V_t^1 \\ V_t^2 \\ \vdots \\ V_t^n \end{bmatrix} \quad (3.48)$$

in which the matrix blocks are replaced by the respective terminal quantities and subscript “t” is added to signify these quantities being referred to the antenna terminals. Now equivalent to 3.47, we have

$$(Z_t^{pp} - Z_L) I_t^p = -Z_t^{p1} I_t^1 - \dots - Z_t^{p(p-1)} I_t^{p-1} - Z_t^{p(p+1)} I_t^{p+1} - \dots - Z_t^{pn} I_t^n + V_t^p \quad (3.49)$$

The measurable quantities in (3.49) are the terminal currents $I_t^1, I_t^2, \dots, I_t^n$. We can remove the voltage dropped across the self-impedance $Z_t^{pp} I_t^p$ of the p th antenna element from the left hand side of (3.49) while also removing that part of coupling voltages and that part of the incident voltage that cause this voltage drop on the self-impedance from the right hand side of (3.49). That is, we obtain (after re-arranging)

$$Z_L I_t^p = Z_\ell^{p1} I_t^1 + \dots + Z_\ell^{p(p-1)} I_t^{p-1} + Z_\ell^{p(p+1)} I_t^{p+1} + \dots + Z_\ell^{pn} I_t^n + U_\ell^p \quad (3.50)$$

where $Z_\ell^{p1} I_t^1 + \dots + Z_\ell^{p(p-1)} I_t^{p-1} + Z_\ell^{p(p+1)} I_t^{p+1} + \dots + Z_\ell^{pn} I_t^n$ is the total coupling voltage across the terminal load Z_L of the p -th antenna element due to the current distributions on the other antenna elements in the array, and U_ℓ^p is the coupling-free voltage across the terminal load Z_L of the p -th antenna element due to the incident field alone. Hence the objective now is to find voltage U_ℓ^p from the measured voltage $Z_L I_t^p$. From (3.50), it is even not necessary to know the self-impedance of the antenna elements. Note that the mutual impedance Z_ℓ^{pq} ($p \neq q$) in (3.50) is defined with the antenna elements connected to their respective terminal loads rather than open-circuited. For example, the mutual impedance Z_ℓ^{pq} in (3.50) is equal to the voltage induced across the terminal load Z_L of the p -th antenna element by a unit current flowing through the terminal load Z_L of the q -th antenna element (with its current distribution being excited by the incident field and the fields scattered from all the other antenna elements in the array).

We see that the mutual impedances so defined have already taken into account the presence of other antenna elements. This is because the current distributions (whose terminal values are those currents in (3.50)) used to calculate the mutual impedances produce a total electric field which satisfies the boundary condition on all the antenna wire surfaces. Hence in the calculation of the mutual impedances Z_ℓ^{pq} , we have to use the actual measured current distribution on the q -th antenna element, i.e., a current distribution which is excited by the SOI and the interferences and which produces an electric field satisfying the boundary condition on all the antenna wire surfaces. If we calculate all the mutual impedances in this way, then we can obtain the voltages U_ℓ^p ($p = 1, 2, \dots, n$) on the antenna terminals with the mutual coupling effect completely removed.

The accuracy of this method depends on the accuracy in the calculation of the mutual impedances Z_ℓ^{pq} which in turn depends on how well we know the current distributions on the antenna elements. As mentioned before, it is impossible to know the current distributions on the antenna elements but only their terminal values on the antenna loads. To solve this problem, we can use an estimated current distribution. But in this case the mutual coupling effect can only be reduced by a certain amount but not completely. For a small antenna such as a dipole antenna, the current distribution is rather stable in the sense that its shape and its phase are relatively unchanged irrespective to the incoming azimuth angle of the incident field if the incident field comes from the horizontal direction. Hence a reasonably accurate estimate of the current distributions on the antenna elements can be obtained by exciting a single antenna with an incoming plane wave from the horizontal direction and taking

the current distribution (after normalized by the value at the antenna terminal) as the estimated current distribution. This current distribution is then used to calculate the mutual impedances. An example of such a current distribution for a dipole antenna is shown in Fig.3.3. The dimensions of the dipole antenna are wire length $\ell = 0.5\lambda$, wire radius $a = \lambda/200$, frequency is at $2.4GHz$ and terminal load $Z_L = 50\ \Omega$. Note that both the relative magnitudes and the phases of the current at different points of the antenna are important in the calculation of the mutual impedances. This current distribution is used to calculate the mutual impedances for all the numerical examples in the following chapters. The result was obtained by using moment method.

Figure 3.3: The estimated normalized current distribution of a single dipole used for the calculation of the mutual impedances.

We see that the phase of this current distribution is very different from an equal-phase sinusoidal function which is typically assumed for the calculation of mutual

impedance in the conventional method [117], [108].

Note that the mutual impedances are needed to be calculated only once if there is no change to the array configuration. If there are no extremely large signals or interferences coming an elevation angle too different from the horizontal direction, this estimated current distribution is actually a good estimate for the measured current distributions on the antenna elements.

In brief, there five steps are used to obtain the compensated Voltages:

- 1.) Obtain the measured voltages across the antenna terminal loads, V_t^k .
- 2.) Calculate the measured terminal currents from the measured voltages, $I_t^k = V_t^k / Z_L$, where($k = 1, 2, \dots, n$).
- 3.) Calculate the mutual impedance Z_t^{ki} using the new method.
- 4.) Using the mutual impedances and measured terminal currents to calculate the coupling voltages due to other antenna elements. i.e. Coupling-voltages = $Z_t^{kl} I_t^i$.
- 5.) Subtract the result in step 4 from the result in step 1 to obtain the compensated voltage U_t^k .

To validate effectiveness of the new method, experimentally measured mutual impedance is need to obtain and compare with calculated one as shown in Figure 3.5. In following sections, the theoretical definition of new mutual impedance and experimental measurement procedure for mutual impedance are presented.

3.4.2 Receiving Mutual Impedance in New Method and Traditional Mutual Impedance

Traditionally, mutual impedance was used to measure the mutual coupling effect. The traditional mutual impedance employs a definition similar to that used for the mutual impedance in circuit analysis. The following description on the traditional mutual impedance are provided for comparison with our newly defined mutual impedance.

To begin a discussion of mutual coupling effects, consider the input impedance, or driving point impedance, of any element in an array of N elements. The relationship between the various currents and voltages are given by a familiar network relationships

$$\begin{aligned}
 V_1 &= I_1 Z_{11} + I_2 Z_{12} + \cdots + I_N Z_{1N} \\
 V_2 &= I_1 Z_{12} + I_2 Z_{22} + \cdots + I_N Z_{2N} \\
 &\vdots \\
 V_N &= I_1 Z_{1N} + I_2 Z_{2N} + \cdots + I_N Z_{NN}
 \end{aligned} \tag{3.51}$$

Where V_n and I_n are the impressed current and voltage in the n th element, Z_{nn} is self-impedance of the n th element when all other elements are open circuited, and Z_{mn} ($= Z_{nm}$ by reciprocity) is the mutual impedance between the m th and n th elements. The mutual impedance Z_{mn} between the two terminal pairs of elements m and n is the open circuit voltage produced at the first terminal pair divided by the current supplied to the second when all other terminals are open circuited, that is,

$$Z_{mn} = \left. \frac{V_m}{I_n} \right|_{I_i=0} \quad (3.52)$$

This equation is valid for all i except $i = n$. The active impedance of an element is the input impedance of that element when all other elements are excited. For example, from (3.51) the active impedance of element 1 is

$$Z_{1,in} = \frac{V_1}{I_1} = Z_{11} + \frac{I_2}{I_1} Z_{12} + \cdots + \frac{I_N}{I_1} Z_{1N} \quad (3.53)$$

We note that the active impedance is not merely the sum of the self-impedance and all the mutual impedance, but depends on the various currents as well.

Next, Let us consider how we might measure the mutual impedance between two antennas. Suppose an antenna when isolated in free space has a voltage V_1 and a current I_1 , so the input impedance is

$$Z_{in} = Z_{11} = \frac{V_1}{I_1} \quad (3.54)$$

If a second antenna is brought into proximity with the first then radiation from the first antenna will induce currents on the second, which in turn will radiate by virtue of that induced current and influence the current on the first antenna. The second antenna may either be excited or unexcited (parasitic), but in any case it has terminal

current I_2 . Then the total voltage at the first antenna is

$$V_1 = Z_{11}I_1 + Z_{12}I_2 \quad (3.55)$$

Similarly, the voltage at the terminals of the second antenna is expressed by

$$V_2 = Z_{21}I_1 + Z_{22}I_2 \quad (3.56)$$

Now suppose the second antenna has a load impedance Z_2 across its terminals such that $V_2 = -Z_2I_2$. We may write (3.56) as

$$-Z_2I_2 = Z_{21}I_1 + Z_{22}I_2 \quad (3.57)$$

Solving for I_2 we obtain

$$I_2 = \frac{-Z_{21}I_1}{Z_{22} + Z_2} = \frac{-Z_{12}I_1}{Z_{22} + Z_2} \quad (3.58)$$

Substituting this into (3.55) and dividing by I_1 , we find that

$$\frac{V_1}{I_1} = Z_{1,in} = Z_{11} - \frac{(Z_{12})^2}{Z_{22} + Z_2} \quad (3.59)$$

This express the input impedance in terms of the two self-impedances (Z_{11}) and (Z_{22}), the mutual impedance (Z_{12}) and the load Z_2 at the unexcited terminals of antenna 2. For example, if the two antenna are half-wave dipoles and if the terminals of the second dipole are open circuited, then $Z_{1,in} = Z_{11}$ because $Z_2 = \infty$. Physically this means that very little current is induced on each of the arms of the second dipole. As a consequence, one could measure the self-impedance of the first dipole in the presence of the second simply by open circuiting the second dipole thereby rendering it non-resonant and reducing the current on it.

The above discussion suggests the equivalent circuit for the coupling between two resonant antennas. Note that if the terminals on the right are open circuited $Z_{1,in} = Z_{11}$.

In general, to determine the mutual impedance between two antennas, the following three measurements are performed.

- 1.) Measure Z_{11} at the terminal of antenna 1 with antenna 2 either removed or open circuited.
- 2.) Measure Z_{22} in a similar manner to that for Z_{11} .
- 3.) Measure $Z_{1,in}$ when antenna 2 is short circuited, that is, $Z_2 = 0$. From (3.59) we

$$\text{find that } Z_{12} = \sqrt{Z_{22}(Z_{11} - Z_{1,in})}$$

All quantities on the right are known from the three measurements and thus Z_{12} can be computed [8].

While traditional mutual impedance give some convenience for the analysis of the mutual coupling problem in antenna arrays [67], it is rather questionable as to whether this circuit concept can give accurate enough results in a field problem such as antenna array analysis. As wireless components become increasingly smaller in size, antenna arrays with small antenna element such as monopoles, dipoles and normal-mode helical antennas will provide a very attractive solution in many applications. In following section, a new mutual impedance, called the receiving mutual impedance is defined.

The two identical antennas are separated along the x-axis by a distance d . An external source coming from a direction of ϕ on the azimuth plane is to excite the antennas. Both antennas are connected to a terminal load Z_L and mounted over a large ground plane. The receiving mutual impedance, denoted by Z_t^{12} , between these two antennas is defined as the ratio of the induced voltage V_1 across the terminal load of antenna 1 to the current I_2 on the terminal load of antenna 2 when the current on antenna 2 is excited by the external source. That is,

$$Z_t^{12} = \frac{V_1}{I_2} \quad (3.60)$$

Here, the induced voltage V_1 on antenna 1 should be solely due to the current distribution on antenna 2 and not due to the external source. The word "receiving" is to distinguish this definition from the conventional definition of mutual impedance in that both antennas are in the receiving mode. Z_t^{12} is defined similarly but with the position of antenna 1 and 2 interchanged. Note that this definition requires an

external excitation source to excite the two antennas. Furthermore, the definition of the receiving mutual impedance includes a terminal load Z_L and hence it also depends on the terminal load connected to the antennas. The advantage of including the terminal load into the receiving mutual impedance is that the loading effect is also taken into account. This is similar to the definition of scattering parameters which also depends on the system impedance. To find the receiving mutual impedance, both experimental and theoretically methods are used. The measuring procedure for new mutual impedance is listed in next section. In the theoretical method, the values of V_1 and I_2 in 3.60 are calculated in an indirect manner by using a simulation tool, FEKO [116]. First, the two antennas are excited by an incoming plane electromagnetic wave and the terminal voltages and currents are calculated, for example V'_1 , V'_2 , I_1 , and I_2 . Secondly, the terminal voltages on the isolated antennas are calculated. This can be done by removing one of the antennas from the array when the terminal voltage on the other is being calculated. This can be done by removing one of the antennas from the array when the terminal voltage on the other is being calculated. For example, the isolated terminal voltages are calculated to be V''_1 and V''_2 . Then according to the superposition principle, the terminal voltage on antenna 1, i.e., V_1 is then given by,

$$V_1 = V'_1 - V''_1 \quad (3.61)$$

Hence Z_t^{12} can be calculated by using 3.60. Z_t^{12} can be calculated similarly.

3.4.3 Measuring Procedure for New Mutual Impedance

The measurement is to be done inside the anechoic chamber. The two antennas must be in a fixed position inside the anechoic chamber during the whole measuring process which is shown in Figure 3.4.

Figure 3.4: The experiment set-up for measuring mutual impedance.

The antennas are labelled as $ANT1$ and $ANT2$. Three voltages need to be measured.

1. Measure the terminal voltage at $ANT1$ with $ANT2$'s terminal connected to a load Z_0 . Denote it as V_1 .
2. Measure the terminal voltage at $ANT2$ with $ANT1$'s terminal connected to a

load Z_0 . Denote it as V_2 .

3. Measure the terminal voltage at $ANT1$ with $ANT2$ remove. Denote it as V'_1 .

Then,

$$V_1 = Z_{12}I_2 + V_{o1} \quad (3.62)$$

$$V'_1 = V_{o1} \quad (3.63)$$

$$I_1 = -\frac{V_1}{Z_0}, I'_1 = -\frac{V'_1}{Z_0}, \text{ and } I_2 = -\frac{V_2}{Z_0} \quad (3.64)$$

From (3.62) (3.63), we have

$$\begin{aligned} (V'_1 - V_1) &= -Z_{12}I_2 \\ &= \text{total open - circuit voltage dropped across} \\ &\text{the terminal of ANT1 due to a current } I_2 \text{ on} \\ &\text{the terminal of ANT2} \end{aligned} \quad (3.65)$$

Hence we have,

$$Z_{12} = \frac{(V'_1 - V_1)}{V_2} Z_0 \quad (3.66)$$

Now the measured quantity in the anechoic chamber is the scattering parameter:

$$S_{21} = \frac{b}{a} \quad (3.67)$$

where a is the square root of the power supplied to the emitting horn and $b = \frac{V}{\sqrt{Z_0}}$ is the square root of the power received by the test antenna with V being the terminal voltage of the test antenna. Thus we can write

$$V_1 = a\sqrt{Z_0}S_{21}^{(1)} \quad (3.68)$$

$$V_1' = a\sqrt{Z_0}S_{21}^{(1)'} \quad (3.69)$$

$$V_2 = a\sqrt{Z_0}S_{21}^{(2)} \quad (3.70)$$

and express (3.66) in terms of the scattering parameters as

$$Z_{12} = \frac{\left(S_{21}^{(1)'} - S_{21}^{(1)}\right)}{S_{21}^{(2)}}Z_0 \quad (3.71)$$

Hence, we can measure the scattering parameter S_{12} instead of measuring the voltage by using the same procedure as for the measurement of voltage, i.e., the 3 steps stated at the beginning. We measure $S_{12}^{(1)}$, $S_{12}^{(1)'}$, $S_{12}^{(2)}$ instead of V_1 , V_1' , V_2 , respectively, and calculate the mutual impedance using (3.71.)

We give measured results of the new mutual impedance to verify the theoretical calculations.

Figure 3.5: The measured and calculated mutual impedances of two monopole antennas.

Figure 3.5, shows the measured and calculated mutual impedance of two monopole antennas. The dimensions of two monopoles are shown in the figure: length = 3.0 cm, wire radius = 0.3 mm and inter element spacing $d = 6.25$ cm (half wavelength at 2.4

GHz). From this figure, we can see that using new method, we can obtain very closely spaced two similar lines, which mean using new method the calculation of the mutual impedances is very accurate.

3.5 Concluding Remarks

In this chapter, two conventional method and the new method for the compensation of mutual coupling effect are introduced. The new method introduces a new definition of mutual impedance which is used to find the coupling-free signal voltages across the antenna terminal loads. The procedures on how to measure and calculate the traditional and new mutual impedances are given out. It will show in the later chapters, that the new method works well in different array configurations and applications even under the influence of the strong interferences.

Chapter 4

Compensation for the Mutual Coupling Effect in ULA with ESPRIT DOA Estimation

The proposed method of compensation for the mutual coupling effect is applied with the ESPRIT direction finding algorithm in array antennas. Studying on two closely spaced arriving signals shows that the high-resolution capability of ESPRIT can be achieved only when the mutual coupling effect is compensated for by using the proposed compensation method. Critical situations with a large signal level difference, with an increased mutual coupling effect resulting from a more compact-size antenna array, and with signals coming from a non-horizontal elevation angle are also studied using the ESPRIT algorithm. Results show that the proposed compensation method, when applied to ESPRIT, is more accurate, more robust, and more flexible than the previous

open-circuit voltage method.

4.1 Introduction

ESPRIT [81] is a well-known high-resolution direction finding algorithm. Compare to MUSIC [88], ESPRIT requires much less computation and storage and does not involve an exhaustive searching procedure. However, notwithstanding these advantages, ESPRIT, which relies on the eigen-structure of the input covariance matrix, requires accurate knowledge of the received signal voltages without interference from other antenna elements. This is almost impossible to meet in any practical antenna arrays due to the mutual coupling effect. Therefore, methods have to be used to reduce (or compensate for) the mutual coupling effect [64]- [75]. Unfortunately, compensation methods for the ESPRIT algorithm are rarely seen in previous studies. Only in [68] and [69], the authors used a method which was suggested first in [64] to compensate for the mutual coupling effect in the direction finding algorithms ESPRIT and MUSIC, respectively. The open-circuit voltages were calculated from the measured voltages through the mutual impedances. This may be termed the open-circuit voltage method. In [75], Hui proposed a new method for the compensation of the mutual coupling effect and demonstrated its superior performance over the open-circuit voltage method in adaptive nulling and direction finding. This method does not involve any unrealistic assumption but relies on the calculation of a new mutual impedance. However, so far the continuing studies in [77] and [74] only revealed the effectiveness of the new method in direction finding using the MUSIC algorithm and in adaptive

nulling using the direct data domain method. The performance of the new method when applied to other adaptive algorithms is still a question needed to be answered. It was in our research work [73] that the new method was first applied to the ESPRIT algorithm and some preliminary results were obtained. ESPRIT uses a different approach from MUSIC and is capable of very high-resolution direction finding. In this chapter, we report an intensive study of applying the new compensate method to the ESPRIT algorithm with a focus on studying the new method under some critical situations. Through a comparison study, we found that the high-resolution capability of the ESPRIT can only be restored by using our new compensation method to compensate for the mutual coupling effect. The cases with a large signal level difference, with an increased mutual coupling effect resulting from a more compact-size antenna array, and with signals coming from a non-horizontal elevation angle are studied for the first time using the ESPRIT algorithm. Results show the robustness and the flexibility of the new compensation method even under some critical signal conditions.

4.2 The Compensation Method and Its Application to ESPRIT

The original idea of the proposed compensation method adopts a more practical approach similar to the open-circuit voltage method [64], requiring only the knowledge of the terminal voltages or currents.

It seeks first to quantify the amount of mutual coupling and then to remove it from

the measured signal voltage. In order to quantify the mutual coupling effect in a more accurate manner, a new mutual impedance concept has been introduced. This new mutual impedance emerges from the following consideration of the mutual coupling effect and is the key parameter in the proposed compensation method. Let us consider an antenna array consisting of n elements, the induced voltage on an antenna terminal load can be considered to be excited by two external sources: the incoming signals and the scattered field due to other antenna elements in the array. The latter part is the cause of the mutual coupling. Thus the voltage, for example, across the k -th antenna terminal load V_t^k of the antenna array can be written as:

$$V_t^k = Z_L I_t^k = U_t^k + Z_t^{k1} I_t^1 + \cdots + Z_t^{k(k-1)} I_t^{k-1} + Z_t^{k(k+1)} I_t^{k+1} + \cdots + Z_t^{kn} I_t^n \quad (4.1)$$

where Z_L is the terminal load, I_t^k ($k = 1, 2, \cdots, n$) is the current on the k -th antenna terminal load, U_t^k is the voltage across the k -th antenna terminal load due to the incoming signals alone (without mutual coupling), and Z_t^{ki} ($i = 1, 2, \cdots, k-1, k+1, \cdots, n$) is the new mutual impedance between the k -th and the i -th antenna elements. That part $Z_t^{k1} I_t^1 + \cdots + Z_t^{k(k-1)} I_t^{k-1} + Z_t^{k(k+1)} I_t^{k+1} + \cdots + Z_t^{kn} I_t^n$ on the right-hand side of (4.1) is the voltage due to mutual coupling which must be removed from the terminal voltage in order to get U_t^k . Apparently, in order to calculate the voltage due to the mutual coupling, the mutual impedance Z_t^{ki} must be known. From (4.1), the meaning of the new mutual impedance Z_t^{ki} is the ratio of the induced voltage across the terminal load Z_L of the k -th antenna element to the exciting terminal current flowing through the terminal load Z_L of the i -th antenna element (with its current distribution being

excited by the incoming signal and the scattered field from all other antenna elements in the array). Obviously, the meaning of the new mutual impedances is different from the conventional definition of the mutual impedance (for example, that in [108]). The most distinguished difference of the new mutual impedance is that it is defined with the antenna terminals connected to a load, rather than open-circuited as in the conventional definition. Secondly, in the new definition of mutual impedance, both antenna elements are in the receiving mode (not connected to a source) whereas in the conventional definition, one antenna is in the receiving mode while the other is in the transmitting mode (connected to a source). The actual calculation of the new mutual impedance requires an estimated current distribution as explained in Chapter 3. Once the mutual impedances are all known, the voltage due to the mutual coupling effect (the second part on the right-hand side of (4.1) can be determined. When considering all elements in the array together, we have the following system of equations relating the measured terminal voltages $\mathbf{V} = [V_t^1, V_t^2, \dots, V_t^n]^T$ to the voltages due to the signals alone $\mathbf{U} = [U_t^1, U_t^2, \dots, U_t^n]^T$ i.e.,

$$\mathbf{ZV} = \mathbf{U} \quad (4.2)$$

where,

$$\mathbf{Z} = \begin{bmatrix} 1 & -\frac{Z_t^{12}}{Z_L} & \cdots & -\frac{Z_t^{1n}}{Z_L} \\ -\frac{Z_t^{21}}{Z_L} & 1 & \cdots & -\frac{Z_t^{2n}}{Z_L} \\ \vdots & \vdots & \ddots & \vdots \\ -\frac{Z_t^{n1}}{Z_L} & -\frac{Z_t^{n2}}{Z_L} & \cdots & 1 \end{bmatrix} \quad (4.3)$$

Hence if Z is completely known, we can determine U from V .

Now let $X(t)$ be the complex voltage vector received at time t from the array. Then

$$\mathbf{x}(t) = \mathbf{v}(t)^T + \mathbf{n}(t)^T \quad (4.4)$$

where $\mathbf{v}(t)$ is the time-domain form of V and $\mathbf{n}(t)$ is the complex voltage vector (in time-domain) of the Gaussian noise. The complex noise voltages on the antenna terminals are assumed to be completely uncorrelated with each other and with the signal voltages. Accounting for the compensation of mutual coupling effect using (4.2) and (4.3) above, the input covariance matrix for the ESPRIT algorithm can be expressed as:

$$\mathbf{R}_{xx} = \mathbf{Z} E \{ \mathbf{x}(t) \mathbf{x}^H(t) \} \mathbf{Z}^H = E \{ \mathbf{u}(t) \mathbf{u}^H(t) \} + \mathbf{Z} E \{ \mathbf{n}(t) \mathbf{n}^H(t) \} \mathbf{Z}^H \quad (4.5)$$

where $E \{ \bullet \}$ represents the time-averaged value of the quantity inside the bracket, “ H ” denotes the conjugate transpose operation, and $u(t)$ is time-domain form of U . We see from (4.5) that the compensation process distorts the noise subspace also, as observed in [69], with the result being $\mathbf{Z} E \{ \mathbf{n}(t) \mathbf{n}^H(t) \} \mathbf{Z}^H$. If perfect sampling can be

assumed in sampling the voltage vector $X(t)$, then the right-hand side of (4.5) can be represented by its frequency-domain equivalence as

$$\mathbf{R}_{xx} = \mathbf{U}\mathbf{U}^H + \sigma^2\mathbf{Z}\mathbf{Z}^H \quad (4.6)$$

where σ^2 is the noise power. Thus we see the input covariance matrix consists of a covariance matrix formed by the compensated voltage vector $\mathbf{U}$ and a distorted noise matrix $\sigma^2\mathbf{Z}\mathbf{Z}^H$. In the following section, computer simulations will be used to demonstrate the effectiveness of this new mutual coupling compensation method when applied to ESPRIT.

4.3 Computer Simulation and Discussions

To demonstrate the compensation effectiveness of the new method for ESPRIT, computer simulation examples are performed on a linear monopole antenna array with seven elements as shown in Fig 4.1. The dimensions of the monopole antenna elements are: length = 3.0 cm and wire radius = 0.3 mm. They are placed over a large ground plane and are connected to a 50 Ω load. The array is aligned along the X axis with the monopole elements parallel to the Z axis. The new mutual impedances Z_t^{ki} of this array have been calculated and shown in Table 4.1, at a frequency of 2.4 GHz and an inter-element spacing $d = 6.25$ cm (half wavelength at 2.4 GHz).

Note that as shown in [74], we even do not need to know the self-impedance of the antennas elements by using the new compensation method. The two incoming signals

(a)

(b)

Figure 4.1: A linear monopole array with seven elements.

are coherent sinusoidal plane waves with vertical polarization, amplitude = 1 V /m, frequency = 2.4 GHz, and a signal-to-noise ratio = 10 dB. Three different kinds of voltages obtained from the array are input to the ESPRIT algorithm for DOA estimations. The first kind is the measured (uncompensated) voltages across the terminal loads of the antenna elements, i.e., the vector V in (4.2) and are obtained by using the moment method. The second kind of voltages is the open-circuit voltages obtained from the measured voltages with the mutual coupling effect being compensated by using the open-circuit voltage method [64]. The third kind of voltages is obtained from the measured voltages with the mutual coupling effect being compensated by using the new method as described in Section II, i.e., the vector U in (4.2). As the two signals are coherent, the spatial smoothing technique [86] has been used to de-correlate them and reconstruct the signal and noise subspaces. We also assume that perfect sampling of the sinusoidal signal sources has been performed. This eliminates the averaging process to obtain the covariance matrix and also removes the error due to imperfect sampling so that the error due to the mutual coupling effect can be studied alone. The detection results are shown in Table 4.2. In our simulations, we gradually decreased

Table 4.1: The mutual impedances calculated by using the new method for a monopole array.

Mutual impedances by the new method	
$Z_t^{12} = Z_t^{21} = Z_t^{23} = Z_t^{32} = Z_t^{34} = Z_t^{43} = Z_t^{45} = Z_t^{54} = Z_t^{56} = Z_t^{65} = Z_t^{67} = Z_t^{76}$	$= 4.0 + j8.7$
$Z_t^{13} = Z_t^{31} = Z_t^{24} = Z_t^{42} = Z_t^{35} = Z_t^{53} = Z_t^{46} = Z_t^{64} = Z_t^{57} = Z_t^{75}$	$= -1.3 - j5.2$
$Z_t^{14} = Z_t^{41} = Z_t^{25} = Z_t^{52} = Z_t^{36} = Z_t^{63} = Z_t^{47} = Z_t^{74} = 0.7 + j3.6$	
$Z_t^{15} = Z_t^{51} = Z_t^{26} = Z_t^{62} = Z_t^{37} = Z_t^{73} = -0.4 - j2.7$	
$Z_t^{16} = Z_t^{61} = Z_t^{27} = Z_t^{72} = 0.3 + j2.2$	
$Z_t^{17} = Z_t^{71} = -0.2 - j1.9$	

the azimuth angular separation between the two incoming signals until they are 90° and 93° as shown in Table 4.2.

Table 4.2: The ESPRIT detection results by using different compensation methods when the azimuth angles of two signals are $\phi_1 = 90^\circ$ and $\phi_2 = 93^\circ$, elevation angle $\theta = 90^\circ$, and $d = 0.5 \lambda$.

Compensation Method	Signal 1		Signal 2	
	Estimated ϕ_1	Error%	Estimated ϕ_2	Error%
New method	90.58°	0.64	93.11°	0.12
Open-circuit voltage Method	91.58°	1.76	81.73°	12.11
Measured voltages	91.49°	1.66	14.10°	84.83

This is to test the effect of mutual coupling on the high-resolution capability of the ESPRIT algorithm. As can be seen, the errors produced by using the open-circuit voltage compensation method are much greater than those by using the new compensation method. This indicates that the compensation for the mutual coupling effect is not effective by using the open-circuit voltage method. Even worse, Signal 2 has been detected to be on the wrong side of Signal 1. On the other hand, by using the new compensation method, we see the errors produced are very small, indicating the effective removal of the mutual coupling effect.

In the first example, Signal 1 and Signal2 are assumed to be of same strength so that the levels of mutual coupling effect due to the incidences of these two signals on the antenna elements are approximately the same. In this example, we increase the strength of Signal 2 gradually while keeping the strength of Signal 1 unchanged as in the above examples. That is, we increase the level of mutual coupling due to Signal 2. This is a critical test for compensation methods as the level of mutual coupling from the stronger signal will ultimately overwhelm the weaker signal and fail the detection

algorithm. However, as the result (with the same signal environment as that in Table 4.2 except that the amplitude of Signal 2 increases from 1 to 20 V/m) shown in Fig. 4.2 shows, the open-circuit voltage method fails ($error = 75\%$) very fast for the detection of Signal 1 (the weaker signal) when amplitude of Signal 2 increases to about $3V/m$. On the other hand, the % error incurred by the new compensation method for the detection of Signal 1 increases to a maximum value of only about 11% when Signal 2's amplitude increases to $20V/m$. This shows that the new compensation method works well with ESPRIT even under a large signal level difference. Note that the detections of the Signal 2 (the stronger signal) are similar for the two compensation methods after its amplitude increases beyond $10V/m$ as the weaker signal has basically been overwhelmed by the mutual coupling effect.

Next we investigate the performance of the compensation method for ESPRIT when the mutual coupling effect is increased by decreasing the inter-element spacing of the array from 0.5λ to 0.3λ . The signal environment is same as that in Table 4.2. The detection results are shown in Table 4.3 for the signal angular separation being 5° and Table 4.4 for the signal angular separation being 7° . It can be seen that when the mutual coupling is strong, both the open-circuit voltage method and uncompensated voltages fail to locate Signal 2 as shown in Table 4.3.

Only by using the new method can both signals be located. When the signal separation is increased to 7° as in Table 4.4, we see that open-circuit voltage method is now able to locate Signal 2. Hence from these two examples, we see that the new compensation method possesses a stronger compensation power even when the mutual

Figure 4.2: ESPRIT detection error at different Signal 2 amplitudes by using different compensation methods. The azimuth angles of the two signals are $\phi = 90^\circ$ AND $\phi = 93^\circ$ and the elevation angle $\theta = 90^\circ$. The array is a linear monopole antenna array with seven elements with element spacing $d = 0.5 \lambda$. The monopole length $= 3.0$ cm and wire radius $= 0.3$ mm. The amplitude of Signal 1 is fixed at $1 \text{ V} / \text{m}$.

Table 4.3: The ESPRIT detection results by using different compensation methods when the azimuth angles of two signals are $\phi_1 = 90^\circ$ and $\phi_2 = 95^\circ$, elevation angle $\theta = 90^\circ$, and $d = 0.3 \lambda$.

Compensation Method	Signal 1		Signal 2	
	Estimated ϕ_1	Error%	Estimated ϕ_1	Error%
New method	91.67°	1.86	93.59°	1.48
Open-circuit voltage Method	92.47°	2.74	No real solution	-
Measured voltages	92.78°	3.09	No real solution	-

coupling effect is increased.

Table 4.4: The ESPRIT detection results by using different compensation methods when the azimuth angles of two signals are $\phi_1 = 90^\circ$ and $\phi_2 = 97^\circ$, elevation angle $\theta = 90^\circ$, and $d = 0.3 \lambda$.

Compensation Method	Signal 1		Signal 2	
	Estimated ϕ_1	Error%	Estimated ϕ_2	Error%
New method	90.50°	0.56	96.49°	0.52
Open-circuit voltage Method	93.18°	3.53	97.34°	0.35
Measured voltages	93.74°	4.15	No real solution	-

In a more realistic consideration, incoming signals may not always come from the 90° elevation angle (the horizontal plane) although the linear monopole array is theoretically only capable of detecting 1D signal direction (on the horizontal plane). We show in this last example the effect of the removal of restriction on the elevation angle of the incoming signals being on the horizontal plane ($\theta = 90^\circ$). The signal environment for this example is same as that in Table 4.2 except that the elevation angle of Signal 2 is changed to $\phi = 85^\circ$, i.e., 5° above the horizontal plane. The detection results are shown in Table 4.5. It can be seen that the new compensate method produces a much smaller detection error than the open-circuit voltage method, especially for Signal 2. This shows the flexibility of the new compensation method even when the signals do not all come from the horizontal direction.

Table 4.5: The ESPRIT detection results by using different compensation methods when the azimuth angles of two signals are $\phi_1 = 90^\circ$ and elevation angle $\theta_1 = 90^\circ$ for signal 1, $\phi_2 = 93^\circ$, $\theta_2 = 85^\circ$ for signal 2, and $d = 0.5 \lambda$.

Compensation Method	Signal 1		Signal 2	
	Estimated DOA	Error%	Estimated DOA	Error%
The new method	90.57°	0.64	93.10°	0.11
Open-circuit voltage Method	91.61°	1.79	84.86°	8.75
Measured voltages	91.48°	1.65	14.05°	84.88

From the above examples, we see that although ESPRIT is capable of producing high resolution estimations of signal directions, the mutual coupling effect has to be compensated for properly before the algorithm can be used and the accuracy of its detection is highly dependent on the effectiveness of the method used to compensate for the mutual coupling effect.

4.4 Concluding Remarks

An effective method has been introduced to compensate for the mutual coupling effect in the ESPRIT direction finding algorithm. The compensation method seeks to quantify the mutual coupling more accurately by using a new mutual impedance. Study of two closely spaced arriving signals shows that the high-resolution capability of ESPRIT can be achieved only when the mutual coupling effect is compensated for by using our new compensation method. Furthermore, the cases with a large signal level difference, with an increased mutual coupling effect resulting from a more compact-size antenna array, and with signals coming from a non-horizontal elevation angle have also been studied using the ESPRIT algorithm. Our results show that the proposed compensation method, when applied to ESPRIT, is more accurate, more robust, and more flexible than the previous open-circuit voltage method.

Chapter 5

Mutual Coupling Compensation for UCA with ML-GA DOA Estimation

The new compensation method for the mutual coupling effect is applied to uniform circular arrays (UCAs) which are employed for two dimensional DOA estimations. A new 2D DOA searching algorithm using the maximum likelihood technique optimized by a genetic algorithm is introduced. This method circumvents the difficulty of dealing with coherent signals in 2D DOA estimations. The new searching algorithm is less computationally demanding than the maximum likelihood method (MLM) and statistically more efficient. Our study shows that the searching algorithm can effectively combine with the new mutual coupling compensation method to obtain very accurate and robust 2D DOA estimations in UCAs. The theory of the new searching algorithm

and its articulation with the compensation method for UCAs is succinctly formulated. Computer simulation examples on several typical scenarios are presented to demonstrate the accuracy and improved performance by using the proposed compensation method and the new searching algorithm.

5.1 Introduction

The performance of an adaptive antenna array is strongly affected by the effect of mutual coupling between antenna elements. Therefore compensation of the mutual coupling effect in adaptive antenna arrays is of critical importance, especially for high resolution direction of arrival (DOA) estimation. To compensate for or to reduce this effect, many efforts have been made before but only for uniform linear array (ULA) [67]- [79]. Unfortunately, ULAs do not provide a solution to most practical situations wherein a 360° field of view is required. On the other hand, uniform circular arrays (UCAs) can be easily used to provide information on DOA of both the azimuth and elevation angles. UCAs are attractive in many practical applications because of its simplicity. For ULAs, a traditional common method to compensate for the mutual coupling effect is through the calculation of the open-circuit voltages as in [67]- [71]. Because of the use of conventional mutual impedance, the method lacks accuracy, especially when the element separation is small. In [63], the authors introduced a minimum norm technique mutual coupling compensation method, which is based on the technique in [67]. They showed that the approach was more accurate than the traditional method using open-circuit voltages. In [74]- [77], the authors proposed a completely new compensation

method for ULAs through the introduction of a new mutual impedance. A dramatically increase in the accuracy of DOA estimations based on MUSIC and ESPRIT has been demonstrated. In this chapter, this method will be extended to compensate for the mutual coupling effect in UCAs for the first time. For UCAs, traditional DOA searching algorithms such as MUSIC does not work with coherent signals because the directional vector lacks a Vandermonde structure. In the current study, we will employ a new searching algorithm which is based on using the maximum likelihood (ML) technique and optimized by a genetic algorithm. This method circumvents the difficulty of dealing with coherent signals in 2D DOA estimations. The new searching algorithm is also less computationally demanding than the maximum likelihood method and statistically more efficient. Through several computer simulation examples, we demonstrate the advantages of combining the new compensation method with the new DOA searching algorithm together for use in UCAs. The accuracy of 2D DOA estimation has been increased in comparison with the traditional compensation method. The ability to new DOA searching algorithm to handle coherent signals by using UCAs is fully tested. In the next section, we will first briefly describe the working of the new compensation method in UCAs and then in Section III, we explain the new 2D DOA searching algorithm. Computer simulation results are in Section IV and conclusions are made in Section V.

Table 5.1: The mutual impedances calculated for the UCA by using new method.

$Z_t^{12} = Z_t^{23} = Z_t^{34} = Z_t^{45} = Z_t^{56} = Z_t^{67} = Z_t^{78} = -0.0817 - j11.7693$
$Z_t^{13} = Z_t^{24} = Z_t^{35} = Z_t^{46} = Z_t^{57} = Z_t^{68} = 7.1851 + j2.1175$
$Z_t^{14} = Z_t^{25} = Z_t^{36} = Z_t^{47} = Z_t^{58} = -0.0625 + j5.9557$
$Z_t^{15} = Z_t^{26} = Z_t^{37} = Z_t^{48} = -2.4977 + j4.9595$

5.2 Coping with Mutual Coupling Effect in UCAs

The UCA geometry is depicted in Fig.5.1. The $N = 8$ identical monopole antenna elements, oriented in z -direction, are uniformly placed along the circumference of a circle of radius $a = 6.25$ cm in the xy plane. The monopole elements are of equal length $l = 3.13$ cm and radius $\alpha = 0.3$ mm $\ll \lambda$, where λ is the operation wavelength. All the monopole elements are loaded with a terminal load $Z_L = 50\Omega$. The spherical coordinate system is used to present the arrival directions of the incoming plane waves which are assumed to come from different directions. All signals are assumed to be plane electromagnetic waves propagating in the 3-D space. Signal frequency is at 2.4 GHz. The origin of the coordinate system is located at the center of the UCA.

The problem of compensation for the mutual coupling effect in UCAs can be formulated in the same way to that for ULAs in chapter 3. To demonstrate the performance of UCAs with the compensation method, computer simulation is carried out with an array of 8 monopole antenna elements in the geometry of an UCA. The monopole antenna array are employed for DOA finding, Figure 5.1 is the side view of the UCA, and Figure 5.2, top view of the array.

The calculated new mutual impedance for the UCA is shown in Table 5.1.

Figure 5.1: Side view of the UCA deployed for direction finding

Figure 5.2: Top view of the UCA deployed for direction finding.

5.3 The ML-EMSGA DOA Estimation Technique For UCAs

A difficult problem in the DOA estimation is for fully correlated signals which are also referred to as the coherent signals. Many popular techniques such as MUSIC [57], minimum variance [89] and minimum norm [88] fail to deal with coherent signals. In case of ULAs, a technique named of spatial smoothing [85], was used to circumvent this difficulty. For the case of UCAs, Mati Wax introduced a preprocess technique [86], which is based on the transformation of the actual array into a virtual array that is amenable to spatial smoothing. These methods are all proposed to circumvent the difficulty of dealing with coherent signals in DOA estimation by using some low-complexity techniques. The maximum likelihood (ML) DOA estimation is a more complex technique compare with MUSIC, minimum variance and minimum norm. The method is to apply the ML principle to the statistics of the observed raw data. The ML method is of practical interest due to an optimal performance compared to other methods, particularly when the SNR is small, the number of samples are small, or the sources are correlated [94]. The estimates of ML method are nearly equal to their true values. Therefore, it may be used as a standard to compare with the other methods. Normally it needs to make an assumption first that the number of signal sources are already known. However, this method is computationally intensive and requires iterative schemes for solutions. Several researchers have proposed various schemes to optimize the log-likelihood function to increase probability of global convergence and computation efficiency. Unfortunately, most of them can not guarantee global convergence in general cases, or are only

applicable to ULAs [98].

Consider an array composed of N identical sensors located on an UCA of radius a . Assume that M narrow band sources, centered around a known wavelength, λ_0 , impinge on the array from M distinct directions $\phi_1, \dots, \phi_M$ asynchronously, with the reference point is set at the center of the array. For the k -th signal source, ϕ_k is measured with respect to the line connecting the reference point with the first array element. The θ_k is the elevation angle measured with respect to the Z axis. The N -dimensional array steering vector $\mathbf{s}(\phi_k, \theta_k)$, $k = 1, \dots, M$, is an array response to a narrow-band signal of wavelength λ_0 arriving from angle (ϕ_k, θ_k) . Then $\mathbf{s}(\phi_k, \theta_k)$ is given by

$$\mathbf{s}(\phi_k, \theta_k) = \begin{bmatrix} e^{-j2\pi \frac{r}{\lambda_0} \sin(\theta_k) \cos(\phi_k)} \\ e^{-j2\pi \frac{r}{\lambda_0} \sin(\theta_k) \cos(\phi_k - \frac{2\pi}{N})} \\ \vdots \\ e^{-j2\pi \frac{r}{\lambda_0} \sin(\theta_k) \cos(\phi_k - \frac{2\pi(N-1)}{N})} \end{bmatrix} \quad (5.1)$$

From this steering vector, we can see that the information of elevation angles for signals are already included. In matrix notation, the $N \times 1$ voltage vector $\mathbf{y}$ received by the array can be expressed by

$$\mathbf{y} = \mathbf{S}\mathbf{x} + \mathbf{n} \quad (5.2)$$

where $\mathbf{n}$ denotes the $N \times 1$ additive noise vector of the array, $\mathbf{x} = [x_1, x_2, \dots, x_m]$ is a complex valued signal amplitude vector. $\mathbf{S}$ is the steering matrix of the array, where $\mathbf{S} = [\mathbf{s}(\phi_1, \theta_1), \mathbf{s}(\phi_2, \theta_2), \dots, \mathbf{s}(\phi_M, \theta_M)]$.

Assuming the number of sources M is 2, and signal and noise are statistically independent. Noise is zero mean gaussian distribution with variance at σ^2 . In direction finding applications, $\mathbf{y}$ represents the voltage observations at all antenna elements in the array. The parameter estimator is concerned with estimating ϕ and θ from $\mathbf{y}$. The estimates of ML method are those that maximize the joint probability density of all the data observations vector $\mathbf{y}$ derived as a function of the unknown parameters.

Let $R_{\mathbf{y}}$ is the obtained covariance matrix of measurable observations,

$$\begin{aligned}
 R_{\mathbf{y}} &= E[\mathbf{y}\mathbf{y}^*] \\
 &= E[(\mathbf{S}\mathbf{x} + \mathbf{n})(\mathbf{S}\mathbf{x} + \mathbf{n})^*] \\
 &= E[\mathbf{S}\mathbf{x}\mathbf{x}^*\mathbf{S}^*] + E[\mathbf{n}\mathbf{n}^*] \\
 &= \mathbf{S}\mathbf{A}\mathbf{S}^* + \sigma^2\mathbf{I}
 \end{aligned} \tag{5.3}$$

where $*$ and $E[\]$ denote matrix conjugate transpose and statistical expectation respectively. $\mathbf{A} = E[\mathbf{x}\mathbf{x}^*]$ is a $M \times M$ matrix that represents the covariance matrix of the source signals, where $\mathbf{A}$ is a positive definite. σ^2 is the noise variance at each sensor.

The ML problem can be considered as solving the following maximization function [101],

$$F_{ML}(\phi, \theta) = \text{tr}[\mathbf{I} - \mathbf{S}(\mathbf{S}^* \mathbf{S})^{-1} \mathbf{S}^*] \hat{R}_{\mathbf{y}} \tag{5.4}$$

where,

$$\hat{R}_{\mathbf{y}} = \frac{1}{N} \sum_{t=1}^N \mathbf{y}(t)\mathbf{y}(t)^* \tag{5.5}$$

For one snapshot of observations, $\hat{R}_{\mathbf{y}} = R_{\mathbf{y}}$. The problem in maximizing $F_{ML}(\phi, \theta)$ is severe. It is a multidimensional maximization problem, and searching function may not convergent, resulting in intense computation load. In this research work, a modified and refined GA, named Emperor Selective Genetic Algorithms (EMS-GA) [109]- [111] is used to find the optimal multi-parameters of the ML fitness function, this efforts result in fast convergence to the global optimum, with much lower computational load than MLM.

5.4 Computer Simulation Results

To demonstrate the new method for the compensation of mutual coupling effect of the UCA, computer simulations are carried out with an array of 8 monopole antenna elements. The signal environment consists of two signals. The measured voltages on the antenna terminals are obtained by the moment method. These voltages are then compensated for the mutual coupling effect by using the open-circuit voltage method [67] and the method in [74] as described previously. The compensated voltages (by using two compensation method) are then input to the ML-EMSGA to detect the azimuth and elevation angles of the signals of interest. The performance of the new compensation method which is conjunction with ML-EMSGA in UCAs are studied. The following examples have been studied in different scenarios to demonstrate the effectiveness of the new compensation method for UCAs.

5.4.1 Two Signals From Horizontal Direction

In the first example, the azimuth angles of two signals are at 60° and 65° , with elevation angle at 90° . The two incoming signals are coherent sinusoidal plane waves and vertically polarized. The simulation results are shown in table 5.2. The comparison are made in this example between the results which are obtained by using the measured voltages across the antenna terminal loads (uncompensated voltages) and the corrected voltages which are corrected from the measured voltages by using the new method and open-circuit voltage method. The difference between the estimates and the real value is denoted as “Error”. Assuming, “ α ” is the error in the estimation of azimuth angle for the signal k , and “ β ” the error in the estimation of elevation angle for the signal k , then the standard deviation of error (STD) in the tables listed below is thus equal to “ $|\sqrt{\alpha^2 + \beta^2}|$ ”. In table 5.3, the azimuth angle between two signals is extended from 5° to 15° . In table 5.4, the separation of two signals is extended to 20° in azimuth angle and elevation angles are same at 90° . For the larger separation of two signals, we found that the detection results by using new compensated method is much better than that obtained by using the open circuit voltage method. The uncompensated voltage can hardly be used to detect the directions of arrival of signals.

In Fig.5.3, the solid line is the error standard deviation of DOA estimates obtained from the uncompensated voltage on the array elements. The dashed line and the dash dot line are the error standard deviation obtained from the compensated voltage by using the open circuit voltage method and new compensation method, respectively. It can be easily found that the standard deviation of error obtained from the new

Table 5.2: The ML-EMSGA detection results by using different compensation methods. Signal 1 ($\phi = 60^\circ$, $\theta = 90^\circ$), signal 2 ($\phi = 65^\circ$, $\theta = 90^\circ$), and $d = 0.5\lambda$

Compensation method		Signal 1		Signal 2	
		ϕ	θ	ϕ	θ
New Method	Detection	60.06	90.32	64.70	90.06
	Error	0.06	0.32	-0.30	0.06
	STD	0.32		0.31	
Open-circuit voltage method	Detection	53.01	101.06	69.04	94.05
	Error	-6.99	11.06	4.04	4.05
	STD	13.08		5.72	
Uncompensated voltage	Detection	60.90	91.48	58.84	123.16
	Error	0.90	1.48	-6.16	33.16
	STD	1.73		33.72	

Table 5.3: The ML-EMSGA detection results by using different compensation methods. Signal 1 ($\phi = 60^\circ$, $\theta = 90^\circ$), signal 2 ($\phi = 75^\circ$, $\theta = 90^\circ$), and $d = 0.5\lambda$

Compensation method		Signal 1		Signal 2	
		ϕ	θ	ϕ	θ
New Method	Detection	59.69	90.01	74.89	90.90
	Error	-0.31	0.01	-0.11	0.90
	STD	0.31		0.90	
Open-circuit voltage method	Detection	58.93	97.24	78.09	96.88
	Error	-1.07	7.24	3.09	6.88
	STD	7.31		7.54	
Uncompensated voltage	Detection	73.51	92.77	75.03	98.51
	Error	13.51	2.77	0.03	8.51
	STD	13.79		8.51	

Table 5.4: The ML-EMSGA detection results by using different compensation methods. Signal 1 ($\phi = 65^\circ$, $\theta = 90^\circ$), signal 2 ($\phi = 80^\circ$, $\theta = 90^\circ$), and $d = 0.5\lambda$

Compensation method		Signal 1		Signal 2	
		ϕ	θ	ϕ	θ
New Method	Detection	60.22	91.81	79.82	90.96
	Error	0.22	1.81	-0.18	0.96
	STD	1.83		0.98	
Open-circuit voltage method	Detection	59.89	99.78	82.53	95.69
	Error	-0.11	9.78	2.53	5.69
	STD	9.78		2.63	
Uncompensated voltage	Detection	63.01	109.23	76.38	90.53
	Error	3.01	19.23	-3.62	0.53
	STD	19.46		3.66

Figure 5.3: Standard deviation of DOA estimates with the signals coming from horizontal direction

compensation method is much smaller than that from open circuit voltage method.

5.4.2 Two Signals From Non-Horizontal Directions

In the second example, we consider signals coming from non-horizontal directions. The azimuth angles for two signals are 60° and 65° , with elevation angles at 60° . The signals are all linearly polarized to match with the monopoles. The direction finding results for this example are shown in table 5.5. Even under this condition, our new method still gives a much better result than the uncompensated voltages and the open circuit voltage method. Our method has no significant effect on the assumption of same current distribution from horizontal direction in calculation of the mutual impedances. The results obtained by using the new method are much more accurate than those obtained by using the uncompensated voltages and the open circuit voltage method. Next, we keep the two signals with same elevation at 60° , and change the azimuth

Table 5.5: The ML-EMSGA detection results by using different compensation methods. Signal 1 ($\phi = 60^\circ$, $\theta = 60^\circ$), signal 2 ($\phi = 65^\circ$, $\theta = 60^\circ$), and $d = 0.5\lambda$

Compensation method		Signal 1		Signal 2	
		ϕ	θ	ϕ	θ
New Method	Detection	60.27	60.10	67.08	59.69
	Error	0.27	0.10	2.08	-0.31
	STD	0.29		2.10	
Open-circuit voltage method	Detection	53.36	65.82	73.74	52.64
	Error	-6.64	5.82	8.74	-7.36
	STD	8.83		11.43	
Uncompensated voltage	Detection	58.38	69.72	81.31	61.02
	Error	-1.62	9.72	16.31	1.02
	STD	9.85		16.34	

Table 5.6: The ML-EMSGA detection results by using different compensation methods. Signal 1 ($\phi = 60^\circ$, $\theta = 60^\circ$), signal 2 ($\phi = 75^\circ$, $\theta = 60^\circ$), and $d = 0.5\lambda$

Compensation method		Signal 1		Signal 2	
		ϕ	θ	ϕ	θ
New Method	Detection	58.77	59.36	74.46	59.79
	Error	-1.23	-0.64	-0.54	-0.21
	STD	1.39		0.58	
Open-circuit voltage method	Detection	60.77	60.59	80.75	59.21
	Error	0.77	0.59	5.75	-0.79
	STD	0.97		5.81	
Uncompensated voltage	Detection	68.54	60.65	68.66	49.01
	Error	8.54	-0.65	-6.34	-10.99
	STD	8.56		12.69	

angle of one signal to 75° and 85° , another one is still at 60° . The direction finding results for these two examples are shown in table 5.6, and table 5.7 respectively.

We also plot a figure 5.4 to show the error standard deviation of DOA estimates from this example. In this figure, we can easily see that the error standard deviation obtained from new compensation method is smallest.

Table 5.7: The ML-EMSGA detection results by using different compensation methods. Signal 1 ($\phi = 60^\circ$, $\theta = 60^\circ$), signal 2 ($\phi = 85^\circ$, $\theta = 60^\circ$), and $d = 0.5\lambda$

Compensation method		Signal 1		Signal 2	
		ϕ	θ	ϕ	θ
New Method	Detection	60.03	59.24	84.49	60.57
	Error	0.03	-0.76	-0.51	0.57
	STD	0.76		0.76	
Open-circuit voltage method	Detection	59.95	60.98	86.22	60.95
	Error	-0.05	0.98	1.22	0.95
	STD	0.98		1.54	
Uncompensated voltage	Detection	57.20	56.38	87.78	55.35
	Error	-2.80	-3.62	2.78	-4.65
	STD	4.58		5.42

Figure 5.4: Standard deviation of DOA estimates with the signals coming from non-horizontal direction

5.5 Conclusions

The new method for compensating the mutual coupling effect in UCAs employed for direction finding is studied together with a new 2D DOA search algorithm. The advantages of combining these two methods together for UCAs have been demonstrated in 2D DOA estimations of coherent signals. The accuracy of DOA estimation is increased in comparison with the traditional compensation method. Simulation results show the robustness and the capability of the new method with EMS-GA in UCA employed for direction finding in the two signal environments.

Chapter 6

Simultaneous Estimation of MCM and DOAs Using Structured LS

A structured Least Square (LS) method for simultaneous estimation of the mutual coupling matrix (MCM) and DOA of signal source is proposed in this chapter. An analytical model of the mutual coupling effect is presented, and its effects are in the form of a complex Toeplitz matrix. The MCM is dependent on the array geometry and its electromagnetic characteristics. The DOAs and MCM can be simultaneously estimated by using a newly proposed method when the observations of at least two different DOAs are available. This method is especially useful for the calibration of ULAs and UCAs. It significantly simplifies the procedure of the calibration of the mutual coupling effect. Simulation results also confirm the efficiency of the proposed method.

Simulations also show the strong dependency of the convergence behavior on the

signal to noise ratio. It have been found that at very low signal-to-noise ratios (SNR), more iteration rounds are needed, e.g. at a SNR value of $0dB$, after more than 50 iterations, the optimal estimates can be obtained.

6.1 Introduction

Using multiple antennas in a base station may be able to exploit the fact that signals arriving from different mobiles follow different paths and arrive at different times. This allows independent measurements of signals superimposed from different mobiles. This, along with the properties of the modulation technique used, allows separation of signals arriving from different mobiles. Thus, by using the measured signals at various elements of the array at the base station, one is able simultaneously to separate all signals. The process is referred to as the blind estimation of co-channel signals. It does not require knowledge of the directions or other parameters associated with mobiles, such as a reference signal or a training sequence, but exploits the temporal structure that might exist in signals inherited from the source of their generation, for example, the modulation techniques used. Many studies on this topic have been reported in the literatures [38]- [40].

Adaptive array signal processing has been greatly advanced, and applied to many fields such as radar, sonar, wireless mobile communication, and so on. The performance of an adaptive antenna array was strongly affected by the electromagnetic characteristics of the antenna array [113]. It was shown that the effects of the mutual coupling between the antenna elements are significant even for large inter-element

spacings (spacing is more than half a wavelength). The effects became more drastic as the inter-element spacing dropped below half a wavelength [67]. The performance of many conventional array signal processing methods degraded significantly when ignoring of the effects of the mutual coupling. Since the effects of the mutual coupling is a crucial problem in antenna array processing, especially in DOA estimation. Some methods [70] - [73] were proposed to eliminate the effects of the mutual coupling in DOA estimation. However, most of them were not so practical in real applications. In [70], the degradation of the performance was observed in the direct data domain algorithms. The method of moment (MOM) is used to analyze and calculate the mutual impedance matrix in antenna array. In [71], the effects of the mutual coupling were calculated by using the open circuit voltages method in adaptive array. In [73], the authors applied a new and very accurate method to calculate the mutual impedance matrix. The new calculation method was based on an estimated current distribution, which carried a direction reference of the incoming signal. It was shown this new calculation method can significantly reduce the mutual coupling effect and thus lead to an more accurate DOA estimation of the DOA finding algorithms. However, in above mentioned methods, the calculation of the mutual coupling matrix are very computation-consuming and time-consuming, because of big computation load for calculating of electromagnetic field of antenna arrays. In this chapter, we propose a new calibration method to simplify the calculation of mutual coupling matrix in DOA estimation applications. This method is quite straightforward to understand and implement. The main advantage of the newly proposed method is that the MCM and DOAs can be simultaneously estimated, if there are at least two testing signals are available. This method is very attractive in

practical applications because of its efficiency and simplicity.

The structure of mutual coupling matrix can be transformed into the product of two matrices. One matrix is a diagonal matrix in which the diagonal elements are the gain of corresponding antenna, another is a matrix whose elements are the mutual coupling coefficient between array elements. In the case of absence of gain and phase mismatching between the array elements, we assume that identical load impedances are terminated at the output of each elements. It has already been known that for some arrays with special geometries, their MCM have special forms. For example, the mutual coupling matrix of ULA is a symmetric Toeplitz matrix, and MCM of UCA is an circular matrix [71]. In order to greatly simplify the estimation procedure, this special property of MCM for ULAs and UCAs can be well exploited in the calculation process. The gain matrix of each antenna is easy to be measured. In this chapter, it is assumed that the gain matrix of the antennas is known.

6.2 Proposed Method for ULA

Consider an array composed of M equally spaced elements located in an uniform linear array. Assume that one narrow band source impinging on the array from θ direction.

It can be proved that for ULA, the MCM denoted by $\mathbf{A}$ is a symmetric toeplitz

matrix [77], which can be represented as

$$\mathbf{A} = \begin{bmatrix} a_1 & a_2 & \cdots & a_N \\ a_2 & a_1 & \ddots & \vdots \\ \vdots & \ddots & \ddots & a_2 \\ a_N & \cdots & a_2 & a_1 \end{bmatrix} = \sum_{i=1}^N a_i \mathbf{E}_i \quad (6.1)$$

where $\mathbf{E}_i$ is a symmetric Toeplitz matrix. Its first row vector is $[0 \cdots 1 \ 0 \cdots 0]$, where the i th element is one, others are all zero.

The signal received by the array is $\mathbf{y}(n)$,

$$\mathbf{y}(n) = x(n)\mathbf{A}\mathbf{s}(\theta) + \mathbf{w}(n) \quad (6.2)$$

where $\mathbf{w}(n)$ denotes the noise vector at each of the loaded antenna element, and $x(n)$ denotes the incident signal on the elements of the array from the direction θ . The $\mathbf{s}(\theta)$ denotes the ideal field pattern of the array (also called the array steering vector) toward direction θ . By using a matrix representation, $\mathbf{s}(\theta)$ becomes,

$$\mathbf{s}(\theta) = \left[1 \ e^{-j\frac{\pi}{2}\sin(\theta)} \ \cdots \ e^{-j\frac{(M-1)\pi}{2}\sin(\theta)} \right]^T \quad (6.3)$$

In the controlled environment of calibration phase, the observed array steering vector (ASV) $\bar{\mathbf{s}}(\theta)$ is easy to be measured using the antenna output voltages. For L different calibration sources, we have L observed ASVs. For each of DOA θ_i ($i =$

$1, \dots, L$), we can obtain the estimated ASV,

$$\bar{\mathbf{s}}(\theta_i) = \mathbf{A}\mathbf{s}(\theta_i) + \mathbf{e}_i, \quad i = 1, \dots, L \quad (6.4)$$

where $\mathbf{e}_i$ is the measurement error.

If the exact θ_i in the experiment are already known, with sufficient number of L experiments, the mutual coupling matrix $\mathbf{A}$ can be estimated by using LS method to solve the following optimization problem,

$$\hat{\mathbf{A}} = \arg \min_{\mathbf{A}} \sum_{i=1}^L \|\bar{\mathbf{s}}(\theta_i) - \mathbf{A}\mathbf{s}(\theta_i)\|_2^2 \quad (6.5)$$

Since there are M unknown parameters in MCM, and with one more experiment performed, one more unknown parameter θ_i will be included in optimization problem. Therefore, if L experiments performed, there are $M + L$ unknown parameters in equation (6.5). For one experiment performed, there are M linear equations, same as (6.3), which can be obtained. In case of L experiments carried out, there are ML linear equations available. Therefore, if $ML > (L + M)$, it is possible to determine the unknown parameters uniquely. That means, If $L \geq M/(M - 1)$ experiments are carried out, we can obtain L independent steering vector $\mathbf{s}(\theta_i)$, then the estimate of mutual coupling matrix $\mathbf{A}$ can be uniquely determined. From the above description, we know that the smallest number for L is 2.

However, for some practical experiments, the direction of signal source θ_i can not be accurately known, or totally unknown. In such cases, the estimates of MCM have

errors as well as the estimated DOAs. In this chapter, we estimate DOAs and MCM simultaneously by using an efficient method.

The problem can be stated as followings. Let us consider a uniformly spaced linear array with the mutual coupling effect among array elements, the mutual coupling matrix $\mathbf{A}$ and direction of arrival of signal sources $\{\theta_i\}$ are all unknown. The following optimization method is used to find the estimates of MCM and DOAs.

$$\begin{aligned} J(\mathbf{A}, \theta_i) &= \sum_{i=1}^L \|\bar{\mathbf{s}}_i - \mathbf{A}\mathbf{s}_i\|_2^2 \\ \{\hat{\theta}_i, \hat{\mathbf{A}}\} &= \arg \min_{\mathbf{A}, \theta_i} J(\mathbf{A}, \theta_i) \end{aligned} \tag{6.6}$$

It can be proved that for fixed DOAs, the optimization problem (6.6) to $\mathbf{A}$ is a quadratic problem. The closed form of the solution is easy to be derived. However, the optimization problem to θ_i is nonlinear and not quadratic, so it is difficult to get its closed form solution. To simplify the optimization problem, we proposed to solve the optimization problem in (6.6) using two steps. Firstly, Let us assume that θ_i is known, with this assumption, we can obtain mutual coupling matrix $\mathbf{A}$. By substituting estimated $\mathbf{A}$ into (6.6), this optimization problem is simplified and only have unknown parameters θ_i . Secondly, if very accurate DOAs are estimated, this estimated value can be used to iteratively estimate the MCM until optimization algorithm is convergent.

Substituting (6.1) into (6.6), we have

$$\begin{aligned}
 J(\mathbf{A}, \theta_i) &= \sum_{i=1}^L \|\bar{\mathbf{s}}_i - \mathbf{A}\mathbf{s}_i\|_2^2 \\
 &= \sum_{i=1}^L \bar{\mathbf{s}}_i^H \bar{\mathbf{s}}_i - \bar{\mathbf{s}}_i^H \mathbf{A}\mathbf{s}_i - \mathbf{s}_i^H \mathbf{A}^H \bar{\mathbf{s}}_i + \mathbf{s}_i^H \mathbf{A}^H \mathbf{A}\mathbf{s}_i \\
 &= \sum_{i=1}^L \bar{\mathbf{s}}_i^H \bar{\mathbf{s}}_i - \sum_{i=1}^L \sum_{k=1}^N a_k \bar{\mathbf{s}}_i^H \mathbf{E}_k \mathbf{s}_i - \sum_{i=1}^L \sum_{k=1}^N a_k^* \mathbf{s}_i^H \mathbf{E}_k \bar{\mathbf{s}}_i \\
 &\quad + \sum_{i=1}^L \sum_{l=1}^N \sum_{k=1}^N a_k^* a_l \mathbf{s}_i^H \mathbf{E}_k \mathbf{E}_l \mathbf{s}_i
 \end{aligned} \tag{6.7}$$

The optimal estimate of $\mathbf{A}$ is obtained by setting the derivative of J with respect to a_k^* to zero

$$\frac{\partial J(\mathbf{A}, \theta_i)}{\partial a_k^*} = - \sum_{i=1}^L \mathbf{s}_i^H \mathbf{E}_k \bar{\mathbf{s}}_i + \sum_{i=1}^L \sum_{l=1}^N a_l \mathbf{s}_i^H \mathbf{E}_k \mathbf{E}_l \mathbf{s}_i = 0 \tag{6.8}$$

We have the linear equations for each k ,

$$\sum_{l=1}^N a_l \left[\sum_{j=1}^L \mathbf{s}_j^H \mathbf{E}_k \mathbf{E}_l \mathbf{s}_j \right] = \sum_{j=1}^L \mathbf{s}_j^H \mathbf{E}_k \bar{\mathbf{s}}_j, \quad k = 1, \dots, N \tag{6.9}$$

With estimated MCM $\hat{\mathbf{A}}$, the DOAs of sources can be estimated using following optimization problem,

$$\hat{\theta}_i = \arg \min_{\theta_i} \sum_{i=1}^L \|\bar{\mathbf{s}}_i - \hat{\mathbf{A}}\mathbf{s}(\theta_i)\|^2 \tag{6.10}$$

It is achieved through using of this iterative optimization method. The followings is a step-by-step description on how to obtain the estimated DOAs in order to obtain the

estimated MCM iteratively.

Step1 Using the observed ASVs to estimate the DOAs, and considering these DOAs as initial DOAs θ_i^0 .

Step2 Using the estimated DOAs $\theta_i^{(m-1)}$ ($m - 1$ th iteration) to construct the ASV $s(\theta_i^{(m-1)})$ at m th iteration.

Step3 Using (6.9) to estimate $\hat{\mathbf{A}}^m$.

Step4 Using (6.10) to estimate θ_i^m with the estimated $\hat{\mathbf{A}}^m$.

Step5 If the algorithm converges, then stop. Otherwise, go to Step 2 to continue.

6.3 Two Techniques for UCA

Here we can solve the same problem in circular array by using two approaches. First, Let's consider an approach which is a commonly used one.

6.3.1 Method I

Consider an array composed of N sensors located on an UCA of radius r . It is assumed that L signal sources are available, which are centered around a known wavelength, say λ_0 , impinge on the array from L distinct directions $\theta_1, \dots, \theta_L$ asynchronously. The reference point is set at the center of the array and θ is measured with respect to the line connecting the reference point with the first array element. The n -th component

of the N -dimensional array ideal steering vector $\mathbf{s}(\theta)$ ($n = 1, \dots, N$), to a narrow-band signal of wavelength λ_0 arriving from angle θ , $\theta \in [-\pi, \pi]$, is given by

$$s_n(\theta) = G_n(\theta) \exp[j \frac{2\pi r}{\lambda_0} \cos(\theta - \frac{2\pi(n-1)}{N})] \quad (6.11)$$

where $G_n(\theta)$ is the complex gain pattern of the n -th element. Suppose the array elements are all identical and isotropic, then $G_n(\theta) = 1$, $n = 1, \dots, N$. The steering vector of this array is then given by

$$S(\theta) = [e^{j2\pi r \cos \theta / \lambda_0}, e^{j2\pi r \cos(\theta - \frac{2\pi}{N}) / \lambda_0}, \dots, e^{j2\pi r \cos(\theta - \frac{2\pi(N-1)}{N}) / \lambda_0}]^T \quad (6.12)$$

From [87], we obtain that the effect of mutual coupling between the array elements is to transform the ideal steering vector $\mathbf{S}(\theta)$ to a different steering vector $\mathbf{AS}(\theta)$, where $\mathbf{A}$ is the matrix of mutual coupling. In matrix notation, the $N \times 1$ vector of received signal by the array can be expressed as followings

$$\mathbf{Y} = \mathbf{XAS}(\theta) + \mathbf{W} \quad (6.13)$$

$\mathbf{W}$ denotes the $N \times 1$ noise vector of the array, $\mathbf{X}$ is the k -th signal source which is received at the array center. The $\mathbf{S}(\theta)$ is the steering vector of the array toward direction of k -th signal source.

As we already known, for the antenna array with circular geometry, from the rotational symmetry, it follows that the matrix $\mathbf{A}$ is circular matrix. Thus, the matrix of

mutual coupling $\mathbf{A}$ is given as below,

$$\mathbf{A} = \begin{bmatrix} a_0 & a_{-1} & \cdots & a_{2-N} & a_{1-N} \\ a_1 & a_0 & a_{-1} & \cdots & a_{2-N} \\ \vdots & \ddots & \ddots & \ddots & \vdots \\ a_{N-2} & \ddots & \ddots & \ddots & \vdots \\ a_{N-1} & a_{N-2} & \cdots & a_1 & a_0 \end{bmatrix} \quad (6.14)$$

where $a_{-k} = a_{N-k}$ for $1 \leq k \leq N-1$. The mutual coupling matrix $\mathbf{A}$, which can also be represented as followings:

$$\mathbf{A} = \sum_{i=0}^{N-1} a_i \mathbf{H}_i \quad (6.15)$$

where $\mathbf{H}_i$ is the i -th power of cyclic permutation matrix. $\mathbf{H}$ is given as below,

$$\mathbf{H} = \begin{bmatrix} 0 & 1 & 0 & \cdots & 0 \\ 0 & 0 & 1 & \cdots & 0 \\ \vdots & \vdots & \vdots & \vdots & \vdots \\ 0 & 0 & 0 & \cdots & 1 \\ 1 & 0 & 0 & \cdots & 0 \end{bmatrix} \quad (6.16)$$

Same as in linear array, we can easily obtain the observed array steering vector $\bar{\mathbf{s}}(\theta)$ using the antenna output voltages. For L different calibration signal sources, we have L observed array steering vectors. For each of DOAs, θ_i ($i = 1, \dots, L$), we can obtain

the estimated ASVs as followings,

$$\bar{\mathbf{s}}(\theta_i) = \mathbf{A}\mathbf{s}(\theta_i) + \mathbf{e}_i, \quad i = 1, \dots, L \quad (6.17)$$

where $\mathbf{e}_i$ is the measurement error. We assume that exact θ_i are already known, with sufficient number of L experiments, the MCM can be estimated by using below optimization equation:

$$\hat{\mathbf{A}} = \arg \min_{\mathbf{A}} \sum_{i=1}^L \|\bar{\mathbf{s}}(\theta_i) - \mathbf{A}\mathbf{s}(\theta_i)\|_2^2 \quad (6.18)$$

6.3.2 Method II

In [90], Davies proposes a method to transform the outputs at the antenna elements of an UCA to a virtual array. The steering vector of the virtual array is Vandermonde matrix, or approximately so. Following the approach outlined above, we shall transform the N dimensional circular array to a M dimensional virtual array, $M < N$. In [93], the transformation matrix T is defined by

$$T = JF \quad (6.19)$$

where J is $M \times M$ diagonal matrix, for $m = 1, \dots, M$,

$$J = \text{diag} \left\{ \frac{1}{j^{m-1-\frac{(M-1)}{2}} \sqrt{N} J_{m-1-\frac{(M-1)}{2}} \left(\frac{2\pi r}{\lambda} \right)} \right\} \quad (6.20)$$

and F is $M \times N$ matrix,

$$F = \frac{1}{\sqrt{N}} \begin{bmatrix} 1 & e^{-j2\pi 1 \cdot (M-1)/2N} & \dots & e^{-j2\pi (N-1) \cdot (M-1)/2N} \\ \vdots & \vdots & \vdots & \vdots \\ 1 & e^{-j2\pi 1 \cdot 2/2N} & \dots & e^{-j2\pi (N-1) \cdot 2/2N} \\ 1 & 1 & \dots & 1 \\ 1 & e^{j2\pi 1 \cdot 2/2N} & \dots & e^{j2\pi (N-1) \cdot 2/2N} \\ \vdots & \vdots & \vdots & \vdots \\ 1 & e^{j2\pi 1 \cdot (M-1)/2N} & \dots & e^{j2\pi (N-1) \cdot (M-1)/2N} \end{bmatrix} \quad (6.21)$$

Let $\widetilde{S}(\theta)$ denote the steering vector of the virtual array,

$$\widetilde{S}(\theta) \cong TS(\theta) \cong JFS(\theta) \quad (6.22)$$

so, $\widetilde{S}(\theta)$ is the M dimensional vector,

$$\widetilde{S}(\theta) = [e^{-j\theta(M-1)/2}, \dots, e^{-j\theta}, 1, e^{j\theta}, \dots, e^{j\theta(M-1)/2}]^T \quad (6.23)$$

Thus we can use $\widetilde{S}(\theta)$ to take the place of $S(\theta)$ in (6.4). By using this preprocess, we can solve the minimization problem as in equation(6.5).

6.4 Numerical Study

In this section, numerical examples are presented to demonstrate the proposed approach. Firstly, we consider an ULA consisting of 8 monopoles. The array is aligned along the X axis with the monopole elements parallel to the Z axis. The dimensions of the monopole elements are: length= 3.0 cm and wire radius= 0.3 mm. They are placed over a large ground plane and connected to a $50\ \Omega$ load and inter-element spacing is $d = 6.25$ cm (half wavelength at 2.4 GHz). The incoming signal is sinusoidal plane wave with vertical polarization. The signal amplitude = 1 V, frequency = 2.4 GHz, and additive thermal noise about -20 dB. In the simulation, the MCM is constructed as a Toeplitz matrix which is obtained from simulations in our previous work [73]. Assuming that we have two observations available, the true DOAs are from azimuth angle of 30° and 40° , respectively. In Fig. 6.1, the DOA estimates at each iteration is shown. Moreover, the estimated MCM elements are shown in Fig. 6.2. The plots are shown that the convergent speed of the estimation algorithm is very fast. After 140 iterations, the algorithm is converged at the optimum point. The estimation errors of DOAs and MCM is very small.

In second example, 4 signals from 5° , 15° , 25° and 35° are impinging on an ULA. The dimension of the ULA is same as the first example in this chapter.

In this example, the plots are shown that the convergent speed of the estimation algorithm is faster compared with the first example. The algorithm is converged at the optimum point after 50 iterations. The converged estimation errors of DOAs and MCM is very small. Compared with the simulation results from first example which are using

Figure 6.1: DOA estimates at each iteration.

Figure 6.2: MCM elements estimated after 150 iterations.

Figure 6.3: DOA estimates at each iteration.

Figure 6.4: MCM elements estimated after 50 iterations.

two observation signals, it was found that with more observation signals available, the algorithm converges much faster.

In third example, 3 signals from 30° , 45° , and 60° are impinging on an UCA. The UCA consisting of 9 monopoles. Array elements are oriented in Z -direction, and uniformly placed along the circumference of a circle of radius of 6.25 cm in the XY plane. The monopole elements are of equal length of 3.0 cm and the element radius = 0.3mm. All the monopole elements are loaded with a terminal load of $50\ \Omega$. The incoming signals are sinusoidal plane wave with vertical polarization, amplitude = 1 V, frequency = 2.4 GHz, and additive thermal noise about -20 dB. In the simulation, the MCM is constructed as a circulant matrix which is obtained from simulation results of our previous work in Chapter 5.

Figure 6.5: DOA estimates at each iteration.

The estimated DOAs and MCM elements are shown in Fig. 6.5 and 6.6, respectively. The figures show that the algorithm converged very fast. The converged estimation errors of DOAs is very small. The MCM from iterative optimization algorithm is a circulant matrix, and very close to true value of mutual coupling matrix.

6.5 Concluding Remarks

A structured least square method is proposed to simultaneously estimate the MCM and the DOAs in the presence of mutual coupling and thermal noise in ULAs and UCAs. It was found that the proposed method can estimate the DOAs and MCM simultaneously when at least two observation signals are available. The proposed method is also

Figure 6.6: MCM elements estimated after 50 iterations.

useful in calibration of ULA and UCA when the DOAs of calibration signal sources are not known. By evaluating three examples, it is shown that the proposed iterative optimization method converges very fast and can obtain very accurate estimation of MCM and DOAs.

Chapter 7

Compensation for Mutual Coupling Effect in Adaptive Nulling

This chapter describes an antenna array that is designed and applied to enhance the reception of signal of interest (SOI) which have been corrupted by co-channel and multipath interference. More specifically, a practical dipole antenna array together with the proposed method for compensation of the mutual coupling effect is applied for the efficient adaptive nulling of interferences in a mobile communication environment. The proposed method for compensation of mutual coupling effect is shown to have a greater ability and flexibility than conventional methods. Numerical examples demonstrate the superior performance of the proposed adaptive array and the proposed compensation method.

7.1 Introduction

Exploiting the spatial dimension of wireless channels through the use of multiple antennas at the transmitter and receiver can obtain significant gains either in energy efficiency or spectral efficiency [4].

There are two of the major problems including the multi-path propagation from transmitter to receiver, and the co-channel interference in mobile communications. With a conventional antenna, all echoes of the signal are added up at the receiver with arbitrary phases, leading to fading, and thus an increased error probability. By using smart antennas, there are two ways that can be used to avoid fading: One way is say that it can process all multi-path components separately, thus avoiding fading. Another possibility is to say that a complex antenna pattern can be adaptively generated which can avoid or at least mitigate fading. For the problem of suppression of the co-channel interference, a smart antenna can put nulls in the direction of the interferers to suppress the co-channel interference.

In adaptive nulling, it has been shown that the mutual coupling effect will significantly affect the depths and the accuracy of the positioning of the nulls towards the interferences. Mutual coupling is a general problem for all practical antenna arrays and it affects almost all functions that the array is to perform. This problem has long been addressed before on other occasions [63]- [72]. In this chapter, we applied the new method to compensate for the mutual coupling effect of a linear dipole antenna array employed in adaptive nulling of interferences.

7.2 Theoretical Analysis

Consider a linear dipole antenna array with n dipole antennas as shown in Fig.7.1.

Figure 7.1: The linear dipole antenna array and the coordinate system.

The dipole antenna elements are assumed to be thin wires of equal length $\ell = 0.5\lambda$ and radius $a = 0.3$ mm, where λ is the operating wavelength. The array is placed along the X -axis with an equal spacing $d = 0.5 \lambda$ between the antenna elements. The antenna elements are oriented with their axes parallel to the Z -axis. All the antenna elements are connected at their centers to a terminal load Z_L .

The array is used to receive an incoming signal of interest (SOI) in the presence of interferences which are assumed to come from different angles from that of the SOI. The SOI and all the interferences are assumed to be plane electromagnetic waves propagating in a 3D space. By using the new compensation method and standard

moment-method analysis [112] which are described in previous chapters, we calculate all the mutual impedances in that way, then we can obtain the voltages on the antenna terminals with the mutual coupling effect completely removed.

7.3 Computer Simulations

In this section, we study by using computer simulations. The performance of a linear dipole array with seven antenna elements employed for adaptive nulling by using the new compensation method. The dimensions of the array are as below. The number of array elements is $n = 7$, element separation $d = 0.5 \lambda$, element length $l = 0.5 \lambda$, element radius $a = \lambda/200$, and terminal load $Z_L = 50\Omega$. The array is deployed in the manner as shown in Fig.7.1. We will compare the improvements obtained with the mutual coupling effect being compensated for by using the new method and the open-circuit voltage method. The adaptive nulling algorithm used in the examples is the direct data domain adaptive algorithm [92] which is simple and easy to implement. The detailed description of this algorithm can be found in [91], [92]. The required inputs to the direct data domain adaptive algorithm are the voltages on the antenna terminal loads and the DOA of the SOI. The measured voltages on the antenna terminal loads are first compensated for the mutual coupling effect and then passed to the adaptive algorithm. The following examples have been studied.

7.3.1 SOI and Interferences Coming from the Horizontal Direction

The signal environment for the first example is shown in Table 7.1.

Table 7.1: The signal environment for the first example.

	Amplitude (V/m)	Azimuth ϕ	Elevation θ
Signal of interest	$1.0\hat{z}$	45°	90°
Interference #1	$1.0\hat{z}$	75°	90°
Interference #2	$1.5\hat{z}$	60°	90°
Interference #3	$2.0\hat{z}$	30°	90°

The SOI and interferences are plane waves linearly polarized in the Z direction. This example has been studied before in [70]. From our study, the result of this example is shown in Fig. 7.2. Adaptive radiation patterns obtained by using four types of voltages have been shown in the Figure.

The dash-line radiation pattern is the result obtained by using the measured voltages across the antenna terminal loads which have not been compensated for the mutual coupling effect. The dotted-line radiation pattern is the result obtained by using the open-circuit voltages derived by the method in [67]. The solid-line radiation pattern is the result obtained by using voltages corrected from the measured voltages by using the new method. The small graph inside Fig.7.2 shows the radiation pattern obtained by using ideal voltages across the terminal loads which are completely free from any mutual coupling effect. Both the ideal voltages and the measured voltages are calculated by the moment method [112] with the current distribution on each antenna being

Figure 7.2: The adaptive radiation patterns obtained by using the measured voltages, the voltages corrected by using open-circuit-voltages method, the voltages corrected by the new method, and the ideal voltages. The signal environment is shown in Table 7.1.

expanded by 20 sinusoidal basis functions and using the Galerkin matching procedure. From Fig.7.2, we see that the (un-compensated) measured voltages can hardly be used to suppress the interferences. The open-circuit voltages can be used to locate the three interferences but the dips are not deep enough. However, by using the compensated (corrected) voltages obtained by the new method, the dips generated at the interference directions are substantially deeper. For the ideal case, the dips are much deeper than all the other cases because this is a theoretically ideal situation in which the antenna elements are completely isolated from each other.

7.3.2 Interferences Coming from Different Elevation Angles

In the second example, we consider interferences coming from different elevation angles with signal environment shown in Table 7.2.

Table 7.2: The signal environment for the second example.

	Amplitude (V/m)	Azimuth ϕ	Elevation θ
Signal of interest	$1.0\hat{z}$	45°	90°
Interference #1	$0.0449\hat{x} + 0.1677\hat{y} + 0.9848\hat{z}$	75°	100°
Interference #2	$1.5(-0.0868\hat{x} - 0.1504\hat{y} + 0.9848\hat{z})$	60°	80°
Interference #3	$2.0(-0.2962\hat{x} - 0.1710\hat{y} + 0.9397\hat{z})$	30°	70°

The azimuth angles remain unchanged as in the first example. The SOI and the interferences are all linearly polarized on the planes perpendicular to their directions of propagation. The adaptive radiation patterns are shown in Fig.7.3.

We notice two differences when compared with the result in the first example. First, all the dips are shallower whether by using measured or corrected voltages. Secondly, the positions of the dips are shifted a small but noticeable amount from the azimuth

Figure 7.3: The adaptive radiation patterns with interferences coming from different elevation angles obtained by using different compensation methods for the mutual coupling effect. The signal environment is shown in Table 7.2.

angles of the interferences. Both of these two phenomena are caused by the elevation angles of the interferences being not all on the horizontal plane ($\theta = 90^\circ$). This results in a propagation constant smaller than the free-space propagation constant in the horizontal direction. That is, the effective wavelengths of the interferences along the horizontal direction are longer than their actual wavelength. However, even under this condition, our new method still gives a much better result than the open-circuit voltage method. The dips obtained by using the new method are much deeper and sharper than those obtained by using the open-circuit voltage method. The reasons that the suppressions are smaller than those in the first example (Fig.7.2) can be explained by the fact that the estimated current distribution is now less accurate to simulate the true (measured) current distributions on the antenna elements which are now excited by incident waves coming from different elevation angles instead of coming from the single horizontal direction.

7.3.3 Interference with Different Strengths

In the third example, we investigate the effect of the strength of one of the interference on the performance of the new method. The signal environment is shown in Table 7.3.

First we study the effect on the adaptive radiation pattern. The absolute value of the amplitude ($|\vec{J}_1|$) of interference #1 at $\phi = 75^\circ$, $\theta = 100^\circ$ is increased from $1V/m$ to $500V/m$ in 4 steps. The result is shown in Fig.7.4.

In this figure, the adaptive radiation patterns with different strengths of interference #1 obtained by using the new method to compensate for the mutual coupling effect is

showed. The signal environment is shown in Table 7.3.

Figure 7.4: The adaptive radiation patterns with different strengths of interference #1 obtained by using the new method to compensate for the mutual coupling effect.

We observe that as the strength of interference #1 is increased, the most affected is the suppression of interference #2 at $\phi = 60^\circ$, $\theta = 85^\circ$. The dip at this position becomes

Table 7.3: The signal environment for the third example.

	Amplitude (V/m)	Azimuth ϕ	Elevation θ
Signal of interest	$1.0\hat{z}$	45°	90°
Interference #1, case(a)	$0.0449\hat{x} + 0.1677\hat{y} + 0.9848\hat{z}$	75°	100°
Interference #1, case(b)	$50(0.0449\hat{x} + 0.1677\hat{y} + 0.9848\hat{z})$	75°	100°
Interference #1, case(c)	$200(0.0449\hat{x} + 0.1677\hat{y} + 0.9848\hat{z})$	75°	100°
Interference #1, case(d)	$500(0.0449\hat{x} + 0.1677\hat{y} + 0.9848\hat{z})$	75°	100°
Interference #2	$1.5(-0.0436\hat{x} - 0.0755\hat{y} + 0.9962\hat{z})$	60°	85°
Interference #3	$2.0(-0.1504\hat{x} - 0.0868\hat{y} + 0.9848\hat{z})$	30°	80°

smaller and smaller and finally disappears. However, the dips at the interference #1 and interference #3 are less affected. The dip at interference #1 becomes deeper while the dip at interference #3 becomes shallower as the strength of interference #1 is increased. Secondly, we obtain the recovered signal strength [92] and the null depth at interference #1 with an increasing strength of interference #1. The results are shown in Fig.7.5.

Figure 7.5: The variations of the recovered signal and the null depth with different strengths of interference #1 obtained with the new method and the open circuit voltage method. The signal environment is same as in Table 7.3 except the continuous varying strength of interference #1.

The signal environment is same as in Table 7.3 except the continuous varying strength of interference #1. We observe that both the recovered signal strength and the null depth obtained by the new method are more stable (less affected) compared with those obtained by the open circuit voltage method.

7.3.4 SOI with Different Intensities at the Elevation Angle of

$$\theta = 120^\circ$$

In the last example (the fourth example), we study the effect of the intensity of the SOI on the performance of the new method. The signal environment is shown in Table 7.4.

Table 7.4: The signal environment for the fourth example.

	Amplitude (V/m)	Azimuth ϕ	Elevation θ
Signal of interest, <i>case(a)</i>	$0.3536\hat{x} + 0.3536\hat{y} + 0.8660\hat{z}$	45°	120°
Signal of interest, <i>case(b)</i>	$50(0.3536\hat{x} + 0.3536\hat{y} + 0.8660\hat{z})$	45°	120°
Interference #1	$0.0449\hat{x} + 0.1677\hat{y} + 0.9848\hat{z}$	75°	100°
Interference #2	$1.5(-0.0436\hat{x} - 0.0755\hat{y} + 0.9962\hat{z})$	60°	85°
Interference #3	$2.0(-0.1504\hat{x} - 0.0868\hat{y} + 0.9848\hat{z})$	30°	80°

The SOI comes from an elevation angle of $\theta = 120^\circ$. Two cases of signal intensities are studied: $|\vec{A}_s| = 1V/m$ and $|\vec{A}_s| = 50V/m$. The result is shown in Fig.7.6.

We see that when the signal intensity is increased by 50 times, the dip at interference #2 becomes shallower (18 *dB* smaller) while the dip at interference #3 becomes deeper (6 *dB* greater). The positions of the dips are also somewhat affected. On the other hand, we found that by using the open-circuit voltage method, a very poor result is obtained even for $|\vec{A}_s| = 1V/m$. Thus it shows that the new method is less sensitive to the change of the elevation angle and the intensity of the SOI than the open-circuit voltage method.

Figure 7.6: The adaptive radiation patterns with different signal intensities obtained by using the new method and the open-circuit voltage method to compensate for the mutual coupling effect. The signal environment is shown in Table 7.4.

7.4 Concluding Remarks

A new method for compensating the mutual coupling effect of a linear adaptive dipole array employed in adaptive nulling is introduced and testified. The new method adopts a realistic approach in that it needs only the measured voltages across the antenna terminal loads and an estimated current distribution for the calculation of the mutual impedances. The mutual impedances are defined and calculated differently from the conventional method and the results are more capable of removing the mutual coupling effect. The new method does not require the knowledge of the elevation angles of the SOI and the interferences and still works if the elevation angles of the SOI and the interferences do not deviate too much from the horizontal direction. This increases the capability of the array to work in three-dimensional signal environments. Computer simulations for a number of rather extreme signal environments have been carried out to testify the robustness and the capability of the new method.

Chapter 8

Conclusions and Recommendations

8.1 Conclusions

In this thesis, we have focused on the problem of effect of mutual coupling in smart antennas and proposed several techniques to compensate for mutual coupling effect in applications of adaptive nulling, DOA estimation and antenna array calibrations. The topic of smart antennas has been introduced, noting that they can reduce cellular interference levels and improve capacity. A number of points have been discussed, and several approaches were proposed to make the adaptive arrays robust to various applications.

Based on investigation and computer simulation results, we summarized this thesis with the following five categorized conclusions:

1. One conclusion of this thesis is relying on the techniques of compensation for

mutual coupling effect. A new method for compensation for the mutual coupling effect is proposed for SA systems. The new compensation method is shown to have a greater ability and flexibility than those previous methods in removing the mutual coupling effect. The new method introduces a new definition of mutual impedance and uses it to find the coupling-free signal voltages across the antenna terminal loads. The procedures on how to measure and calculate the mutual impedance are also given out. Measurement results for the newly defined mutual impedance in the method were provided.

2. The new compensation method has been applied to an ULA and constructed a mutual impedance matrix which is used to compensate for mutual coupling among the elements, scattering effect from surroundings. The results showed that the new method 1) reduced the errors in direction of arrival estimates by using MUSIC and ESPRIT estimation algorithms; 2) increased the probability of resolving two signal sources as their angular separation decreased to zero; 3) increased accuracy to DOA estimates in the case of fully coherent signal sources.
3. We applied three techniques of compensation for the mutual coupling effect in the UCA for direction finding . The proposed methods applied to UCA in conjunction with ML-EMSGA DOA estimation algorithm are shown to have a greater ability and flexibility in removing the mutual coupling effect. By using ML-EMSGA can obtain much more accurate estimates of elevation angles and azimuth angles of signal of interests in circular arrays and at same time can lower the computation load of ML. Simulation results showed that the robustness and the capability of the proposed method with ML-EMSGA in UCA employed for direction finding.

4. The proposed method for compensating for the mutual coupling effect in a linear adaptive dipole array employed in adaptive nulling is simulated and analyzed. The new method does not require the knowledge of the elevation angles of the SOI and the interferences. It still works if the elevation angles of the SOI and the interferences do not deviate too much from the horizontal direction. This increases the capability of the array to work in three-dimensional signal environments. Computer simulations for a number of rather extreme signal environments have been carried out to testify the robustness and the capability of the proposed method.
5. A structured least square method is proposed to simultaneously estimate the mutual coupling matrix and the DOAs in the presence of mutual coupling and thermal noise. The geometry of the antenna array are ULA and UCA. It was found that the proposed method can estimate the DOA and MCM simultaneously when at least two observations signals are available. The proposed method is also useful in calibration of ULA and UCA when the DOAs of calibration sources were not known. By evaluating three examples, it was shown that the proposed method converged very fast and is very effective.

8.2 Recommendations for Further Research

Based on the techniques and schemes developed in this thesis, the scope of the current work may be further extended.

In this thesis, we have simulated and modelled the ULA and UCA, which are designed for different applications. Previous studies have shown a greater ability and flexibility of the proposed technique in removing the mutual coupling effect. Further investigation can be done into algorithms with a less computational complexity and faster convergence to achieve commercial implementations of the techniques with SA.

In DOA estimations, five algorithms have been studied in ULAs and UCAs. As shown in the simulation study, the performance of direction finding degrades in a coherent signals environment. Different DOA estimation algorithms are adapted to different array geometries. Many of them totally fail in coherent signal environment. Extra investigations will be carried out to improve the estimation accuracy. New techniques should be exploited to cope with the problems of coherent signals in a non-uniform spaced array too.

As multiple input and multiple output systems represent an economic way to increase the channel capacity and realize high data-rate broadband wireless transmissions, further work will be carried out to search for efficient DOA estimation algorithms for more complicated and useful array geometries and the array with different type of antennas. Also the application and utilizations of the new proposed compensation method in mobile and satellite communications will be developed.

Finally, smart antenna technology is a vast research area. New algorithms and techniques will evolve with the demands of future generations of wireless communications.

Author's Publications

Journal Papers

1. T. T. Zhang and H. T. Hui, "DOA estimation by using ESPRIT with mutual coupling effect compensated by a new method," *IEE Proc. Science, Measurement and Technology*, vol. 151, pp. 476-478, 2004.
2. T. T. Zhang, H. T. Hui and Y. L. Lu, "Compensation for the mutual coupling effect in ESPRIT DOA estimations," *IEEE Transactions on Antennas and Propagation*, vol. 53, no. 4, pp. 1552-1556, Apr. 2005.
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Conference Papers

1. T. T. Zhang, Y. L. Lu and H. T. Hui, "Simultaneous estimation of mutual coupling matrix and DOAs using structured least square method," *Proc. IASTED International Conference on Antennas, Radar, and Wave Propagation*, Jul. 2005.
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