1

PM-1 3½"x4" PHOTOGRAPHIC MICROCOPY TARGET
NBS 1010a ANSI/ISO #2 EQUIVALENT

PRECISIONSM RESOLUTION TARGETS

PIONEERS IN METHYLENE BLUE TESTING SINCE 1974

10000 COUNTY ROAD 3, BURNING WOOD, WV 26037 USA
TEL: 812 485 7857 FAX: 812 485 7857 TLA: 040000000

National Library
of Canada

Acquisitions and
Bibliographic Services Branch

395 Wellington Street
Ottawa, Ontario
K1A 0N4

Bibliothèque nationale
du Canada

Direction des acquisitions et
des services bibliographiques

395, rue Wellington
Ottawa (Ontario)
K1A 0N4

Your file - Votre référence

Our file - Notre référence

NOTICE

The quality of this microform is heavily dependent upon the quality of the original thesis submitted for microfilming. Every effort has been made to ensure the highest quality of reproduction possible.

If pages are missing, contact the university which granted the degree.

Some pages may have indistinct print especially if the original pages were typed with a poor typewriter ribbon or if the university sent us an inferior photocopy.

Reproduction in full or in part of this microform is governed by the Canadian Copyright Act, R.S.C. 1970, c. C-30, and subsequent amendments.

AVIS

La qualité de cette microforme dépend grandement de la qualité de la thèse soumise au microfilmage. Nous avons tout fait pour assurer une qualité supérieure de reproduction.

S'il manque des pages, veuillez communiquer avec l'université qui a conféré le grade.

La qualité d'impression de certaines pages peut laisser à désirer, surtout si les pages originales ont été dactylographiées à l'aide d'un ruban usé ou si l'université nous a fait parvenir une photocopie de qualité inférieure.

La reproduction, même partielle, de cette microforme est soumise à la Loi canadienne sur le droit d'auteur, SRC 1970, c. C-30, et ses amendements subséquents.

Adaptive Multiple-Beam Antenna Arrays for Indoor Wireless Communications

by

Pimpin Kwala, B. Eng.

A thesis submitted to
the Faculty of Graduate Studies and Research
in partial fulfillment
of the requirements for the degree of

Master of Engineering

Ottawa-Carleton Institute for Electrical Engineering
Faculty of Engineering
Department of Systems and Computer Engineering
Carleton University
Ottawa, Ontario
K1S 5B7, Canada

June 29, 1992

© copyright

1992, Pimpin Kwala

National Library
of Canada

Acquisitions and
Bibliographic Services Branch

395 Wellington Street
Ottawa, Ontario
K1A 0N4

Bibliothèque nationale
du Canada

Direction des acquisitions et
des services bibliographiques

395, rue Wellington
Ottawa (Ontario)
K1A 0N4

Voici l'un de vos exemplaires

Voici l'un de vos exemplaires

The author has granted an irrevocable non-exclusive licence allowing the National Library of Canada to reproduce, loan, distribute or sell copies of his/her thesis by any means and in any form or format, making this thesis available to interested persons.

L'auteur a accordé une licence irrévocable et non exclusive permettant à la Bibliothèque nationale du Canada de reproduire, prêter, distribuer ou vendre des copies de sa thèse de quelque manière et sous quelque forme que ce soit pour mettre des exemplaires de cette thèse à la disposition des personnes intéressées.

The author retains ownership of the copyright in his/her thesis. Neither the thesis nor substantial extracts from it may be printed or otherwise reproduced without his/her permission.

L'auteur conserve la propriété du droit d'auteur qui protège sa thèse. Ni la thèse ni des extraits substantiels de celle-ci ne doivent être imprimés ou autrement reproduits sans son autorisation.

ISBN 0-315-79816-5

The undersigned recommend to the Faculty of Graduate Studies
and Research acceptance of the thesis

**“Adaptive Multiple-Beam Antenna Arrays
for Indoor Wireless Communications”**

submitted by
Pimpin Kwala, B. Eng.
in partial fulfillment of the requirements for
the degree of Master of Engineering

Thesis Supervisor

Chairman, Department of Systems and Computer Engineering

Ottawa-Carleton Institute for Electrical Engineering
Department of Systems and Computer Engineering
Faculty of Engineering
Carleton University
July 31, 1992

Abstract

This thesis investigates some of the important issues involved in using an *adaptive multiple-beam antenna* (MBA) array for *indoor wireless* communications.

A model to characterize the individual path signal of an indoor multipath is derived for multiple-beam forming. The *linear* (1-D) and the *ring* (2-D) arrays are studied. The ring array is preferred for a multipath environment with *cochannel interference* (CCI) because of its ability to form unique spatial beams. *Adaptive* multiple-beam forming using pilot signal training with *least mean square* (LMS) and *recursive least square* (RLS) algorithms is investigated. The LMS algorithm has a slower rate of convergence, but is less complex. A 20-element ring array is shown to track a moving path in steps of 5° . With CCI, nulls as deep as 6 dB is formed in the CCI path direction.

The *symbol error rate* (SER) performance of the adaptive array is evaluated for a QPSK signalling scheme. The SER improves with larger number elements, but deteriorates with increasing Doppler frequency. A 2-element (5-element) array has a 3 dB (7 dB) minimum improvement over a single element antenna for 10^{-3} SER. With 5-element array, a dominant path has a gain of about 1 dB over equal energy paths. In the presence of a CCI, the array performance is degraded by as much as 11 dB (3 dB) for 3 dB (10 dB) *symbol-to-interference* ratio (SIR).

*To:
My parents,
for their love,
care and dedications ...*

Acknowledgments

I wish to thank my thesis supervisor, Dr. A. U. H. Sheikh, for the privilege to work under his guidance and suggesting the thesis topic. My deepest gratitude is due to him for his advice, patience, understanding and encouragement during the course of this research.

I am also grateful to the Friends of India for the generous Indira Gandhi Memorial Fellowship in support of my graduate study and research. This research was also supported by Mitel Corporation and Carleton University.

Special appreciation and thanks to my fellow former and current graduate students in particular, Edward Pham and Shahram Golastaneh for their help and many enlightening discussions and suggestions. Thanks are also due to the staff members in the System and Computer Engineering department, for their continuing support throughout my stay in the department.

Table of Contents

Abstract	iii
Acknowledgments	v
List of Tables	x
List of Figures	xi
List of Abbreviations and Acronyms	xv
Lists of Mathematical Symbols and Notation	xviii
CHAPTER 1 Introduction	1
1.1 Problem Definition	2
1.2 Thesis Objectives	3
1.3 Thesis Organization	3
CHAPTER 2 Survey of Previous and Related Work	5
2.1 Directional (Sectorized) Antennas	5
2.2 Multiple-Beam Antennas	6
2.3 Adaptive (Optimum) Combining	8
2.4 Adaptive Antenna Arrays	9
CHAPTER 3 Indoor Wireless Channel Characteristics	11
3.1 Indoor Multipath Propagation	11
3.2 Ray-Theoretic Path Signal Model	13
3.3 Statistical Indoor Received Signal Model	17
3.3.1 Time Domain Statistical Received Signal Model	17
3.3.1.1 Long-Term Fading (Shadowing) Model	17
3.3.1.2 Short-Term Fading Model	18

3.3.1.3 Doppler Frequency Shift (Fading Rate)	20
3.3.2 Combined Multipath Signal Envelope Power and Phase	21
3.3.3 Received Signal Envelope Correlation and Power Spectrum	22
3.4 Conventional Multipath-Distortion Correction Techniques	27
3.4.1 Diversity Techniques	28
3.4.2 Adaptive Equalization Techniques	29
3.4.3 Spread Spectrum Techniques	29
3.5 Indoor and Vehicular Cellular Characteristics	30
CHAPTER 4 Multiple-Beam Antenna Arrays	32
4.1 Antenna Array Structures	32
4.1.1 Linear Antenna Array	32
4.1.2 Ring Antenna Array	35
4.2 Adaptive Array Narrowband Beam-Forming	39
4.2.1 Antenna Array Radiation Pattern	42
4.3 Single-Path Beam-Forming with Known DOA	42
4.3.1 Linear Array Single-Path Beam-Forming	42
4.3.2 Ring Array Single-Path Beam-Forming	43
4.3.3 Linear and Ring Array Beamwidth Comparison	48
4.4 Multiple-Path Beam-Forming with Known DOA's	48
4.4.1 Linear Array Multiple-Path Beam-Forming	50
4.4.2 Ring Array Multiple-Path Beam-Forming	50
CHAPTER 5 Adaptive Multiple-Beam Forming	54
5.1 Optimum Solution for Beam-Forming in a Multipath Environment	54
5.2 Adaptive Algorithms for Multiple-Beam Forming	56
5.2.1 Least Mean Square Algorithm	56
5.2.2 Recursive Least Square Algorithm	58

5.3 Adaptive Beam-Forming Simulations	60
5.3.1 Simulation Procedures	60
5.3.2 Algorithm Rate of Convergence	63
5.3.3 Adaptive Single-Path Beam-Forming	67
5.3.3.1 Adaptive Single-Path Tracking	68
5.3.4 Adaptive Multiple-Path Beam-Forming without CCI	73
5.3.5 Adaptive Multiple-Path Beam-Forming with CCI	76
5.4 Multipath Signal Reception Using Antenna Arrays	77
CHAPTER 6 Adaptive MBA Array System Performance	81
6.1 System Model	81
6.1.1 Transmitter Subsystem	83
6.1.1.1 Desired and CCI Signal	83
6.1.2 Transmit and Receive Filtering	84
6.1.3 Narrowband Multipath Channel	85
6.1.4 Additive White Gaussian Noise (AWGN)	86
6.1.5 Adaptive MBA Array	87
6.1.6 Receiver Subsystem	87
6.2 Simulation Procedures	89
6.2.1 Simulation Assumptions	92
6.3 QPSK SER Performance over an AWGN Channel	93
6.4 Adaptive Array SER Performance without CCI	94
6.4.1 Single-Path Channel without CCI	95
6.4.2 Three-Path Channel without CCI	102
6.5 Adaptive Array SER Performance with CCI	105
6.5.1 Single-Path Channel with CCI	106

CHAPTER 7 Conclusions and Recommendations	116
7.1 Summary and Conclusions	116
7.2 Recommendations for Further Research	117
APPENDIX A General Antenna Array Theory	118
APPENDIX B Optimum Weight Solution for Beam-Forming in a Multipath Environment	122
APPENDIX C System Performance Simulation Documentation	126
C.1 SER Simulation Operating Scenarios	126
C.2 Transmit and Receive Filters	129
C.3 Desired Signal, Interference and Noise Power Calculations	130
Bibliography	136

List of Tables

Table 3.1	Indoor vs. vehicular normalized fading rate	21
Table 3.2	Combined multipath signal envelope power and phase simulation parameters	21
Table 3.3	Indoor vs. vehicular cellular characteristics	30
Table 5.1	LMS and RLS algorithm computational complexity comparison . . .	60
Table 5.2	LMS and RLS algorithm rate of convergence simulation parameters .	63
Table 5.3	LMS and RLS algorithm rate of convergence for a 10–element ring array with a signal path	65
Table 5.4	Omnidirectional vs. null-pattern initialization simulation parameters .	67
Table 5.5	Adaptive single-path tracking simulation parameters	70
Table 5.6	Adaptive multiple-path beam-forming without CCI simulation parameters	73
Table 5.7	Adaptive multiple-path beam-forming with CCI simulation parameters	76
Table 6.1	Dibit to QPSK phase and QPSK complex symbol mapping with Gray coding	83
Table 6.2	SER simulation parameters for single-path channel with AWGN . . .	95
Table 6.3	SNR deviations from ideal curves for a 5–element array in a single-path channel with AWGN	102
Table 6.4	SER simulation parameters for three-path channel with AWGN . . .	103
Table 6.5	SER simulation parameters for single-path channel with AWGN and one CCI	106
Table 6.6	SNR deviations from ideal curves for a 5–element array in a single-path channel with AWGN and one CCI	115
Table 6.7	Performance degradations with CCI for a 5–element array in a single-path channel with AWGN and one CCI	115

List of Figures

Figure 2.1	Conceptual multiple-beam reflector antenna	7
Figure 3.1	Multipath rays in a typical office environment	12
Figure 3.2	Phasor representation of multipath signals	14
Figure 3.3	Time evolution of a combined multipath signal envelope power, # paths = 3, a) $f_{dm}T = 0.005$, b) $f_{dm}T = 0.0005$	23
Figure 3.4	Time evolution of a combined multipath signal phase, # paths = 3, a) $f_{dm}T = 0.005$, b) $f_{dm}T = 0.0005$	24
Figure 3.5	Time evolution of a combined multipath signal envelope power, # paths = 8, a) $f_{dm}T = 0.005$, b) $f_{dm}T = 0.0005$	25
Figure 3.6	Time evolution of a combined multipath signal phase, # paths = 8, a) $f_{dm}T = 0.005$, b) $f_{dm}T = 0.0005$	26
Figure 4.1	An M-element linear array structure with an incident l^{th} path plane wave	33
Figure 4.2	An M-element ring array structure with an incident l^{th} path plane wave	36
Figure 4.3	An M-element adaptive array narrowband beam-former	40
Figure 4.4	Array pattern for a single path, polar plot, linear array, $d = \frac{\lambda}{2}$, a) DOA = 90° , b) DOA = 45° , c) DOA = 0°	44
Figure 4.5	Array pattern for a single path, linear plot, linear array, # elements = 5, $d = \frac{\lambda}{2}$, DOA = 90°	45
Figure 4.6	Array pattern for a single path, polar plot, ring array, $d = \frac{\lambda}{2}$, a) DOA = 90° , b) DOA = 45° , c) DOA = 0°	46
Figure 4.7	Array pattern for a single path, linear plot, ring array, # elements = 5, $d = \frac{\lambda}{2}$, DOA = 90°	47
Figure 4.8	Mainlobe beamwidth variation vs. azimuthal angle, $d = \frac{\lambda}{2}$, a) linear array, b) ring array	49

Figure 4.9	Array pattern for three paths, polar plot, linear array, $d = \frac{\lambda}{2}$, DOA = 45°, 90° and 135°, a) # elements = 10, b) # elements = 20	51
Figure 4.10	Array pattern for three paths, polar plot, ring array, $d = \frac{\lambda}{2}$, DOA = 45°, 90° and 135°, a) # elements = 10, b) # elements = 20	52
Figure 5.1	Adaptive multiple-beam forming simulation flowchart	61
Figure 5.2	Single path squared error time evolution, ring array, # elements = 10, LMS, training = 300 symbols, DOA = 90°, a) SNR = 10 dB, b) SNR = 20 dB	64
Figure 5.3	Single path squared error time evolution, ring array, # elements = 10, LMS and RLS, training = 300 symbols, DOA = 90°, SNR = 20 dB	66
Figure 5.4	Converged array pattern for a single path, ring array, # elements = 10, LMS, $\mu = 0.05$, training = 20 symbols, DOA = 90°, SNR = 10 dB, a) polar plot, b) linear plot	69
Figure 5.5	Converged array pattern for a single-path tracking from 30° to 45° in steps of 5°, ring array, LMS, $\mu = 0.05$, training = 20 symbols, SNR = 10 dB, a) # elements = 10, b) # elements = 20	71
Figure 5.6	Converged array pattern for a single-path tracking from 30° to 45° in steps of 5°, ring array, # elements = 20, LMS, pilot training followed by decision-directed training, $\mu = 0.05$, training = 20 symbols, SNR = 10 dB	72
Figure 5.7	Converged array pattern for four paths, polar plot, ring array, # elements = 20, LMS, $\mu = 0.001$, training = 20 symbols, DOA = 0°, 90°, 180° and 270°, SNR = 10 dB	74
Figure 5.8	Converged array pattern for four paths, polar plot, ring array, # elements = 20, LMS, $\mu = 0.001$, training = 20 symbols, DOA = 0°, 15°, 60° and 75°, SNR = 10 dB	75
Figure 5.9	Converged array pattern for two paths, # elements = 10, LMS, $\mu = 0.005$, training = 40 symbols (total), desired DOA = 45° & 135°, SNR = 10 dB, CCI DOA = 90°, SIR = 0 dB, a) polar plot, b) linear plot	78

Figure 5.10	Converged array pattern for two paths, ring array, # elements = 10, LMS, $\mu = 0.005$, training = 40 symbols (total), desired DOA = 45° & 135° , SNR = 10 dB, CCI DOA = 135° , SIR = 0 dB, a) polar plot, b) linear plot	79
Figure 6.1	Indoor wireless cellular communication system model	82
Figure 6.2	QPSK decision region	88
Figure 6.3	SER simulation flowchart	90
Figure 6.4	Frame structure used for simulation	91
Figure 6.5	SER vs. desired SNR, # desired paths = 1, desired path DOA = 90°	97
Figure 6.6	SER vs. desired SNR, # desired paths = 1, desired path DOA = 0° , a) $f_{dm}T = 0.005$, b) $f_{dm}T = 0.0005$	98
Figure 6.7	Converged radiation pattern, # desired paths = 1, desired path DOA = 0° , a) $f_{dm}T = 0.005$, b) $f_{dm}T = 0.0005$	99
Figure 6.8	SER vs. desired SNR, # desired paths = 1, desired path DOA = 135° , a) $f_{dm}T = 0.005$, b) $f_{dm}T = 0.0005$	100
Figure 6.9	SER vs. desired SNR, # desired paths = 1, desired path DOA = 225° , a) $f_{dm}T = 0.005$, b) $f_{dm}T = 0.0005$	101
Figure 6.10	SER vs. desired SNR, # desired paths = 3, $f_{dm}T = 0.0005$, a) desired path DOA = 0° , 120° and 240° , b) desired path DOA = 0° , 15° and 180°	104
Figure 6.11	SER vs. desired SNR, # desired paths = 1, desired path DOA = 0° , # CCI paths = 1, CCI path DOA = 120° , $f_{dm}T = 0.0005$, a) SIR = 3 dB, b) SIR = 10 dB	108
Figure 6.12	Converged radiation pattern, # desired paths = 1, desired path DOA = 0° , # CCI paths = 1, CCI path DOA = 120° , $f_{dm}T = 0.0005$, a) SIR = 3 dB, b) SIR = 10 dB	109
Figure 6.13	SER vs. desired SNR, # desired paths = 1, desired path DOA = 0° , # CCI paths = 1, CCI path DOA = 120° , $f_{dm}T = 0.0005$	110

Figure 6.14	SER vs. desired SNR, # desired paths = 1, desired path DOA = 135°, # CCI paths = 1, CCI path DOA = 120°, $f_{dm}T = 0.0005$, a) SIR = 3 dB, b) SIR = 10 dB	111
Figure 6.15	SER vs. desired SNR, # desired paths = 1, desired path DOA = 135°, # CCI paths = 1, CCI path DOA = 120°, $f_{dm}T = 0.0005$	112
Figure 6.16	SER vs. desired SNR, # desired paths = 1, desired path DOA = 225°, # CCI paths = 1, CCI path DOA = 120°, $f_{dm}T = 0.0005$, a) SIR = 3 dB, b) SIR = 10 dB	113
Figure 6.17	SER vs. desired SNR, # desired paths = 1, desired path DOA = 315°, # CCI paths = 1, CCI path DOA = 120°, $f_{dm}T = 0.0005$, a) SIR = 3 dB, b) SIR = 10 dB	114
Figure A.1	An M-element general array structure with an incident l^{th} path plane wave	119
Figure C.1	Portable moving around a base station	127
Figure C.2	Square root raised cosine (SRRC) filter normalized frequency magnitude response, $\alpha = 0.25$	131
Figure C.3	Square root raised cosine (SRRC) filter normalized impulse response, $\alpha = 0.25$	132
Figure C.4	Raised cosine (RC) filter normalized frequency magnitude response, $\alpha = 0.25$	133
Figure C.5	Raised cosine (RC) filter normalized impulse response, $\alpha = 0.25$	134

List of Abbreviations and Acronyms

ACI	- Adjacent Channel Interference
AWGN	- Additive White Gaussian Noise
BPSK	- Binary Phase Shift Keying
CCI	- Cochannel Interference
CDMA	- Code Division Multiple Access
CELP	- Code-Excited Linear Prediction
CIR	- Channel Impulse Response
CT	- Cordless Telephone
<i>dB</i>	- decibel
DECT	- Digital European Cordless Telephone
DFE	- Decision Feedback Equalizer
DOA	- Direction of Arrival
DQPSK	- Differentially encoded Quadrature Phase Shift Keying
DSP	- Digital Signal Processing
FDMA	- Frequency Division Multiple Access
FFT	- Fast Fourier Transform
FIR	- Finite Impulse Response
<i>GHz</i>	- 10^9 Hertz
GMSK	- Gaussian filtered Minimum Shift Keying
GSM	- Global System for Mobile Telecommunications
<i>Hz</i>	- Hertz (cycle/s)
IEEE	- Institute of Electrical and Electronics Engineers
ISI	- Intersymbol Interference
<i>kbps</i>	- 10^3 bits per second

<i>ksps</i>	- 10^3 symbols per second
LAN	- Local Area Network
LMS	- Least Mean Square
LOS	- Line-of-Sight
MBA	- Multiple-Beam Antenna
<i>MHz</i>	- 10^6 Hertz
MMSE	- Minimum Mean Square Error
<i>ms</i>	- 10^{-3} seconds
MSE	- Mean Square Error
<i>Msp</i>	- 10^6 symbols per second
MUSIC	- Multiple Signal Classification
<i>mW</i>	- 10^{-3} Watt
<i>ns</i>	- 10^{-9} seconds
PBX	- Private Branch Exchange
PCS	- Personal Communication Services
<i>pdf</i>	- probability density function
PRMA	- Packet Reservation Multiple Access
PSK	- Phase Shift Keying
PSTN	- Public Switched Telephone Network
QPSK	- QuadriPhase-Shift Keying
RC	- Raised Cosine
RLS	- Recursive Least Squares
RMS	- Root Mean Squared
SER	- Symbol Error Rate (Probability of Symbol Error)
SIR	- Signal to Interference Ratio
SMI	- Sample Matrix Inverse

SNR	- Signal to Noise Ratio
SRRC	- Square Root Raised Cosine
TDL	- Tapped Delay Line
TDMA	- Time Division Multiple Access
WIN	- Wireless In-building Network
WSS	- Wide Sense Stationary

Lists of Mathematical Symbols and Notation

The following symbols and notational conventions are used throughout the thesis. In some cases, the same symbol is to denote different quantities, but the distinction should be obvious from the context of the discussion.

$\mathbf{a}$	- general vector
$\hat{\mathbf{a}}$	- unit vector of $\mathbf{a}$
$\mathbf{a}^T(t)$	- transpose of $\mathbf{a}$
$\hat{\mathbf{a}}_l(t)$	- l^{th} path unit direction vector
$\mathbf{A}$	- general matrix
$\mathbf{A}^T$	- transpose of $\mathbf{A}$
$\mathbf{A}^H$	- Hermitian (conjugate transpose) of $\mathbf{A}$
$\mathbf{A}^{-1}$	- matrix inverse of $\mathbf{A}$
c	- speed of light (3×10^8 m/s)
$c[k]$	- sampled training signal
$\mathbf{D}_{in}$	- n^{th} interference direction manifold matrix
$d[k]$	- sampled desired signal
$\mathbf{d}_l$	- l^{th} path direction manifold vector
$\mathbf{D}_s$	- desired signal direction manifold matrix
$E(.)$	- expected value of the quantity within parenthesis
$e[k]$	- sampled adaptive combiner output error signal
$erfc(.)$	- complementary error function
E_s	- energy per symbol
$\exp(.)$	- natural exponential function

f	- frequency variable
f_c	- carrier frequency
f_d	- Doppler frequency shift
f_{dl}	- l^{th} path Doppler frequency shift
f_{dm}	- maximum Doppler frequency shift
$f_{e_m}(\theta, \phi)$	- radiation pattern of the m^{th} antenna element
$F(\theta, \phi)$	- array factor
$\mathbf{g}[k]$	- Kalman gain vector
$g_l(t)$	- l^{th} path medium loss
$G(\theta, \phi)$	- normalized radiation pattern
$H(f)$	- frequency response of a filter
$h(t)$	- impulse response of a filter
$h(\tau; t)$	- time-varying channel impulse response
$\mathbf{I}$	- identity matrix
$\mathbf{I}_{lm}(\mathbf{r}_m, t)$	- excited current phasor at the m^{th} element due to the l^{th} path plane wave
$Im(.)$	- imaginary part of the quantity in parenthesis
$i_{nl}[k]$	- sampled l^{th} path signal of the n^{th} interference
$i_{nl}(t)$	- n^{th} interference l^{th} path signal
$I_o(.)$	- modified Bessel function of the first kind of zero order
j	- imaginary number, square root of negative one
$J[k]$	- MSE function
$J_{\min}$	- minimum MSE
$J_o(.)$	- ordinary Bessel function of zero order
k	- used as discrete time index

k	- wave number ($2\pi/\lambda$)
$\mathbf{K}[k]$	- correlation matrix of weight error
l	- used as path index
L	- number of desired signal paths
$\ln(.)$	- natural logarithm
L_n	- number of n^{th} interfering signal paths
$\log_{10}(.)$	- logarithm to the base 10
m	- local signal mean
M	- number of array elements
$m(t)$	- received signal local mean component
N	- number of cochannel interferers
$\frac{N_a}{2}$	- AWGN power spectral density
$\mathbf{P}$	- cross-correlation between the received vector and desired response
r	- radius of ring array
$\mathbf{R}$	- correlation matrix of received vector
$Re(.)$	- real part of the quantity in parenthesis
R_{lm}	- distance from the l^{th} path source to the m^{th} element,
$\mathbf{r}_m$	- position vector of the m^{th} array element
$r(t)$	- received signal envelope
$S(f)$	- received signal power spectrum
$s[k]$	- sampled desired signal
$s(t)$	- desired signal
$s_l(t)$	- l^{th} path desired signal
t	- used as continuous time index
T	- symbol period

$v_m[k]$	- sampled complex AWGN process at the m^{th} array element
v_p	- speed of portable
$\mathbf{w}$	- adaptive combiner complex weight vector
$x[k]$	- sampled received signal at an element
$\mathbf{x}[k]$	- sampled antenna array output vector
$x(t)$	- general received fading signal
$x_c(t)$	- received signal in-phase component
$x_o(t)$	- short term fading component
$x_s(t)$	- received signal quadrature component
$y[k]$	- sampled adaptive combiner output signal
$\alpha_l(t)$	- l^{th} path loss
$\gamma_l(t)$	- overall l^{th} path reflection coefficient phase
δ_{lm}	- phase delay at m^{th} antenna element due to l^{th} path signal
$\delta(t)$	- Kronecker delta function
$\delta(\tau)$	- Dirac impulse (delta) function
λ	- carrier wavelength
μ	- LMS step size (convergence parameter)
π	- 3.141592654 . . .
ϕ	- azimuth angle
$\phi(t)$	- received signal phase
$\phi_l(t)$	- l^{th} path direction of travel in azimuth
$\psi_l(t)$	- l^{th} path random (total) phase delay
ρ_l	- overall l^{th} path reflection coefficient magnitude
$\rho_{x_1 x_2}$	- cross correlation between two received signals
σ_d^2	- average desired reference signal power

σ_i^2	- average interfering signal power
σ_m	- standard deviation of the local signal mean (log scale)
σ_n^2	- real and imaginary component noise variance
σ_s^2	- average desired signal power
σ_r^2	- received signal power at an element
σ_v^2	- complex noise variance
σ_z	- standard deviation of the local signal mean (linear scale)
$\tau_l(t)$	- l^{th} path propagation time delay
θ	- elevation angle
$\theta_l(t)$	- l^{th} path direction of travel in elevation
$\nabla_{\mathbf{w}}$	- gradient operator with respect to $\mathbf{w}$
$\sum_a^b$	- summation operator from a to b

Chapter 1

Introduction

The past decade has witnessed a phenomenal growth in the *wireless (mobile)* communications, driven by the overwhelming demand for people to communicate *anywhere, anytime*. The trend is to integrate the wireline services (PSTN and PBX) with the wireless services (paging, cellular and cordless systems) into a universal, location and network independent *personal communication services* (PCS) [1–3]. PCS is a service concept aiming to bring communications to the individual via a *personal* telephone number supported by worldwide *distributed intelligent* networks. The *indoor* wireless PCS growth is evidenced by the widespread adoption of the British CT2 [4] and the Canadian CT2Plus standards for the wireless PBX and the public *telepoint*¹ systems.

The challenge to provide an efficient and reliable indoor wireless PCS is enormous. The indoor wireless channel is plagued by *time-varying multipath* propagation (from reflections and diffractions at surrounding obstacles) that creates signal fading. *Flat fading* occurs when the time delay spread of the multipath is negligible. When the time delay spread is in excess of $\frac{1}{10}$ of a symbol period, signal dispersion or *intersymbol interference* (ISI) appears as *frequency selective fading* [5–7]. In addition to fading distortions, there are system and environmental noise accompanied by *adjacent* and *cochannel interference* (ACI and CCI) that arise from frequency adjacency and frequency reuse respectively in the indoor cellular systems [8].

Various *diversity* techniques have been suggested for correcting flat fading [5, 6] to improve system reliability. *Adaptive linear* and *decision feedback equalizers* (DFE) can be used to mitigate the frequency selective fading and interference at high data rate under slow and fast fading conditions [9–11]. Other schemes such as *directional (sectorized) antennas*, *power control* and *dynamic channel allocation* have also been proposed to reduce the effect of CCI [8].

¹ Telepoints are base stations located in public places such as shopping malls, train stations and airports.

The indoor wireless system is also operating under severe spectrum constraints. Low bit rate *speech coders* such as CELP [12] and VSELP [13] coders can be used to alleviate spectrum limitations. Bandwidth efficient *digital modulation* techniques such as, DQPSK (in second-generation North American IS-54 system [13]) and GMSK (in DECT [14] and European Global System for Mobile Telecommunications (GSM) systems [15]), have also been proposed. *Cell splitting* and *frequency reuse* in cellular systems [16] also increase spectral utilization, albeit at the expense of elevated CCI. *Multiple access* schemes such as, FDMA [17], TDMA [17], CDMA [18] and PRMA [19], have been implemented to improve spectral utilization efficiency.

In spite of the above efforts to improve system reliability under spectrum constraints, *adaptive multiple-beam-antenna* (MBA) arrays still have not received until recently, much attention in indoor wireless communications. Adaptive arrays have been used successfully in satellite and radar communications. Among their benefits are the flexibility to handle rapid *beam steering* for high gain and the *interference nulling* capability for protection against CCI or jamming.

Possible reasons for the slowness in applying adaptive arrays for wireless communications are the prohibitive array size for the portables and the lack of low-cost signal processing systems. A 10-element linear array, for example, has a length of 1.5 m ($\frac{\lambda}{2}$ interelement spacing at 900 MHz). This array could be accommodated at the base stations, but not at the portables. However, the shift towards higher operating frequency (1.7–2.3 GHz band allocation for PCS [20]), and the availability of microstrip array implementations coupled with inexpensive VLSI/DSP hardware have opened up a new vista for adaptive array applications in indoor wireless communications.

1.1 Problem Definition

Multipath fading and CCI are two major impairments in the indoor cellular environment. To date, most wireless communication research has been focused on techniques such as, diversity combining, adaptive equalization and channel assignment strategies to mitigate the channel impairments. These techniques overlook the fact that the desired

and interfering signals arrive at the receiver from different path directions.

This thesis proposes an adaptive MBA array that utilizes the path *direction of arrival* (DOA) information to enhance the desired signal reception. In principle, the adaptive array is trained to create “*beams*” in the desired path directions and *nulls* towards the unwanted CCI paths. Beams and nulls are created with periodic *pilot signal* and *decision-directed* training that continuously adapt the array to changing environment.

Adaptive arrays offer improved system reliability through enhanced signal reception and reduced CCI. A higher system capacity could also be attained from a smaller cochannel reuse distance permitted by reduced CCI.

1.2 Thesis Objectives

The objectives of this thesis are:

- Reexamine the multipath propagation characteristics of the indoor wireless channel using *ray-theoretic* principles for multiple-beam forming.
- Investigate the *linear* and *ring* array, with focus on the ring array, for *multiple-beam forming* in an indoor multipath fading environment with CCI.
- Evaluate the *symbol error rate* (SER) system performance of the array for an indoor channel in the presence of CCI through simulation.

1.3 Thesis Organization

This thesis is organized as follows:

Chapter 2 provides a brief survey of related work in: *directional (sectorized) antennas, multiple-beam antennas, adaptive (optimum) combining* and *adaptive antenna arrays*.

In chapter 3, the indoor wireless multipath characteristics are discussed. A simple ray-theoretic model is developed to characterize the received signal on a single path. A review of the statistical indoor channel model, conventional multipath correction techniques and a comparison of the indoor with the vehicular cellular systems are also included.

Chapter 4 presents the linear and ring antenna arrays. Beam-forming using both arrays for known multiple-path DOA's are investigated and compared.

In chapter 5, the *least mean square* (LMS) and *recursive least square* (RLS) adaptive algorithms for beam-forming in an unknown multipath environment are discussed. Simulation results are included to demonstrate the adaptive multiple-beam forming with different operating conditions.

Chapter 6 describes a symbol by symbol system performance simulation using the adaptive array for an indoor cellular system. Simulation results are presented in terms of SER under various scenarios.

The conclusions and suggestions for further research are provided in chapter seven.

Chapter 2

Survey of Previous and Related Work

This chapter provides a brief survey of related work dealing with *directional (sectorized) antennas*, *multiple-beam-antenna (MBA) arrays*, *adaptive (optimur..) combining*, and *adaptive antenna arrays*.

2.1 Directional (Sectorized) Antennas

A directional (sectorized) antenna is a fixed pattern antenna with enhanced gain in a certain sector and reduced response outside the sector. Since multipath signals propagate in different spatial directions, Lee [8] suggested using sectorized antennas to reduce signal fading and CCI by constraining the spatial signal coverage. This idea has been proven in cellular systems that use 120° or 60° sectorized antennas to reduce CCI. Compared with an omnidirectional antenna that have six principal cochannel interferers in the first tier of a seven-cell reuse pattern, a 120° sector antenna has only two principal interferers, and a 60° sectors has only one [8].

Lee [21] also experimented the use of a directional linear array on a mobile unit to reduce signal fading. The results showed a marked reduction in the fading rate when using directional antenna as compared to using omnidirectional antennas. Furthermore, the fading rate was found to decrease with narrower beamwidth. The experiment results corroborate the idea of using narrowbeam to reduce signal fading.

Ikegami and Yoshida [22] also attempted to isolate the individual waves in the multipath of an urban mobile environment with directional antennas. The results revealed close agreements between the experimentally observed principal waves and those predicted with geometrical ray principles. It was also found that the urban mobile radio propagation is dominated by a limited number of wave paths.

Recently, a system consisting of a six sectored fixed-beam antennas at both the base stations and mobile stations has been proposed by Motorola [23] for wireless in-building

network (WINTM). Sophisticated protocols are used to select the “best” sector pair that has the “smallest” interference for radio communications at the start of and during the communications. This scheme is designed for a near-static wireless LAN environment, where fading rate is very slow.

The concept of a fixed sectorized antenna can be extended to an adaptive multiple narrowbeam antenna that only responds to the desired signal paths and rejects the unwanted CCI.

2.2 Multiple-Beam Antennas

This section briefly reviews the multiple-beam antenna (MBA) in radar and satellite communications for use in wireless communications.

Conventional multiple beams can be created in three different ways. The first technique that involves using separate antennas for each beam pointing in different directions, is the simplest. It suffers from beamwidth limitation associated with a single antenna, and fixed beam directions.

A second method to create multiple beams commonly used in satellite and troposcatter communications is to use a reflector antenna with multiple feeds (Figure 2.1). This method also has beamwidth limitation, and the feed orientations have to be mechanically changed to alter the beam directions.

In radar communications, multiple fixed beams can also be generated using a linear array with *beam-forming matrices* such as *Butler matrix* [24]. Butler matrix feed uses hybrid junctions and fixed-phase shifters to generate the necessary phase shifts for each beam in the radiation pattern. The number of beams, which also equals the number of input ports and elements, is fixed (a power of 2) by the network.

The above MBA systems are less than satisfactory for a dynamic wireless environment because of the fixed number of beams and fixed beam positions. An adaptive MBA system that uses spatially distributed antenna elements with adaptive combining is proposed instead.

Figure 2.1 Conceptual multiple-beam reflector antenna

2.3 Adaptive (Optimum) Combining

Adaptive (optimum) combining has been suggested for mitigating the distortions due to flat fading and CCI. It is a technique for combining the received signals from diversity antennas to enhance the output SNR. The conventional adaptive combining does not link the *beam-forming* and *null-steering* concepts to signal reception and interference rejection.

Optimum combining was first discussed by Bogachev and Kiselev [25] using space diversity antennas. They derived the theoretical SNR and probability curves under Rayleigh fading conditions.

Subsequently, Winters [26] studied the performance of optimum combiner at base stations using PSK modulations under slow flat Rayleigh fading condition. From theoretical analysis, he demonstrated that optimum combining is superior to maximal ratio combining by several *dB*'s, even when the number of interferers is larger than the number of diversity antennas.

Vaughan [27], on the other hand, discussed optimum combining for diversity antennas at the mobiles. Array branch signals, rather than discrete spatial sources, are used in his analysis. In contrast with Winters who suggested the LMS algorithm, Vaughan proposed a rather complicated sample matrix inversion (SMI) algorithm using covariance matrices.

The performance of diversity and multiple branch MMSE receiver for combating the signal fading and CCI was also investigated by Kennedy *et al.* [28]. Bit error curves were given for different channel models and interference conditions. But again, no effort was made to study the beam pattern formation of the multiple antenna receivers. They also presented a matched filter bound to the performance of MMSE receiver on a set of mobile channels.

Hanna *et al.* [29] studied the LMS adaptive combiner for a QPSK system using two linear antenna elements. The simulation results conform with Winter's analysis.

The work by Kohno *et al.* [30] showed how temporal and spatial filtering can

be used in a directional diversity manner to compensate for the multipath distortion in a mobile environment. Their approach, in which the directional information of the signal paths was exploited to improve the output SNR, departed from the conventional approach.

In this thesis, the directional information of the signal paths is exploited for multiple-beam forming and adaptive combining to enhance the desired signal reception.

2.4 Adaptive Antenna Arrays

The adaptive antenna array is an established field of research. But, work on the adaptive antenna array for a multipath fading environment is still fairly new. An adaptive antenna array consists of a set of spatially distributed antenna elements with adaptive combining. The array exploits potential differences in the direction of arrival (DOA), phase, spectral characteristics (power and frequency), modulation characteristics, and polarization of the desired and the interfering signals for signal discrimination.

Adaptive interference cancellation in non-fading environment has been studied extensively by many researchers since the introduction of *adaptive sidelobe canceller* by Applebaum [31] and *pilot based beam-former array* by Widrow [32, 33]. Widrow [32, 33] introduced the adaptive pilot signal adaptive array for interference rejection in a non-fading environment. He used pilot signal training to fashion the beam pattern to produce a high gain beam towards the desired signal and nulls towards the interfering signals.

Multiple-beam base-station antenna arrays were initially considered by Davies *et al.* [34] to increase the number of simultaneous users in the cellular systems. It was suggested that each mobile is tracked in azimuth by a narrowbeam for both mobile-to-base and base-to-mobile transmissions. Because of the directive nature of the beams, the mean received interference power could be much less than that using omnidirectional base station antennas.

A *null-steering* adaptive antenna system was proposed by Marcus and Das [35] for land mobile environment. Squelch tone injection (FM system) was suggested for steering nulls towards cochannel sites. From the analysis, it was shown that by having

20 dB nulls towards CCI, cochannel reuse distance can be reduced to increase the system capacity by a factor of about 2.8. Their theoretical work, however, only dealt with a linear array with two elements in the absence of multipath effects.

In contrast to null-steering, Swales *et al.* [36, 37] proposed steering radiation pattern maxima towards desired signals from the base stations. A single beam steering from the base station is used to track a single mobile. But, in actual indoor environment there are usually many propagation paths that might require multiple-beam reception.

An adaptive array was again proposed at the base stations by Anderson *et al.* [38]. A scheme based on the complex *high resolution direction finding* technique was suggested for separating the various received signals. The scheme was found to have great potentials for rejecting CCI, but at a price of computational complexity.

The above survey indicates that very little work has been done on the applications of adaptive MBA arrays for wireless communications in general, and indoor wireless communications in particular. Thus, an investigation into the issues involved in the use of adaptive MBA arrays for indoor wireless communications is needed.

The next chapter provides the background on the indoor wireless channel characteristics for multiple-beam forming.

Chapter 3

Indoor Wireless Channel Characteristics

The indoor wireless cellular characteristics are examined in this chapter. The indoor multipath propagation is presented using *ray-theoretic* principles. The statistical channel models and conventional multipath-distortion correction techniques are also reviewed. A comparison of the indoor and the vehicular cellular characteristics is also provided.

3.1 Indoor Multipath Propagation

Indoor environments can be office buildings, malls, hospitals, airports, factories or any enclosed environments. Measurements have been conducted for these environments at 800–900 MHz [39–41], 1–4 GHz [42–44] and 60 GHz [39] to characterize the channel for system design and evaluation. These measurements indicate that the indoor channel characteristics depend on many factors such as, operating frequency, data symbol rate and building structure and type.

An example of indoor multipath propagation in a typical office environment is shown using geometrical rays in Figure 3.1. A signal from the base-station antenna (on the ceiling) propagates to the portable antenna via several propagation paths because of *specular reflections* and *diffractions*² from walls, partitions and nearby furniture. In general, the number of significant paths depends on the structure and layout of the scatterers and obstacles. Ignoring low-energy paths due to *diffuse reflections*³, high order (multiple) reflections and diffractions, there are usually only a small number of paths arriving at the portable and satisfying *Snell's* law of reflection.

Differences in the propagation time delays produce a time delay spread profile in the received path signals. Significant delay spread (greater than $\frac{1}{10}$ of a symbol period)

² Diffractions occur when incoming waves graze or hit sharp edges, corners, or vertices of obstacles.

³ Diffuse reflection by a rough surface scatter the energy in other directions as well as the desired direction, thus reducing the field strength in the desired direction.

Figure 3.1 Multipath rays in a typical office environment

are usually produced by distant scatterers. Local scatterers normally create paths of negligible mutual delays or delay spread. Because of the randomness in the scattering, each path delay $\tau_l(t)$ can be modelled as a random process. For *narrowband* systems, the mutual path delays are negligible relative to the symbol period.

Each individual path of the multipath has a *path loss*, *phase delay*, and *direction of arrival* (DOA) at the receiver. The signals from various path DOA's combine like *phasors* when received by an *omnidirectional* antenna. A strong or a weak resultant signal is produced from constructive or destructive combining at different instants (cf. Figure 3.2a, b and c). The interaction between the path signals creates a *spatial wave pattern* with maxima and minima. In a stationary environment, the spatial wave pattern is fixed, because the individual path loss, relative phase delay and path DOA are all constant. Thus, a portable can always "tune" to a strong signal by positioning itself at a pattern maximum.

A roving portable in a stationary surrounding experiences fluctuations in the signal amplitude and phase, a phenomenon known as fading, as it moves through the static spatial standing wave pattern. If the surroundings also change (e.g., people moving around), the spatial wave pattern will undulate.

To sum up, the indoor wireless channel is dominated by a few significant propagation paths with different path losses, phase relationships and path DOA's. When the path characteristics vary with time, signal fading is produced. Fade as deep as -40 dB has been observed within buildings [40]. When a fade occurs, the SNR drops dramatically, generating error bursts that is perceptible as intermittence in the wireless connectivity.

The next section presents a ray-theoretic path signal model for multiple-beam forming.

3.2 Ray-Theoretic Path Signal Model

Based on ray tracing [45], a simple ray-theoretic model can be developed for the individual path signal of the multipath described above. Referring to Figure 3.1, a signal propagation path is defined as a sequence of rays (could be just one ray) from

Figure 3.2 Phasor representation of multipath signals

the transmitter to the receiver. Each propagation path suffers a path medium loss (g) and a time delay (τ), which could be time-dependent. When a *narrowband bandpass* signal⁴ $s(t)\exp(j2\pi f_c t)$ transmitted, the complex lowpass equivalent of the received l^{th} path signal $s_l(t)$ can be expressed as

$$\begin{aligned} s_l(t)\exp(j2\pi f_c t) &= s(t - \tau_l(t)) \rho_l \exp(j\gamma_l) g_l(t) \exp(j2\pi(f_c + f_{dl})(t - \tau_l(t))) \\ s_l(t) &= s(t - \tau_l(t)) \rho_l g_l(t) \exp(j(\gamma_l - 2\pi(f_c + f_{dl})\tau_l(t) + 2\pi f_{dl}t)) ; \\ l &= 1, 2, \dots, L \end{aligned} \quad (3.1)$$

where

f_c is the carrier frequency,

$\tau_l(t)$ is the l^{th} path propagation time delay,

$\rho_l \exp(j\gamma_l)$ is the complex composite reflection coefficient⁵ along the l^{th} path,

$g_l(t)$ is the l^{th} path medium loss,

f_{dl} is the l^{th} path *Doppler frequency shift*⁶, to be defined in Eq. (3.19),

and L is the number of propagation paths.

The total phase delay $\psi_l(t)$

$$\psi_l(t) = \gamma_l - 2\pi(f_c + f_{dl})\tau_l(t) \quad (3.2)$$

whose variation is dominated by the carrier, can be assumed to have uniform distribution in $(0-2\pi)$. This could be realistic for high carrier frequency, because a small change in the time delay produces a significant change in the carrier phase.

The signal time variation in an indoor environment is usually quite small⁷, when compared to that in a vehicular environment. The indoor RMS time delay spread τ_{rms} has been reported by Ganesh *et al.* [46] as 20–40 ns for line-of-sight (LOS) and 40–60 ns for obstructed LOS propagation. Thus, for a symbol rate ($\frac{1}{T}$) not exceeding 1 Msps

⁴ f_c much larger than one-half the bandwidth of the bandpass signal [7].

⁵ $\rho_l \exp(j\gamma_l) = \prod_{p=1}^{P_l} \rho_{lp} \exp(j\gamma_{lp})$ where P_l is the number of reflectors along the l^{th} path.

⁶ Positive value implies the transmitter and receiver are approaching each other.

⁷ Walking pace is of the order of 1–2 m/s.

($T > 1 \mu s$), it is safe to assume $s(t - \tau_l(t)) \approx s(t)$. Incorporating these assumptions in Eq. (3.1), the complex lowpass equivalent⁸ of the l^{th} path signal can be rewritten in as

$$s_l(t) = s(t) \alpha_l(t) \exp(j\psi_l(t)) \exp(j2\pi f_{dl}t) ; \quad l = 1, 2, \dots, L \quad (3.3)$$

where

$\alpha_l(t)$ is the l^{th} path loss, which is a combination of the magnitude of the composite reflection coefficient and the medium loss, i.e. $\alpha_l(t) = \rho_l g_l(t)$,

and $\psi_l(t)$ is the uniformly distributed ($0-2\pi$) random phase delay.

The DOA of each path at the receiver is defined by a unit vector $\hat{\mathbf{a}}_l(t)$ as

$$\hat{\mathbf{a}}_l^T(t) = [\theta_l(t), \phi_l(t)]; \quad l = 1, 2, \dots, L \quad (3.4)$$

where

superscript T denotes transposition,

$\theta_l(t)$ is the l^{th} path time-dependent DOA in elevation,

and $\phi_l(t)$ is the l^{th} path time-dependent DOA in azimuth.

The path DOA at the receiver typically changes by a fraction of a degree in a frame period of 10 ms or less for a typical indoor environment (see appendix C.1).

A *discrete multipath channel impulse response* (CIR) model can be derived from the above path signal model with a finite number of statistically independent paths, to characterize the indoor wireless channel. For a channel with L independent paths, a general expression for the time-varying CIR $h(\tau; t)$ can be written as

$$h(\tau; t) = \sum_{l=1}^{L(t)} \delta(\tau - \tau_l(t)) \alpha_l(t) \exp(j(\gamma_l - 2\pi(f_c + f_{dl})\tau_l(t) + 2\pi f_{dl}t)) \quad (3.5)$$

where

$\delta(\tau)$ is the unit impulse signal,

and $L(t)$ is the number of propagation paths which may also vary with time.

⁸ Note: From here on, all signals, unless specified, are assumed as complex lowpass equivalents.

Using the narrowband assumptions, the CIR can be simplified as

$$h(\tau; t) = \delta(\tau) \sum_{l=1}^{L(t)} \alpha_l(t) \exp(j\psi_l(t)) \exp(j2\pi f_{dl}t) \quad (3.6)$$

For a large number of paths, the above CIR model leads to the conventional statistical model, which is the subject of the next section.

3.3 Statistical Indoor Received Signal Model

Statistical models are used for the indoor wireless systems because of their simplicity and reasonable accuracy at predicting the average behavior and performance of such systems. The models employ statistical concepts such as *probability distribution function (pdf)*, *signal mean* and *variance*, to characterize the received signal time variation [5, 6].

3.3.1 Time Domain Statistical Received Signal Model

The indoor signal environment is qualitatively similar to the outdoor vehicular signal environment in terms of *shadowing* and *multipath fading*. In general, the received indoor signal envelope time variation consists of a stationary *short-term fading* component, superimposed on a slow varying (*long-term fading*) mean signal component [6]. Let $x(t)$ be the received signal, then

$$x(t) = m(t) x_o(t) \quad (3.7)$$

where

$m(t)$ is the long-term fading mean signal (local signal mean),
and $x_o(t)$ is the short-term fading signal.

3.3.1.1 Long-Term Fading (Shadowing) Model

The local signal mean $m(t)$ (over the time to cover short distances (40λ)) of the received fading signal $x(t)$, exhibits a slow variation called *shadowing* over large distances (greater than 200λ) due to variations in the "distant" surroundings [6]. Expressed in *dB*,

the local signal mean variation can be modelled by a *normal* distribution with *pdf* [6]

$$p_M(m) = \frac{1}{\sqrt{2\pi} \sigma_m} \exp \left[-\frac{(m - \bar{m})^2}{2\sigma_m^2} \right] \quad (3.8)$$

where

m is the local signal mean (in *dB*),

and σ_m is the standard deviation of the local signal mean (log. scale).

The local signal mean and standard deviation depend on the carrier frequency and the “distant” surroundings. In general, the local signal mean is a function of the distance d between the transmitter and the receiver.

On a linear scale, the local signal mean follows a *log-normal* distribution with *pdf*

$$p_Z(z) = \frac{\log_{10} e}{\sqrt{2\pi} z \sigma_z} \exp \left[-\frac{(20 \log_{10} z - \bar{z})^2}{2\sigma_z^2} \right] \quad (3.9)$$

where

z is the local signal mean (linear scale), defined as $z = 10^{\frac{m}{20}}$,

and σ_z is the standard deviation of the local signal mean (linear scale).

3.3.1.2 Short-Term Fading Model

For the short-term fading model, the followings are usually assumed:

- number of propagation paths is large ($L > 8$),
- and path DOA's are independent and equally distributed in the azimuthal plane.

Using an unmodulated carrier signal of frequency f_c , the real valued received signal can be obtained from the CIR model of Eq. (3.6) as

$$\begin{aligned} r(t) &= \sum_{l=1}^{L(t)} \text{Re}(\alpha_l(t) \exp(j\psi_l(t)) \exp(j2\pi f_{dl}t) \exp(j2\pi f_c t)) \\ &= \sum_{l=1}^{L(t)} \alpha_l(t) \cos(2\pi f_c t + \vartheta_l(t)) \end{aligned} \quad (3.10)$$

where $\vartheta_l(t) = \psi_l(t) + 2\pi f_{dl}t = \psi_l(t) + 2\pi f_{dm} \cos(\phi_l)t$.

The above equation can be rewritten in terms of the *in-phase* and the *quadrature* components, such that

$$x(t) = x_c(t) \cos(2\pi f_c t) - x_s(t) \sin(2\pi f_c t) \quad (3.11)$$

where

$$x_c(t) = \sum_{l=1}^{L(t)} \alpha_l(t) \cos(\vartheta_l(t)) \quad (3.12)$$

$$x_s(t) = \sum_{l=1}^{L(t)} \alpha_l(t) \sin(\vartheta_l(t)) \quad (3.13)$$

For large L ($L > 8$), both $x_c(t)$ and $x_s(t)$ are *independent Gaussian processes* by virtue of the *Central Limit Theorem* [47]. The envelope $r(t)$, defined as

$$r(t) = \sqrt{x_c^2(t) + x_s^2(t)} \quad (3.14)$$

can be shown to have a *Rayleigh* distribution with *pdf*

$$p_R(r) = \begin{cases} \frac{r}{\sigma^2} \exp\left(-\frac{r^2}{2\sigma^2}\right) & ; r \geq 0 \\ 0 & ; r < 0 \end{cases} \quad (3.15)$$

where σ^2 is the Rayleigh variance (power) parameter ($\text{var}(r) = \frac{(4-\pi)\sigma^2}{2}$) [6].

The phase $\phi(t)$, defined as

$$\phi(t) = \arctan\left(\frac{x_s(t)}{x_c(t)}\right) \quad (3.16)$$

has a uniform distribution in the interval $(0, 2\pi)$ with *pdf* [6]

$$p_\Phi(\phi) = \begin{cases} \frac{1}{2\pi} & ; 0 \leq \phi \leq 2\pi \\ 0 & ; \text{otherwise} \end{cases} \quad (3.17)$$

Rayleigh distribution has been found to fit measurement data from many indoor building structures with heavy clutter [42, 48].

When there is a predominant signal or LOS component (such as in *microcellular* structure [49]), a *Rician* distribution (Eq. (3.18)) [6] is more appropriate for modelling the signal envelope variation.

$$p_R(r) = \frac{r}{\sigma^2} \exp\left(-\frac{r^2 + A^2}{2\sigma^2}\right) I_0\left(\frac{Ar}{\sigma^2}\right) ; r \geq 0 \quad (3.18)$$

where

σ^2 is the diffuse (Rayleigh) multipath variance (power) parameter,

A is the direct component signal strength,

and $I_0(.)$ is the modified *Bessel* function of the first-kind of zero order.

Rician model, with different K^9 values, has also been found to fit measurement data from many building structures [41, 50] with LOS components and light clutter.

3.3.1.3 Doppler Frequency Shift (Fading Rate)

When the portable is in motion, the path signal experiences a Doppler shift whose frequency determined by the spatial angle between the path DOA and the portable direction of motion. In the azimuthal plane, the Doppler frequency f_d is

$$\begin{aligned} f_d &= f_{dm} \cos(\phi) \\ &= \left(\frac{v_p}{c}\right) f_c \cos(\phi) \end{aligned} \quad (3.19)$$

where

f_{dm} is the maximum Doppler frequency,

ϕ is the angle between the path DOA and the portable direction of motion,

v_p is the speed of the portable,

c is the speed of light,

and f_c is the carrier frequency.

Doppler frequency f_d is also used to characterize the rate of signal fading in a time-varying multipath channel. The *normalized fading rate* $f_d T$ (T is the symbol period) is a better measure of how fast the channel varies relative to a symbol period. As shown

⁹ $K = \frac{A^2}{2\sigma^2}$, ratio of the power of the direct signal component to the power of the diffused signal component

in the following table, for the same carrier frequency of 10 GHz and data symbol rate of 100 kbps, the maximum normalized fading rate for the indoor system is 0.0005 and that for the vehicular system is 0.005.

	indoor	vehicular
carrier frequency, f_c	10 GHz	10 GHz
portable speed, v_p	1.5 m/s	54 km/h
max. Doppler frequency, f_{dm}	50 Hz	500 Hz
symbol rate, $1/T$	100 kbps	100 kbps
max. normalized fading rate, $f_{dm}T$	0.0005	0.005

Table 3.1 Indoor vs. vehicular normalized fading rate

3.3.2 Combined Multipath Signal Envelope Power and Phase

The time variation of the combined multipath signal as experienced by an omnidirectional antenna is simulated using with a three-path and an eight-path channel models. The model from Eq. (3.3) is used to represent the individual path signal, with the path “loss” (path signal strength), phase delay and DOA chosen randomly as shown in the following table:

	case #1	case #2
# paths	3	8
path “loss”, α_l (normalized)	0.58, 0.47, 0.67	0.60, 0.11, 0.14, 0.41, 0.36, 0.42, 0.22, 0.32
path phase delay, ψ_l (rad)	1.35, 5.71, 4.27	3.93, 3.48, 4.70, 3.36, 3.13, 5.60, 6.00, 0.21
path DOA, ϕ_l (azimuth)	89.5°, 169.7°, 309.9°	1.1°, 32.8°, 47.0°, 57.5°, 76.6°, 158.9°, 257.3°, 303.1°

**Table 3.2 Combined multipath signal envelope
power and phase simulation parameters**

Two maximum normalized Doppler fading rate, $f_{dm}T = 0.005$ for the vehicular system and $f_{dm}T = 0.0005$ for the indoor system are used.

Figure 3.3 displays the received signal envelope power for case #1 (three-path model) under two fading conditions. The fast fading signal envelope power in Figure 3.3a shows a faster variation than the slow fading signal envelope power in Figure 3.3b. Also, some periodicity is apparent in the signal envelope power variation for the three-path model (see Figure 3.3a).

The corresponding signal phase for the three-path model (case #1) is plotted in Figure 3.4. It exhibits a smooth phase variation with periodic phase “discontinuities”. The phase variation is also faster under fast fading conditions.

With an eight-path model (case #2), the received signal envelope power (see Figure 3.5) varies more randomly, mimicking those found in most indoor wireless environment. The corresponding signal phase, plotted in Figure 3.6, also exhibits similar smooth phase variation with periodic “discontinuities”.

The fade depth, which depends on the phase relationship between the path signals, is independent of the fading rate. Fade level as deep as -40 dB is observed in the signal envelope power figures for case #1 and case #2 under different fading conditions.

In general, the signal time variation changes from one realization to another with different set of simulation parameters. The periodicity observed in the three-path channel simulation could be explained in terms of the standing wave pattern. A more “regular” standing wave pattern is obtained from a three-path channel as compared to an eight-path channel. When the portable cuts through the more “regular” standing wave pattern, the “regularity” appears as the periodicity in the signal envelope power and phase plot.

3.3.3 Received Signal Envelope Correlation and Power Spectrum

The *autocorrelation coefficient* of the received signal envelope is an important concept in antenna diversity and combining. For an antenna array, the cross-correlation between the received signals $x_1(t)$ and $x_2(t)$ at two antennas separated by a distance d

Figure 3.3 Time evolution of a combined multipath signal envelope power, # paths = 3, a) $f_{dm}T = 0.005$, b) $f_{dm}T = 0.0005$

Figure 3.4 Time evolution of a combined multipath signal phase, # paths = 3, a) $f_{dm}T = 0.005$, b) $f_{dm}T = 0.0005$

Figure 3.5 Time evolution of a combined multipath signal envelope power, # paths = 8, a) $f_{dm}T = 0.005$, b) $f_{dm}T = 0.0005$

Figure 3.6 Time evolution of a combined multipath signal phase, # paths = 8, a) $f_{dm}T = 0.005$, b) $f_{dm}T = 0.0005$

can be shown to be

$$\rho_{r_1 r_2}(d) = J_0^2\left(\frac{2\pi}{\lambda}d\right) \quad (3.20)$$

where λ is the carrier wavelength and $J_0(\cdot)$ is the zeroth-order *Bessel* function. The auto-correlation function has a first null at $d = 0.378\lambda$ and a value less than 0.2 for $d > 0.378\lambda$. Thus, a separation of 0.5λ will normally ensure uncorrelated received signals, provided the multipath signals has a uniform distributed angle of arrival [5].

The *power spectrum* $S(f)$ of the received signal envelope can be derived by taking the Fourier transform of the autocorrelation function. For an omnidirectional receiver antenna, it has the form

$$S(f) \propto \begin{cases} \left(1 - \left(\frac{f-f_c}{f_{dm}}\right)^2\right)^{-\frac{1}{2}} & ; |f| \leq f_{dm} \\ 0 & ; |f| > f_{dm} \end{cases} \quad (3.21)$$

where f_c is the carrier frequency and f_{dm} is the maximum Doppler frequency shift. If directional antennas are used, the received power spectrum will be modified by the antenna pattern [5].

The models presented so far are for narrowband flat fading case¹⁰. A different model such as the FIR based model with independent fading taps has to be used [7] for frequency selective fading¹¹. Frequency selective fading produces ISI and is a main cause for irreducible error rate in indoor wireless communications.

3.4 Conventional Multipath-Distortion Correction Techniques

With proper considerations on the basic differences between the outdoor vehicular and the indoor environments (see section 3.5), most conventional multipath-distortion correction techniques for the outdoor vehicular environment can also be used for the indoor environment. In the next sections, three principal multipath-distortion correction

¹⁰ The channel frequency response is "flat" and has a linear phase in the bandwidth of interest.

¹¹ Different portions of the bandwidth experience different channel response.

techniques: *diversity*, *adaptive equalization* and *spread spectrum* techniques, are briefly reviewed.

3.4.1 Diversity Techniques

Diversity techniques exploit the fact that the probability of having all received signals fade simultaneously is very much smaller than the probability of any signal fades independently to improve the received signal quality. The independent fading signals can be obtained from *antenna*, *time* and *frequency* diversity [5, 6].

Antenna diversity includes *space*, *angle* and *polarization* diversity. Space diversity is the simplest and most economical technique for overcoming signal fading due to local scattering. It uses a set of two or more antennas spaced more than a quarter wavelength apart to produce uncorrelated received signals at each antenna (see section 3.3.3) [5, 6].

Various combining schemes can be used to produce a "best" signal from several independently fading signals. Combining techniques such as, *selection*, *switched*, *equal-gain*, *maximal ratio* and *optimum* combining can be used with space diversity using *pre-detection* or *post-detection* (baseband) configuration [5, 6].

In selection diversity, the antenna that has the "best" signal is selected. Switched diversity is a degenerate form of selection diversity with two antennas and is simpler to implement. In switched diversity, a "good" antenna is selected at the beginning and its signal is compared with a fixed threshold from time to time. When the signal drops below the threshold, the system switches to the other antenna [5, 6].

Equal gain combining combines all signals with equal weights after phase alignments. Maximal ratio combining, however, uses weights that are proportional to the signal to noise ratio of each antenna, i.e. the strongest get the most weight. Maximal ratio combining performs slightly better (approx. 1 dB) than equal gain combining. Pre-detection combining, where the detection circuitry is placed before the combiner, has better performance than post-detection combining [6].

Diversity techniques work best when the diversity mechanism can provide the required independent fading versions of the same signal.

3.4.2 Adaptive Equalization Techniques

Adaptive equalization can be used to mitigate the slow frequency selective fading (ISI) generated by large delay spread [10]. The adaptive equalizer attempts to recreate the multiple echoes with a *tapped delay line* (TDL) structure, and adjusts them appropriate weights to cancel the original ISI components.

The adaptive equalizer is very effective in reducing the ISI, especially when the RMS delay spread of the multipath exceeds one symbol period. Adaptive equalizer can be combined with diversity antennas to mitigate the joint effect of fast fading and frequency selectivity [9, 11].

Adaptive equalization techniques, however, require expensive hardware for implementation, which makes them less attractive for low-cost systems.

3.4.3 Spread Spectrum Techniques

Spread spectrum techniques have been gaining popularity lately in wireless systems because of its potentials for high system capacity and excellent interference protection [18]. There are in principle two kinds of spread spectrum, *direct-sequence* (DS) and *frequency-hopping* (FH) spread spectrum (SS).

In a direct-sequence spread spectrum (DS/SS), a data bit is chopped into a series of smaller fragments called *chips* by correlating the data bit with a known random chip pattern. The same chip pattern is used at the receiver to undo the spreading effect introduced at the transmitter. The spreading and de-spreading of each data bit essentially protect the bit from ISI corruption. The price paid is the increase in signal bandwidth as a result of spectrum spreading.

Frequency-hopping spread spectrum (FH/SS), however, does not change the original bit pattern, but moves the carrier frequency in a random fashion according to a predetermined chip pattern. Frequency-hopping spread spectrum is less popular than direct-sequence spread spectrum for commercial wireless applications because of the complexity involved in implementing frequency shifting.

Since most wireless systems adopt the cellular structure, a comparative discussion of the indoor and vehicular cellular characteristics is provided in the next section.

3.5 Indoor and Vehicular Cellular Characteristics

Because of the environmental differences, indoor cellular systems differ markedly from vehicular cellular systems, even though they share similar frequency reuse architecture. Their system characteristics are compared in the table below:

Indoor Cellular	Vehicular Cellular
small cells (upto 500 m)	large cells (upto 20 km)
low power (upto 100 mW)	high power (1-10 W)
reuse same frequencies in adjacent cells (CT2)	frequency planning for reuse in adjacent cells
inexpensive base station	costly base station
low speed user (1-2 m/s)	high speed user (20-100 km/h)
slower fading for same data rate	faster fading for same data rate
small time delay spread (< 130 ns)	large time delay spread (~10 μ s)

Table 3.3 Indoor vs. vehicular cellular characteristics

Indoor cellular systems typically use small cells because of the shorter range and the high traffic intensity. *Picocells* and *microcells* not greater than 500 m in radius have been proposed [51]. Three dimensional cellular grids are normally used to cover multistory buildings.

With smaller cell size, low power (1–10 mW) portable handsets can be used, that not only save battery power but also create less interference. Also, with reduced interference, there is a potential to reuse the same frequencies more often in adjacent cells.

Base stations for the indoor cellular systems are less expensive and can be mounted on convenient structures such as ceilings and walls, depending on the structure of the

environment. The vehicular cellular base stations are more expensive because of the cost of land, building or supporting structure and antenna size.

The indoor cellular systems involve slowly moving pedestrians (1–2 *m/s*) whereas the vehicular systems cater for users in high-speed vehicles (20–100 *km/h*). Because of the slow motion, the Doppler spread for the indoor system has been reported as 0.3– 6.1 *Hz* at 900 *MHz* [52], whereas that for the outdoor falls in the range of 50–100 *Hz* for the same carrier frequency. Thus, an indoor user experiences a slower channel fading variation compared to a vehicular user for similar data transmission rate.

The multipath time delay spread for an indoor environment does not exceed 130 *ns* [43], which is smaller than the approximately 10 μ s [53]) delay spread for an outdoor vehicular environment. Adaptive equalizers may not be needed for the indoor channel unless the data rate is high (in excess of 1 *Mbps*).

Because of the presence of walls and partitions, special considerations have to be taken to accommodate the attenuation due to these obstacles in the indoor environment.

As a chapter summary, a model that incorporates the path loss, time (phase) delay and Doppler frequency shift, has been derived to characterize the received signal from a single path of an indoor multipath. The model is shown to generate typical indoor signal fading with as few as eight signal paths. The review of conventional multipath-distortion correction techniques indicate that antenna space diversity could be the most practical and cost-effective technique for indoor wireless environment. Adaptive equalizers could also be integrated with the antenna diversity receiver for higher data transmission rates.

The adaptive *multiple-beam antenna* (MBA) array and its applications for the indoor wireless communications form the subject of discussion for the rest of the thesis. The next chapter first presents the antenna array systems for multiple-beam forming.

Chapter 4

Multiple-Beam Antenna Arrays

This chapter introduces the *linear* and the *ring* arrays. *Narrowband* beam-forming for *single* and *multiple paths* with *known* path directions are discussed through examples. A simple technique to create *multiple beams* is also given.

4.1 Antenna Array Structures

The antenna array elements can be arranged in several ways to create various array radiation pattern for different purposes. In linear (1-D) array, the elements are arranged with *uniform* or *nonuniform* spacing on a straight line. In two-dimensions, the elements can be distributed on a planar grid such as a ring, concentric rings, triangular, rectangular, or hexagonal grid. The elements can also be put on a cylindrical or a spherical surface in three-dimensions.

This thesis considers a uniform linear (1-D) and a ring (2-D) arrays with emphasis on the ring array. The background theory of a general antenna array is given in appendix A. The received signal vector of a linear and a ring array are derived in the following two sections.

4.1.1 Linear Antenna Array

A linear antenna array with M uniformly spaced identical antenna elements placed along the x-axis is shown in Figure 4.1. The interelement spacing d is chosen as $\frac{\lambda}{2}$ ¹². The element output signals of an unweighted linear array due to an l^{th} path plane wave are derived as follows.

First, it is noted that the plane wave arrives at each element of the array with progressive time delays (with respect to the first element). This time delay produces a

¹² $d \leq \frac{\lambda}{2}$, so as to avoid *grating lobes* or multiple peaks from under-sampling in the spatial domain.

Figure 4.1 An M -element linear array structure with an incident l^{th} path plane wave

relative phase delay between the element output signals. A vector approach is taken to derive the relative phase delay.

Referring to Figure 4.1, the position vector $\mathbf{r}_m$ of the m^{th} element with respect to the reference (first) element is

$$\mathbf{r}_m = -(m-1)\frac{\lambda}{2}\hat{\mathbf{x}} \quad ; m = 1, 2, \dots, M \quad (4.1)$$

The DOA unit vector $\hat{\mathbf{a}}_l$ of the l^{th} path is

$$\begin{aligned} \hat{\mathbf{a}}_l &= \cos(\theta_l)\cos(\phi_l)\hat{\mathbf{x}} + \cos(\theta_l)\sin(\phi_l)\hat{\mathbf{y}} + \sin(\theta_l)\hat{\mathbf{z}} ; \\ l &= 1, 2, \dots, L \end{aligned} \quad (4.2)$$

The relative phase delay δ_{lm} between the m^{th} element output signal and the reference element output signal can be obtained by substituting Eq. (4.1) and Eq. (4.2) in the relative phase delay expression of Eq. (A.5), yielding

$$\begin{aligned} \delta_{lm} &= -(m-1)k\frac{\lambda}{2}\cos(\theta_l)\cos(\phi_l) \\ &= -(m-1)\pi\cos(\theta_l)\cos(\phi_l) ; \\ l &= 1, 2, \dots, L ; \quad m = 1, 2, \dots, M \end{aligned} \quad (4.3)$$

With only azimuthal DOA (elevation angle $\theta_l = 0^\circ$), the relative phase delay reduces to [54]

$$\delta_{lm} = -(m-1)\pi\cos(\phi_l) ; l = 1, 2, \dots, L ; \quad m = 1, 2, \dots, M \quad (4.4)$$

Let $x_{lm}(t)$ denote the noiseless output signal due to the l^{th} path signal excitation from the m^{th} element of an M -element array, then¹³

$$\begin{aligned} x_{lm}(t) &= s_l(t)\exp(j\delta_{lm}) ; \\ l &= 1, 2, \dots, L ; \quad m = 1, 2, \dots, M \end{aligned} \quad (4.5)$$

¹³ Strictly speaking should be proportional to, as described in appendix A

where $s_l(t)$ is the l^{th} path signal.

The noiseless output signal from each element of the M -element array can be conveniently put in a vector form such that

$$\mathbf{x}_l(t) = s_l(t) \mathbf{d}_l ; l = 1, 2, \dots, L \quad (4.6)$$

where

$\mathbf{x}_l(t)$ is the received signal vector, defined as $\mathbf{x}_l^T(t) = [x_{l1}(t) \cdots x_{lm}(t) \cdots x_{lM}(t)]$ (superscript T denotes transposition),

and $\mathbf{d}_l$ is the l^{th} path *direction manifold* (vector), defined as

$$\begin{aligned} \mathbf{d}_l &= \begin{bmatrix} \exp(j\delta_{l1}) \\ \vdots \\ \exp(j\delta_{lm}) \\ \vdots \\ \exp(j\delta_{lM}) \end{bmatrix} \\ &= \begin{bmatrix} 1 \\ \vdots \\ \exp(-j(m-1)\pi \cos(\phi_l)) \\ \vdots \\ \exp(-j(M-1)\pi \cos(\phi_l)) \end{bmatrix} \end{aligned} \quad (4.7)$$

$$l = 1, 2, \dots, L$$

A linear array of M elements with $\frac{\lambda}{2}$ interelement spacing has a total linear dimension of $(M-1)\frac{\lambda}{2}$.

4.1.2 Ring Antenna Array

When a linear antenna array is bent around a circle, a ring antenna array is obtained as in Figure 4.2. The ring array interelement separation is assumed uniform and equal to $\frac{\lambda}{2}$.

Figure 4.2 An M -element ring array structure with an incident l^{th} path plane wave

For the derivation of the received signal vector, the center of the ring is taken as the reference. The position vector $\mathbf{r}_m$ of the m^{th} element is¹³

$$\begin{aligned}\mathbf{r}_m &= r \cos(\varphi_m) \hat{\mathbf{x}} + r \sin(\varphi_m) \hat{\mathbf{y}} \\ &= r \cos\left(\frac{2\pi}{M}(m-1)\right) \hat{\mathbf{x}} + r \sin\left(\frac{2\pi}{M}(m-1)\right) \hat{\mathbf{y}} ; \\ m &= 1, 2, \dots, M\end{aligned}\quad (4.8)$$

where r is the ring radius and M is the number of elements in the array.

The DOA unit vector $\hat{\mathbf{a}}_l$ of the l^{th} path is similar to that of a linear array, i.e.,

$$\begin{aligned}\hat{\mathbf{a}}_l &= \cos(\theta_l) \cos(\phi_l) \hat{\mathbf{x}} + \cos(\theta_l) \sin(\phi_l) \hat{\mathbf{y}} + \sin(\theta_l) \hat{\mathbf{z}} ; \\ l &= 1, 2, \dots, L\end{aligned}\quad (4.9)$$

Putting Eq. (4.8) and Eq. (4.9) in Eq. (A.5), the relative phase delay δ_{lm} between the m^{th} element and the reference element, after simplification, is

$$\begin{aligned}\delta_{lm} &= k r \left(\cos(\theta_l) \cos\left(\phi_l - \frac{2\pi}{M}(m-1)\right) \right) ; \\ l &= 1, 2, \dots, L ; \quad m = 1, 2, \dots, M\end{aligned}\quad (4.10)$$

An M -element ring array with $\lambda/2$ interelement separation has a radius r of

$$r = \frac{\lambda}{4 \sin\left(\frac{\pi}{M}\right)} \quad (4.11)$$

or for large M ¹⁴,

$$r = \frac{M\lambda}{4\pi} \quad (4.12)$$

Thus, the diameter of an M -element ring array is about π factor smaller than the length of an M -element linear array.

Substituting the ring radius expression into Eq. (4.10) produces a new expression for the relative phase delay δ_{lm} as

$$\delta_{lm} = \frac{\pi}{2 \sin\left(\frac{\pi}{M}\right)} \left(\cos(\theta_l) \cos\left(\phi_l - \frac{2\pi}{M}(m-1)\right) \right) \quad (4.13)$$

¹³ 1st element is on the positive x-axis.

¹⁴ $M \geq 10$, such that $\sin\left(\frac{\pi}{M}\right) \approx \frac{\pi}{M}$.

which can be further simplified for large M as

$$\delta_{lm} = \frac{M}{2} \left(\cos(\theta_l) \cos\left(\phi_l - \frac{2\pi}{M}(m-1)\right) \right) \quad (4.14)$$

For the azimuthal DOA ($\theta_l = 0^\circ$), the relative phase delay reduces to

$$\delta_{lm} = \frac{M}{2} \cos\left(\phi_l - \frac{2\pi}{M}(m-1)\right) \quad (4.15)$$

Similar to the linear array, the received signal vector of an M -element ring array due to the l^{th} path signal excitation is

$$\mathbf{x}_l(t) = s_l(t) \mathbf{d}_l ; l = 1, 2, \dots, L \quad (4.16)$$

where $\mathbf{d}_l$ is the l^{th} path direction manifold for a ring array such that

$$\mathbf{d}_l = \begin{bmatrix} \exp\left(j \frac{\pi}{2 \sin(\frac{\pi}{M})} \cos(\theta_l) \cos(\phi_l)\right) \\ \vdots \\ \exp\left(j \frac{\pi}{2 \sin(\frac{\pi}{M})} \cos(\theta_l) \cos\left(\phi_l - \frac{2\pi}{M}(m-1)\right)\right) \\ \vdots \\ \exp\left(j \frac{\pi}{2 \sin(\frac{\pi}{M})} \cos(\theta_l) \cos\left(\phi_l - \frac{2\pi}{M}(M-1)\right)\right) \end{bmatrix} ; \quad (4.17)$$

$$l = 1, 2, \dots, L$$

In practice, the identical element radiation pattern assumption may not always be true because of the presence of *mutual couplings* between the array elements that vary from the center to the periphery of the array. In general, the closer the interelement spacing, the more serious is the mutual coupling effects on the total radiation pattern. A good discussion of the mutual coupling effects on adaptive arrays is given in [55].

In subsequent discussions, it is assumed that:

- each antenna element is *omnidirectional*¹⁵,

¹⁵ Omnidirectional means having a uniform radiation pattern in the horizontal (azimuthal) plane.

- mutual couplings are negligible,
- and antenna operation is in the *far field* (Fraunhofer zone)¹⁶.

4.2 Adaptive Array Narrowband Beam-Forming

The unweighted antenna array can be coupled with an *adaptive combiner* to form an M -element *narrowband beam-former* as depicted in Figure 4.3. The received signal is first frequency translated and sampled. The noiseless k^{th} sampled outputs of the array elements due to the l^{th} path signal excitation is

$$\mathbf{x}_l[k] = s_l[k]\mathbf{d}_l; \quad l = 1, 2, \dots, L \quad (4.18)$$

where

$\mathbf{x}_l[k]$ is the received signal vector, defined as $\mathbf{x}_l^T[k] = [x_{l1}[k] \cdots x_{lm}[k] \cdots x_{lM}[k]]$ ($x_{lm}[k]$ is the output signal from the m^{th} element),

and $\mathbf{d}_l$ is the l^{th} path direction manifold given in Eq. (4.7) and Eq. (4.17) for linear and ring array respectively.

Each element output signal $x_{lm}[k]$ is scaled by an adjustable complex weight w_m of the adaptive combiner that modifies the received signal amplitude and phase prior to combining. Denoting the weights by a complex vector $\mathbf{w}$, $\mathbf{w}^T[k] = [w_1[k] \cdots w_m[k] \cdots w_M[k]]$, the combiner output signal $y[k]$ ¹⁷ can be written as

$$\begin{aligned} y[k] &= \sum_{l=1}^L \mathbf{w}^H[k] \mathbf{x}_l[k] \\ &= \mathbf{w}^H[k] \sum_{l=1}^L s_l[k] \mathbf{d}_l \end{aligned} \quad (4.19)$$

where superscript H denotes Hermitian operation and L is the number of signal paths.

The output signal $y[k]$ can be maximized by phase aligning the individual element received signal of each element prior to combining. For a single path ($L = 1$) channel,

¹⁶ In indoor operation, the presence of conducting structures in the vicinity of the array could potentially alter the total radiation pattern. This issue is not considered in this thesis.

¹⁷ ignoring the effect of receive filter.

Figure 4.3 An M -element adaptive array narrowband beam-former

this can be accomplished by setting

$$\mathbf{w} = \mathbf{d} \quad (4.20)$$

which produces a maximum output

$$y_{max}[k] = Ms[k] \quad (4.21)$$

With a proper (optimum) choice of weight vector, the adaptive antenna array output SNR can be made to exceed that of a single antenna. Let σ_v^2 be the noise power at each antenna element, the combiner output noise power then is $M\sigma_v^2$ (the noise combine incoherently). The optimum combiner output SNR can be derived as

$$\begin{aligned} SNR_O &= 10\log_{10}\left(\frac{|y[k]|^2}{M\sigma_v^2}\right) \\ &= 10\log_{10}\left(\frac{M^2|s[k]|^2}{M\sigma_v^2}\right) \\ &= SNR_I + (10\log_{10}M) \end{aligned} \quad (4.22)$$

Thus, optimum combining provides a $10\log_{10}M$ dB improvement in the output SNR. For a 2-element array ($M = 2$), there is a 3 dB gain in the output SNR. For 10 elements, the gain could be as high as 10 dB.

Since the signal path DOA is normally unknown, the direction manifold $\mathbf{d}$ and hence the optimum weight $\mathbf{w}$ can not be pre-determined. A pilot training signal $d[k]$ can be used to train the array (adjust the weights) with an adaptive algorithm to create beams in the desired path directions. The weights are usually adjusted to minimize the mean-square of the error signal or $\langle e^2[k] \rangle = \langle |d[k] - y[k]|^2 \rangle$. Adaptive beam-forming is further discussed in the next chapter.

4.2.1 Antenna Array Radiation Pattern

The main concern of an antenna array for indoor wireless applications is the sensitivity of the array to signals arriving from different path directions. The *array radiation pattern* is a directional power pattern which is defined as the spatial variation of the relative sensitivity of the array to signals of a specified frequency arriving from different directions [24]. Expressed mathematically, the array radiation pattern (normalized) $G(\theta, \phi)$ (in dB) is

$$G(\theta, \phi) = 20 \log_{10} \left(\frac{|\mathbf{w}^H \mathbf{d}(\theta, \phi)|}{M} \right) \text{ (in } dB) \quad (4.23)$$

The normalized array radiation pattern is computed from the array output due to a unit amplitude signal from different angle of arrivals for a fixed (converged) weight vector.

The “*peak*” responses in the array pattern are referred to as “*beams*” in this thesis. The *half-power beamwidth* is defined as the angle between two adjacent directions (in the azimuthal plane) on a beam with radiation intensity 3 dB below the maximum intensity of that beam. The simulations in the following sections involve array patterns in the azimuthal plane ($\theta = 0^\circ$).

4.3 Single-Path Beam-Forming with Known DOA

This section discusses the beam-forming with the linear and the ring array for a single path with known DOA. The weight vectors are computed from Eq. (4.7) and Eq. (4.17) for a linear and a ring array respectively. The linear and the ring array beamwidth at different azimuthal DOA's are compared for arrays with different number of elements.

4.3.1 Linear Array Single-Path Beam-Forming

An array pattern (polar plot) with a beam at 90° DOA is shown in Figure 4.4a for a linear array with 2, 5, 10 and 20 elements. The linear array is positioned along the horizontal line (x-axis) in the azimuthal plane. In three-dimensions, the array pattern

of a linear array is a *figure of revolution* about the axis of the array. This *cylindrical symmetry* appears in the plane polar plot as beam mirror images with respect to the array axis. Subsidiary sidelobes are unnoticeable in the polar plot. The mainlobe beamwidth is noted to decrease as the number of array elements increases from 2 to 20.

Figure 4.4b shows the array pattern for a different path DOA at 45° . For a 2-element array, the sidelobe level is rather prominent. The mainlobe beamwidth is similarly observed to decrease with increasing number of array elements.

The array pattern for a path DOA at 0° (endfire direction) is plotted in Figure 4.4c. Comparing the three plots, it can be deduced that the mainlobe beamwidth increases monotonically from the broadside (90°) direction to the endfire (0°) direction.

To evaluate the sidelobe level, a linear plot of a 5-element array pattern with main beam at 90° DOA is depicted in Figure 4.5. It is noted that the highest sidelobe level is about 13 dB below the peak of the mainlobe.

4.3.2 Ring Array Single-Path Beam-Forming

Array patterns from a 5, 10 and 20-element array for a single path DOA at 90° , 45° and 0° are plotted in Figure 4.6a, b and c respectively. A 2-element ring array is not shown because it is just another linear array discussed previously.

Similar to the linear array, the mainlobe beamwidth is observed to decrease as the number of array elements increases from 5 to 20 elements. Comparing the three plots in Figure 4.6, it can be deduced that the mainlobe beamwidth stays relatively constant from the broadside to the endfire direction. Because of the discrete element distribution, the beam pattern in general, is not exactly symmetrical with respect to the DOA axis (Figure 4.6b), unless the DOA happens to be parallel with the position vector of one of the ring element (Figure 4.6a and c).

Figure 4.7 shows a linear plot of a 5-element ring array pattern for a path DOA at 90° . It is seen that the highest sidelobe level is about -5 dB, which is higher than that of a linear array (-13 dB).

a)

b)

c)

Figure 4.4 Array pattern for a single path, polar plot, linear array, $d = \frac{\lambda}{2}$, a) DOA = 90°, b) DOA = 45°, c) DOA = 0°

Figure 4.5 Array pattern for a single path, linear plot, linear array, # elements = 5, $d = \frac{\lambda}{2}$, DOA = 90°

a)

b)

c)

Figure 4.6 Array pattern for a single path, polar plot, ring array, $d = \frac{\lambda}{2}$, a) $\text{DOA} = 90^\circ$, b) $\text{DOA} = 45^\circ$, c) $\text{DOA} = 0^\circ$

Figure 4.7 Array pattern for a single path, linear plot, ring array, # elements = 5, $d = \frac{\lambda}{2}$, DOA = 90°

4.3.3 Linear and Ring Array Beamwidth Comparison

The variation of the mainlobe beamwidth at different azimuthal DOA's is plotted in Figure 4.8a for a linear array and in Figure 4.8b for a ring array, for different number of array elements. For a linear array, it can be deduced that the beamwidth shrinks dramatically from the endfire (0°) to the broadside (90°) position. The beamwidth is relatively constant in the vicinity of the broadside direction. For a ring array, the beamwidth is noted to be fairly constant as the beam swings from 0° to 360° . Both figures indicate the narrowing of beamwidth with increasing number of array elements (wider aperture).

A 10-element linear array would have a mainlobe beamwidth of about 10° in the broadside region. With a 10 GHz carrier frequency, the array length is 13.5 cm, which is still too large to fit in a handheld portable. A 10-element ring array would produce a mainlobe beamwidth of about 25° . The ring diameter can be calculated from Eq. (4.11) as 4.9 cm, which can fit quite nicely in a handheld portable. Thus, for low frequency of operation (smaller than 10 GHz), the array is recommended at the base station.

4.4 Multiple-Path Beam-Forming with Known DOA's

Beam-forming for a single path with known DOA has been demonstrated in the previous sections. In a multipath environment, a multiple-beam system is needed to capture the individual path signals. The conventional multiple-beam concept refers to the collection of main beams from each array pattern created by a *multi-port* feed system, such as the one described in section 2.2. As defined in section 4.2.1, multiple beams in this thesis refer to the multiple "peak" power responses of an array pattern created by a single port (phased-array feed).

One simple technique to create a multiple beam is to use a *linear combination* of weight vectors, each of which is the weight vector for the individual beam, i.e.,

$$\mathbf{w} = \sum_{l=1}^L c_l \mathbf{w}_l \quad (4.24)$$

Figure 4.8 Mainlobe beamwidth variation vs. azimuthal angle, $d = \frac{\lambda}{2}$, a) linear array, b) ring array

where c_l is a constant that can be set to $1/L$ for all l 's.

As an example, if w_{45} , w_{90} and w_{135} denote the weight vector for a 45° , 90° and 135° beam respectively, then a normalized weight vector to generate beams for path DOA's at 45° , 90° and 135° is $w = \frac{(w_{45} + w_{90} + w_{135})}{3}$. The multiple-beam forming using this technique is demonstrated in the next sections.

4.4.1 Linear Array Multiple-Path Beam-Forming

The number of beams that can be formed by an array depends on the beamwidth and the separation between the path DOA's. Without loss of generality, an array pattern for path DOA's at 45° , 90° and 135° is shown in Figure 4.9a for a 10-element linear array and in Figure 4.9b for a 20-element linear array.

As indicated previously, larger aperture (larger number of elements) produces smaller beamwidth or finer resolution. Also, the sensitivity of the beams seems to be better with larger number of array elements (cf. Figure 4.9a and b).

4.4.2 Ring Array Multiple-Path Beam-Forming

As in the linear array case, an array pattern for path DOA's at 45° , 90° and 135° is shown in Figure 4.10a for a 10-element ring array. As seen, a 10-element ring array is incapable of forming the three beams. Since the multiple-beam pattern is generated from a linear combination of individual beam patterns, it is conjectured that some sidelobes from other individual array patterns could have diminished the desired main beam of an array pattern (when the number of elements is small). Thus, the intended beam at 90° DOA is missing from Figure 4.10a. The sidelobe levels in the ring array pattern are also noted to be higher than those in the linear array pattern of the previous section.

A 20-element ring array pattern for the same path DOA's is plotted in Figure 4.10b, showing the three beams at the path DOA's. From the two examples, it can be concluded that the number of elements needed to form a given number of beams in a ring array is relatively high, when compared to a linear array.

a)

b)

Figure 4.9 Array pattern for three paths, polar plot, linear array, $d = \frac{\lambda}{2}$, DOA = 45°, 90° and 135°, a) # elements = 10, b) # elements = 20

a)

b)

Figure 4.10 Array pattern for three paths, polar plot, ring array, $d = \frac{\lambda}{2}$, DOA = 45°, 90° and 135°, a) # elements = 10, b) # elements = 20

As a chapter summary, the ring array which has the capacity to direct unique beams to any points in space is preferred for a multipath environment with CCI than the linear array which suffers from cylindrical symmetry in its array pattern. The ring array also has the advantage of constant beamwidth in the plane of the ring, whereas the linear array beamwidth varies from the broadside to the endfire direction. In addition, the diameter of a ring array is about π factor smaller than the length of a similar linear array. Thus, the subsequent discussion is focused on the ring array. The ring array, however, requires a larger number of elements for a given beamwidth.

The above simulations confirm the feasibility of beam-forming for multiple paths with known DOA's. The next chapter discusses beam-forming for multiple desired paths with unknown DOA's that are commonly found in an indoor wireless environment.

Chapter 5

Adaptive Multiple-Beam Forming

As mentioned, the unknown time-varying characteristics of an indoor multipath channel require an adaptive technique to train the array to search for and form beams on the desired signal paths. In this chapter, the *optimum* weight solution for beam-forming in a multipath environment and the LMS/RLS adaptive algorithms in searching the optimum solutions are first discussed. Adaptive beam-forming simulations and results are then presented for an indoor multipath with CCI.

5.1 Optimum Solution for Beam-Forming in a Multipath Environment

The array beam-forming problem in a multipath environment can be formulated as a multivariable *minimum mean-square-error* (MMSE) problem, whose optimum solution can be obtained using standard optimization techniques.

Consider again the M -element adaptive array of Figure 4.3 in section 4.2. Incident on the array are a desired signal, and N cochannel interfering signals¹⁸. The desired signal has L propagation paths, and the n^{th} interference has L_n propagation paths¹⁹.

The received signal at each element consists of L desired path signals, $\sum_{n=1}^N L_n$ interference path signals and the system noise. Using matrix notations, the sampled array received signal vector $\mathbf{x}[k]$ at the k^{th} instant can be expressed as

$$\mathbf{x}[k] = \mathbf{D}_s \mathbf{s}[k] + \sum_{n=1}^N \mathbf{D}_{in} \mathbf{i}_n[k] + \mathbf{v}[k] \quad (5.1)$$

where

$\mathbf{x}^T[k] = [x_1[k] \cdots x_m[k] \cdots x_M[k]]$ is the array received signal vector (super-script T denotes transposition),

¹⁸ There are usually one or two dominant interferers, and N is typically smaller than 6.

¹⁹ All path signals are assumed to impinge on all elements of the array.

$\mathbf{D}_s = [\mathbf{d}_1 \cdots \mathbf{d}_l \cdots \mathbf{d}_L]$ is the desired signal direction manifold matrix, whose column manifold vectors are defined in section 4.1.2,

$\mathbf{s}^T[k] = [s_1[k] \cdots s_l[k] \cdots s_L[k]]$ is the desired path signal vector, whose elements are defined in Eq. (3.3) of section 3.2,

$\mathbf{D}_{in} = [\mathbf{d}_{n1} \cdots \mathbf{d}_{nl} \cdots \mathbf{d}_{nL_n}]$ is the n^{th} interfering signal direction manifold matrix, whose column manifold vectors are similarly defined in section 4.1.2,

$\mathbf{i}_n^T[k] = [i_{n1}[k] \cdots i_{nl}[k] \cdots i_{nL_n}[k]]$ is the n^{th} interference path signal vector, whose elements are modelled like the desired path signal but with different transmitted symbol information and parameters,

and $\mathbf{v}[k] = [v_1[k] \cdots v_m[k] \cdots v_M[k]]$ is the antenna noise vector with uncorrelated elements.

The received signal vector $\mathbf{x}[k]$ is multiplied by an adjustable complex weight vector $\mathbf{w}[k]$, and combined to form an output signal $y[k] = \mathbf{w}^H[k]\mathbf{x}[k]$ (superscript H denotes Hermitian operation). The weights are adjusted (during training) to minimize the MSE $J[k]$, defined as

$$J[k] = E(e[k]e^*[k]) \quad (5.2)$$

where

$e[k] = d[k] - y[k]$ is the adaptive combiner output error signal, which is the difference between the desired reference signal $d[k]$ and the adaptive combiner output $y[k]$,

superscript $*$ denotes complex conjugate operation,

and $E(\cdot)$ is the expectation operator.

The MSE function is a *multivariable* function of the weight vector whose optimum weight vector $\mathbf{w}_{opt}$ or *Wiener* solution [54] can be obtained by solving (see appendix B)

$$\mathbf{R}\mathbf{w}_{opt} = \mathbf{P} \quad (5.3)$$

where

$\mathbf{R} = E(\mathbf{x}[k]\mathbf{x}^H[k])$ is the *autocorrelation matrix* of the received signal vector $\mathbf{x}[k]$, and $\mathbf{P} = E(\mathbf{x}[k]d^*[k])$ is the *cross-correlation vector* between the received signal vector $\mathbf{x}[k]$ and the desired reference signal $d[k]$.

Eq. (5.3) is a system of linear equations which can be solved if the matrix $\mathbf{R}$ is nonsingular. Furthermore, if the matrix $\mathbf{R}$ is also well-conditioned, numerical solution techniques can be used. The $\mathbf{R}$ and $\mathbf{P}$ matrices can be estimated if the multipath channel characteristics and the noise correlation are known. Since this information is usually unavailable, the LMS and the RLS adaptive algorithms (to be described in the next section) are used to search for the optimum solution.

5.2 Adaptive Algorithms for Multiple-Beam Forming

Various adaptive algorithms can be used to search for the optimum solution for beam-forming in a multipath environment. Two most commonly used algorithms: the *least mean square* (LMS) and the *recursive least square* (RLS) adaptive algorithms, are discussed in the next two sections.

5.2.1 Least Mean Square Algorithm

The complex LMS algorithm was first formulated by Widrow [56] for searching the minimum point of a *quadratic* MSE function. It is based on the *steepest descent* technique that adjusts the weight vector $\mathbf{w}$ in a direction opposite (negative) to the *gradient* of the MSE function or towards the minimum MSE (optimum weight) point. The weight vector $\mathbf{w}$ is updated iteratively such that

$$\mathbf{w}[k] = \mathbf{w}[k-1] - \frac{1}{2}\mu_k \nabla_{\mathbf{w}} J[k] \quad (5.4)$$

where

μ_k is the LMS step size that determines the amount of adjustment towards the optimum weight,

and $\nabla_{\mathbf{w}}$ is the gradient operator with respect to $\mathbf{w}$.

The factor $\frac{1}{2}$ is included in the above equation so that the final equation will have only one arbitrary constant (μ_k).

Since the MSE function is usually unknown, a noisy instantaneous estimate of the gradient (Eq. (B.18) of appendix B)

$$\nabla_{\mathbf{w}} J[k] \approx 2e^*[k]\mathbf{x}[k] \quad (5.5)$$

is used in the weight update equation (Eq. (5.4)), yielding

$$\mathbf{w}[k] = \mathbf{w}[k-1] + \mu_k e^*[k]\mathbf{x}[k] \quad (5.6)$$

which is the weight update equation for the complex LMS algorithm.

To summarize, the LMS algorithm equations are:

$$\begin{aligned} y[k] &= \mathbf{w}^H[k]\mathbf{x}[k] \\ e[k] &= d[k] - y[k] \\ \mathbf{w}[k] &= \mathbf{w}[k-1] + \mu_k e^*[k]\mathbf{x}[k] \end{aligned} \quad (5.7)$$

Initial condition:

$$\mathbf{w}^T[0] = [w_1[0] \ \cdots \ w_m[0] \ \cdots \ w_M[0]] \quad (5.8)$$

The initial weight vector $\mathbf{w}^T[0]$ is usually set to zero vector for simplicity. The step size parameter μ ²⁰ determines the rate of convergence, the converged MSE and the stability of the algorithm. In general, a small value for μ produces a small MSE but with slow convergence. Conversely, a large value for μ (within a limit) produces a larger MSE but results in a faster convergence.

The choice of μ value depends on the number of array elements and the multipath channel characteristics through the *eigenvalue spread* of the autocorrelation matrix $\mathbf{R}$. In general, a large eigenvalue spread produces a slower convergence, and a small spread produces a faster convergence. For stable operation, μ is constrained to [54, 32]

$$0 < \mu < \frac{2}{M\sigma_x^2} \quad (5.9)$$

²⁰ Subscript k is dropped for convenience.

where M is the number of array elements and σ_x^2 is the power (variance) of the received signal $x[k]$ at an element.

As a rule of thumb, the inverse of the product of the number of array elements and the received signal power at an element is commonly used for the step size μ , i.e.,

$$\mu = \frac{1}{M\sigma_x^2} \quad (5.10)$$

It can be shown that the LMS algorithm converges to the optimum weight solution in a statistical sense [54]. In practice, because of the presence of noise and CCI, the "converged" weight vector wanders in the neighborhood of the optimum weight. The LMS algorithm can also track the input signal with slow changing statistics.

The LMS weight update equation does not require the pertinent correlation functions ($\mathbf{R}$ and $\mathbf{P}$) or any matrix inversion. Its implementation in software or hardware is also very simple. The LMS performance, however, depends on the input signal characteristics and the number of unknown weights (variables). It also tends to have a slow rate of convergence. A faster algorithm based on the *Kalman* filtering theorem is presented in the next section.

5.2.2 Recursive Least Square Algorithm

As pointed, the LMS algorithm suffers from a slow rate of convergence because it does not make full use of all the information available at the input of the combiner. The RLS algorithm attempts to remedy this [54].

A full derivation of the RLS algorithm can be found in [54]. The RLS algorithm weight update equations are summarized below:

$$\begin{aligned} \mathbf{g}[k] &= \mathbf{K}[k-1]\mathbf{x}[k] \left(\mathbf{x}^H[k]\mathbf{K}[k-1]\mathbf{x}[k] + J_{min} \right)^{-1} \\ \epsilon[k] &= d[k] - \mathbf{w}^H[k-1]\mathbf{w}[k] \\ \mathbf{w}[k] &= \mathbf{w}[k-1] + \mathbf{g}[k]\epsilon^*[k] \\ \mathbf{K}[k] &= \mathbf{K}[k-1] - \mathbf{g}[k]\mathbf{x}[k]\mathbf{K}[k-1] \end{aligned} \quad (5.11)$$

where

$\mathbf{g}[k]$ is the Kalman gain vector,

$\mathbf{K}[k]$ is the correlation matrix of the weight-error vector,

$\mathbf{x}[k]$ is the received signal vector,

J_{min} is the minimum MSE,

$e[k]$ is the combiner output error signal,

$\mathbf{w}[k]$ is the weight vector,

and $d[k]$ is the desired reference signal.

Assumption

$$J_{min} = 0.001 \text{ to } 0.01 \quad (5.12)$$

Initial conditions

$$\begin{aligned} \mathbf{w}^T[0] &= [0 \cdots 0 \cdots 0], \\ \mathbf{K}[0] &= c\mathbf{I}, \quad c > 0 \end{aligned} \quad (5.13)$$

Note from Eq. (5.11) that the weight update equation of the RLS algorithm employs a Kalman gain vector $\mathbf{g}[k]$ which is computed from the current and previous received signal information. This will result in a faster convergence as shown in the simulations. The LMS algorithm, however, only uses the current received signal information for weight update (Eq. (5.6)).

The LMS and RLS algorithms require the calculation of an error signal, and the computation of a new set of M tap weights for each output sample. A comparison of the total number of complex calculation per output sample for the two algorithms is shown in the following table:

Algorithm	Number of complex operations per output sample		
	Multiplications	Divisions	Additions
LMS	$2M + 1$	0	$2M$
RLS	$2M^2 + 3M$	$M^2 + 2M$	$2M^2 + 2M - 1$

Table 5.1 LMS and RLS algorithm computational complexity comparison

It can be seen that the RLS algorithm is computationally more expensive than the LMS algorithm.

In practice, the choice of adaptation algorithm is determined by:

- rate of convergence and tracking ability,
- operational signal environment, such as the availability of a priori signal path DOA information, etc.,
- and implementation complexity and cost.

The next section presents beam-forming simulations with the adaptive algorithms under various operating scenarios.

5.3 Adaptive Beam-Forming Simulations

A simulation software was written in C programming language [57] to implement the adaptive ring array beam-forming. The simulation procedures is described in the next section.

5.3.1 Simulation Procedures

A block diagram flowchart of the simulation procedure is sketched in Figure 5.1.

The simulation starts by reading the following simulation parameters from a file:

- number of array elements,
- number of CCI (0– N),
- *signal-to-noise* ratio (SNR),
- *signal-to-interference* ratio (SIR) for each CCI,
- number of desired and CCI signal paths,

Figure 5.1 Adaptive multiple-beam forming simulation flowchart

- path DOA's (in azimuth only),
- training length (in number of symbols),
- and algorithm specific parameters such as μ and J_{min} .

After initialization, a pilot training signal $c[k]$ consisting of a pseudorandom BPSK (± 1) sequence²² is transmitted. The transmitted signal is shaped by a square root raised cosine (SRRC) filter with 25% roll-off factor (see appendix C.2), and then modified by a narrowband multipath channel that creates multiple path signals with different path loss, phase delay and path DOA.

The signal at each element is multiplied by a ring array factor that represent the phase delay due to path DOA. *Additive white Gaussian noise* (AWGN) and CCI are then added at each element. The SNR (see appendix C.3) at an array element is defined as

$$\begin{aligned} SNR &= 10\log_{10}\left(\frac{\sigma_s^2}{\sigma_v^2}\right) \\ &= 10\log_{10}\left(\frac{|c[k]|^2}{\sigma_v^2}\right) \end{aligned} \quad (5.14)$$

where

σ_s^2 is the variance of the training signal,

$|c[k]|^2$ is the average power of the training signal,

and σ_v^2 is the variance of the noise samples.

The n^{th} CCI SIR (see appendix C.3) is similarly defined as

$$\begin{aligned} SIR &= 10\log_{10}\left(\frac{\sigma_{i_n}^2}{\sigma_{i_n}^2}\right) \\ &= 10\log_{10}\left(\frac{|c[k]|^2}{|i_n[k]|^2}\right) \end{aligned} \quad (5.15)$$

where $\sigma_{i_n}^2$ is the variance of the n^{th} CCI and $|i_n[k]|^2$ is the average power of the n^{th} CCI.

The corrupted signal is passed through a matched SRRC receive filter with 25% roll-off factor.

²² BPSK is used for simplicity.

The reference signal $d[k]$ is set equal to the training sequence $c[k]$ for training the array to look in the pilot signal DOA's with the LMS/RLS algorithm. The resulting array pattern (defined in section 4.2.1) is plotted at the end of the training period.

5.3.2 Algorithm Rate of Convergence

The adaptive algorithm rate of convergence is an important issue in applying the adaptive array for an indoor wireless system where only a very short time is allowed for training. In this section, the rate of convergence of the LMS and the RLS algorithms with zero weight vector initialization is investigated for beam-forming for a single path. The rate of convergence is measured in terms of the instantaneous squared error $e^2[k]$ (squared error from one realization). The simulation are given in the following table:

	experiment #1 (LMS)	experiment #2 (LMS & RLS)
# elements	10	10
# desired paths	1	1
desired path DOA	90°	90°
SNR	10 dB	20 dB
CCI	none	none
simulation length	300 symbols	300 symbols

Table 5.2 LMS and RLS algorithm rate of convergence simulation parameters

In the first experiment, the time evolution of the instantaneous squared error $e^2[k]$ for the LMS algorithm is plotted in Figure 5.2a for four different step size μ 's at 10 dB SNR. The required number of iterations for convergence is defined as the number of iterations required to reach a "flat" slope squared error level. This number for different step size μ 's is entered in Table 5.3. It is seen that the larger the step size μ , the faster is the rate of convergence. For a single-path beam-forming system, a step size of 0.05 would produce a reasonable fast rate of convergence.

Figure 5.2 Single path squared error time evolution, ring array, # elements = 10, LMS, training = 300 symbols, DOA = 90° , a) SNR = 10 dB, b) SNR = 20 dB

LMS step size, μ	# iterations for convergence
0.001	~230
0.005	~50
0.01	~25
0.05	~8
RLS	~5

**Table 5.3 LMS and RLS algorithm rate of convergence
for a 10-element ring array with a signal path**

As a reminder (see section 5.2.1), the step size μ value to produce a stated number of iterations for convergence depends on the number of array elements and the total power of the incoming signals.

The rate of convergence is also tested in a second experiment for a different SNR value (20 dB). Comparing Figure 5.2a (SNR = 10 dB) with Figure 5.2b (SNR = 20 dB), it can be concluded that the rate of convergence is independent of the SNR. However, with higher SNR, the curves are smoother and the converged squared errors become smaller.

The above findings for the LMS algorithm in beam-forming are in line with the theoretical predictions [58], that state the LMS rate of convergence is inversely proportional to the step size μ . Also, the larger the step size μ , the larger is the variance of the converged weight.

The rate of convergence of the LMS and the RLS algorithms for single-path beam-forming are shown in Figure 5.3. It is seen that the RLS algorithm has a fairly short (approx 5 iterations) convergence time as compared with the LMS algorithm with different step size μ 's.

These results also agree with the theoretical predictions which says that the RLS algorithm requires about one tenth the number of iterations for the LMS algorithm for convergence. Furthermore, the RLS algorithm is inherently stable, whereas the LMS algorithm stability depends on the step size μ [54].

Figure 5.3 Single path squared error time evolution, ring array, # elements = 10, LMS and RLS, training = 300 symbols, DOA = 90° , SNR = 20 dB

Although the RLS algorithm offers fast rate of convergence, it is computationally expensive. Unless a very fast rate of convergence is required, the LMS algorithm would be sufficient and easier to implement. Also, in view of the lengthy simulation time, only the LMS algorithm is used in the subsequent simulations.

5.3.3 Adaptive Single-Path Beam-Forming

Adaptive beam-forming for a single path is first investigated using a ring array. At the start of training, the ring array can be initialized with an omnidirectional (initial weights all set to $\frac{1}{M}$) or a null (initial weights all set to 0) pattern²³. The performance of the two weight initialization scheme is studied as part of the single-path beam-forming experiments, with the simulation parameters set as follows:

	experiment #1 null-pattern initialization	experiment #2 omnidirectional initialization
# elements	10	10
# desired paths	1	1
desired path DOA	90°	90°
SNR	10 dB	10 dB
CCI	none	none
training length	20 symbols	20 symbols

Table 5.4 Omnidirectional vs. null-pattern initialization simulation parameters

Figure 5.4a depicts two converged array patterns, one for the null-pattern initialization and the other for the omnidirectional-pattern initialization in polar plot. The converged main beam points towards the 90° DOA, with a beamwidth of approximately 25°. The converged patterns for both experiments with 20 symbols of training are

²³ Null pattern also means no signal reception at start.

fairly similar, except the omnidirectional-pattern initialization produces slightly higher (approx. 3 dB) sidelobes than the null-pattern initialization (Figure 5.4b).

The sidelobe level usually varies slightly from one realization (simulation) to the other, possibly due to the presence of AWGN. But, in general it can be said that null-pattern initialization would probably be better for array pattern convergence performance. This could be attributed to the fact that the null-pattern initialization starts out from scratch and attempts to form beam pattern with training, while the omnidirectional-pattern initialization strives to restrict response to the path directions by creating notches in other directions of the initial omnidirectional pattern. The converged array pattern would be the same if the optimum solution is unique and sufficient training time is allowed (which varies from case to case).

The subsequent simulations use zero weight vector (null-pattern) initialization for simplicity.

5.3.3.1 Adaptive Single-Path Tracking

Since the portable could be moving around, which causes the signal path DOA's to change, some sort of signal path tracking is needed. Moreover, it would also be desirable to be able to switch beams off/on when old signal paths disappear or new signal paths emerge.

To simplify matters, a single-path tracking is investigated using the adaptive array. The array is first trained to point a beam at 30° path DOA. To simulate the change in the path DOA, the DOA angle is varied from 30° to 45° in equal steps of 5°²⁴. The tracking is simulated by keeping the initial weight for each new training cycle continuous (not reset to zero) from the previous training cycle. The simulation parameters are shown in the following table:

²⁴ such that the discrete beam is visible with a 10-element array and a 20-element array.

Figure 5.4 Converged array pattern for a single path, ring array, # elements = 10, LMS, $\mu = 0.05$, training = 20 symbols, DOA = 90°, SNR = 10 dB, a) polar plot, b) linear plot

	experiment #1	experiment #2
# elements	10	20
# desired paths	1	1
desired path DOA	varies from 30° to 45° in steps of 5°	varies from 30° to 45° in steps of 5°
SNR	10 dB	10 dB
CCI	none	none
training length	20 symbols each cycle	20 symbols each cycle

Table 5.5 Adaptive single-path tracking simulation parameters

Figure 5.5a shows the discrete beam progression (pointed by arrow) with a 10-element array from the first experiment. A reasonably good tracking is observed, although the beam does not move exactly in locked step with the signal path. It is important to point out that when the number of elements is small (smaller than 5), some broadening effect could result. This is generally due to the non-unique weight solutions obtained from the adaptive search algorithm. The beam broadening is inconsequential in practice, because if there is any interference or noise present, the broadening will be subdued.

A similar path tracking result with a 20-element array (experiment #2) is shown in Figure 5.5b. The beam now is narrower, and is following the movement of the pilot signal path with greater accuracy.

In practice, the path DOA could still be changing during the data transmission after the training period. Since the pilot signal is not available during the data transmission, the decision data could be used instead in a decision-directed mode to continue beam adaptation. This scenario is simulated by first sending a pilot signal to train the array to form a beam at a 30° path DOA. The incoming signal path DOA is then varied from 30° to 45° in steps of 5°, while the array is now trained in a decision-directed mode. The plot in Figure 5.6 shows the beam progression with a 20-element array. The beam still tracks the incoming signal path fairly closely. Thus, a reasonably good path-tracking performance is also achievable via decision-directed training.

a)

b)

Figure 5.5 Converged array pattern for a single-path tracking from 30° to 45° in steps of 5° , ring array, LMS, $\mu = 0.05$, training = 20 symbols, SNR = 10 dB, a) # elements = 10, b) # elements = 20

Figure 5.6 Converged array pattern for a single-path tracking from 30° to 45° in steps of 5° , ring array, # elements = 20, LMS, pilot training followed by decision-directed training, $\mu = 0.05$, training = 20 symbols, SNR = 10 dB

A 5° step has been used in the simulations. In practice, the change in DOA is usually very small from one frame to the next (see appendix C.1), so a slow or infrequent path tracking might be sufficient.

5.3.4 Adaptive Multiple-Path Beam-Forming without CCI

To evaluate the adaptive array pattern for multipath signal reception in the absence of CCI, a case with four equal strength pilot signal paths from four different directions is simulated using the following simulation parameters:

	experiment #1	experiment #2
# elements	20	20
# desired paths	4	4
desired path DOA	0°, 90°, 180° & 270°	0°, 15°, 60° & 75°
SNR for all paths	10 dB	10 dB
CCI	none	none
training length	20 symbols	20 symbols

Table 5.6 Adaptive multiple-path beam-forming without CCI simulation parameters

In the first experiment, a pilot signal with four clearly separated signal paths is simulated. Figure 5.7 shows the four-path array pattern created with a 20-element ring array²⁵. It is observed that the maximum gain for the four beams are not all equal, probably because of the presence of added noise.

In the second experiment, the four pilot signal paths are spaced closely in two pairs (0° & 15°, 60° & 75°). The converged array pattern is plotted in Figure 5.8 which shows only two beams. The proximity of the pilot signal paths has caused the closely separated adjacent beams to coalesce. The converged beam also appears to point midway between its two constituent adjacent paths.

²⁵ 20 elements for a resolution beamwidth of approximately 12°.

21 of/de 2

PM-1 3½"x4" PHOTOGRAPHIC MICROCOPY TARGET
NBS 1010a ANSI/ISO #2 EQUIVALENT

PRECISIONSM RESOLUTION TARGETS

PIONEERS IN METHYLENE BLUE TESTING SINCE 1974

METED

10000 COUNTY ROAD 8, BURNING WOOD, WV 26037 USA
TEL: 812 485 7151 FAX: 812 485 7057 TLA: 04/0000000

Figure 5.7 Converged array pattern for four paths, polar plot, ring array, # elements = 20, LMS, $\mu = 0.001$, training = 20 symbols, DOA = 0° , 90° , 180° and 270° , SNR = 10 dB

Figure 5.8 Converged array pattern for four paths, polar plot, ring array, # elements = 20, LMS, $\mu = 0.001$, training = 20 symbols, DOA = 0°, 15°, 60° and 75°, SNR = 10 dB

These simulations demonstrate the feasibility of multiple-path beam-forming using the adaptive array with pilot signal training.

5.3.5 Adaptive Multiple-Path Beam-Forming with CCI

In this section, the effect of the CCI on the array pattern is investigated in two experiments, with the simulation parameters set as follows:

	experiment #1	experiment #2
# elements	10	10
# desired paths	2	2
desired path DOA	45° & 135°	45° & 135°
SNR for all desired paths	10 dB	10 dB
# CCI	1	1
# CCI paths	1	1
CCI path DOA	90°	135°
SIR	0 dB	0 dB
training length	20 symbols before, and 20 symbols after CCI injection	20 symbols before, and 20 symbols after CCI injection

Table 5.7 Adaptive multiple-path beam-forming with CCI simulation parameters

The simulation is first performed without CCI for a two equal strength desired paths at 45° and 135° DOA's. The array is then retrained with the previous two desired paths in the presence of CCI.

In the first experiment a CCI is introduced from one of the sidelobe direction of the first converged array pattern. The final array pattern is plotted in Figure 5.9. Referring to the polar plot in Figure 5.9a, the original sidelobe at the 90° DOA (vertical arrow) of the first converged pattern without CCI (solid curve in figure) is suppressed (dotted curve in figure) after the introduction of CCI coming from that direction. This is expected,

because the training algorithm attempts to minimize the output MSE. A linear plot of the same pattern in Figure 5.9b shows approximately 6 dB attenuation in the CCI path.

If the CCI path happens to coincide with one of the desired signal paths, then that particular desired signal path will be suppressed only if it is not the only desired signal path. The second experiment simulates this scenario with a CCI path that coincides with one of the desired signal path at 135° DOA. Figure 5.10a shows a polar plot of the array pattern with CCI (dotted curve) and without CCI (solid curve). A linear plot of the same pattern (Figure 5.10b.) shows approximately 5 dB attenuation in the CCI path.

Thus, it can be concluded that the adaptive MBA array is very adept at excluding the CCI paths.

To complete the discussion, a general comment on the use of antenna arrays to receive multipath signals is discussed in the next section.

5.4 Multipath Signal Reception Using Antenna Arrays

One approach to multipath signal reception using antenna arrays is to divide the task into three stages. The first stage is to estimate the number of desired signal paths. Next, the DOA of each path is determined. Finally, multiple beams are created at the desired paths to intercept the desired signal.

The number of desired signal paths can be estimated using *parameter estimation* methods based on *Akaike's information-theoretic criterion* or *minimum description length criterion* [54].

Several *spectral estimation* methods such as the conventional FFT method [59] and the superresolution *multiple signal classification* (MUSIC) [54] method, can be used to estimate the path DOA's. The FFT method is simpler, but suffers from high sidelobes, biased estimate and poor resolution. The superresolution method gives excellent resolution at the expense of intensive computations.

The technique described in section 4.4 could be used to create multiple beams with a ring array.

Figure 5.9 Converged array pattern for two paths, # elements = 10, LMS, $\mu = 0.005$, training = 40 symbols (total), desired DOA = 45° & 135° , SNR = 10 dB, CCI DOA = 90° , SIR = 0 dB, a) polar plot, b) linear plot

Figure 5.10 Converged array pattern for two paths, ring array, # elements = 10, LMS, $\mu = 0.005$, training = 40 symbols (total), desired DOA = 45° & 135° , SNR = 10 dB, CCI DOA = 135° , SIR = 0 dB, a) polar plot, b) linear plot

The multiple-beam reception of multipath signal in an indoor wireless environment is faced with the following issues:

- changing number of signal paths,
- and time varying direction of arrivals.

In view of the above issues, the adaptive technique with pilot signal training proposed in this thesis is preferable. The adaptive technique will identify the number of desired signal paths, their DOA's and form beams at those DOA's, as demonstrated in the previous sections.

The adaptive arrays can be deployed at base stations or portables. For the portable, the structure shown in Figure 4.3 can be used directly. At the base station, a unique weight vector is needed for each portable so as to be able to uniquely identify each portable's desired signal paths.

The symbol error rate (SER) system performance of the adaptive array for an indoor wireless channel is the subject of the next chapter.

Chapter 6

Adaptive MBA Array System Performance

This chapter describes a system performance simulation of the adaptive array for an indoor wireless channel in the presence CCI. The system performance is evaluated in terms of the *symbol error rate* (SER) for a QPSK signalling scheme. The system model and simulation procedures are first introduced, followed by a presentation and discussion of the simulation results.

6.1 System Model

An indoor wireless cellular system with an adaptive MBA array receiver front-end is modelled in Figure 6.1. The model consists of one desired user, and N potential cochannel users. Each user generates a stream of *binary* data which is QPSK²⁶ modulated. A *square root raised cosine* (SRRC) filter is used to shape the data symbols before transmission through the indoor wireless channel. In the block diagram, the indoor wireless channel is represented by a narrowband multipath channel block.

An adaptive MBA array is used to intercept the desired multipath signals. The desired path signals are corrupted by *additive white Gaussian noise* (AWGN) and CCI from simultaneous cochannel users. When the array is deployed at a base station, the desired signal is the signal from a portable located within the base station home cell, and the CCI signals are from cochannel cell portables. Similarly, for array at the portable, the desired and CCI signals originate from the portable home cell base station and the cochannel cell base stations respectively.

The received signal is filtered by a matched SRRC receive filter and demodulated to yield an estimated output signal. The output signal is also feedback to the array, to continue the adaptation in a decision-directed mode, after the pilot signal training mode.

²⁶ QPSK is chosen for its simplicity.

Figure 6.1 Indoor wireless cellular communication system model

The blocks in the diagram are discussed in greater detail in the following sections.

6.1.1 Transmitter Subsystem

The transmitter subsystem comprises a user binary data source and a $\frac{\pi}{4}$ -shifted *quadrature-phase-shift keying* (henceforth referred to as QPSK) modulator. User generated bits are combined in pairs (*dibit*), and encoded using *Gray* coding scheme into a stream of QPSK symbols $s(t)$ of the form

$$s(t) = \frac{1}{\sqrt{2}} \exp(j\phi(t)); \quad kT \leq t \leq (k+1)T, \quad k = \dots, -1, 0, 1, \dots \quad (6.1)$$

where $\phi(t)$ is the carrier phase that assumes one of the four equally likely values ($\frac{\pi}{4}$, $\frac{3\pi}{4}$, $\frac{5\pi}{4}$ and $\frac{7\pi}{4}$) for QPSK modulation. The dibit to the QPSK phase and the QPSK complex symbol mapping with Gray coding is tabulated as follows:

Input dibit $kT \leq t \leq (k+1)T$	QPSK phase $\phi(t)$	QPSK complex symbol $s(t)$
11	$\frac{\pi}{4}$	$\frac{1}{\sqrt{2}}(+1 + j)$
01	$\frac{3\pi}{4}$	$\frac{1}{\sqrt{2}}(-1 + j)$
00	$\frac{5\pi}{4}$	$\frac{1}{\sqrt{2}}(-1 - j)$
10	$\frac{7\pi}{4}$	$\frac{1}{\sqrt{2}}(+1 - j)$

Table 6.1 Dibit to QPSK phase and QPSK complex symbol mapping with Gray coding

6.1.1.1 Desired and CCI Signal

As mentioned, the system model includes one desired user and N cochannel users. The desired and cochannel user data bit rate is set at 200 *kbps*. so as to give a QPSK symbol rate of 100 *ksps*²⁷. For discussion purposes, the focus is on the QPSK signal instead of the user binary data.

²⁷ A symbol rate of 100 *ksps* or less will satisfy the narrowband criterion of appendix C.1.

The desired QPSK signal $s(t)$ is assumed to have zero-mean and is mutually uncorrelated, i.e.,

$$\begin{aligned} E(s(t)) &= 0 \\ E(s(t_1)s^*(t_2)) &= \sigma_s^2 \delta(t_1 - t_2) \end{aligned} \quad (6.2)$$

where

$E(.)$ is the expectation operator,

t_1 and t_2 denotes two time instants,

superscript $*$ denotes complex conjugate operation,

σ_s^2 is the variance (power) of the desired signal which is set to unity in the simulation,

and $\delta(t)$ is the *Kronecker delta* function, defined as

$$\delta(t) = \begin{cases} 1 & ; t = 0 \\ 0 & ; t \neq 0 \end{cases} \quad (6.3)$$

The CCI QPSK signals $i_n(t)$ also have the same properties and are independent of each other, i.e.,

$$\begin{aligned} E(i_n(t)) &= 0 \\ E(i_{n_1}(t_1)i_{n_2}^*(t_2)) &= \sigma_{i_n}^2 \delta(t_1 - t_2) \delta(n_1 - n_2) ; \quad n = 1, \dots, N \end{aligned} \quad (6.4)$$

where

n_1 and n_2 denotes two CCI's,

and $\sigma_{i_n}^2$ is the variance (power) of the n^{th} CCI signal.

Furthermore, the desired signal and the CCI signals are assumed to be uncorrelated,

$$E(s(t_1)i_n^*(t_2)) = 0 ; \quad n = 1, 2, \dots, N \quad (6.5)$$

6.1.2 Transmit and Receive Filtering

SRRC filters with 25% roll-off factor were chosen as transmit and receive filters to minimize the *intersymbol interference* (ISI). An overall *Nyquist raised cosine* (RC)

filtering is obtained by splitting the filtering function equally between the transmitter and the receivers, i.e.,

$$\begin{aligned} |H_{tx}(f)| &= |H_{rx}(f)| = |H_{SRRC}(f)| \\ |H_{tx}(f) \otimes H_{rx}(f)| &= |H_{RC}(f)| \end{aligned} \quad (6.6)$$

where

- $H_{tx}(f)$ is the transmit filter frequency response,
- $H_{rx}(f)$ is the receive filter frequency response,
- $H_{SRRC}(f)$ is the SRRC filter frequency response,
- $H_{RC}(f)$ is the RC filter frequency response,
- $\otimes$ denotes convolution operation,
- and $|\cdot|$ is the magnitude operator.

A $\frac{T}{2}$ spaced digital SRRC FIR filter with 61 coefficients was used in the simulation. The filter energy was normalized to unity such that filtering does not change the per symbol signal and interference energy²⁸.

The SRRC and RC filters are explained in greater detail in appendix C.2.

6.1.3 Narrowband Multipath Channel

The narrowband indoor multipath channel model from section 3.2 is used to represent the indoor wireless channel. A narrowband system is assumed so as to avert any ISI that might arise from time delay spread.

As described in section 3.2, each path signal is characterized by a path loss, a phase delay and a DOA that may change with time. The discrete (sampled) l^{th} path desired signal from Eq. (3.10) is

$$s_l[k] = s[k] \alpha_l[k] \exp(j\psi_l[k]) \exp(j2\pi f_{dl}kT) ; \quad l = 1, 2, \dots, L \quad (6.7)$$

where

- $\alpha_l[k]$ is the l^{th} sampled path loss,
- $\psi_l[k]$ is the uniform random phase delay ($0-2\pi$),

²⁸ The output signal energy is equal to the input signal energy times the filter energy

f_{dl} is the l^{th} path Doppler frequency shift,

and T is the symbol period.

The indoor multipath channel split the signal energy among the independent signal paths such that the total path energy equals the signal energy, or

$$|s[k]|^2 = \sum_l |s_l[k]|^2 \quad (6.8)$$

where

$|s[k]|^2$ is the signal power,

and $|s_l[k]|^2$ is the l^{th} path signal power.

The signal path DOA is defined by the unit vector $\hat{\mathbf{a}}_l(k)$ as

$$\hat{\mathbf{a}}_l^T[k] = [\theta_l[k], \phi_l[k]]; \quad l = 1, 2, \dots, L \quad (6.9)$$

where

superscript T denotes transposition,

$\theta_l[k]$ is the time dependent direction of travel in elevation,

and $\phi_l[k]$ is the time dependent direction of travel in azimuth.

In the simulation, the path loss α_l and the path DOA are fixed for each frame duration (see Appendix C.1). The path DOA is also constrained in the azimuthal plane ($\theta_l = 0^\circ$). The phase delay ψ_l is randomly chosen for every frame but also fixed for each frame duration. The Doppler frequency shift f_{dl} is determined by the vehicle speed and the path DOA.

The desired and CCI signal are subject to similar narrowband multipath propagation conditions.

6.1.4 Additive White Gaussian Noise (AWGN)

A complex *additive white Gaussian noise* (AWGN) $v_m(t)$ is used to model the system noise at the m^{th} array element. The AWGN is added to the received signal at each element of the array. The noise has zero mean and is uncorrelated both in time and

space (between array elements), i.e.,

$$\begin{aligned} E(v_m(t)) &= 0; \quad m = 1, 2, \dots, M \\ E(v_{m_1}(t_1)v_{m_2}^*(t_2)) &= N_0 \delta(t_1 - t_2) \delta(m_1 - m_2) \end{aligned} \quad (6.10)$$

where

m_1 and m_2 denotes two array elements,

and $\frac{N_0}{2}$ is the power spectral density of the A*VGN.

The noise $v_m(t)$ is also assumed to be independent with the desired and interference signals, i.e.,

$$\begin{aligned} E(v_m(t_1)s^*(t_2)) &= 0 \\ E(v_m(t_1)i_n^*(t_2)) &= 0; \quad m = 1, 2, \dots, M, \quad n = 1, 2, \dots, N \end{aligned} \quad (6.11)$$

6.1.5 Adaptive MBA Array

The adaptive MBA array is discussed in detail in chapter 4. The simulation uses a ring array with an LMS training algorithm, both in the pilot-signal and the decision-directed training.

6.1.6 Receiver Subsystem

The receiver subsystem consists of a QPSK demodulator which extracts the signal phase information from the received data symbols. The information is passed through a slicer decision device that determines the transmitted symbol amongst the four possible symbol combinations. The QPSK signal space decision region for the slicer is depicted in Figure 6.2.

The operation of the decision device can be summarized by the following expression:

$$\hat{s}[k] = \frac{1}{\sqrt{2}} (\text{sign}(\text{Re}(y[k])) + j \text{sign}(\text{Im}(y[k]))) \quad (6.12)$$

where

$\hat{s}[k]$ is the estimated QPSK symbol,

$y[k]$ is the receiver filter output,

Figure 6.2 QPSK decision region

Re(.) is the real part operator,

$Im(.)$ is the imaginary part operator,
and $sign(.)$ is the sign function, defined as

$$sign(x) = \begin{cases} 1 & ; x \geq 0 \\ -1 & ; x < 0 \end{cases} \quad (6.13)$$

6.2 Simulation Procedures

The simulation software was written in C programming language [57] and run on SUN workstations using the UNIX operating systems. The simulation procedure is outlined in a flowchart in Figure 6.3.

The simulation starts by reading in the following simulation parameters from a file:

- number of array elements,
- number of CCI (0– N),
- SNR, and SIR for each CCI,
- number of desired and CCI signal paths,
- path DOA's (in azimuth only),
- path energy ratios,
- training and frame length (in number of symbols),
- and LMS step size (μ) parameter.

After initialization, the simulation proceeds with the transmission of information frames from the desired transmitter to the adaptive MBA array receiver. Each transmitted frame is partitioned into a training segment of length N_t symbols and a data segment of length N_d symbols, with a total length of N_f symbols (Figure 6.4).

The training sequence consisting of a pseudorandom QPSK sequence is added to the beginning of each frame at the transmitter. The QPSK training sequence is unique for each portable so that the desired signal paths can be identified by the array.

The antenna weight coefficients are updated once every symbol period according to the complex valued LMS algorithm (Eq. (5.4)). After training, the array can either freeze the weight values or continue with a decision-directed mode (symbol decisions

Figure 6.3 SER simulation flowchart

Figure 6.4 Frame structure used for simulation

are used in place of the desired signal) while it processes the actual transmitted data until the next training cycle. Since the multipath channel has varying statistics, it is deemed better to use decision-directed mode after training. The risk is the occurrence of an incorrect decision may severely impair the system performance.

The number of symbols in error is obtained by counting each estimated symbol that differs from the corresponding transmitted symbol in each frame. The simulated SER is defined as

$$SER = \frac{\sum_{i=1}^{N_{fr}} N_{ei}}{N_{fr} N_d} \quad (6.14)$$

where

N_{ei} is the number of symbols in error in the i^{th} frame,

N_{fr} is the number of frames,

and N_d is the number of data symbols per frame.

The simulated SER is a statistical estimate of the average probability of symbol error. This estimate generally improves with larger number of transmitted symbols ($N_{fr}N_d$). For reasonable accuracy, the simulation was run to count about 100 errors for each SER point, which gives a SER standard deviation of no more than 10 percent. In view of the lengthy computation time, the simulation was performed only up to a minimum SER of 10^{-5} . Since the SER curves were obtained by counting the number of symbols in error, they may not be very accurate whenever the probability of error is low (in the range of 10^{-4} and 10^{-5}).

6.2.1 Simulation Assumptions

The following assumptions were used for the simulations:

- array elements are identical and omnidirectional,
- mutual couplings are negligible,
- all path signals arrive at all array elements²⁹,

²⁹ In some rare cases, because of the finite size of a few reflectors the incoming rays might impinge only on a few array elements.

- signals are narrowband and ISI is negligible,
- ACI is negligible after receive filtering,
- perfect carrier recovery and symbol and frame synchronization,
- adaptation is performed every symbol period, T ,
- after pilot-signal training, the combiner weights are continuously updated in a decision-directed mode
- sampling is once every symbol period,
- path loss and DOA are fixed for each frame,
- and path phase delay is randomly chosen from a uniform distribution in $(0-2\pi)$ and is fixed for each frame.

The SER simulation results without and with CCI are presented in the following sections.

6.3 QPSK SER Performance over an AWGN Channel

As a benchmark for the simulation results, the SER performance of a QPSK modulation over a non-fading AWGN channel using an array receiver is first discussed.

The *theoretical probability of symbol error* P_s ³⁰ of a QPSK signalling over a non-fading AWGN channel with a single antenna receiver is [7]

$$P_s = \text{erfc}\left(\sqrt{\frac{E_s}{2N_o}}\right) - \frac{1}{4}\text{erfc}^2\left(\sqrt{\frac{E_s}{2N_o}}\right) \quad (6.15)$$

where

E_s is the symbol energy,

$\frac{N_o}{2}$ is the *power spectral density* of the AWGN,

and $\text{erfc}(\cdot)$ is the *complementary error function*, defined as

$$\text{erfc}(t) = \frac{2}{\sqrt{\pi}} \int_t^{\infty} \exp(-x^2) dx \quad (6.16)$$

³⁰ SER is the empirical estimate of the probability of symbol error

Since the simulation results are presented as SER vs. SNR³¹ curves, the probability of symbol error can also be expressed in terms of SNR as

$$P_s = \operatorname{erfc} \left(\sqrt{\frac{10^{\frac{\text{SNR}}{10}}}{2}} \right) - \frac{1}{4} \operatorname{erfc}^2 \left(\sqrt{\frac{10^{\frac{\text{SNR}}{10}}}{2}} \right) \quad (6.17)$$

The theoretical achievable SNR gain at a given SER when using an M -element array over a non-fading AWGN channel is derived in section 4.2 as $10\log_{10}(M)$ dB. Thus, the probability of symbol error with M -element array P_{sM} is

$$P_{sM} = \operatorname{erfc} \left(\sqrt{\frac{10^{\frac{\text{SNR} + 10\log_{10}(M)}{10}}}{2}} \right) - \frac{1}{4} \operatorname{erfc}^2 \left(\sqrt{\frac{10^{\frac{\text{SNR} + 10\log_{10}(M)}{10}}}{2}} \right) \quad (6.18)$$

In the following discussions, the ideal SER curves refer to the theoretical SER curves obtained from the above equation for a non-fading AWGN single-path channel. The corresponding equations for a multipath channel can also be derived, in principle, with different assumptions.

6.4 Adaptive Array SER Performance without CCI

In this section, the adaptive array SER performance is simulated for an indoor channel with AWGN only (no CCI). Two channel models are considered, one with single path and the other with three signal paths. The simulations were conducted for several desired signal path DOA's with various SNR values and different number of array elements. The simulation scenarios and results are discussed in the following sections.

³¹ SNR (in dB) is defined in Eq. (C.12)

6.4.1 Single-Path Channel without CCI

For single-path channel simulation, a desired signal is sent from a given path DOA to the adaptive array. The signal includes the path loss, phase delay and Doppler frequency shift. The simulation parameters are given as follows:

		case #1	case #2
$f_{dm}T$		0.005	0.0005
# elements		1, 2 & 5	1, 2 & 5
# desired paths		1	1
desired path DOA	experiment #1	-	90°
	experiment #2	0°	0°
	experiment #3	135°	135°
	experiment #4	225°	225°
SNR		0-20 dB	0-20 dB
CCI		none	none
frame (training) length		100 (20) symbols	1000 (20) symbols
LMS step size (μ)		0.07	0.05

Table 6.2 SER simulation parameters for single-path channel with AWGN

In the first experiment, the simulation software was first tested with a path signal arriving from a direction of 90° that has zero Doppler frequency ($\cos(90^\circ) = 0$ in Eq. (3.19). Figure 6.5 shows the simulated and the ideal SER performance curves for experiment #1. As mentioned, the ideal curves are for a non-fading AWGN single-path channel from Eq. (6.18). The simulated curves agree with the ideal curves, verifying the correctness of the simulation software. The curves display the characteristic “waterfall” drop with increasing SNR values, and the SER improves significantly with increased

number of array elements. Specifically, there is an approximately 3 dB and 7 dB gain over one element for a 2 and 5-element array respectively.

The second experiment involves a single path coming from a 0° DOA, simulated for two maximum normalized Doppler frequencies: the low Doppler ($f_{dm}T = 0.0005$) reflects a typical indoor wireless channel and the high Doppler case ($f_{dm}T = 0.005$) represents a typical vehicular wireless channel (see section 3.3.1.3). The simulation results are plotted in Figure 6.6.

It can be viewed from the figure that the presence of Doppler frequency degrades the simulated SER performance, when compared with the ideal SER curves. For example, at an SER of 10^{-3} , there is an approximately 5 dB loss for 1 element, 3 dB loss for 2 elements and 2 dB loss for 5 elements in the high Doppler simulated curves, in comparison with the corresponding ideal curves (Figure 6.6a). For low Doppler (Figure 6.6b), the loss is about 0.5 dB throughout for 1, 2 and 5 elements.

The converged antenna radiation patterns at the end of training period are also plotted in Figure 6.7 for the two Doppler cases. The polar plot shows a strong beam pointed in the direction of the 0° DOA. Apparently, there is little difference between the radiation patterns for the two cases.

To test the effect of path DOA on the SER performance, a different path DOA (135°) is simulated in experiment #3. Figure 6.8 shows the resulting SER curves again for two Doppler cases. The simulated SER curves are slightly better than that of a 0° DOA in experiment #2, because the 135° DOA gives rise to a smaller Doppler than the 0° DOA ($\cos(\phi)$ factor in the Doppler frequency formula). In fact, the worst Doppler conditions are from the 0° and the 180° DOA's.

Because of symmetry, the 225° path DOA should have the same SER performance as the 135° path DOA. This conjecture is tested in experiment #4, the results of which are shown in Figure 6.9. Comparing the curves in Figure 6.9 with the corresponding curves in Figure 6.8, it is seen that the SER of the 225° DOA is almost similar to that of the 135° DOA, as expected from symmetry.

Figure 6.5 SER vs. desired SNR, # desired paths = 1, desired path DOA = 90°

Figure 6.6 SER vs. desired SNR, # desired paths = 1, desired path DOA = 0° , a) $f_{dm}T = 0.005$, b) $f_{dm}T = 0.0005$

a)

b)

Figure 6.7 Converged radiation pattern, # desired paths = 1, desired path DOA = 0° , a) $f_{dm}T = 0.005$, b) $f_{dm}T = 0.0005$

**Figure 6.8 SER vs. desired SNR, # desired paths = 1,
desired path DOA = 135°, a) $f_{dm}T = 0.005$, b) $f_{dm}T = 0.0005$**

Figure 6.9 SER vs. desired SNR, # desired paths = 1, desired path DOA = 225°, a) $f_{dm}T = 0.005$, b) $f_{dm}T = 0.0005$

It can be inferred from the above simulations that the SER performance depends on the Doppler frequency and the path DOA (which indirectly determines the Doppler frequency). In most cellular systems, a 10^{-2} SER threshold is acceptable. For better performance, a SER of 10^{-3} might be preferable. The SNR deviations of the simulated curves from the corresponding ideal curves at 10^{-3} SER are collected in the following table for a 5-element array with various path DOA's:

desired path DOA	SNR deviation from ideal curve (in dB) at SER = 10^{-3}	
	$f_{dm}T = 0.005$	$f_{dm}T = 0.0005$
90°	~0.0	~0.0
0°	~2.0	~0.4
135°	~1.2	~0.3
225°	~1.2	~0.3
315°	~1.2	~0.3

Table 6.3 SNR deviations from ideal curves for a 5-element array in a single-path channel with AWGN

As seen, the SER is less sensitive to the path DOA at low than at high Doppler frequency.

6.4.2 Three-Path Channel without CCI

To evaluate the performance of the adaptive array over an indoor multipath channel, a three-path channel model is chosen for simplicity. The simulation parameters for the experiments involving a three-path channel corrupted by AWGN are specified as follows:

		case #1 with equal energy paths	case #2 with one dominant energy path
$f_{dm}T$		0.0005	0.0005
# elements		1, 2 & 5	1, 2 & 5
# desired paths		3	3
desired path DOA	experiment #1	0°, 120° & 240°	0°, 120° & 240°
	experiment #2	0°, 15° & 180°	0°, 15° & 180°
SNR		0-20 dB	0-20 dB
CCI		none	none
frame (training) length		500 (20) symbols	500 (20) symbols
LMS step size (μ)		0.05	0.05

Table 6.4 SER simulation parameters for three-path channel with AWGN

Two cases were simulated for each experiment, one case represents equal energy paths and the other with one dominant energy path. For the equal energy path case, the total signal energy is divided equally ($\frac{1}{3}$) among the three paths. For the one dominant energy path case, the 0° path is chosen as the dominant path with 70% of the total energy, and each of the remainder two paths is assigned 15% of the total energy.

The first experiment involves three widely separated path DOA's (0°, 120° and 240°). The experiment results for the equal path and the dominant path cases are both plotted in Figure 6.10a. As seen, the SER improves with increasing number of elements. The SER performance disparity between the dominant path case and the equal path case is smaller for 5-elements than for 1 or 2 elements. Also, the SER performance for the dominant path case is better than that of the equal path case, when the number of elements is 5 or larger. This could be attributed to the increased ability of the larger element array (larger aperture) to pick out the dominant path.

Figure 6.10 SER vs. desired SNR, # desired paths = 3, $f_{dm}T = 0.0005$, a) desired path DOA = 0° , 120° and 240° , b) desired path DOA = 0° , 15° and 180°

The results of the second experiment that uses a different set of DOA's is shown in Figure 6.10b. In this experiment, two DOA's are closely spaced (0° and 15°) and the third (180°) is clearly separated from the first two. Similar to the first experiment results, the difference between the dominant path SER performance and that of the equal path is smaller for 5-elements than for 1 or 2 elements. The dominant path SER performance is again better than the equal path SER performance.

Comparing the SER performance for the 5-element array in both experiments, it can be deduced that for large number of elements, the SER performance is less sensitive to the path DOA's and energy distributions among the signal paths. A dominant path in a multipath improves the SER performance by as much as 1 dB over the equal energy path with 5-element array.

In the next section, the effect of CCI on the adaptive array SER performance is investigated.

6.5 Adaptive Array SER Performance with CCI

It is well known that adaptive arrays are very effective at cancelling directional interference [32, 33]. An array of M elements is capable of creating $M-1$ nulls, thus rejecting $M-1$ uncorrelated interference [60]. The adaptive array SER performance is simulated for an indoor channel with AWGN and CCI in this section. In view of the simulation constraints, only a single-path channel and one CCI is considered. The simulation scenarios and results are discussed in the following section.

6.5.1 Single-Path Channel with CCI

For single-path channel with AWGN and one CCI simulation, a desired signal and a CCI are transmitted from two given directions to the adaptive array. Without loss of generality, the CCI path direction is fixed as 120° . Two SIR cases were considered, one with low SIR (3 dB) and the other with high SIR (10 dB). Also, the same maximum normalized fading rate ($f_{dm}T = 0.0005$, for a typical indoor channel) is used for both the desired signal and CCI³². The simulation parameters for the experiments are given in the following table:

		case #1	case #2
SIR		3dB	10 dB
desired signal & CCI $f_{dm}T$		0.0005	0.0005
# elements		1, 2 and 5	1, 2 and 5
# desired path		1	1
desired path DOA	experiment #1	0°	0°
	experiment #2	135°	135°
	experiment #3	225°	225°
	experiment #4	315°	315°
SNR		0-20 dB	0-20 dB
# CCI path		1	1
CCI path DOA		120°	120°
frame (training) length		1000 (20) symbols	1000 (20) symbols
LMS step size (μ)		0.05	0.05

**Table 6.5 SER simulation parameters for
single-path channel with AWGN and one CCI**

³² Different Doppler combinations could also be used.

The results of the first experiment for low and high SIR are plotted in Figure 6.11. The desired path is from 0° which is clearly isolated from the CCI path (120°).

In general, the low SIR case (Figure 6.11a) exhibits poorer performance than the high SIR case (Figure 6.11b). At low SIR, the single-element array is incapable of detecting the desired signal. The single-element array SER improves with higher SIR (Figure 6.11b). With 1 & 2 elements, the SER is limited by CCI. The SER for low and high SIR are comparable with 5 elements. This could be due to null forming in the CCI path direction (120° arrow in Figure 6.12).

It can be inferred that the SER performance is more sensitive to the SIR for small number of elements (1 & 2), and less sensitive to the SIR for larger number of elements (5). This point is shown more clearly in Figure 6.13, where the SER curves for the low and high SIR are plotted on the same graph.

In the second experiment, the effect of having a closely spaced desired signal path and CCI path is tested. The desired path from 135° , which is separated from the CCI path DOA (120°) by 15° . The results for the low SIR case is plotted in Figure 6.14a and that for the high SIR case is plotted in Figure 6.14b. It is noted that the SER performance has deteriorated considerably when compared with that of the 0° desired path of the first experiment. This is because the adaptive array can not form a dominant response towards the desired path, and at the same nulling out the closely spaced CCI.

Similar to the first experiment, the low SIR case (Figure 6.14 a) exhibits poorer performance than the high SIR case (Figure 6.14b). The SER curves for the two cases are also plotted in Figure 6.15 for comparison purposes. Comparing Figure 6.15 with Figure 6.13, it can be inferred that the SER sensitivity towards the SIR also depends on the spacing between the desired and CCI paths. Specifically, when the desired and CCI paths are closely spaced, the SER curves are more sensitive to the SIR values.

The results of experiment #3 and #4 are shown in Figure 6.16 and Figure 6.17 respectively. The figures also give the same conclusions, i.e., the SER is more sensitive to SIR for small number element than for large number element array, and the sensitivity is also dependent on the spacing between the desired and CCI paths.

Figure 6.11 SER vs. desired SNR, # desired paths = 1, desired path DOA = 0° , # CCI paths = 1, CCI path DOA = 120° , $f_{dm}T = 0.0005$, a) SIR = 3 dB, b) SIR = 10 dB

a)

b)

Figure 6.12 Converged radiation pattern, # desired paths = 1, desired path DOA = 0° , # CCI paths = 1, CCI path DOA = 120° , $f_{dm}T = 0.0005$, a) SIR = 3 dB, b) SIR = 10 dB

Figure 6.13 SER vs. desired SNR, # desired paths = 1, desired path DOA = 0° , # CCI paths = 1, CCI path DOA = 120° , $f_{dm}T = 0.0005$

Figure 6.14 SER vs. desired SNR, # desired paths = 1, desired path DOA = 135° , # CCI paths = 1, CCI path DOA = 120° , $f_{dm}T = 0.0005$, a) SIR = 3 dB, b) SIR = 10 dB

Figure 6.15 SER vs. desired SNR, # desired paths = 1, desired path DOA = 135°, # CCI paths = 1, CCI path DOA = 120°, $f_{dm}T = 0.0005$

Figure 6.16 SER vs. desired SNR, # desired paths = 1, desired path DOA = 225° , # CCI paths = 1, CCI path DOA = 120° , $f_{dm}T = 0.0005$, a) SIR = 3 dB, b) SIR = 10 dB

Figure 6.17 SER vs. desired SNR, # desired paths = 1, desired path DOA = 315° , # CCI paths = 1, CCI path DOA = 120° , $f_{dm}T = 0.0005$, a) SIR = 3 dB, b) SIR = 10 dB

As a summary, the SNR deviations of the simulated curves from the corresponding ideal curves at 10^{-3} SER are collected in the following table for a 5-element array with various path DOA's:

desired DOA	SNR deviation from ideal curve (in dB) at SER = 10^{-3}	
	$f_{dm}T = 0.0005$ SIR = 3 dB	$f_{dm}T = 0.0005$ SIR = 10 dB
0°	-1.3	-0.8
135°	-11.0	-3.2
225°	-1.2	-0.9
315°	-0.5	-0.3

Table 6.6 SNR deviations from ideal curves for a 5-element array in a single-path channel with AWGN and one CCI

It is seen from the above table that higher SIR gives better SER performance (smaller deviation). Also the SER performance deteriorates as the desired path approaches the CCI path (135° case). The performance degradation due to one CCI compared with no CCI is summarized in the following table for a 5-element array with various path DOA's:

desired DOA	Performance degradation with CCI (in dB) at SER = 10^{-3}	
	$f_{dm}T = 0.0005$ SIR = 3 dB	$f_{dm}T = 0.0005$ SIR = 10 dB
0°	-0.9	-0.4
135°	-10.7	-2.9
225°	-0.9	-0.6
315°	-0.2	-0.0

Table 6.7 Performance degradations with CCI for a 5-element array in a single-path channel with AWGN and one CCI

The same conclusions as before can be made from the above table.

Chapter 7

Conclusions and Recommendations

7.1 Summary and Conclusions

The major findings and contributions of this thesis are summarized in this section.

A model to characterize the signal from a single path of an indoor multipath has been derived using ray-theoretic concepts. It incorporates the individual path characteristics such as, path loss, time (phase) delay and Doppler frequency shift. The model is shown to generate the typical indoor signal fading with as few as eight signal paths.

Linear and ring arrays are considered for multiple-beam forming to capture the desired path signals and nulling to suppress the CCI. The ring array which has the ability to form unique spatial beams is preferred for a multipath environment with CCI than the linear which suffers from cylindrical symmetry in its array pattern. As opposed to a ring array which has a constant beamwidth in the plane of the ring, the linear array beamwidth varies from the broadside to the endfire direction. Also, the diameter of a ring array is about π factor smaller than the length of a similar linear array. The ring array, however, requires a larger number of elements for a given beamwidth. A linear weight vector combination technique is also suggested for multiple-beam creation.

Adaptive beam-forming using pilot signal training with the LMS and the RLS algorithms has been investigated. The LMS algorithm rate of convergence is shown to depend on the chosen step size μ parameter and is slower than the RLS algorithm, but it is computationally simpler. The simulations verify the ring array's capability to fashion strong beams in the desired signal path DOA's and nulls in the CCI path DOA's. A 20-element ring array is shown to track a moving signal path (in steps of 5°) with good accuracy. With CCI, nulls as deep as 6 dB could be formed in the CCI path direction. When the CCI path coincides with a desired path, a 5 dB null is formed only if that path is not the only desired path.

SER system performance of the adaptive array over an indoor channel has been simulated for a narrowband coherent QPSK signalling scheme. The SER performance is found to improve with greater number of array elements. A 2-element (5-element) array has a minimum improvement of 3 dB (7 dB) over a single element antenna at a 10^{-3} SER. The SER also depends on the Doppler frequency and the path DOA. It is less sensitive to the path DOA at low than at high Doppler. A dominant path in a multipath improves the SER performance by as much as 1 dB over the equal energy path with a 5-element array. In the presence of CCI, the array performance is degraded by about 11 dB for low SIR (3 dB) and 3 dB for high SIR (10 dB).

The adaptive MBA array can be deployed at base stations or portables, and may be implemented using microstrip array technology for higher operating frequency. A frequency greater than 5 GHz (5-element ring diameter = 5.1 cm) is recommended when used at the portables. At lower frequencies, use at base stations is more feasible. With the ability to form beams towards desired signal paths and nulls against unwanted CCI paths, the adaptive arrays hold great promises for indoor wireless communications.

7.2 Recommendations for Further Research

To expand on the thesis work, the followings are suggested for further research:

- Effect of the time delay spread on the wideband adaptive array performance.
- Comparison of various adaptive algorithms such as the normalized LMS and constant modulus algorithm for adaptive multiple-beam forming.
- Analysis of the adaptive array performance with other modulation schemes such as FSK, MSK, GMSK, QAM, etc..
- Study of the reference signal generation techniques, and the potential for integration with spread spectrum techniques.
- Examination of the issues involved in the implementation of the adaptive array.
- Investigation of the use of coding and other strategies to reduce the training overhead.

Appendix A General Antenna Array Theory

This appendix provides the background theory of a general antenna array. In specific, the received signal at each array element due to a signal from the l^{th} path of an indoor multipath is derived. The concept of *array factor* and *pattern multiplication theorem* are also introduced.

A general antenna array consisting of M spatially distributed identical antenna elements with the same orientation is depicted in Figure A.1. A narrowband plane wave from the l^{th} path (defined in Eq. (3.3)) impinges on the array elements.

The induced current phasor $\mathbf{I}_{lm}(\mathbf{r}_m, t)$ at the m^{th} element due to the l^{th} path plane wave signal is [24]

$$\mathbf{I}_{lm}(\mathbf{r}_m, t) = K_m s_l(t) f_{e_m}(\theta, \phi) \frac{\exp(-jkR_{lm})}{4\pi R_{lm}} ;$$

$$l = 1, 2, \dots, L ; \quad m = 1, 2, \dots, M \quad (\text{A.1})$$

where

$\mathbf{r}_m$ is the position vector from the origin of the coordinate system to the m^{th} element,

K_m is a conversion factor,

$s_l(t)$ is the signal from the l^{th} path³³,

$f_{e_m}(\theta, \phi)$ is the m^{th} element radiation pattern,

θ is the elevation angle,

ϕ is the azimuth angle,

k is the wave number ($2\pi/\lambda$),

R_{lm} is the distance from the l^{th} path “source” to the m^{th} element,

L is the number of signal paths,

and M is the number of elements in the array.

³³ Transit time across the array dimension is too small to affect the received signal at different elements.

Figure A.1 An M -element general array structure with an incident l^{th} path plane wave

Assuming that the l^{th} path "source" is sufficiently removed from the array³⁴, the signal rays are essentially in parallel. The distance R_{lm} in the denominator of the above equation can then be approximated by R_{l1} . A first order approximation, $R_{lm} \approx R_{l1} - \hat{\mathbf{a}}_l \cdot \mathbf{r}_m$, is also used in the exponential function because a small change in distance results in a significant change in phase delay. Incorporating these approximations in Eq. (A.1) and adding the individual current phasor contributions from each element with equal weighting (weight = 1), the total current excitation $\mathbf{I}_l(\mathbf{r}, t)$ due to the l^{th} path signal is

$$\mathbf{I}_l(\mathbf{r}, t) = s_l(t) \frac{\exp(-jkR_{l1})}{4\pi R_{l1}} \sum_{m=1}^M K_m f_{e_m}(\theta, \phi) \exp(jk \hat{\mathbf{a}}_l \cdot \mathbf{r}_m); \quad (A.2)$$

$l = 1, 2, \dots, L$

where $\hat{\mathbf{a}}_l$ is the l^{th} path unit direction vector (defined in Eq. (3.4) of section 3.2) and $\mathbf{r}_m$ is the m^{th} element position vector.

Since the elements are electrically identical and have the same spatial orientation ($K_m = K$ and $f_{e_m}(\theta, \phi) = f_e(\theta, \phi)$ for all m), the element radiation pattern can be factorized from the summation, yielding

$$\mathbf{I}_l(\mathbf{r}, t) = \left(K s_l(t) \frac{\exp(-jkR_{l1})}{4\pi R_{l1}} f_e(\theta, \phi) \right) F(\theta, \phi); \quad (A.3)$$

$l = 1, 2, \dots, L$

which says that the total induced current is a product of a field component that depends on the array element and the received signal, and an *array factor* $F(\theta, \phi)$ which is a function of the array geometry. The array factor is defined as the radiation pattern of an unweighted antenna array with isotropic elements, or mathematically

$$F(\theta, \phi) = \sum_{m=1}^M \exp(jk \hat{\mathbf{a}}_l \cdot \mathbf{r}_m) \quad (A.4)$$

³⁴ For frequencies in the GHz or above, the wavelength and hence the interelement spacing is small enough to warrant this assumption.

The above result is also known as the *pattern multiplication principle* in antenna array theory [61]. It states that the radiation pattern of an array of identical antenna elements is the product of the element radiation pattern and the array factor. The pattern multiplication principle has been derived for a receiving array. By virtue of the *antenna reciprocity theorem* [24], the results are equally applicable to a transmitting array.

The term in the exponential function of the array factor, denoted by δ_{lm} where

$$\delta_{lm} = k \hat{\mathbf{a}}_l \cdot \mathbf{r}_m ; \quad l = 1, 2, \dots, L ; \quad m = 1, 2, \dots, M \quad (\text{A.5})$$

is the relative phase delay between the m^{th} element current phasor and the reference (first element) current phasor. The received signal $x_{lm}(t)$ at the m^{th} element due to the l^{th} path signal is then

$$\begin{aligned} x_{lm}(t) &\propto s_l(t) \exp(j\delta_{lm}) ; \\ l &= 1, 2, \dots, L ; \quad m = 1, 2, \dots, M \end{aligned} \quad (\text{A.6})$$

The relative phase delay depends on the array geometry.

Appendix B Optimum Weight Solution for Beam-Forming in a Multipath Environment

A derivation of the *optimum weight solution* for beam-forming in a multipath environment is given in this appendix. The derivation proceeds by defining the error signal to be minimized as a function of the unknown weight vector. The optimum weight solution is then obtained by minimizing the *mean-square* value of the error signal.

From section 5.1, the error signal $e[k]$ is defined as the difference between the desired reference signal $d[k]$ and the adaptive combiner output signal $y[k]$, or

$$e[k] = d[k] - y[k] \quad (\text{B.1})$$

The adaptive combiner output signal $y[k]$ is a weighted linear combination of the array element signals,

$$y[k] = \mathbf{w}^H[k] \mathbf{x}[k] \quad (\text{B.2})$$

where

$\mathbf{w}^T[k] = [w_1[k] \cdots w_m[k] \cdots w_M[k]]$ is the unknown weight vector (superscript T denotes transposition),

superscript H denotes Hermitian operation,

and $\mathbf{x}[k] = \mathbf{D}_s \mathbf{s}[k] + \sum_{n=1}^N \mathbf{D}_n \mathbf{i}_n[k] + \mathbf{v}[k]$ is the received signal vector defined in Eq. (5.1).

The *mean-square error* (MSE) $J[k]$ is defined as

$$J[k] = E(e[k]e^*[k]) \quad (\text{B.3})$$

where $E(\cdot)$ is the expectation operator.

Substituting the expression for the error signal $e[k]$ and the output signal $y[k]$ into the above equation, yields

$$J[k] = E\left(d[k]d^*[k] - \mathbf{w}^H[k]\mathbf{x}[k]d^*[k] - d[k]\mathbf{x}^H[k]\mathbf{w}[k] + \mathbf{w}^H[k]\mathbf{x}[k]\mathbf{x}^H[k]\mathbf{w}[k]\right) \quad (\text{B.4})$$

The *variance* of the desired signal is assumed constant, and is defined as

$$\sigma_d^2 = E(d[k]d^*[k]) \quad (\text{B.5})$$

The *cross-correlation vector* between the received signal vector $\mathbf{x}[k]$ and the desired response $d[k]$ is denoted by $\mathbf{P}$, where

$$\mathbf{P} = E(\mathbf{x}[k]d^*[k]) \quad (\text{B.6})$$

The *autocorrelation matrix* of the received vector $\mathbf{x}[k]$ is denoted by $\mathbf{R}$, where

$$\mathbf{R} = E(\mathbf{x}[k]\mathbf{x}^H[k]) \quad (\text{B.7})$$

Using the above definitions, the MSE can be rewritten as

$$\begin{aligned} J[k] &= \sigma_d^2 - \mathbf{w}^H[k]\mathbf{P} - \mathbf{P}^H\mathbf{w}[k] + \mathbf{w}^H[k]\mathbf{R}\mathbf{w}[k] \\ &= \sigma_d^2 - 2\text{Re}\left(\mathbf{w}^H[k]\mathbf{P}\right) + \mathbf{w}^H[k]\mathbf{R}\mathbf{w}[k] \end{aligned} \quad (\text{B.8})$$

which is a *real valued quadratic function* of the complex vector variable $\mathbf{w}$ ($\text{Re}(\cdot)$ is the real part of the quantity in parenthesis).

If the received signal and the desired signal are *jointly stationary*, the MSE function can be visualized as an *M-dimensional paraboloid error performance surface* with a unique *minimum point* (optimum weight solution). The adaptive algorithm has the task of seeking this minimum point with no prior information. If the signal is *nonstationary*, the adaptive algorithm has the added task of tracking the shifting minimum point.

The optimum weight vector $\mathbf{w}_{opt}$ for the beam-forming in a multipath environment can be obtained by applying the standard optimization technique, that is, by equating the *gradient* of the MSE function to zero. The gradient function $\nabla_{\mathbf{w}} J[k]$ in vector notation is defined as

$$\nabla_{\mathbf{w}} J[k] = \frac{\partial J[k]}{\partial \text{Re}(\mathbf{w})} + j \frac{\partial J[k]}{\partial \text{Im}(\mathbf{w})} \quad (\text{B.9})$$

or in vector column form as

$$\nabla_{\mathbf{w}} J[k] = \begin{bmatrix} \frac{\partial J[k]}{\partial \text{Re}(w_1)} + j \frac{\partial J[k]}{\partial \text{Im}(w_1)} \\ \vdots \\ \frac{\partial J[k]}{\partial \text{Re}(w_m)} + j \frac{\partial J[k]}{\partial \text{Im}(w_m)} \\ \vdots \\ \frac{\partial J[k]}{\partial \text{Re}(w_M)} + j \frac{\partial J[k]}{\partial \text{Im}(w_M)} \end{bmatrix} \quad (\text{B.10})$$

The partial derivative with respect to the real part of the m^{th} weight coefficient can be written as

$$\begin{aligned} \frac{\partial J[k]}{\partial \text{Re}(w_m)} &= E \left(\frac{(\partial e[k])}{\partial \text{Re}(w_m)} e^*[k] \right) + E \left(e[k] \frac{(\partial e^*[k])}{\partial \text{Re}(w_m)} \right) \\ &= -E(x_m[k] e^*[k]) + E(e[k] (-x_m^*[k])) \\ &= -2\text{Re}(E(x_m[k] e^*[k])) \\ m &= 1, 2, \dots, M \end{aligned} \quad (\text{B.11})$$

Similarly, the partial derivative with respect to the imaginary part of the m^{th} weight coefficient is

$$\begin{aligned} \frac{\partial J[k]}{\partial \text{Im}(w_m)} &= E \left(\frac{(\partial e[k])}{\partial \text{Im}(w_m)} e^*[k] \right) + E \left(e[k] \frac{(\partial e^*[k])}{\partial \text{Im}(w_m)} \right) \\ &= E(j x_m[k] e^*[k]) + E(e[k] (-j x_m^*[k])) \\ &= -2\text{Im}(E(x_m[k] e^*[k])) \\ m &= 1, 2, \dots, M \end{aligned} \quad (\text{B.12})$$

Combining the partial derivatives with respect to the both the real and the imaginary parts of the weight coefficients, and putting them in the vector form of Eq. (B.9), gives

$$\begin{aligned}\nabla_{\mathbf{w}} J[k] &= -2\text{Re}(E(\mathbf{x}[k]e^*[k])) - j2\text{Im}(E(\mathbf{x}[k]e^*[k])) \\ &= -2E(\mathbf{x}[k]e^*[k])\end{aligned}\quad (\text{B.13})$$

Substituting the expression for the error signal from Eq. (B.1), the gradient function can be reformulated as

$$\begin{aligned}\nabla_{\mathbf{w}} J[k] &= -2E\left(\mathbf{x}[k]\left(d^*[k] - \mathbf{x}^H[k]\mathbf{w}\right)\right) \\ &= -2\mathbf{P} - 2\mathbf{R}\mathbf{w}\end{aligned}\quad (\text{B.14})$$

Setting the above expression to zero, and after rearranging yields

$$\mathbf{R}\mathbf{w} = \mathbf{P} \quad (\text{B.15})$$

which is the *Wiener-Hopf equation*.

If the autocorrelation matrix $\mathbf{R}$ is nonsingular, the optimum weight vector $\mathbf{w}_{\text{opt}}$ can be found as

$$\mathbf{w}_{\text{opt}} = \mathbf{R}^{-1}\mathbf{P} \quad (\text{B.16})$$

where superscript $^{-1}$ denotes matrix inversion.

Equation B.16 represents the optimum or *Wiener* weight solution for the beam-forming in a multipath environment. It can be solved explicitly to yield an optimum weight $\mathbf{w}_{\text{opt}}$, provided $\mathbf{P}$ and $\mathbf{R}$ are known or can be estimated easily.

The *minimum* MSE is found by substituting the optimum weight $\mathbf{w}_{\text{opt}}$ in Eq. (B.3),

$$J_{\min} = \sigma_d^2 - \mathbf{P}^H \mathbf{w}_{\text{opt}} \quad (\text{B.17})$$

In the absence of prior information, the average gradient function in Eq. (B.14) can be estimated by an *unbiased instantaneous estimate*

$$\nabla_{\mathbf{w}} J[k] \approx -2\mathbf{x}[k]e^*[k] \quad (\text{B.18})$$

which appears in many real-time adaptive algorithms such as the LMS algorithm.

Appendix C System Performance Simulation Documentation

C.1 SER Simulation Operating Scenarios

The adaptive MBA array system performance simulation in chapter 6 was conducted for the following operating scenarios.

Consider a portable moving at a constant speed v_p at radius distance r around a base station (Figure C.1) in an indoor environment. The normal walking pace is fairly slow, typically in the range of 0–2.0 m/s . Using an average walking speed v_p of 1.5 m/s , and a radius distance r of 5 m , the rate of angular displacement is

$$\frac{\delta\phi}{\delta t} = \frac{v_p}{r} = \frac{1.5 \text{ m/s}}{5 \text{ m}} = 0.3 \text{ rad/s} (17^\circ/\text{s}) \quad (\text{C.1})$$

In a frame time of 10 ms , the angular distance covered by the portable is

$$\Delta\phi = \frac{1.5 \text{ m/s} * 0.01 \text{ s}}{5 \text{ m}} = 0.003 \text{ rad} = 0.17^\circ \quad (\text{C.2})$$

This is quite a small angular displacement, hence, it is reasonably safe to assume that the path DOA and loss is fixed in a 10 ms time frame interval.

The operating carrier frequency was chosen as 10 GHz ³⁵, such that a 5–element ring array will have a diameter of 2.552 cm to fit in a handheld portable. From Eq. (3.19), the maximum Doppler frequency f_{dm} is

$$f_{dm} = \frac{v_p}{c} f_c = \frac{1.5 \text{ m/s}}{3 \times 10^8 \text{ m/s}} \times 10^{10} \text{ Hz} = 50 \text{ Hz} \quad (\text{C.3})$$

For narrowband system, the maximum time delay spread τ_{max} should be less than one tenth of a symbol period T . A good approximation of the maximum path delay in a typical indoor environment is 300 m , which gives rise to a maximum time delay of

$$\tau_{max} = \frac{300 \text{ m/s}}{3 \times 10^8 \text{ m/s}} = 1 \mu\text{s} \quad (\text{C.4})$$

³⁵ Future indoor wireless is forecasted to operate in GHz range.

Figure C.1 Portable moving around a base station

A symbol rate of less than 100 *k*sps will meet the above narrowband criterion (non-frequency selectivity) in an indoor multipath environment. With 100 *k*sps symbol rate, the maximum normalized Doppler fading frequency $f_{dm}T$ can be calculated as 0.0005. For the same operating conditions, the corresponding maximum normalized Doppler fading frequency for an outdoor vehicular system moving at a speed of 54 *km/h* is 0.005 (see section 3.3.1.3). The SER simulations were performed at these two normalized Doppler fading rate to test the effect of the Doppler rate on the SER performance.

Since periodic training requires special time slot allocation for training symbols, it is desirable to have a short training sequence with a long data sequence for high system throughput efficiency. But, the longer the time period between training, the more variation the path signal attributes (path loss, phase delay, and DOA) experience. To assume a fixed loss, phase delay and DOA within a time frame, it is necessary to have short frame length with frequent training. Thus, there is a compromise to be made between higher throughput efficiency and better system performance.

The system throughput depends on the number of training symbols and the frame length. The percentage of training overhead is

$$\%Overhead = \frac{100N_t}{N_t + N_d} \% \quad (C.5)$$

and the throughput efficiency can be defined as

$$\%Throughput = \frac{100N_d}{N_t + N_d} \% \quad (C.6)$$

For a 20 symbol training symbols in a 1000 symbols total frame length (10 *ms* frame at 100 *k*sps), the system throughput is 98 percent. For a shorter total frame length of 500 symbols (5 *ms* frame), the system throughput reduces to 96 percent.

C.2 Transmit and Receive Filters

Square root raised cosine (SRRC) filters were used as transmit and receive filters in the simulation model of chapter 5 and 6. The SRRC receive filter was matched to the SRRC transmit filter to have an overall filter response of a *raised cosine* (RC) filter.

The frequency response $H_{SRRC}(f)$ of a SRRC filter is

$$H_{SRRC}(f) = \begin{cases} \sqrt{T} & ; |f| \leq \frac{1}{2T}(1 - \alpha) \\ \sqrt{\frac{T}{2} \left(1 - \sin \left[\frac{\pi T}{\alpha} \left(|f| - \frac{1}{2T} \right) \right] \right)} & ; \frac{1}{2T}(1 - \alpha) < |f| \leq \frac{1}{2T}(1 + \alpha) \\ 0 & ; |f| > \frac{1}{2T}(1 + \alpha) \end{cases} \quad (C.7)$$

or equivalently, in time domain $h_{SRRC}(t)$ is [62]

$$h_{SRRC}(t) = \begin{cases} \frac{4\alpha}{\pi} + (1 - \alpha) & ; t = 0 \\ \frac{\alpha}{\sqrt{2}} \left[\left(1 - \frac{2}{\pi} \right) \cos \left(\frac{\pi}{4\alpha} \right) + \left(1 + \frac{2}{\pi} \right) \sin \left(\frac{\pi}{4\alpha} \right) \right] & ; t = \pm \frac{T}{4\alpha} \\ \frac{(1 - \alpha) \text{sinc} \left[(1 - \alpha) \frac{t}{T} \right] + \frac{4\alpha}{\pi} \cos \left[(1 + \alpha) \frac{\pi t}{T} \right]}{\left[1 - \left(\frac{4\alpha t}{T} \right)^2 \right]} & ; t \neq \frac{T}{4\alpha} \end{cases} \quad (C.8)$$

where

$$\text{sinc}(x) = \frac{\sin(\pi x)}{\pi x},$$

α is the roll-off factor of the filter, (0 to 1.0),

and T is the symbol period.

The cascade of two SRRC filters produces a raised cosine filter with a frequency response $H_{RC}(f)$ of

$$H_{RC}(f) = \begin{cases} T & ; |f| \leq \frac{1}{2T}(1 - \alpha) \\ \frac{T}{2} \left\{ 1 - \sin \left[\frac{\pi T}{\alpha} \left(|f| - \frac{1}{2T} \right) \right] \right\} & ; \frac{1}{2T}(1 - \alpha) < |f| \leq \frac{1}{2T}(1 + \alpha) \\ 0 & ; |f| > \frac{1}{2T}(1 + \alpha) \end{cases} \quad (C.9)$$

that satisfies the *Nyquist* criteria for zero ISI [7].

The normalized frequency response and normalized time response of a 25 % roll-off SRRC filter used in the simulation is shown in Figure C.2 and Figure C.3 respectively.

The raised cosine normalized frequency and time response obtained by convolving the two transmit and receive SRRC filters are shown in Figure C.4 and Figure C.5 respectively.

The SRRC filter impulse response coefficients were designed by performing an inverse FFT [59] (in MATLAB) on the corresponding frequency response samples.

C.3 Desired Signal, Interference and Noise Power Calculations

This appendix section describes the desired path, CCI path and noise power calculation implemented in the simulation.

The average desired signal power σ_s^2 is normalized to unity for simplicity. It is further divided into several desired signal paths according to a pre-determined ratio such that the total signal path energy equal unity.

For the n^{th} CCI, the *signal-to-interference* ratio (SIR_n) is defined as

$$\begin{aligned} SIR_n &= 10 \log_{10} \left(\frac{\sigma_s^2 \sum_u h_{tx}^2[u]}{\sigma_{in}^2 \sum_u h_{tx}^2[u]} \right) \\ &= 10 \log_{10} \left(\frac{\sigma_s^2}{\sigma_{in}^2} \right) \end{aligned} \quad (C.10)$$

where σ_{in}^2 is the n^{th} average interference power.

The required interference power is calculated from a given SIR_n as

$$\sigma_{in}^2 = \frac{1}{10^{SIR_n/10}} \quad (C.11)$$

**Figure C.2 Square root raised cosine (SRRC) filter
normalized frequency magnitude response, $\alpha = 0.25$**

**Figure C.3 Square root raised cosine (SRRC)
filter normalized impulse response, $\alpha = 0.25$**

Figure C.4 Raised cosine (RC) filter normalized frequency magnitude response, $\alpha = 0.25$

Figure C.5 Raised cosine (RC) filter normalized impulse response, $\alpha = 0.25$

The average *signal-to-noise* ratio (SNR) per transmitted symbol of the desired signal at the input to the adaptive combiner is defined as

$$\begin{aligned}
 SNR &= 10\log_{10}\left(\frac{E_s}{N_0}\right) \\
 &= 10\log_{10}\left(\frac{\sigma_s^2 \sum_u h_{tx}^2[u]}{\sigma_v^2}\right) \\
 &= 10\log_{10}\left(\frac{\sigma_s^2}{\sigma_v^2}\right)
 \end{aligned} \tag{C.12}$$

where

E_s is the energy per symbol,

$\frac{N_0}{2}$ is the power spectral density of the AWGN,

σ_s^2 is the variance (power) of the desired signal which is set to unity in the simulation,

$h_{tx}[u]$ is the filter impulse response which has unit energy,

and σ_v^2 is the complex noise variance.

For a desired SNR, the required complex noise variance is calculated as

$$\sigma_v^2 = \frac{1}{10^{SNR/10}} \tag{C.13}$$

Since the noise has zero mean, the complex noise power equal its variance σ_v^2 . The real and imaginary part of the complex noise each has the same variance of σ_n^2 . For complex narrowband noise, the total noise power is $N_o = \sigma_v^2 = 2\sigma_n^2$, and the component noise variance σ_n^2 can be expressed in terms of the SNR as

$$\sigma_n^2 = \frac{1}{2 \times 10^{SNR/10}} \tag{C.14}$$

Bibliography

- [1] O. Eriksson and P. Bjömdal, "Personal Communications: a Possible Evolution in the Near Future of Today's Digital Mobile Telecommunications Systems and Intelligent Networks," in *Proc. of the Fourth Nordic Sem. on Digital Mobile Radio Communications*, (Oslo, Norway), pp. 8.3.1–8.3.7, Jun. 1990.
- [2] D. C. Cox, "Personal Communications: a Viewpoint," *IEEE Commun. Mag.*, vol. 28, no. 11, pp. 10–20, Nov. 1990.
- [3] R. Singer and D. Irwin, "Personal Communication Services: The Next Technological Revolution," *IEEE Commun. Mag.*, vol. 29, no. 2, pp. 62–66, Feb. 1991.
- [4] R. S. Swain and D. W. J. Holmes, "The Digital Cordless Telecommunications Common Air Interface," *British Telecom Technical Journal*, vol. 8, no. 1, Jan. 1990.
- [5] J. W. C. Jakes, ed., *Microwave Mobile Communications*. New York: John Wiley and Sons, 1974.
- [6] W. C. Y. Lee, *Mobile Communication Engineering*. New York: McGraw-Hill, 1982.
- [7] J. Proakis, ed., *Digital Communications*. New York: John Wiley and Sons, 1974.
- [8] W. C. Y. Lee, *Mobile Cellular Telecommunication Engineering*. New York: McGraw-Hill, 1990.
- [9] P. Monsen, "MMSE Equalization of Interference on Fading Diversity Channels," *IEEE Trans. on Commun.*, vol. COM-32, no. 1, pp. 5–12, Jan. 1984.
- [10] T. A. Sexton and K. Pahlavan, "Channel Modeling and Adaptive Equalization of Indoor Radio Channels," *IEEE J. Sel. Areas in Commun.*, vol. SAC-7, no. 1, pp. 114–121, Jan. 1989.
- [11] N. W. K. Lo, D. D. Falconer, and A. U. H. Sheikh, "Adaptive Equalization and Diversity Combining for Mobile Radio Using Interpolated Channel Estimation," *IEEE Trans. on Vehicular Tech.*, vol. 40, no. 3, pp. 636–645, Aug. 1991.

- [12] R. D. de Iacovo and D. Sereno, "CELP Coding at 6.55 Kbit/s for Digital Mobile Radio Communications.," in *Proc. Globecom 90*, (San Diego, Ca.), pp. 538–541, Dec. 1990.
- [13] E. I. A., "Cellular System: Dual-Mode Mobile Station — Base Station Compatibility Standard," Tech. Rep. IS-54, Electronic Industries Association, Jan. 1991.
- [14] H. Ochser, "DECT — Digital European Cordless Telecommunications," in *Proc. VTC 89*, (San Francisco, Ca.), pp. 718–721, May 1989.
- [15] B. Mallinder, "An Overview of the GSM System," in *Proc. of the Third Nordic Sem. on Digital Mobile Radio Communications*, (Copenhagen, Denmark), pp. 3.1.1.–3.1.13, Sep. 1988.
- [16] V. H. MacDonald, "The Cellular Concept," *Bell Sys. Tech. J.*, vol. 58, no. 1, pp. 15–42, Jan. 1979.
- [17] W. C. Y. Lee, "Comparison of Spectrum Efficiency Between FDMA and TDMA in Digital Cellular," in *Proc. VTC 90*, (San Francisco, Ca.), pp. 165–168, May 1989.
- [18] M. Kavehrad and G. Bodeep, "Design and Experimental Results for a Direct-Sequence Spread Spectrum Radio Using Differential Phase-Shift Keying Modulation for Indoor Wireless Communications," *IEEE J. Sel. Areas in Commun.*, vol. SAC-5, pp. 815–823, Jun. 1987.
- [19] D. J. Goodman and S. X. Wei, "Efficiency of Packet Reservation Multiple Access," *IEEE Trans. on Vehicular Tech.*, vol. 40, no. 1, pp. 170–176, Feb. 1991.
- [20] K. A. Wimmer and J. B. Jones, "Global Development of PCS," *IEEE Commun. Mag.*, vol. 30, no. 6, pp. 22–27, Jun. 1992.
- [21] W. C. Y. Lee, "Preliminary Investigation of Mobile Radio Signal Fading Using Directional Antennas on the Mobile Unit.," *IEEE Trans. on Vehicular Commun.*, vol. VC-15, no. 2, pp. 8–15, Oct. 1966.
- [22] F. Ikegami and S. Yoshida, "Analysis of Multipath Propagation Structure in Urban Mobile Radio Environments," *IEEE Trans. on Antennas Propag.*, vol. AP-28,

- no. 4, pp. 531–537, Jul. 1980.
- [23] T. A. Freeburg, "Enabling Technologies for Wireless In-Building Network Communications — Four Technical Challenges, Four Solutions," *IEEE Commun. Mag.*, no. 4, pp. 58–64, Apr. 1991.
 - [24] R. E. Collin, *Antennas and Radio Wave Propagation*. New York: McGraw-Hill, 1985.
 - [25] V. I. Bogachev and I. G. Kiselev, "Optimum Combining of Signals in Space-Diversity Reception," *Telecommun. Radio Eng.*, vol. 34/35, no. 10, pp. 83–85, Oct. 1980.
 - [26] J. H. Winters, "Optimum Combining in Digital Mobile Radio with Co-channel Interference," *IEEE J. Sel. Areas in Commun.*, vol. SAC-2, no. 4, pp. 528–539, Jul. 1984.
 - [27] R. Vaughan, "On Optimum Combining at the Mobile," *IEEE Trans. on Vehicular Tech.*, vol. VT-37, no. 4, pp. 181–188, Nov. 1988.
 - [28] M. Clark, W. Kennedy, M. Shafi, and L. Greenstein, "MMSE Diversity Combining for Wide-Band Digital Cellular Radio," in *Proc. Globecom 90*, (San Diego, Ca.), pp. 495–499, Dec. 1990.
 - [29] S. A. Hanna, M. El-Tanany, and S. A. Mahmoud, "An Adaptive Combiner for Co-channel Interference Reduction in Multi-User Indoor Radio Systems," in *Proc. VTC 91*, (St. Louis, Mo.), pp. 222–227, May 1991.
 - [30] R. Kohno, H. Imai, and S. Pasupathy, "Adaptive Array Antenna Based on Combination of Spatial and Temporal Filtering for Channels with Multipath Distortion," in *EURASIP Fifth European Signal Conference Proc.*, (Barcelona), pp. 729–732, Sep. 1990.
 - [31] "Special Issues on Active and Adaptive Antennas," *IEEE Trans. on Antennas Propag.*, vol. AP-12, no. 2, Mar. 1964.
 - [32] B. Widrow and S. D. Stearns, *Adaptive Signal Processing*. Englewood Cliffs, NJ: Prentice-Hall, 1985.

- [33] B. Widrow, P. E. Mantey, L. J. Griffiths, and B. B. Goode, "Adaptive Antenna Systems," in *Proc. of the IEEE*, vol. 55, no. 12, pp. 2143–2159, Dec. 1967.
- [34] W. S. Davies, R. J. Lang, and E. Vinnal, "The Challenge of Advanced Base-Station Antennas for Future Cellular Radio Systems," in *IEEE International Workshop — Digital Mobile Radio*, (Melbourne, Australia), Mar. 1987.
- [35] M. J. Marcus and S. Das, "The Potential Use of Adaptive Antennas to Increase Land Mobile Frequency Reuse," in *IEE 2nd International Conference on Radio Spectrum Conservation Techniques*, vol. 224, (Birmingham, U. K.), pp. 113–117, Sep. 1983.
- [36] S. C. Swales, M. A. Beach, and D. J. Edwards, "Multi-beam Adaptive Base-Station Antennas for Cellular Land Mobile Radio Systems," in *Proc. VTC 89*, (San Francisco, Ca.), pp. 341–348, 1989.
- [37] S. C. Swales, M. A. Beach, D. J. Edwards, and J. P. McGeehan, "The Performance Enhancement of Multibeam Adaptive Base-Station Antennas for Cellular Land Mobile Radio Systems," *IEEE Trans. on Vehicular Tech.*, vol. 39, no. 1, pp. 56–67, Feb. 1990.
- [38] S. Anderson, M. Millnert, M. Viberg, and B. Wahlberg, "An Adaptive Array for Mobile Communication Systems," *IEEE Trans. on Vehicular Tech.*, vol. VT-40, no. 1, pp. 230–236, Feb. 1991.
- [39] S. Alexander and G. Pulgiese, "Cordless Communications within Buildings: Results of Measurements at 900 MHz and at 60 GHz," *British Telecom Tech. J.*, pp. 99–105, 1983.
- [40] S. J. Patsiokas, B. K. Johnson, and J. L. Dailing, "Propagation of Radio Signals Inside Buildings at 150, 450 and 850 MHz," in *Proc. VTC 86*, (Dallas, Tx.), pp. 66–71, May 1986.
- [41] R. J. C. Bultitude, "Measurements, Characterization and Modeling of Indoor at 800/900 MHz Radio Channels for Digital Communications," *IEEE Commun. Mag.*, vol. 25, no. 6, pp. 5–12, Jun. 1987.

- [42] J. Horikoshi, T. Tanaka, and T. Morinaga, "1.2 GHz Band Wave Propagation Measurements in a Concrete Building for Indoor Radio Communications," *IEEE Trans. on Vehicular Tech.*, vol. VT-35, no. 4, pp. 146–152, Nov. 1983.
- [43] D. A. Hawbaker and T. S. Rappaport, "Indoor Wideband Radiowave Propagation Measurements at 1.3 GHz and 4.0 GHz," *Electronic Letters*, vol. 26, no. 21, pp. 1800–1802, Oct. 1990.
- [44] M. Heath, "Propagation Measurements at 1.76 GHz for Digital European Cordless Telecommunications," in *Proc. Globecom 90*, (San Diego, Ca.), pp. 1007–1012, Dec. 1990.
- [45] J. W. McKown and R. L. Hamilton, "Ray Tracing as A Design Tool for Radio Networks," *IEEE Network Mag.*, vol. 5, no. 6, pp. 27–30, Nov. 1991.
- [46] R. Ganesh and K. Pahlavan, "Effects of Traffic and Local Movements on Multipath Characteristics of an Indoor Radio Channels," *Electronic Letters*, vol. 26, no. 12, pp. 810–812, Jun. 1990.
- [47] A. Papoulis, *Probability, Random Variables and Stochastic Processes*. New York: McGraw-Hill, 1965.
- [48] P. Yegani and C. D. McGillem, "A Statistical Model for the Factory Radio Channel," *IEEE Trans. on Commun.*, vol. COM-39, no. 10, pp. 1445–1454, Oct. 1991.
- [49] Y. D. Yao and A. U. H. Sheikh, "Outage Probability Analysis for Microcell Mobile Radio Systems with Co-channel Interferers in Rician/Rayleigh Fading Environment," *Electronic Letters*, vol. 26, no. 13, pp. 864–866, Jun. 1990.
- [50] P. Yegani and C. D. McGillem, "A Statistical Model for Line-of-Sight (LOS) Factory Radio Channel," in *Proc. VTC 89*, (San Francisco, Ca.), pp. 496–503, May 1989.
- [51] D. Akerberg, "Properties of a TDMA Picocellular Office Communication System," in *Proc. VTC 89*, (San Francisco, Ca.), pp. 186–191, May 1989.
- [52] S. J. Howard and K. Pahlavan, "Doppler Spread Measurements of Indoor Radio

- Channel,” *Electronic Letters*, vol. 26, no. 2, pp. 107–108, Jan. 1990.
- [53] R. W. Lorenz, “Impact of Frequency-Selective Fading on Digital Land Mobile Radio Communication at Transmission Rates of Several Hundred kbit/s,” *IEEE Trans. on Vehicular Tech.*, vol. 35, no. 3, pp. 122–128, Aug. 1987.
 - [54] S. Haykin, *Adaptive Filter Theory*. Englewood Cliffs, New Jersey: Prentice-Hall, 1991.
 - [55] I. J. Gupta and A. A. Ksienski, “Effect of Mutual Coupling on the Performance of Adaptive Arrays,” *IEEE Trans. on Antennas Propag.*, vol. AP-31, no. 5, pp. 785–791, Sep. 1983.
 - [56] B. Widrow, J. McCool, and M. Ball, “The Complex LMS Algorithm,” in *Proc. of the IEEE*, vol. 63, no. 4, pp. 719–720, Apr. 1975.
 - [57] C. Schirmer, *Programming in C for Unix*. Chichester, England: Ellis Horwood Ltd., 1987.
 - [58] B. Widrow, J. McCool, M. G. Larimore, and C. R. Johnson, “Stationary and Nonstationary Learning Characteristics of the LMS Adaptive Filter,” in *Proc. of the IEEE*, vol. 64, no. 8, pp. 1151–1162, Aug. 1976.
 - [59] A. V. Oppenheim and R. W. Shafer, *Discrete Time Signal Processing*. Englewood Cliffs, NJ: Prentice-Hall, 1989.
 - [60] R. A. Monzingo and T. W. Miller, *Introduction to Adaptive Arrays*. New York: John Wiley and Sons, 1980.
 - [61] R. S. Elliott, *Antennas Theory and Design*. Englewood Cliffs, NJ: Prentice-Hall, 1981.
 - [62] M. Despinic and D. Taylor, “A Simulation Study of ACI and TCM in M-QAM Digital Microwave Radio,” Master’s thesis, Faculty of Engineering, McMaster University, CRL Report No. 196, Hamilton, Ontario, Canada, Oct. 1989.

END

2|4 0|5 93

FIN