INFORMATION TO USERS

This manuscript has been reproduced from the microfilm master. UMI films the text directly from the original or copy submitted. Thus, some thesis and dissertation copies are in typewriter face, while others may be from any type of computer printer.

The quality of this reproduction is dependent upon the quality of the copy submitted. Broken or indistinct print, colored or poor quality illustrations and photographs, print bleedthrough, substandard margins, and improper alignment can adversely affect reproduction.

In the unlikely event that the author did not send UMI a complete manuscript and there are missing pages, these will be noted. Also, if unauthorized copyright material had to be removed, a note will indicate the deletion.

Oversize materials (e.g., maps, drawings, charts) are reproduced by sectioning the original, beginning at the upper left-hand corner and continuing from left to right in equal sections with small overlaps. Each original is also photographed in one exposure and is included in reduced form at the back of the book.

Photographs included in the original manuscript have been reproduced xerographically in this copy. Higher quality 6" x 9" black and white photographic prints are available for any photographs or illustrations appearing in this copy for an additional charge. Contact UMI directly to order.

UMI

A Bell & Howell Information Company
300 North Zeeb Road, Ann Arbor MI 48106-1346 USA
313/761-4700 800/521-0600

Characterization of 29.5 GHz Broadband Indoor Radio Channels Using a Steerable Receive Antenna

by

Gamantyo Hendrantoro, 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

April 1997

© copyright

1997, Gamantyo Hendrantoro

National Library
of Canada

Acquisitions and
Bibliographic Services

395 Wellington Street
Ottawa ON K1A 0N4
Canada

Bibliothèque nationale
du Canada

Acquisitions et
services bibliographiques

395, rue Wellington
Ottawa ON K1A 0N4
Canada

Your file *Votre référence*

Our file *Notre référence*

The author has granted a non-exclusive licence allowing the National Library of Canada to reproduce, loan, distribute or sell copies of this thesis in microform, paper or electronic formats.

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

L'auteur a accordé une licence non exclusive permettant à la Bibliothèque nationale du Canada de reproduire, prêter, distribuer ou vendre des copies de cette thèse sous la forme de microfiche/film, de reproduction sur papier ou sur format électronique.

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

0-612-22120-2

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

Characterization of 29.5 GHz Broadband Indoor Radio Channels Using a Steerable Receive Antenna

submitted by

Gamantyo Hendrantoro, B.Eng.

in partial fulfillment of the requirements for
the degree of Master of Engineering

Professor David Falconer, thesis supervisor

Professor Mohammed El-Tanany, thesis supervisor

Professor Samy Mahmoud, chair, Department of Systems and Computer Engineering

Carleton University

Date: May 26, 1997

Abstract

The work reported in this thesis sought to investigate the characteristics, in terms of parameters such as the channel delay spread and the received power, of broadband millimeter-wave indoor radio channels when at both the transmitter and the receiver a directional antenna is employed. It was also of great interest to study the effect of the variation in reception angle on these characteristics. The result of this study is significant in designing transmission systems suitable for broadband mm-wave wireless communication systems deployed indoors.

Four rooms in several buildings on the Carleton University campus were selected as experimental sites, one of which was a hallway while the rest were of rectangular shape. A spread-spectrum based channel measurement system was adopted, using a 1023-bit pseudonoise sequence generated at 200 Mbps and transmitted over a 29.5 GHz carrierwave using a biphasic modulation process. A 30° -beamwidth horn antenna was employed at the transmitter, while the receive antenna was a programmable phased-array antenna of 8° beamwidth. Field measurements were accomplished by acquiring complex envelope samples of the received signal and the received power for each of eleven reception angles, that range from -30° to 30° with 6° spacings in between, programmed on the receive antenna controller. The channel impulse response estimates were obtained afterward by crosscorrelating the measured complex envelope and a replica of the original sequence.

The main results from the analyses include that with directional antennas of the same types as ours at both sides of the link, (i) high suppression of interferences coming from angles different than that of the direct path is achieved while making the received power sensitive to antenna misalignment, (ii) multipath delay spread is influenced by the room dimension along which the transmission link exists while its angle-dependent behavior is affected by the antenna characteristics, (iii) an angular difference of 6° between two branches of an angle diversity system is sufficient to yield a low correlation, and (iv) the average received power decreases as the inverse n^{th} power of the propagation range, where the exponent n is on the order of 0.93, when both antennas are aligned.

Acknowledgments

I would particularly like to thank my supervisors, Prof. David Falconer for introducing me to the opportunity of using the CTR measurement system for my thesis research, for his views on the conduct of my study and for the discussion on the experimental results, and Prof. Mohammed El-Tanany for his invaluable advice in technical and practical aspects of my research and for his highly constructive comments on the writing of my thesis.

I would like to thank Dr. Robert Bultitude for the discussion on the experimental results on the distance-power law. I would also like to thank Dr. Yves Roy who developed the original broadband mm-wave wireless measurement system, helped me to understand how it works, and suggested possible ways to modify it to suit the purpose of my research. The help of Prof. M. Lecours of Laval University and Simon Damphousse, who supplied the phased-array antenna radiation patterns, is also much appreciated. Thanks to fellow student László Házy for helping me transport the measurement setup during the first stages of the field measurements and for sparking very fruitful and mind-enriching discussions on many subjects related to our respective theses. The assistance of Guowu Zheng, who helped move the experimental setup during the second stage of the field measurements and, together with Alberto Ginesi, assumed the roles of people in motion during the temporal variation measurements, is greatly acknowledged as well. Thanks also to Wirawan, Endroyono and Dr. M. Salehudin of ITS Indonesia for their opinions on the subject of radio wave reflection. Thanks are also due to Budi Santoso for the use of his digital scanner.

Greatest appreciations are due to people in the Department of Systems and Computer Engineering and people in the Laboratories of Portable Radio Communications and Communication System Simulation of Carleton University, for their help and assistance. The measurement system was realized within the framework of the Canadian Institute for Telecommunications Research. My study in the Carleton University's MEng program was fully financed by the Sepuluh Nopember Institute of Technology, Surabaya, Indonesia, and the Indonesian Ministry of Education and Culture through its Higher Education Project, from the ADB loan #1253-INO. In Canada, the administration of the scholarship funding was well organized by the World University Service of Canada. Highest appreciations are therefore extended to all of them for their perpetual support and encouragement.

Table of Contents

Abstract.....	iii
Acknowledgments.....	iv
Table of Contents	v
List of Figures.....	ix
List of Tables	xii
List of Acronyms and Abbreviations	xiii
List of Mathematical Symbols and Notations.....	xvi
 Chapter 1 Introduction.....	 1
1.1 Background	1
1.2 Problem Statement.....	2
1.3 Thesis Significance	3
1.4 Summary of Results.....	3
1.5 Thesis Organization.....	5
 Chapter 2 Theoretical Background.....	 7
2.1 Multipath Radio Channels.....	7
2.1.1 Mathematical Modeling.....	7
2.1.2 Multipath Dispersion.....	8
2.1.3 Multipath Fading	10
2.1.4 Microdiversity Techniques	12
2.1.5 Power-Distance Relation.....	13
2.2 Radio Channel Measurement Techniques.....	14
2.2.1 Types of Channel Measurements	14
2.2.2 Swept-Frequency Method	15
2.2.3 Pulse Sounding Method.....	15
2.2.4 SS-Based Method	16
2.3 Relevant Statistical Tools	18
2.3.1 Correlation function.....	18

2.3.2 Linear regression	19
2.3.3 Kolmogorov-Smirnov Test	20
Chapter 3 Literature Review and Thesis Problem	21
3.1 Review of Relevant Recent Research Projects	21
3.1.1 Introduction	21
3.1.2 Millimeter Wave Indoor Radio Measurements and Characterizations	21
3.1.3 The Effects of Directional Antennas on Channel Characteristics	25
3.1.4 Radio Wave Arrival Angle and Angle Diversity	27
3.2 Thesis Problem Statement	29
Chapter 4 Measurement System	31
4.1 Introduction	31
4.2 Measurement Apparatus	31
4.2.1 System Overview.	31
4.2.2 Transmitter	38
4.2.2.1 Baseband Stage	38
4.2.2.2 IF Carrierwave Generation	39
4.2.2.3 IF Stage	41
4.2.2.4 LO Carrierwave Generation	41
4.2.2.5 RF Stage	42
4.2.2.6 Transmit Antenna	44
4.2.3 Receiver.	44
4.2.3.1 LO Carrierwave Generation	44
4.2.3.2 RF Stage	45
4.2.3.3 IF Stage	45
4.2.3.4 Baseband Stage	47
4.2.4 Adaptive Array Antenna	48
4.2.4.1 Antenna Components	48
4.2.5 Data Acquisition and Storage	49
4.2.6 Calibration	51

4.2.6.1 Calibration Setup	51
4.2.6.2 Calibration Curve	52
4.3 Experimental Procedure	53
4.4 Measurement Locations	56
4.5 Measurement System Limitations	57
Chapter 5 Measurement Results, Analyses and Discussions	61
5.1 Data Processing	61
5.2 Temporal Variation	66
5.3 Delay Spread Distribution	70
5.3.1 Delay Spread Distribution over All Measurements	72
5.3.2 Delay Spread Distribution for Various Experiment Locations	74
5.3.3 Delay Spread Distribution for Different Reception Angles	79
5.4 Angle-Dependent Received Power	83
5.5 Angular Correlation Function	88
5.6 Power-Distance Relation	91
5.7 Link Budget Analysis	105
Chapter 6 Conclusions	108
6.1 Thesis Summary	108
6.1.1 Summary of Experiment	108
6.1.2 Summary of Results	109
6.2 Recommendations for Future Work	111
List of References	113
Appendix A Floor Plans of Measurement Locations	118
Appendix B Photographs of Measurement Setup	120
Appendix C Calibration Curve	122
Appendix D Antenna Radiation Patterns	123
Appendix E Data Analysis Program Listings	126
E.1 Introduction	126
E.2 Programs for Data Processing	126

E.3 Program for Temporal Variation Analysis	128
E.4 Program for Delay Spread Distribution Computation.	129
E.5 Programs for Different-Angle Interference Analysis	136
E.6 Programs for Angle Correlation Function Computation.	137
E.7 Programs for Power-Distance Relation Analysis	141

List of Figures

Figure 4.1: Simplified block diagram of the experimental setup	32
Figure 4.2: Block diagram of the complete experimental setup	33
Figure 4.3: Block diagram of calibration setup	52
Figure 4.4: Data acquisition flowchart	55
Figure 4.5: Room 4038 of Minto Building	57
Figure 4.6: Structure of measurement data collection	58
Figure 5.1: Top view of a transmit-receive antenna configuration with a pictorial description of angular and other parameters used in subsequent discussions	63
Figure 5.2: Magnitudes of impulse response estimates normalized to have equal energy, with the associated reception angle and static rms delay spread values printed on each graph	64
Figure 5.3: Relative received power (dB) vs. reception angle ($^{\circ}$)	65
Figure 5.4: Magnitude in dB of average time autocorrelation function of the channel impulse response estimate for 15° pointing angle	70
Figure 5.5: Complementary cumulative distribution functions of maximum and rms delay spreads for different threshold levels, computed over all measurements	73
Figure 5.6: Complementary cumulative distribution functions of maximum and rms delay spreads for various experiment locations, with threshold levels at 20 and 30 dB below the power delay profile peak	75
Figure 5.7: Complementary cumulative distribution functions of maximum delay spread normalized to the corresponding room size, with threshold levels at 10, 20 and 30 dB below the power delay profile peak	78
Figure 5.8: Complementary cumulative distribution functions of the maximum and rms delay spreads for various reception angles, with threshold levels being set at 20 and 30 dB below the power delay profile peak	80

Figure 5.9: Complementary cumulative distribution functions of the maximum delay spread measured in individual experimental rooms, with the threshold level being set at 30 dB below the power delay profile peak..	82
Figure 5.10: A possible setting of network of base station and user terminals with directional antennas where interference is caused by signals with different angle of arrival.	83
Figure 5.11: Average θ -to- 0° received power ratio $\mu_R(\theta)$, with superimposed vertical lines showing range of mean $\pm$ standard deviation $\sigma_R(\theta)$	84
Figure 5.12: Two possible transmit and receive antennas settings, (a) without and (b) with the presence of a grating lobe on the receive antenna horizontal pattern..	85
Figure 5.13: Angular autocorrelation functions averaged over all measurements, (a) involving pointing angles within the $\pm 30^\circ$ range and (b) involving only those in the $\pm 12^\circ$ range.	90
Figure 5.14: Scatter plots of relative received power versus distance for various values of reception angles, on each of which the associated linear curve fit is superimposed.	95
Figure 5.15: Linearly fitted curves of relative received power (dB) vs. propagation distance (m) in logarithmic scale for all reception angles ($^\circ$).	96
Figure 5.16: Plots of experimental cumulative distribution functions of deviations in dB of the measured received powers from the associated linear-fit means for different reception angles, along with the theoretical normal CDFs of the same means and variances.	97
Figure 5.17: Results of simulations for deterministic settings	102
Figure 5.18: Histograms of the simulation results for the case of randomized floor and walls.	103
Figure 5.19: Simulated received power for the distance range of 3–10 meters, with randomly varying positions of floor and walls.	104
Figure A.1: Floor plan of the 4 th floor of Carleton University’s Minto CASE Building, with shaded areas indicating experimental sites.	118
Figure A.2: Floor plan of the 4 th floor of Carleton University’s Architecture Building, with the shaded area indicating the room chosen as the experimental site.	119

Figure B.1: Transmitter cart that carries the PC that controls the whole measurement process, the transmitting subsystems (including the modulator and the upconverter), the receiver's demodulator, the data acquisition subsystems, and the transmit horn antenna.	120
Figure B.2: Receiver cart that carries the receive phased-array antenna, the PC that controls the antenna phase-shifters, and the receiver's downconverter.. . . .	121
Figure C.1: Calibration curve for power measurement. The vertical axis shows the reading from power meter #2 (see figure 4.2(b)), whereas the horizontal shows the associated actual power level detected at the output of the receive antenna (the input to the receiver's RF stage) relative to a reference level of -16 dBm.	122
Figure D.1: E plane of the receive array antenna radiation pattern, at pointing angles of (a) 0° , (b) 10° , and (c) 20°	123
Figure D.2: H plane of the receive array antenna radiation pattern for pointing angles of 0° and 30°	124
Figure D.3: E plane of the transmit horn antenna radiation pattern [12].. . . .	125
Figure D.4: H plane of the transmit horn radiation pattern [12].	125

List of Tables

Table 4.1: List of operating parameters for selected components	35
Table 5.1: Statistics of channel maximum and RMS delay spreads over all measurements	73
Table 5.2: List of $\hat{n}(\theta)$, $\hat{P}(1, \theta)$, $\hat{\sigma}_{\varepsilon(\theta)}$, $D_{max}(\theta)$, $N(\theta)$, and the 1 st percentiles of experimental and theoretical Gaussian CDFs of the random error	98
Table 5.3: Link budget calculation	106

List of Acronyms and Abbreviations

2-DPSK.....	binary differential phase shift keying
AGC	automatic gain control
ALC.....	automatic leveling control
ATM	asynchronous transfer mode
BER.....	bit error rate
BPF	bandpass filter
BPSK	binary phase shift keying
CCDF	complementary cumulative distribution function
CDF.....	cumulative distribution function
CITR.....	Canadian Institute for Telecommunications Research
CW	continuous carrierwave
dB	decibel
dBi.....	dB of antenna gain relative to isotropic gain
dBm.....	dB of power relative to one milliwatt
DC.....	direct current
DFT.....	discrete-time fourier transform
dir-dir	directional-to-directional
ft	feet
GHz.....	gigahertz
GS/s.....	gigasamples per second
horn-omni	horn-to-omnidirectional
Hz	hertz
<i>I</i>	in-phase
IDFT	inverse discrete-time fourier transform
IF.....	intermediate frequency
K	Kelvin
kHz	kilohertz
LAN	local area network
LNA	low-noise amplifier

LO..... local oscillator
 LOS..... line-of-sight
 LPF..... lowpass filter
 M sequence..... maximal-length sequence
 m..... meters
 mA..... milliamperes
 MB..... megabytes
 Mbps..... megabits per second
 MHz..... megahertz
 mm..... millimeters
 mm-wave..... millimeter-wave
 μ s..... microseconds
 ms..... milliseconds
 mV..... millivolts
 NRZ..... non return-to-zero
 ns..... nanoseconds
 omni..... omnidirectional
 omni-horn..... omnidirectional-to-horn
 omni-omni..... omnidirectional-to-omnidirectional
 PC..... personal computer
 PLO..... phase-locked oscillator
 PN..... pseudonoise
 PSK..... phase shift keying
 PWG..... pseudonoise waveform generator
 Q ..... quadrature
 RAM..... random access memory
 RF..... radio frequency
 rms..... root mean square
 Rx..... receiver
 s..... seconds
 SNR..... signal-to-noise power ratio

SS ..... spread spectrum

Tx ..... transmitter

V ..... volts

W..... watts

List of Mathematical Symbols and Notations

a, b	coefficients of linear regression
a_i	i^{th} element in a set of observed values of a
$a_n(t)$	time-varying amplitude of the n^{th} multipath component
$\hat{a}$	estimate of a
B_c	bandwidth of coherence
c	speed of light ($= 2.99792458 \times 10^8$ m/s)
$D_{\max}$	maximum difference between experimental and hypothesized CDFs
D_α	maximum allowable difference between experimental and hypothesized CDFs for $1-\alpha$ confidence level
d	propagation distance
d_b	break point distance
$d(L, l, p)$	propagation distance at L, l, p
$F(L, l, p, \theta, \tau)$	power delay profile at L, l, p, θ and at excess delay τ
$F(\tau)$	power delay profile value at excess delay τ
f_c	carrier frequency
$f_{D\max}$	maximum Doppler frequency shift
f_s	sampling rate
G_T	transmit antenna gain
G_R	receive antenna gain
h_1	transmit antenna height
h_2	receive antenna height
$h(\tau), h(t, \tau)$	lowpass equivalent channel impulse response
$\hat{h}(L, l, p, \theta, \tau)$	lowpass equivalent channel impulse response estimate at L, l, p, θ
$\hat{h}(nT_s, kT_a, a), \hat{h}_n(k, a)$	n^{th} component of the k^{th} impulse response estimate collected in the a^{th} acquisition time
$\hat{h}(L, l, p, k\theta_s, nT_s), \hat{h}_n(L, l, p, k)$	n^{th} component of the impulse response estimate collected at

	location L, l, p , at the k^{th} reception angle
k	Boltzmann's constant ($= 1.380662 \times 10^{-23}$ J/K)
L	measurement location or room
$L(t)$	time-varying number of multiple paths
l	Tx-Rx arrangement in a room
N	number of chips, number of elements in a sequence
N_o	single-sided noise power spectral density
$N(\theta)$	number of measurements at angle θ
n	multiple component index, propagation exponent
$\hat{n}(\theta)$	estimated propagation exponent at angle θ
$P(1)$	free space average received power at 1 meter distance
$P(d)$	received power at distance d
P_e	probability of error
$P(L, l, p, \theta)$	received power at L, l, p, θ
P_T	transmitted power
$P(t)$	time-varying received power
$\hat{P}(1, \theta)$	estimated received power at 1 meter distance and angle θ
$\bar{P}(d)$	average received power at distance d
$PM(\theta)$	power margin for spatial variation at angle θ
p	position of transmitter in the same Tx-Rx arrangement
R	symbol rate
R_c	chip rate
$R_{xy}(m)$	complex discrete-time crosscorrelation function of $x(n)$ and $y(n)$
$R_{xy}(\tau)$	complex continuous-time crosscorrelation function of $x(t)$ and $y(t)$
$r_l(k)$	k^{th} element of the measured I sequence raw samples
$ScYInc$	incremental value of the oscilloscope's Y axis
$ScYOrigin$	origin of the oscilloscope's Y axis
$ScYRef$	reference value of the oscilloscope's Y axis
$s(L)$	size of room L

T	noise temperature
T_a	inter-acquisition period in temporal variation measurement
T_c	chip period
T_p	pulse period
T_s	symbol period
$x((n-k))_m$	circular shift operation on the m -element sequence $x(n)$
t	input impulse application time
t_c	coherence time
v	velocity
W	overall width of sampled signal spectrum
$\Gamma(\phi)$	reflection coefficient at incidence angle ϕ
Δr	differential rate in sliding correlator receiver
Δt	correlation time lag
$\Delta \theta$	angular difference in angular correlation function
$\delta(\bullet)$	Dirac delta function
ε	difference between measured and estimated powers
ε_r	relative permittivity
$\phi_n(t)$	time-varying phase of the n^{th} multipath component
λ	wavelength
$\mu_R(\theta)$	ratio of received power at θ to that at zero, averaged over all measurements
$\mu_Y(x)$	mean of Y as a function of x
π	3.141592654
θ	reception angle, receive antenna pointing angle
θ_s	angular spacing
σ^2	variance
$\sigma_R(\theta)$	standard deviation of the ratio of received power at θ to that at zero
σ_ε^2	variance of difference between $P(d)$ and $\bar{P}(d)$
τ	excess delay time, correlation time lag
τ_{max}	maximum delay spread

$\tau_{max}(L, l, p, \theta, \upsilon)$	maximum delay spread at L, l, p, θ , with threshold υ
$\tau_n(t)$	time-varying excess delay time of the n^{th} multipath component
$\tau_{norm}(L, \upsilon)$	maximum delay spread at L with threshold υ , normalized to the size of room L
τ_p	pulse width
τ_{rms}	rms delay spread
$\tau_{rms}(L, l, p, \theta, \upsilon)$	rms delay spread at L, l, p, θ , with threshold υ
$\bar{\tau}$	mean excess delay
υ	threshold in dB below power delay profile peak
ϕ	incidence angle
$\{a_i\}$	set/sequence of values of a
$\sum_{n=n_1}^{n_2}$	summation over a range of n from n_1 to n_2
$\int_a^b$	integration from a to b
$\otimes$	continuous- or discrete-time convolution operator
$\otimes_N$	N -point discrete-time circular convolution operator
$ \bullet $	absolute value
$\langle \bullet \rangle_a$	averaged over all possible values of a

Chapter 1

Introduction

1.1 Background

Exponential increases in demands for multimedia communications, along with recent advances in broadband communication technology, have driven the development of communication technology toward broadband applications. Exploration into broadband communications has even been furthered in many independent research projects to make use of the ATM (asynchronous transfer mode) to transport high speed data over indoor LANs (local area networks). Future applications made possible by the deployment of such systems include video conferencing, high resolution image file transfer, and so on.

Following a very different route, the communications community has also witnessed an overwhelmingly fast progress in mobile and wireless communication technology. User terminal mobility and elimination of necessity to install costly physical links required by users to access services are two of the benefits eminent from the use of such technology. The adoption of the cellular concept for spectrum conservation into this system has made it even more attractive to be selected as the ultimate method of modern communications.

A more potentially beneficial application comes into the picture when researchers start to contemplate combining these two technologies. Broadband communication among several multimedia terminals connected to a broadband LAN may now be carried out without having to build the expensive wiring between the terminals and the network. Independent investigations [1] into the feasibility of realizing such technology, together with standardization efforts [2], have been going on for several years in many places all over the world.

Research activities being done in the CITR (Canadian Institute for Telecommunications Research) projects have resulted in a system architecture proposed for a 160-Mbps wireless access to an ATM LAN operating at radio frequencies as high as millimeter wave bands, that is, frequencies in the range of 20-60 GHz [3]. One major technical challenge in realizing such a system is how to deal with the radio aspect, a problem of how to combat the deleterious effects of time-variant multipath fading and dispersion prominent in any indoor environment due to a

number of static or moving obstacles, reflectors and scatterers, and how to suppress interference from unwanted signals. In its attempt to solve the problem, the CITR team has proposed the use of an antenna diversity system with selection combining, utilising a number of directional antennas in both base and user stations aimed at different directions such that a 360° coverage is achieved for both stations. An antenna system very similar to this has also been employed in the Motorola's ALTAIR system [4].

The use of such techniques, commonly called angular diversity, with directional antennas has several purposes. Besides combating the destructive impacts of multipath fading, it is also aimed to reduce the effects of multipath dispersion and interfering signals by establishing a narrow radio path between two directional antennas, one at each of the transmitter and receiver.

1.2 Problem Statement

Despite a considerable number of endeavors in the past to understand such antenna systems more deeply, there are questions left unanswered regarding the effectiveness of such systems to achieve the expected goals. Previously impossible efforts to answer these problems have been made possible by the availability of an adaptive array antenna designed by a Laval University research team, as well as the final realization of a 29.5-GHz broadband radio channel measurement system at Carleton University. The integration of both systems are expected to yield a deeper insight into the use of directional antennas for angular diversity in indoor broadband mm-wave communication systems.

The problems this thesis seeks to solve are, therefore,

- a. How differences in the type of measurement environment and in the direction of reception may affect the delay spread measures of indoor radio channels.
- b. What level of interference coming from a direction different than that of the desired signal appears at the receiver in a directional-to-directional antenna setting.
- c. How misalignment between two directional antennas affects the quality of the received signal in terms of reduction in the received power.
- d. What degree of correlation can be achieved from two-antenna diversity systems, in which two antennas are pointed in different directions.

- e. How the received power is related to the transmitter-receiver separation range in an indoor setting with directional antennas, and how antenna misalignment affects this relation.

1.3 Thesis Significance

Accurate answers to the above-stated problems would certainly be very useful in weighing the performance of angular diversity techniques with directional antennas in broadband mm-wave indoor settings. Outcomes on interference level and misalignment effects could be used further to devise a better directional-antenna diversity system that would produce a better performance in terms of the received signal quality. In the light of such objectives, an exemplary link budget design is given at the end of the analysis in this thesis.

1.4 Summary of Results

The measurements were performed in four different rooms representing various possible environments in which an indoor wireless communication may take place. The measurement system is a spread spectrum-based one that employs a 1023-bit pseudonoise sequence transmitted in binary PSK form at 200 Mbps over a 29.5 GHz carrier wave. Directional antennas, a horn and a phased-array antenna, were used at the transmitter and the receiver, respectively. Measurements at different directions of reception were accomplished by making use of the receive antenna pointing angle programming capability.

Before reviewing the results of data analyses, it should always be kept in mind that due to some limitations of the measurement system, namely the phase controller hysteresis and the grating lobe appearance of the adaptive array antenna as well as the limited buffer and slow speed data transfer of the sampling oscilloscope, some results must be carefully interpreted. These limitations also prevented the acquisition of data needed to observe how angular diversity can reduce the required power received under multipath fading conditions. The results are summarized below.

- a. The temporal variations of the impulse response estimates are very small provided that people movements are restricted outside the region covered by the antennas' main lobes, such that results of the measurement over eleven discrete reception angles stretching over an 8-second period can be used in comparing the channel characteristics among those angles.

- b. From the measured power delay profiles it is found that for any receive antenna pointing angle, a predominant power contribution comes from the direct path component.
- c. Over a variety of rooms and antenna arrangements, the 99th percentiles of the maximum delay spread (referred to also in many literature as the multipath spread) and the rms delay spread of the channel are approximately 393 ns and 60 ns, respectively, for a threshold level set to 30 dB below the individual impulse response peaks.
- d. The maximum delay spread of the channel impulse response is influenced by the dimension of the room along which the measurement link is set up, referred to herein as the dominant dimension, and the threshold applied to the measured impulse response. A very simplified theoretical calculation of even-order reflection relative magnitude tends to support the conjecture that the measured maximum delay spreads are likely to take on values equivalent to the even multiples of the dominant dimension of the room, meaning that even-order reflections off the walls perpendicular to the dominant dimension may be amongst the dominant modes of radio propagation in rectangular rooms.
- e. No significant differences in the maximum and rms multipath spreads result from the use of different reception angles. However, some exceptions occur, such as a tendency of some reception angles to produce small delay spreads, which is possibly attributed to some unknown characteristics of the receive array antenna.
- f. The 0° reception angle, a situation in which antennas at both sides are aligned, does not necessarily yield the smallest delay spread, which perhaps is due to the presence of large even-order reflections propagating in parallel with the measurement link and entering the receive antenna through its main lobe.

The following findings which contain some numerical results, in particular those given in points (g)–(i), are probably unique to our measurement system that used a 30°-beamwidth transmit antenna and an 8°-beamwidth steerable receive array antenna.

- g. An interference signal coming from a direction outside the receive antenna main lobe will be suppressed by 10 dB relative to the desired signal of 0° reception that experiences the maximum antenna gain.

- h. As large a power drop as 6 dB from the 0° reception level may occur due to the antenna misalignment by as small as 6° .
- i. An angle difference of as small as 6° yields a correlation of 0.1 between two branches of an angular diversity system. Smaller correlations are obtained by increasing the differential angle between these two branches.
- j. The average received power in dB is a linearly decaying function of the logarithmic propagation distance in meters, with distance exponents varying for different reception angles from 0.31 to 1.41. For 0° receive antenna pointing angle, the distance exponent is found to be 0.93, far less than the initially-expected free space exponent of two. Reasons for these results are suggested in section 5.6 of this thesis, accompanied by their computer-simulated verifications.
- k. Deviations around the distance-dependent received power median, attributed to spatial variations in multipath structure, are lognormally distributed for some of the reception angles, including 0° . The measured standard deviation ranges from 0.78 to 3.49 dB for various reception angles, with the smallest existing at 0° angle. To account for this spatial variation in a link budget calculation, a power margin of 8 dB is included.

1.5 Thesis Organization

Following this introductory chapter, theoretical background essential in understanding and solving the thesis problems will be overviewed. It includes descriptions on time-varying multipath indoor radio channels, channel measurement techniques and necessary statistical methods to be used in later analyses. The third chapter reviews recent investigations being done on the same issue, the results of which became the starting point of the thesis research. In the same chapter, the thesis problems are formulated in more detail, including the methods taken to solve them.

Chapter four elaborates measurement apparatuses and procedures employed throughout the experiment. Block by block description of transmitter, receiver and data acquisition systems is given, completed by a list of steps taken during each of measurement sessions and a description on the experimental sites. Limitations of the measurement system are also explained. The fifth chapter, which makes up the core of this thesis, discusses the processing and analysis of

measurement data to arrive at solutions to the thesis problems. These solutions are then summarized in the last chapter, followed by recommendations of necessary refinement for further experiment or future topics that still remain unanswered by this research.

Chapter 2

Theoretical Background

2.1 Multipath Radio Channels

2.1.1 Mathematical Modeling

Electromagnetic waves emitted from a radio transmitter will undergo a variety of treatments by the propagation medium before they reach the destination receiver. Some portion of the transmitted energy will travel along a direct imaginary line connecting the transmit antenna to the receive antenna. Some other portions will experience such mechanisms as reflection, diffraction and scattering by structures and objects in the vicinity of the radio link. Consequently, the resultant signal detected by the receiver is in essence the sum of all individual components that arrive through different paths with different gains, phase shifts and arrival times. As the positions of either the transmitter, the receiver, or the reflecting structures change, there must also be changes in the resultant received signal.

Further reasoning on the basis of the previously stated phenomena leads to the adoption of linear time-varying finite-impulse response filter in modeling the lowpass equivalents of mobile radio channels. In such model, a radio channel baseband equivalent is assumed to have a finite length impulse response consisting of a train of replicas of the transmitted, hypothetical impulse waveform, i.e. the input to the linear filter. Each replica shows a unique amplitude, phase and time delay with respect to the impulse transmission time. Hence, the lowpass equivalent channel response $h(t, \tau)$ can be expressed mathematically as follows [5], [6]

$$h(t, \tau) = \sum_{n=1}^{L(t)} a_n(t) e^{j\phi_n(t)} \delta(\tau - \tau_n(t)) \quad (2.1)$$

where $L(t)$ is the time-varying number of multiple paths, while $a_n(t)$, $\phi_n(t)$ and $\tau_n(t)$ are the time-varying amplitude, phase and excess delay, respectively, of the n^{th} path component, and $\delta(\cdot)$ is the Dirac delta function.

In reality, as it is impossible to generate a real impulsive waveform, pulse responses serve

as approximates of the channel impulse responses in describing radio channel characteristics. With respect to the inverse of the width of the transmitted pulse, radio communication systems can be classified into two groups. If the receiver system bandwidth is much larger than the inverse of the pulse width, the receiver will be able to resolve individual received pulses such that equation (2.1) applies. This system is referred to as wideband or broadband. If the receiver bandwidth is too narrow to resolve the individual received pulses, a narrowband system results, in which the $L(t)$ of equation (2.1) equals unity.

It is readily seen from equation (2.1) that the characteristics of a radio channel vary along the axes of observation time τ and impulse transmission time t . Variations in amplitude and phase of the channel response along each of these two time axes take two separable forms; the one caused by the variation in transmission time t manifests itself in multipath fading or Doppler spread, whereas the other, which results from the changes along the observation time τ , is recognized in the form of multipath dispersion or frequency selectivity.

The noiseless baseband output $y(t, \tau)$ from a radio channel having lowpass equivalent channel impulse response $h(t, \tau)$ when a lowpass signal $x(t, \tau)$ modulates the transmitted signal is

$$\begin{aligned} y(t, \tau) &= h(t, \tau) \otimes x(t, \tau) \\ &= \sum_{n=1}^{L(t)} a_n(t) e^{j\phi_n(t)} x(\tau - \tau_n(t)) \end{aligned} \quad (2.2)$$

where $\otimes$ denotes a convolution operator.

2.1.2 Multipath Dispersion

To quantify the effects of multipath dispersion onto the received signal, it is necessary to compute the power delay profile, also known as the delay power spectrum or the multipath intensity profile, of the channel. To derive the profile, individual power profiles are first computed from the individual impulse responses as follows

$$F(t, \tau) = |h(t, \tau)|^2 = \sum_{n=1}^L |a_n(t, \tau)|^2 \delta(\tau - \tau_n(t)) \quad (2.3)$$

The functions are next averaged over some period of time, or equivalently over some distance when the receive terminal is moving, to yield the time (or space) averaged power delay profile as follows

$$\bar{F}(\tau) = \langle F(t, \tau) \rangle_t \quad (2.4)$$

Some measures of the channel multipath dispersion can then be determined from this power delay profile. The most useful ones are the channel mean excess delay $\bar{\tau}$, the rms delay spread τ_{rms} and the maximum excess delay τ_{max} (also known as the multipath spread or the maximum delay spread). The maximum excess delay is simply the time interval from the first component arrival to the point at which the profile drops below an observer-defined threshold value, ν dB below the profile peak. The mean excess delay is the first moment of power delay profile, whereas the rms delay spread is the square root of the profile second central moment. Both can be expressed mathematically as follows after setting the arrival time of the first path to 0 (i.e. $\tau_1=0$)

$$\bar{\tau} = \frac{\sum_{n=1}^L \tau_n \bar{F}(\tau_n)}{\sum_{n=1}^L \bar{F}(\tau_n)} \quad (2.5)$$

$$\tau_{rms} = \sqrt{\frac{\sum_{n=1}^L \tau_n^2 \bar{F}(\tau_n)}{\sum_{n=1}^L \bar{F}(\tau_n)} - \left(\frac{\sum_{n=1}^L \tau_n \bar{F}(\tau_n)}{\sum_{n=1}^L \bar{F}(\tau_n)} \right)^2} \quad (2.6)$$

When the values of τ_{max} and τ_{rms} are computed from $F(\tau)$ of equation (2.3) instead of $\bar{F}(\tau)$ of (2.4), they are commonly referred to as the static maximum and rms delay spreads, respectively. From the above definitions, it is apparent that these parameters depend on the threshold value ν .

Another measure of dispersive channels is the coherence bandwidth B_c , an important parameter in determining the most appropriate channel model used in any analysis. A signal dwelling a spectrum narrower than this coherence bandwidth is considered to suffer from

frequency-flat fading, while that residing in a wider band than the coherence bandwidth undergoes frequency-selective fading. Two signals separated in the frequency domain by less than B_c Hz will experience correlated fading. The exact value of B_c is difficult to measure, but it is known to be inversely proportional to the channel rms delay spread. A practical expression is as follows for a fading correlation coefficient of 0.5 [7]

$$B_c \approx \frac{1}{5\tau_{rms}} \quad (2.7)$$

where τ_{rms} is obtained from equation (2.6).

2.1.3 Multipath Fading

Multipath fading results from the vectorial addition of all individual complex components of the received signal. In terms of total received power, this phenomenon will produce different effects on narrowband and broadband signals. For narrowband signals with a constant envelope of unity (i.e. $x(\tau) = 1$ in equation (2.2)), the total received power $P(t)$ is

$$P(t) = \left| \sum_{k=1}^{L(t)} a_n(t) e^{j\phi(t)} \right|^2 \quad (2.8)$$

whereas for broadband signals with an envelope in the form of a unit impulse signal (i.e. $x(\tau) = \delta(\tau)$ in equation (2.2)), it is

$$P(t) = \sum_{k=1}^{L(t)} |a_n(t)|^2 \quad (2.9)$$

From equations (2.8) and (2.9), several issues can be observed. Firstly, as a radio channel changes, i.e. there are variations with time t in the values of $L(t)$, $a_n(t)$ and $\phi_n(t)$, the received power will also change with time. Secondly, since the received power for a narrowband system is the squared magnitude of the vector sum of incoming signal components, it fluctuates more significantly than does the broadband received power, which is the sum of powers of individual components. Accordingly, narrowband systems are more prone to deep fades than their broadband counterparts are.

Another kind of channel modeling arises from the rate at which the channel varies with respect to the transmission signaling interval. This rate of change is denoted by the value of coherence time t_c , the maximum timespan within which two points in time exhibit correlated fading phenomena.

Suppose a digital transmission system employs a symbol period of T_s . The channel is regarded fast or time-selective fading if T_s is far greater than t_c , and thus the whole period of a single symbol sees varying fades [6]. The opposite situation is when T_s is much less than t_c such that the channel fading is practically constant over several symbol periods. Under this condition, the channel is said to be slowly fading. The usual assumption underlying performance analyses of wireless indoor communication systems is that the channel is slowly varying [5]. This assumption needs to be revisited as we proceed to broadband mm-wave applications

Another parameter that quantifies the channel temporal variation is the Doppler shift, a shift in frequency of the received signal from the originally transmitted frequency due to motions of the user terminal, reflecting objects or scatterers. The presence of multiple paths gives rise to a frequency spread made up of the aggregate of individual paths having their own Doppler shifts, the peak value f_{Dmax} of which depends on the user terminal velocity v as denoted below

$$f_{Dmax} = \frac{vf_c}{c} \quad (2.10)$$

where f_c is the transmitted carrier frequency, and c is the speed of light. The maximum Doppler shift f_{Dmax} has an inversely proportional relation with the coherence time t_c , the exact form of which is indeterminate, but can be approximated in practice by the following [7]

$$t_c = \frac{0.423}{f_{Dmax}} \quad (2.11)$$

The occurrence of multipath fading changes the asymptotic function relating the signal-to-noise ratio (SNR) at the receiver to the average bit or symbol error rate after detection from an exponential form to an inverse proportion [6]. As fading degrades the received signal quality in terms of average bit error rate (BER), microscopic diversity (or microdiversity) has been widely and successfully used as a means to combat the short-term fading induced by multipath situations.

2.1.4 Microdiversity Techniques

Microscopic diversity has proved to be a powerful means to mitigate the deleterious effects of short term multipath fading on the error rate performance. It makes use of multiple input channels that fade independently to be combined in such a way that the detected BER can be suppressed. There are several well-known methods of achieving independent fading channels as well as techniques to combine their outputs [6], [8].

Frequency diversity is a method of obtaining channels with highly uncorrelated fades by transmitting the same information signal at different bands with a separation of more than the channel coherence bandwidth. A similar technique applied in time domain, that is by sending the same message at points in time more than the coherence time apart, results in time diversity. The third method called space diversity is performed by installing two or more antennas with at least one half wavelength distance in between for applications in the microwave band [8]. Two practically uncorrelated channels can also be acquired by using two perpendicular antenna polarizations, a system referred to as polarization diversity. A somewhat unique type of diversity for broadband systems is realized in the form of a rake receiver that captures resolvable signal components arriving with different delays and makes use of all of them as the inputs to the combiner. Finally, a less used technique of diversity is the one called angular diversity, in which antennas are aimed at different directions and are expected to catch multipath components coming from these directions. This technique has recently been proposed for CITR broadband wireless indoor system [3] and also used in Motorola's ALTAIR system [4].

There is as well more than one way to combine the signals coming out from these uncorrelated fading channels. The optimum manner, in the sense that it minimizes the bit error rate (BER), is the maximal-ratio combining. In such a system, before the signals enter a summer, each signal is cophased and scaled in proportion to the ratio of its root-mean-square value to the mean of noise square so that a channel with a high signal-to-noise power ratio (SNR) is weighted higher than that having a low SNR, resulting in a maximum possible value of the combiner output SNR that is equal to the sum of SNR's on all branches [8],[9]. As such, as the number of uncorrelated fading channels employed increases, the SNR-BER curve moves closer toward that for non-fading condition [6].

Other possible ways are suboptimum [8]. Among them is the practically simplest method called the equal gain combining in which, unlike the maximal-ratio method, all channels are weighted equally. The second, referred to as selection combining, is accomplished by periodically checking the SNR levels on all channels and picking the channel with highest SNR to be the combiner output. The third is the so-called scanned diversity combining, in which a switch is employed to stay at a branch with a currently good SNR and moves to one of other branches through scanning only after the current branch shows a drop in SNR below a previously specified threshold.

In reality, as multiple channels that are completely uncorrelated are difficult to establish, correlated channels are made use of as the branches of the diversity system. This is supported by empirical and theoretical findings that as high a correlation coefficient among multiple branches as 0.7 is adequate to produce the desired improvement in system performance under a fading condition [8],[10].

2.1.5 Power-Distance Relation

Over the years several models describing the relationship between received power and propagation distance have been proposed and empirically backed by researchers. This section presents brief descriptions of four of the models, summarized primarily from references [5] and [7].

The first, most widely adopted model is the inverse power model, in which the average received power is described to follow the inverse n^{th} power of the propagation distance. The relation can be expressed mathematically as

$$P(d) = P(1) - 10n\log(d) + \epsilon \quad (2.12)$$

in which $P(d)$ denotes the received power in dBm at distance d meter, $P(1)$ denotes the received power in dBm at 1 meter range in free space, n is the distance exponent, and ϵ represents the random difference between the measured level and the estimated average power obtained from the first two terms of the equation, which from many experiments has been found to be lognormally distributed [7]. The difference ϵ may indicate the influence of variables other than the distance, such as the measurement errors and spatial variations, on the received power.

The other models evolve from the previously stated by slight modifications or additions of some terms. The second model is similar to the previous except in the application of a distance-variant propagation exponent n . One such model developed by Ericsson [7] employs multiple break points, between each successive pair of which a different distance exponent is applied. In addition, this model describes the difference in dB between the estimated and measured powers to be uniformly, instead of normally, distributed.

The third model derives from the first by the incorporation of additional dB terms in equation (2.12) to account for attenuations by blocking structures such as walls, partitions or floors. The last model applies $n=2$, the free-space distance exponent, to equation (2.12), and adds a logarithmic linear attenuation measured in dB/meter.

2.2 Radio Channel Measurement Techniques

2.2.1 Types of Channel Measurements

Radio channel measurement techniques can be classified into several types according to the class of measured channel and the parameters to be quantified. Based on the channel types, there are two distinctive methods; one is used for narrowband systems, the other for broadband.

Narrowband radio channel measurements are performed simply by sending an unmodulated carrier wave from the transmitter at frequency of interest, and detecting its envelope at the receiver. An ideal static channel yields a constant received envelope. A time-varying multipath channel, however, results in a constantly fluctuating envelope, a phenomenon we know as fading. To be capable of capturing the fastest variations in the signal envelope, the measurement system must be able to sample it above the Nyquist rate associated to f_{Dmax} [11]. This system is realized by means of a sampling oscilloscope connected to the receiver output, or a network analyzer to compare the variations in the received CW signal with a reference.

Broadband measurements are typically done by a broadband complex envelope modulating the carrier frequency at the band of interest. For a broadband power measurement, the broadband envelope may take the form of pseudonoise sequence modulated pulses [12]. The power measurements are then accomplished simply by acquiring the received power value from a spectrum analyzer or a power meter reading after necessary amplifications, down conversion, and

filtering.

More sophisticated techniques are required to collect estimates of the channel lowpass equivalent impulse responses. There are three commonly used methods of the channel impulse response measurement. The first one, the swept-frequency method, deals with frequency domain measurements to obtain the channel frequency response or transfer function, and to obtain its time domain impulse response through inverse discrete-time Fourier transform (IDFT). The other two are working in the time domain, one of which, called the pulse sounding method, is implemented by the periodic transmission of probing pulses, whereas the other makes use of spread spectrum (SS) concept, thus explaining the name, the SS-based method.

2.2.2 Swept-Frequency Method

This method is implemented by a vector network analyzer controlling the transmitted frequency swept over the band of interest in discrete steps [7], [11]. With the help of an S-parameter test set, channel responses are collected, each for a single discrete transmitted frequency, making up the discrete-frequency channel transfer function. With an overall sampled spectrum of W Hz and sweep steps of f_s Hz, the band-limited baseband equivalent channel impulse response estimates obtained from IDFT will have a maximum measurable spread of $1/f_s$ sec and delay resolution of $1/W$ sec. The amount of noise and delay resolution can further be adjusted by applying an appropriate window on the discrete frequency response before executing the IDFT.

One of the disadvantages of using this method is the need for hardwiring between transmitter and receiver for synchronized frequency sweep purpose. Another shortcoming is owing to the fact that one complete frequency response over the desired spectrum needs acquisition time equal to the entire frequency sweep time, within which some changes may already take place in the measured environment, making the results somewhat invalid.

2.2.3 Pulse Sounding Method

Periodic pulses of width τ_p and period T_p are sent over a carrier waveform [7]. The receiver incorporates a $2/\tau_p$ Hz wide bandpass filter and an envelope or square law or coherent-

type detector. Delay resolution achieved by this technique is equal to the transmitted pulse width τ_p , whereas the maximum measurable excess delay is as large as the inter-pulse time T_p .

Advantages of using this method are the availability of immediate results of power profile estimates if noncoherent-type detector is used and the relatively low price of required equipment. However, its weaknesses include the low susceptibility against noise and interference due to the use of a wideband receiver filter, the acquisition triggering difficulty owing to the problem in detecting the first arriving pulse under a non line-of-sight condition, and the unavailability of channel phase information in the absence of a coherent-type detector. Also, all employed apparatuses must have a large peak-to-average power ratio in order to elicit optimum results.

2.2.4 SS-Based Method

Sliding Correlator Receiver System

In this technique, prior to transmission the carrier signal energy is spread in frequency by mixing it with an N -chip pseudonoise (PN) sequence of rate R_c chips/sec, equivalent to chip period T_c of $1/R_c$ sec [7], [11]. At the receiver, the received PN-modulated signal is despread by the same PN sequence, only this time at a rate lower than R_c , a system called sliding correlator. This system incorporates a despreading mixer, a PN sequence generator at $R_c + \Delta r$, and a narrowband filter, the output of which then enters demodulation and acquisition processes. When an incoming sequence waveform with the same phase as is the receiver PN sequence goes through a chain of despreading processes, a maximum energy output pulse is detected and acquired. Signal components resulting from multiple paths arrive at different times and each will produce a detected maximum energy pulse after the above processes. Due to the inherent uncorrelated property of the employed PN sequence and its shifted versions, the acquisition system will be able to distinguish these individual pulses. As such, a complete NT_c sec channel impulse response is collected after $NT_c R_c / \Delta r$ sec processing with a delay resolution of $2T_c$.

A comparison study [13] has shown that two most commonly used methods, namely the SS-based with sliding correlator and the swept-frequency methods, have similar accuracies in estimating the measured impulse response functions while possessing balanced practical strengths

and weaknesses. Therefore, the choice of which method to adopt depends more on the equipment present availability.

Off-Line Computation of Crosscorrelation

A slightly different method from the above-described is realized by performing the crosscorrelation computation process after the completion of received waveform acquisition processes, thus eliminating the sliding correlator from the measurement system and reducing the effective acquisition time for a complete response to only NT_c sec. The off-line computation procedure is explained as follows. Suppose the input to the channel, that is the complex envelope of the transmitted signal, is $x(t)$ with an autocorrelation function

$$R_{xx}(\tau) = \int_{-\infty}^{\infty} x(t)x^*(t-\tau)dt \quad (2.13)$$

The channel output, i.e. the complex envelope of the received signal, $y(t)$ is obtained by convolving $x(t)$ with the baseband equivalent channel impulse response $h(t)$ as follows

$$y(t) = \int_{-\infty}^{\infty} h(\lambda)x(t-\lambda)d\lambda \quad (2.14)$$

The output-input crosscorrelation is, hence,

$$\begin{aligned} R_{yx}(\tau) &= \int_{-\infty}^{\infty} y(t)x^*(t-\tau)dt \\ &= \int_{-\infty}^{\infty} h(\lambda) \int_{-\infty}^{\infty} x(t-\lambda)x^*(t-\tau)dt d\lambda \\ &= \int_{-\infty}^{\infty} h(\lambda)R_{xx}(\tau-\lambda)d\lambda \end{aligned} \quad (2.15)$$

From the last form of equation (2.15), it is apparent that provided a PN sequence $x(t)$ having an impulse-like autocorrelation function $R_{xx}(\tau)$ can be generated, the output-input crosscorrelation function $R_{yx}(\tau)$ estimates the desired lowpass equivalent channel impulse response. The required correlation property is satisfied by using a maximal-length PN sequence also referred to as the M sequence for the signal complex envelope.

2.3 Relevant Statistical Tools

Some statistical tools to be used in later chapters are reviewed in the following. They include the correlation function, linear regression, and Kolmogorov-Smirnov goodness-of-fit test.

2.3.1 Correlation function

Correlation function of two random signals measures the degree of similarity between them. For discrete-time energy signals, the crosscorrelation function $R_{xy}(m)$ of two complex sequences $x(n)$ and $y(n)$ is obtained as follows [6], [14]

$$\begin{aligned} R_{xy}(m) &= \sum_{n=-\infty}^{\infty} x(n)y^*(n-m) \\ &= x(m) \otimes y^*(-m) \quad , m = 0, \pm 1, \pm 2, \dots \end{aligned} \quad (2.16)$$

where $y^*(n)$ denotes the conjugate of $y(n)$, and $\otimes$ denotes a discrete-time convolution operator. Several important properties of this function include

- $R_{xy}(m) = R_{yx}^*(-m)$
- $|R_{xy}(m)| \leq \sqrt{R_{xx}(0)R_{yy}(0)}$

For periodic signals, which make up a subset of power signals, the crosscorrelation sequence becomes [6]

$$\begin{aligned} R_{xy}(m) &= \frac{1}{N} \sum_{n=0}^{N-1} x(n)y((n-m))_N \\ &= \frac{1}{N} [x(m) \otimes_N y(-m)] \quad , m = 0, 1, \dots, N-1 \end{aligned} \quad (2.17)$$

where $y((n-m))_N$ represents circular shift operation on an N -point sequence $y(n)$, and $\otimes_N$ denotes an N -point circular convolution operator.

If $y(n) = x(n)$, the function becomes the autocorrelation sequence $R_{xx}(m)$ of $x(n)$. All properties apply for this function as they do for the crosscorrelation function.

2.3.2 Linear regression

In statistics, a regression equation is used to predict the characteristics of relationship between a controlled independent variable X and an uncontrolled, measured dependent variable Y [15]. The estimate of this relationship is derived from pairs of samples (x_i, y_i) obtained from experiment. The term linear regression refers to the regression of Y on x , for which a linear relation is assumed to exist between x and the mean of Y , i.e.

$$\mu_Y(x) \approx a + bx \quad (2.18)$$

where $\mu_Y(x)$ is the mean of Y as a function of x . The regression coefficients, a and b , are estimated from empirical data by $\hat{a}$ and $\hat{b}$. The estimated response $\hat{y}$ for any x is in turn obtained from

$$\hat{y} = \hat{a} + \hat{b}x \quad (2.19)$$

The difference between the estimate $\hat{y}_i$ and the observed output y_i for a given input value $\hat{x}_i$ is referred to as the error ϵ_i . By minimizing the sum of these squared errors, a technique called least-square method, the coefficient estimates are found as follows

$$\hat{b} = \frac{N \sum_{i=1}^N x_i y_i - \left(\sum_{i=1}^N x_i \right) \left(\sum_{i=1}^N y_i \right)}{N \sum_{i=1}^N x_i^2 - \left(\sum_{i=1}^N x_i \right)^2} \quad (2.20)$$

$$\hat{a} = \frac{\sum_{i=1}^N y_i}{N} - \hat{b} \frac{\sum_{i=1}^N x_i}{N} \quad (2.21)$$

where (x_i, y_i) is the i^{th} of N empirical data pairs.

The random error variable ϵ reflects the influence of other, uncontrolled independent variables [15]. It, however, is also affected by the measurement accuracies. If high accuracies can be achieved during the measurements, then the estimate can elicit knowledge of the impacts of the uncontrolled variables on the measured quantity. By further assuming that ϵ_i 's are samples of N

independent gaussian random variables having identical mean 0 and variance σ^2 , the estimate $\hat{\sigma}^2$ of the variance can be computed as in the following

$$\hat{\sigma}^2 = \frac{\sum_{i=1}^N \varepsilon_i^2}{N-2} \quad (2.22)$$

The loss of 2 in degree of freedom is due to the use of estimates $\hat{a}$ and $\hat{b}$, in place of their actual values a and b , in computing the errors. The estimated variance illustrates the level of variation due to the uncontrolled variables. The underlying assumption that these random variations are normally distributed, however, needs to be proved to guarantee the validity of the conclusion drawn on the error variations.

2.3.3 Kolmogorov-Smirnov Test

Kolmogorov-Smirnov one-sample test [15] is one of several techniques that can be used to statistically evaluate the goodness of fit of a theoretical distribution to that constructed from the measured data. It is based on the maximum allowable absolute difference D_α between corresponding values of the observed and the hypothesized cumulative distribution functions to achieve a specified significance level α or confidence level $1-\alpha$. If the maximum absolute difference D_{max} between these two distributions is still less than D_α , the hypothesized distribution is considered to describe well the measured data with $1-\alpha$ confidence level.

Some values of D_α have been tabulated in [15] and [16] for some frequently assumed values of α . For large data size N , D_α is in inverse proportion to the square root of N , e.g. for $\alpha=0.01$ (0.99 confidence level), $D_\alpha \approx \pm 1.63/(\sqrt{N})$ [16].

Chapter 3

Literature Review and Thesis Problem

3.1 Review of Relevant Recent Research Projects

3.1.1 Introduction

This section is mainly aimed to construct a picture of recent research progress in the field pertinent with the topic of this thesis. The research outcomes are presented to provide comparisons with our results to be examined in later chapters. Also, important points are exposed that distinguish the works done for this thesis from those reported in the recent papers so as to emphasize the significance of our results.

3.1.2 Millimeter Wave Indoor Radio Measurements and Characterizations

Since the inception of endeavors aiming to the realization of broadband indoor wireless communication networks, a number of research activities have been undertaken to characterize the indoor radio channel, in particular at millimeter wave band over which abundant spectrum is available. Those efforts have accomplished narrowband and broadband experiments to empirically learn the channel behavior in terms of short-term fading, delay spread and power-distance relation at various frequencies within the band of our interest. Some of them have as well made convenient use of computer simulations to attain the same goal.

Short-Term Fading

Although distribution fit and even envelope fading measurements are not to be attempted in our work, it is useful to acquire a background of what has been discovered by researchers on this topic.

Short-term fading for narrowband systems at 21.6 GHz and 37.2 GHz has been studied by Kalivas *et al* and reported in [17] and [18]. They found that for a narrowband system the cumulative distribution function (CDF) of the envelope fading follows the Rayleigh distribution when two omnidirectional antennas are used at both ends of the link, regardless of whether a two-

branch space diversity with selection combining is applied or not. However, when two horns are employed for the diversity branches at the receiver the CDF becomes closer to a Rician CDF. They also found that the presence of people movement in the environment contributes little in a statistical sense to changes in the fading characteristic. Another narrowband result for 60 GHz band has been published by Tharek and McGeehan [19]. Aside from similar results on the distribution fit, they reported a quite different impact of people movement nearby on the envelope fading CDF, which according to them starts to depart from the Rayleigh distribution. Bultitude *et al* [20] found from their 40 GHz narrowband measurements that the 90th percentile of maximum Doppler shift at the band is around 240 Hz, which matches the expected maximum velocity of people motion inside buildings.

Envelope fading measurements for broadband systems have been performed by Bultitude *et al* [20] and Licht [12]. Bultitude *et al*, working in 40 GHz band, revealed that the frequency of deep fades reduces as the signal spectrum becomes wider and that the fading envelope obeys a Rician distribution for half the time and other distributions, including Rayleigh, for the rest of the time. Nevertheless, they were not able to accurately determine the experimental envelope distribution tails because of insufficiency in the sampling rate. Licht, on the other hand, employed an adequately high sampling rate in his 29 GHz-band experiments and, hence, managed to accurately measure the channel fades and construct their distribution functions. He revealed that, using two horn antennas at both ends of the link, the terminal motion or people movement in the vicinity gives rise to shallow fades in the order of 2-8 dB, whereas fades as deep as 15-25 dB may occur when a person moves across the line-of-sight path. His conclusions include that for two terminals with directional antennas large fade margin is required due primarily to deep shadowing fades rather than shallower multipath fades, and that fade depth is also dependent on the closeness of the shadowing person or object to either the transmit or receive antenna.

Power-Distance Relation

Measurements have been independently conducted by Tharek and McGeehan at 60 GHz [19], Kalivas *et al* at 21.6 GHz and 37.2 GHz [17],[18], Smulders and Wagemans at 42 GHz and 58 GHz [21]–[23], Bultitude *et al* at 40 GHz [20], and Talbi and Delisle at 37.2 GHz [24], to investigate how the received power and the propagation distance are related in an indoor setting at

the millimeter wave band. Each of these efforts have adopted the inverse n^{th} power model to describe the relationship. General results are as follows.

Corridor measurements in all cases resulted in values of distance exponent n less than the free space exponent of 2, with waveguiding effect of the hallway being asserted as the cause. Single-room measurements, in which both terminals were under line-of-sight condition in the same room, also yielded distance exponents lower than 2 (with the finding of Bultitude *et al* being an exception), but still higher than those for corridors. In this situation, waveguiding effects are thought to be somewhat lessened by the presence of in-room objects that might intercept reflections off the walls, floor and ceiling, or create new rays, thus causing a slight departure from a pure waveguide-like environment. When a large number of such specular reflections are precluded, n might climb up to above 2, as was found by Tharek and McGeehan in a furnished and carpeted room. Smulders and Wagemans particularly found that by employing omnidirectional antennas at both ends and positioning them such that their radiation patterns counterbalance the distance-dependent path loss, near-zero exponents can be found and thus single-room cells with practically uniform coverage can be achieved. The base antenna was installed on the ceiling at the center of the room so that users at closer distances experienced smaller path losses while at the same time encountering smaller antenna gain due to the 8-shaped vertical pattern of both the user and base antennas. The reverse applies for users at farther distances.

For measurements in a large room, Bultitude *et al* applied a power-versus-distance curve consisting of two segments separated by a break point at 25 meter distance. The shorter distance segment exhibited an exponent n of 1.5, whereas the other segment possessed an exponent of 4. When an experimental distribution function was constructed for spatial variations of the received power around the mean curve, they discovered that the standard deviation thereof was 4–7 dB. Talbi and Delisle concluded from their own results that the distance exponent increases with the increasing size of the room and volume of furnishings. They also found that the standard deviation of variations around the local mean becomes larger as the number of reflecting objects increases in the room under consideration.

As for experiments in which the transmitter and the receiver were positioned in separate rooms, values of n were found to be larger than 2. The large exponents are usually attributed to

additional attenuations by walls and partitions which, according to [18], is around 3.7 dB/wall at 21.6 GHz and 5.4 dB/wall at 37.2 GHz.

The above works have each determined the value of n in every individual indoor environment selected. The resultant value of n , therefore, applies only in the corresponding room or environment. In our experiment, a single value of n is to be derived from the ensemble of power-distance data pairs collected in various rooms with both transmit and receive terminals located under a line-of-sight condition in the same room. In this way, a common curve relating the average received power to the corresponding propagation distance applies for any room; the variations around it reflect the influence of undertaking the measurements in various rooms and at various positions in each room. Our experimental results will also be distinct in that a value of n is to be determined for each reception angle, thus permitting the study of the effect of variations in the propagation distance on the average received power when the antennas of both terminals are misaligned.

Channel Delay Spread

As has been reported in research papers to date, the millimeter wave broadband indoor radio channel rms delay spread may reach up to 98 ns [23], whereas the largest reported maximum delay spread at 15 dB threshold is approximately 400 ns for horn-omni configuration and 700 ns for omni-omni which are the 90th percentiles of the entire results in [20]. Other important results involving the channel delay spread measures are on the impact of room size, antenna type, or other parameters on the rms or maximum delay spread. Of particular interest in our research is the effects of different room sizes and reception angles on the maximum and rms delay spread distributions when the measurements are performed in a number of different rooms employing directional receive and transmit antennas.

The influence of the room size on the rms delay spread has received different views from researchers. Davies *et al* reported in [25] that the rms delay spread is larger for larger rooms. Kajiwara [26], through simulations and measurements at 94 GHz and adopting the square root of room area as a measure of the room size, arrived at a conclusion that the rms delay spread is in proportion to the room size and is less sensitive to the room size when directional antennas with a smaller beam are employed. Discrepancies to the former result occur when two rooms with

identical area but different diagonals are considered. Contrary to the above results, Smulders and Wagemans [23] discovered that the rms delay spread is dependent on the surrounding wall reflectivity rather than the room size. Bultitude *et al* [20] discovered that the rms spread, found to be uncorrelated with the propagation range, is in inverse proportion to the received power for which a faster downslope is found in larger buildings. In our work, an effort will be attempted to relate the rms and maximum delay spreads to the room size represented by the room length.

Bultitude *et al* [20] discovered that the 90th percentiles of the rms delay spread are practically the same for both the omni-omni and horn-omni configurations. The same observation holds for the maximum spread values when a 10 dB threshold is applied. However, when a transmit horn antenna was used and data were collected outside the horn main beam, values of rms and maximum delay spreads larger than those for omni-omni setting are observed. Also, when lower threshold is applied, the omni-omni configuration produces an increasing 90th percentile of the maximum delay spread. Our work seeks to learn the effect of variations in reception angle on the distributions of the maximum and rms delay spreads when two directional antennas are utilized. For such a narrowbeam receive antenna as the one used in our experiments, a nonzero reception angle thus means a scenario in which the transmit antenna is not illuminated by the receive antenna main beam.

3.1.3 The Effects of Directional Antennas on Channel Characteristics

This section is devoted to the overview of what researchers have found on the impacts of employing directional antennas on such mm-wave channel characteristics as fading, received power, and delay spread. In general, the use of directional antennas at either the receiver or the transmitter with perfect alignment is thought to produce fewer multipath components coming from directions outside the antenna beamwidths, as will be seen in the following review, which brings about some consequences on the measured envelope fades and delay spreads.

Kalivas *et al* [18] found that the use of horn antennas in a hallway link gives a greater probability of deep fades. According to Licht's finding in [12], a horn-horn configuration necessitates a larger fade margin than an omni-horn does. He further investigated the effects of misalignment between two directional antennas and found out that the antenna misalignment

reduces the fade depth when the fade is caused by the LOS path interception by moving people, but slightly increases the depth of fades induced by partial blocking of the radio link. He also found that when people movement occurs outside the region illuminated by the antennas, no fades are detected.

Smulders and Vervuurt [27] performed geometrical optics-based simulations to explore the characteristics of received power when three different combinations of two antennas of different types (directional and omnidirectional) are considered. They found that dir-dir (directional-to-directional) combination gives the best performance in terms of the amount of power delivered to the receiving end. However, this combination is more sensitive to antenna mispointing and the presence of obstacles than the other combinations are. Using a 20° -beamwidth transmit horn and a 30° -beamwidth receive horn, Licht [12] found that on average a 15° misalignment reduces the received power by 0.5 dB, but an average power reduction of 8.2 dB results when the antennas are 30° misaligned. In regard to the distance-dependent nature of the mean received power, Talbi and Delisle [24] found that the introduction of a directional antenna at the transmitter produces a distance exponent n closer to 2 and generally smaller variations around the mean power-distance curve than in the omni-omni case.

In general, results of simulations and field measurements are in agreement on that antennas with a higher horizontal directivity (or a narrower horizontal beam) yield smaller rms delay spreads, as has been reported by Smulders and Vervuurt [27], Manabe *et al* [28], and Kajiwara [26] for various frequencies in the millimeter-wave band. Smulders and Vervuurt found that this fine performance in delay spread degrades when an obstacle is present or some degree of antenna misalignment occurs. Their simulation results show that with 10° antennas at both ends of the link, a misalignment of merely 5° is enough to enlarge the worst-case rms delay spread from 1 ns to more than 40 ns. Guerin's empirical results [29] show that when the transmit and receive antennas are misaligned, the direct ray is attenuated, giving a way to some delayed, reflected rays to exhibit an increase in their power contribution. Bultitude *et al* [20] also empirically found that when a transmit horn is used, a larger 90th percentile of maximum spread is observed when the data are collected outside the horn beamwidth-illuminated region than when the data are collected inside the region.

In the work to be performed for this thesis, comparisons of channel characteristics when directional antennas are in use and when they are not, are not attempted. Instead, this research studies how an antenna misalignment in a directional-to-directional antenna setting affects the received power and the distributions of the maximum and rms delay spreads. In the same manner a study can be undertaken on the received power of interfering signals arriving at a base station directional antenna from any direction while the base antenna main beam is pointing at a specific, communicating user terminal, a situation likely to occur in the conduct of indoor wireless communications. Some additional measurements are also performed to test the presence of any temporal variation during the experiment using directional antennas.

3.1.4 Radio Wave Arrival Angle and Angle Diversity

Attempts have been made by researchers to study how the environment affects the angle of arrival of significant signal components arriving from the transmitter, especially at frequencies in the millimeter-wave band. Such attempts have been accomplished in the past for UHF indoor environments [30], [31], as well as for mm-wave outdoor links [32]. Manabe *et al* [33] have performed some measurements indoors utilizing a transmit horn antenna and a rotating highly directional receive antenna. Although angles of arrival of specular components are not specifically addressed, their measurement results show the following phenomena. The direct path component appears consistently at reception angles within the 50° range around the LOS direction with varying amplitude. Also, some high multipath spikes are detected consistently with angle-varying magnitudes over ranges of up to 40° wide around the directions from where large reflections are expected. Smaller spikes, however, appear only at certain angles over smaller ranges of reception angle. Guerin [29] found for an empty room that when a directional transmit antenna is in use and pointed toward a rotating directional receive antenna, the direct path component is predominant whereas other large rays come from wall surfaces that reside within the transmit antenna illuminated region. However, when the transmit antenna is pointed away from the receiver toward a wall surface, two almost-equally dominant components, the direct path component and the first order reflection off the pointed wall, are detected. When the same room is furnished, he found similar results except for directions at which original paths are blocked by new objects and those at which new objects create new reflections or diffractions. He also discovered that the power-

angle profile may change drastically when an object obstruct the direct ray.

The acquired knowledge of how multiple components may arrive at the receiver antenna from different directions, each being induced by propagation over the direct LOS path or by reflections off a variety of structures and objects, has encouraged researchers to embark upon investigation on the effectiveness of angle diversity technique employing multiple directional antennas facing different directions in an indoor environment. The underlying assumption is that since these incoming components are affected by different reflecting objects, their occurrences must be independent from one another. Hence, if several of these independent components can be captured by different antennas, a multiple-branch diversity can be achieved. The most recent and most relevant work on the angle diversity technique is Licht's [12], in which he explored the use of two 30° -beamwidth horns as two receiving branches with selection combining, each separated in pointing direction from the other by 30° , at 29 GHz. He found that low correlation coefficients, in the order of 0.1-0.2, are obtained between these two branches as long as the envelope fades originate only from the terminal movement or people movement in such a way that the LOS path is intact. However, when deep fades due to brief shadowing by moving people are considered, correlation coefficients as high as 0.5 are observed since the brief obstruction blocks the direct paths of the two branches causing fades of similar patterns on both of them. The knowledge of these correlation coefficients, unfortunately, can not be conveniently used in the calculation of outage probability since these coefficients measure only the similarity of power fluctuations on the branches, neglecting the difference in average received power between them. Licht's experiments also include the computation of the diversity gain, which he found to be as low as 1-2 dB. He asserted that this low gain is owing to the high correlation between branches when a brief shadowing induces deep fades and the large difference in mean received power between branches that may outweigh the fade depth especially when one of the receive antennas is aligned with the transmit antenna.

The experiments performed in our research are intended to find the average angular correlation function, computed from a large number of position-specific angle-dependent impulse response estimates, as a function of reception angle difference and excess delay. Our purpose is to study the correlation of received signal levels arriving from two different directions at any excess delay and to find the angle difference that yield low correlation for most of excess delays.

3.2 Thesis Problem Statement

Having cited and analyzed research results pertinent to the work done for this thesis, we elaborate herein the significance of results obtained from our experiments. The outcomes of our research may offer the answers to some problems which, after the efforts discussed above have been accomplished, remain unclear. The following discussion may reveal the significance of our results.

- a. Although some work has been done on the influence of directional antenna pointing direction on the statistics of the rms and maximum delay spreads, our research attempts to take a more detailed look at this issue, in particular at 29.5 GHz band, by using a larger number of reception angles, a total of eleven ranging from -30° to 30° with 6° steps. In addition, the expected outcomes are obtained from an ensemble of impulse response estimates collected in a variety of rooms, rather than in a single room, producing a more general result for indoor environments.
- b. Because of the differences in results of investigations on the relationship between the room size and the channel delay spread, measured by its maximum or rms value, this issue needs further examination by introducing the room length to represent its size, and normalizing the channel delay spread measures to this length. The explanation to this adoption of the room length to represent the room size is to be given later.
- c. A more detailed view is also deemed necessary on the effect of directional antenna misalignment on the received power, achieved by programming the discrete reception angles as mentioned above on the receive antenna. The result of this exercise is also useful in quantifying the power of an interference signal arriving at the directional receive antenna from a direction other than the direct LOS line. The information obtained from this discussion is important in predicting the interference level a specific user-base link may suffer from in an indoor system that employs directional antennas at the base station.
- d. Correlation of the channel impulse responses obtained for different pointing angles is also paid closer attention. Not only is the received signal amplitude correlation presented as a function of difference in pointing angle, but it is also computed for each excess delay.
- e. A general result for power-distance relationship is sought in our work by determining the

distance exponent n of the inverse power model using the entire result of measurements in all rooms selected for our experimental sites. A single curve representing the distance-dependent average received power hence applies for any room, with variations around it reflect the global (room to room) and local spatial variations. As such, a common link budget calculation can be shared by all rooms by incorporating the same value of n and the same power margin to account for the spatial variations. This procedure may then be applied for rooms of similar sizes.

To achieve the above-specified goals, experimental sessions are to be performed utilizing the 29.5-GHz broadband radio channel measurement system developed at Carleton University and the programmable adaptive array antenna designed by Laval University researchers. Steps taken during these sessions include:

1. Preparations, consisting of the measurement system calibration, the measurement sites determination, and the design of data gathering procedure and storage.
2. Data collection, consisting solely of field measurement activities in the determined rooms.
3. Data processing and analysis, consisting of the generation of impulse response estimates from the collected raw data and the execution of subsequent processes on workstations to produce the desired results.

Chapter 4

Measurement System

4.1 Introduction

Discussed in this chapter is a detailed description of the measurement system, encompassing the complete measurement setup, the procedures taken for calibration and measurement, and the locations chosen to provide measurement environment. The discussion is completed with a description of the measurement system limitations that may bring some effects on the subsequent analysis of the acquired data.

4.2 Measurement Apparatus

4.2.1 System Overview

The whole experimental setup was in essence an integration of two independently-developed systems. The main system was the broadband millimeter-wave radio channel measurement system developed at Carleton University [34], while the later-added part was the programmable phased array antenna designed by researchers from Laval University [35].

The broadband wireless measurement setup consisted of a transmitter, in which a 29.5 GHz carrier wave was biphas-modulated using a 200 Mbps PN sequence, and a receiver, which recorded the complex envelope of the received signal. The transmitter consisted of a pseudonoise waveform generator outputting periodic replicas of a 1023-chip M-sequence, a set of equipment generating 2.5 GHz IF and 27 GHz LO waves, a BPSK modulator operating at IF, an upconverter to 29.5 GHz RF, a power amplifier, and a 32°-beamwidth horn antenna. The BPSK modulator was implemented in reality by utilizing one of two input lines of an IQ modulator. The upper half of figure 4.1 describes the major constituents of the transmitter.

The receiver, as illustrated in the lower part of figure 4.1, incorporated the 8°-beamwidth 8-element programmable phased array antenna, integrated to the existing setup due mainly to its computer-controlled pointing angle programming capability. Following the antenna were an LNA, a downconverter to 2.5 GHz, an IQ demodulator and a sampling oscilloscope. Despite the

use of common sources for IF and RF references, no phase recovery was attempted at the receiver, leaving the channel phase information preserved in the recorded complex envelope. Two separate personal computers were employed to handle the data acquisition and the array antenna direction steering. These computers communicated with one another to synchronize the acquisition and angle adjustment processes. The acquired data were sampled at 1 GS/s, which with 200-Mbps 1023-bit PN sequence resulted in 5115 samples for a one-period acquisition.

The entire setup was divided into two parts constructed on two separate movable carts to allow facile transportation of the equipment and flexible arrangement of transmitter and receiver positions. Running between these carts were three cables carrying the 13.5 GHz reference for receiver down conversion, the received IF signal to be demodulated, and the inter-computer serial communication signals. Figure 4.2 illustrates the block-by-block details of the experimental system used in this work. Part (a) of the figure depicts the components making up the bigger blocks of the transmitter, while part (b) does so for the receiver. A detailed system description is given in subsequent sections, starting with the transmitter, receiver, array antenna, and data acquisition systems, followed by the calibration procedure. A list of important operating parameters of selected components, arranged in the same order as their appearance in later

Figure 4.1: Simplified block diagram of the experimental setup

Figure 4.2: Block diagram of the complete experimental setup

Figure 4.2: Block diagram of the complete experimental setup (Continued)

sections, are given in table 4.1.

Table 4.1: List of operating parameters for selected components

Name	Type	Code (see fig 4.2)	Parameters	
Pseudonoise Waveform Generator		a1	Output rate	200 Mbps NRZ
			Output level	“-1” -500 mV
				“+1” -1.075 V
Synthesizer	HP8656B	a2	Output frequency	200 MHz
			Output level	0 dBm
Lowpass Filter	SBLP-300	a4, h2, h3	3-dB bandwidth	300 MHz
Synthesizer	Q04101	b1	Output frequency	1.25 GHz
			Output level	4.2 dBm
Amplifier	ZFL2000	b2	Gain	20 dB
			Operating band	10–2000 MHz
			1-dB compression	16 dBm
			Noise figure	7 dB
Frequency Doubler	PMCXKAQ3	b3	Input frequency	1.25 GHz
			Output frequency	2.5 GHz
			Output level	1.1 dBm
Bandpass Filter	ABP2500- 400	b4, c3, g2	Centre frequency	2.5 GHz
			3-dB bandwidth	400 MHz
Amplifier	AMC180-10	b5	Gain	10 dB
			Operating frequency	2.5 GHz
IQ Modulator	VMM20B- 2500	c1	LO frequency	2.5 GHz
Amplifier	ZHL1042J	c4	Gain	25 dB
			Operating band	10–4200 MHz
			1-dB compression	20 dBm
			Noise figure	4.5 dB

Table 4.1: List of operating parameters for selected components (Continued)

Name	Type	Code (see fig 4.2)	Parameters			
Synthesizer	HP8673D	d1	Output frequency	13.5 GHz		
			Output level	6 dBm		
Amplifier	HP8349B	d3	Gain	15 dB		
			Operating band	2–20 GHz		
			Noise figure	13 dB max.		
Frequency Doubler	HP83554A	d4	Input frequency	13.5 GHz		
			Output frequency	27.5 GHz		
Upconverter	47471H- 2130	e1	LO frequency	27 GHz		
			IF frequency	2.5 GHz		
			RF frequency	29.5 GHz (USB)		
			IF and RF bandwidth	4.4 GHz (15% of RF)		
			RF output power	4 dBm min.		
Amplifier	AMF 6B275300	e2	Gain	23 dB		
			Operating band	27.5–30 GHz		
			1-dB compression	19 dBm min.		
Horn Antenna	NARDA 48127		Gain	15.3 dBi		
			Polarization	Horizontal		
			3-dB Beamwidth	Horizontal	32°	
				Vertical	30°	
			Height	1.9 m		
Amplifier	AFD120180	d5, d7	Gain	20 dB		
			Operating band	12–18 GHz		
			1-dB compression	20 dBm min.		
Cable	Suhner Sucoflex 104	d6	Overall insertion loss (for total length of 18 m)	@12 GHz	18.7 dB	
				@15 GHz	21.1 dB	

Table 4.1: List of operating parameters for selected components (Continued)

Name	Type	Code (see fig 4.2)	Parameters	
Frequency Doubler	K2000N	d8	Input frequency	13.5 GHz
			Output frequency	27 GHz
			Conversion loss	10 dB
Low Noise Amplifier	JS326003000	f1	Gain	24 dB
			1-dB compression	5 dBm
Down- converter	47411H	g1	LO frequency	27 GHz
			RF frequency	29.5 GHz
			IF frequency	2.5 GHz (LSB)
			Conversion loss	5.5 dB max.
			Noise figure	8 dB max.
Amplifier	AFT4233	g4	Gain	37 dB
			Operating frequency	2.5 GHz
			1-dB compression	-19 dBm
AGC Amplifier	HP8347	g7	Output level	0 dBm
			Operating band	100 kHz–3 GHz
			Input level	-14 dBm max.
			Noise figure	15 dB
IQ Demodulator	IQM20B- 2500	h1	LO frequency	2.5 GHz
			Bandwidth	250 MHz (10% of f_0)
			Conversion loss	12 dB max.
			RF input power level	0 dBm nom.
Amplifier	ZFL500LN	h6, h7	Gain	24 dB
			Operating band	0.1–500 MHz
			1-dB compression	5 dBm
			Noise figure	2.9 dB

Table 4.1: List of operating parameters for selected components (Continued)

Name	Type	Code (see fig 4.2)	Parameters
Phased-Array Antenna			Gain (incl. phase shifter circuits, connectors and combiners) 11.25 dB
			Polarization Horizontal
			3-dB Beamwidth Horizontal 8°
			Vertical 15°
			Overall sweep range -45° to 45°
			Useful range (in the absence of grating lobe) -15° to 15°
			Height 1.9 m

4.2.2 Transmitter

4.2.2.1 Baseband Stage

The sole task of the baseband stage was to generate an analog waveform of the pseudonoise sequence which would become the envelope of the transmitted signal. The role was assumed primarily by a pseudonoise waveform generator (PWG) that generated periodic replicas of a 1023-bit maximal-length sequence at 200 Mbps. The sequence was created by invoking a Matlab program on a PC and loaded into the PWG just before any measurement session was embarked upon. The 200 MHz clock was supplied by an HP synthesizer programmed to suit the purpose. After passing through a DC block and an LPF, the output of the PWG was ready to modulate the IF carrier in the modulator. The following are detailed specifications of the baseband stage components.

Pseudonoise Waveform Generator (a1)

The PWG generated the pseudonoise waveform used to biphase-modulate the carrier wave. It contained a 2^{17} bit counter and a $2^{17} \times 16$ bit RAM. A 1023-chip maximal-length sequence was manually loaded into it from PC#1's parallel port prior to activation of modulation

parts. Since the sequence was only $2^{10}-1$ bits long, a train of 128 sequences of the same form plus 128 all-zero filler bits were programmed to fill in the available memory. The analog output was derived from 10-bit samples of the loaded digital data, with a voltage level of -500 mV representing bit “+1” of the output PN sequence, and -1.075 V representing bit “-1”. The M sequence had been generated in advance by a Matlab program.

Besides the 200-Mbps pseudonoise-modulated NRZ rectangular pulses, the PWG also provided two triggering signals needed to start the display of acquired one-period sequence on the oscilloscope screen and the sampling process of the received signals. Both were digital signals of levels 0 and +5 V.

200 MHz Synthesizer HP8656B (a2)

This 0.1–990 MHz HP Signal Generator provided a 0-dBm 200-Mhz clock input for the pseudonoise sequence generation in the pseudonoise waveform generator (PWG). It took its 10 MHz external reference source from synthesizer Q04101 (block b1). Personal computer #1 took control of the initialization of the synthesizer via an HPIB bus network.

DC Block SMCBLK721N (a3)

Since the PWG output waveform contained a DC component that the IQ modulator could not accept, this DC block was incorporated into the baseband stage.

Lowpass Filter SBLP-300 (a4)

The LPF had a 3-dB bandwidth of *300 MHz*.¹ For a 200-Mbps rectangular pulses, this quantity was equivalent to the spectral width of the main lobe and one half of the first side lobe.

4.2.2.2 IF Carrierwave Generation

This block generated the IF carrier shared by both the transmitter and the receiver. The main signal source was indeed the oscilloscope internal 100 MHz clock source, which was used by the Q04101 synthesizer to produce a 1.25 GHz reference. This reference was in turn doubled to yield the desired 2.5 GHz IF carrier wave. After bandpass filtering and amplification, the IF

1. Quantities printed in italic are used consistently throughout this chapter to present component data taken from the manufacturer's data sheets. These do not include programmable parameters, such as output signal frequency and level of a synthesizer, or component data obtained from measurement.

carrier wave was split into two branches, one heading for the modulator while the other for the demodulator.

1.25 GHz Synthesizer Q04101 (b1)

This synthesizer was a phase-locked oscillator (PLO) synthesizer that can be programmed by adjusting an attached DIP switch to produce a certain output frequency along with a reference output. By feeding on a 100 MHz square wave available from an internal clock source of the oscilloscope (see later description of the oscilloscope), this synthesizer supplied two references for the measurement system: a 10 MHz signal from its reference output port for the 100-MHz clock generation and a 1.25 GHz, later doubled to 2.5 GHz, from its synthesizer output line. The 10 MHz signal went into the 100-MHz synthesizer (block a1), whereas the *4.2-dBm* 1.25-GHz wave entered the frequency doubler (b3) after some amplification (b2).

20 dB Amplifier ZFL2000 (b2)

Prior to frequency doubling, the 1.25 GHz was *20-dB* amplified to provide a 17.5 dBm input to the frequency doubler.

Frequency Doubler PMCXKAQ3 (b3)

The frequency doubler raised the signal frequency to 2.5 GHz and output it at 1.1 dBm power level.

Bandpass Filter ABP2500-400 (b4)

Harmonic frequencies that might have appeared during the preceding amplification and frequency doubling processes were stopped by this BPF which has a bandwidth of *400 MHz* and a centre frequency of *2.5 GHz*.

10 dB Amplifier AMC180-10 (b5)

With a *-0.9 dBm* input and *10 dB* gain, the amplifier gave a *9.1-dBm* 2.5-GHz output that would later be shared by a pair of IQ modulator and demodulator, assuring exactly the same frequency being used for modulation and demodulation purposes.

Power Splitter ZFSC2-11 (b6)

This is a 3-dB power splitter that distributed the 2.5 GHz reference to the IQ modulator

and demodulator. The splitter introduced an insertion loss of 1 dB (beyond the 3-dB division loss) and an isolation of 20 dB .

4.2.2.3 IF Stage

The IF stage started with the IQ modulator which was used to generate the BPSK signal at 2.5 GHz IF. It was followed by a BPF and an amplifier each aimed to bandlimit the modulated signal and to amplify its power so as to drive the power amplifier in the RF stage.

IQ Modulator VMM20B-2500 (c1)

The modulator was used to spectrally shift up the baseband pseudonoise signal to IF band. Since BPSK was desired for modulation scheme, only one of the modulator two input lines was used, while the unused was simply terminated by a $50\text{-}\Omega$ dummy load. The used line obtained as its input the filtered pseudonoise waveform from the PWG. The modulator fed on a 4-dBm 2.5-GHz LO signal input from the power splitter (b6).

10 dB Attenuator (c2)

This attenuator provided a matching pad between the IQ modulator and subsequent elements. It had an operating range of *DC to 4.0 GHz*.

Bandpass Filter ABP2500-400 (c3)

The BPF had a 3-dB bandwidth of *400 MHz* and a *2.5-GHz* center frequency.

25 dB Amplifier ZHL1042J (c4)

The medium power amplifier raised up the power level of modulated IF signal to fit in the input range of the upconverter. With an input at -12 dBm and a gain of 25 dB , it fed the upconverter with an approximate power of 14 dBm . Its operating frequency spanned from *10 MHz through 4.2 GHz*.

4.2.2.4 LO Carrierwave Generation

Similar to the IF generation, the LO carrier generated by this stage was shared by both the transmitter and the receiver for up- and downconversion to and from the 27 GHz RF. The central

part of this stage was a synthesizer programmed to produce a 13.5 GHz reference signal. This signal was split into two and sent to both the upconverter and the downconverter. The frequency doubling that produced the actual LO of 27 GHz was performed just before the signal entered the up- or downconverter.

13.5 GHz Synthesizer HP8673D (d1)

This is a 0.05–26.0 GHz HP Synthesized Signal Generator, the initialization of which was controlled by PC#1 through GPIB connection to output a 6-dBm wave at 13.5 GHz, from which the 27 GHz LO signal was derived.

Power Splitter HP11667A (d2)

The 13.5 GHz signal was split into two for the transmitter's upconversion and receiver's downconversion purposes. The 6-dB splitter had an insertion loss of 1.8 dB.

15 dB Amplifier HP8349B (d3)

The 2.0–20.0 GHz HP Microwave Amplifier raised the split 13.5 GHz signal power by 15 dB. The minimum output power for a 5-dBm input was 19 dBm leveled across the range of 2.0–18.6 GHz. Its noise performance was described by a less than 13 dB noise figure.

Frequency Doubler HP83554A (d4)

The 26.5–40 GHz HP Source Module served as a frequency doubler for the 13.5 GHz wave. It, together with the HP8673D signal generator and HP8349B microwave amplifier, constituted the millimeter-wave source unit that multiplied the source frequency to generate mm-wave frequencies. The input frequencies could be selected from the range provided by the signal generator, 13.25–20 GHz in our case. The minimum and maximum input power levels were +17 dBm and +27 dBm, respectively. The maximum output power was 8 dBm inside the spectral region of 26.5 through 37.2 GHz.

4.2.2.5 RF Stage

An upconverter that shifted the modulated IF signal up in frequency domain to RF began this stage. A power amplifier followed to boost the signal power. Before radiation by the antenna, a directional coupler was inserted, the second branch of which was connected to a power meter.

Upconverter 47471H-2130 (e1)

Upconversion from 2.5 GHz IF to 29.5 GHz RF took place inside this component. The converter incorporated a built-in bandpass filter at its mixer output, passing only the upper-sideband sum signal and eliminating the need for an external BPF.

It possessed a typical *10 dBm* 1-dB output compression point for a 17 dBm LO input and accepted a minimum LO level of *13 dBm*. The IF input power was typically 17 dBm with a spectral width identical to that of the RF. The LO must differ in frequency from the RF by at least 2.0 GHz (2.5 GHz was taken for IF in our case). A *30-dB* isolation was imposed on the LO and RF ports by the converter. Approximations for conversion loss, conversion gain and noise figure were respectively *5 dB*, *33 dB* and *5 dB*.

23 dB Amplifier AMF 6B275300 (e2)

The amplifier operated in the *27.5-30 GHz* range with a gain of *23 dB* and 1-dB gain compression output of *19 dBm*.

Isolator HPR365A (e3)

This isolator was used to mitigate the impedance mismatch between two subsequent blocks and to isolate its input line from reflections on the output line. Minimum isolation ratio provided by this component was *25 dB*, while its maximum insertion loss was *1.6 dB*.

Directional Coupler HPR752D (e4)

This coupler supplied a *20-dB* reduced version of the actual transmitted RF signal before the transmit antenna. The power-reduced signal went to a power meter used for monitoring purpose.

Power Meter HP435B and Power Sensor HPR8486A (e5)

The power meter gave a reading 20 dB below the actual transmitted power. Its function was mainly pertinent to monitoring purposes, such as to check if everything worked properly and to know if the whole setup was sufficiently warmed up. The power meter was set to the -10 to 0 dBm power range and a 92% calibration factor in accordance with the sensor data list. With the actual transmit power of about 18 dBm, the after-warmup normal reading was approximately -2

dBm.

4.2.2.6 Transmit Antenna

For the transmit antenna, a 32°-beamwidth 15.3 dBi-gain NARDA 48127 horn was employed. Its E- and H-plane radiation patterns had previously been measured and made use of for a recent master's thesis [12] on channel measurements at the frequency band identical to that presently under investigation. The patterns are reprinted in Appendix D of this thesis. The antenna was mounted on the top of the cart carrying the transmitter at a height of 1.9 m and installed in such a manner that it emitted a horizontally polarized radiowave.

4.2.3 Receiver

4.2.3.1 LO Carrierwave Generation

The LO signal was generated by the same circuitry as that generating the transmitter's LO signal. The difference, however, was that the signal must travel along an 18 meter cable from the transmitter's cart to the receiver's, hence necessitating the insertion of several amplifiers before the signal entered the frequency doubler at the other end of the cable.

20 dB Amplifier AFD120180 (d5)

One branch of the 13.5 GHz power splitter went to this amplifier, which took a -1.7 dBm input and applied a *20 dB* gain before sending it over the long cable to the downconverter located on the other cart. This amplifier was a medium power amplifier operating in spectral range of *12 to 18 GHz* with 1-dB gain compression output of *20 dBm*.

18 m Cable Suhner Sucoflex 104 (d6)

The amplified 13.5-GHz signal was delivered over this cable to the receiver's downconverter. According to the available manufacturer's data sheet, the overall insertion loss introduced by the 18 m cable was *21.1 dB* at 15 GHz and *18.7 dB* at 12 GHz. This, by interpolation, led to an insertion loss estimate of 19.9 dB for 13.5 GHz frequency.

20 dB Amplifier AFD120180 (d7)

This amplifier applied a 20 dB amplification to the cable-attenuated 13.5 GHz reference signal. It was a medium power amplifier of the same type as that used in front of the long cable.

Frequency Doubler K2000N (d8)

The signal was doubled in frequency to yield the actual LO frequency of 27 GHz . The doubler introduced a conversion loss of 10 dB .

Isolator 45140H2100 (d9)

Like other isolators employed in the measurement system, this isolator assumed its role in mitigating any mismatch and increasing the input line resistance to any reflection occurring on its output.

4.2.3.2 RF Stage

The RF stage consisted of low-noise amplification applied to the output of the receive array antenna. The amplification role was assumed by a single LNA. Its exact noise performance specification was unfortunately unavailable when this thesis was being written.

24 dB Low Noise Amplifiers JS326003000#1 (f1)

The amplifier had a 24 dB gain and a 1-dB compression point output of 5 dBm . Hence in order for this amplifier to work properly in its linear range, its input must be guaranteed to never exceed -20 dBm ($\approx 5\text{ dBm} - 24\text{ dB}$). As a result, the maximum allowable output from the receive antenna and its combining network is -20 dBm .

4.2.3.3 IF Stage

Frequency downconversion from RF to IF was performed in this stage. The received modulated IF was next prepared for transmission over a 10 meter cable back to the other cart (the same cart that conveyed the transmitter), where the demodulation took place. Following the cable was an amplifier with an automatic gain control having a dual purpose of supplying a fixed-power input to the demodulator and producing impulse response estimates of equal energy.

Downconverter 47411H (g1)

The low-noise amplified off-air RF and the frequency-doubled cable-carried LO signals were next fed into the downconverter that employed no bandpass filtering after the mixing process, thus providing a double-sideband output and necessitating the exertion of an auxiliary post-conversion BPF. Properties of the converter included an 8-dB maximum noise figure, a 20-dB minimum LO-to-RF isolation and a 20-dB minimum RF-to-IF gain.

Bandpass Filter ABP2500-400 (g2)

The BPF had a bandwidth of 400 MHz centered at 2.5 GHz . It was useful to pass only the desired difference signal from the downconverter and stop any other spectral component, because the downconverter did not incorporate any post-mixing filtering.

10 dB Attenuator (g3)

The 10-dB DC-to- 12.4 GHz range attenuator was used as a matching pad between the downconverter and the next amplifier.

37 dB Amplifier AFT4233 (g4)

The use of this power amplifier was compulsory since the relatively low power IF signal resulting from the downconversion process had to travel over a 10 meter cable back to the other cart for demodulation and acquisition. A 37-dB power amplification was applied to the input signal, which was required to have a maximum level of -19 dBm corresponding to the amplifier 1-dB compression point at 2.5 GHz .

10 m Cable (g5)

Due to relatively low input power and relatively high attenuation per unit of length introduced by this cable, its length was limited to only 10 meter to maintain a detectable signal at its output. This limitation would in turn limit the transmitter-receiver spacing to 10 meters at most, precluding any measurement performed for distances beyond 10 meters.

Power Splitter HP11667C (g6)

The splitter introduced an insertion loss of 6.0 dB .

Automatic Gain Control HP8347 (g7)

The 100 kHz - 3.0 GHz RF Amplifier HP8347 was set to operate with its automatic leveling

control (ALC) being activated, and thus herein was always referred to as automatic gain control (AGC). By doing so, the output power was leveled across the entire operating frequency span of this amplifier at a certain adjustable value, which in the case of our system was 0 dBm. The ALC activation was required due in part to the necessity to have all the times a 0-dBm nominal input to the IQ demodulator. Another reason to leave the ALC on was to guarantee the recording of received sequences by the oscilloscope that yielded identical-energy impulse response estimates. With the automatic leveling operation running, the maximum allowable input power was -14 dBm. The amplifier had a 22 -dBm 1-dB compression point and a 15 -dB noise figure.

4.2.3.4 Baseband Stage

The output of the AGC went into the IQ demodulator, which output two waveforms, namely the I and Q sequence waveforms. These waveforms went through the same lowpass filtering and amplification before arriving at the acquisition stage inside the oscilloscope.

IQ Demodulator IQM20B-2500 (h1)

The demodulator fed on the AGC 0-dBm output and produced two baseband outputs, the I and Q sequences. It also made use of a 3.3-dBm LO signal coming from 2.5 GHz reference generating blocks, identical to that used by the IQ modulator. The required RF power level was 0 dBm, achieved by the use of the 0-dBm leveled-output AGC. A bandwidth of 250 MHz (10% of LO frequency) was provided and a maximum of 12 dB RF-to- I or Q conversion loss was introduced by the demodulator.

Lowpass Filters SBLP-300 (h2 and h3)

These LPF's had a 3-dB bandwidth of 300 MHz.

12 dB Attenuators (h4 and h5)

Each of these attenuators was a cascade of two 6-dB DC-4.0 GHz attenuators, intended to provide a matching pad between the demodulator and the next amplifier and to place the signal power inside the linear operating range of the amplifier.

24 dB Amplifiers ZFL500LN (h6 and h7)

Each of these amplifiers had a 24-dB gain, 0.1-to-500 MHz operating spectrum, a 5 dBm

maximum output power and a 2.9-dB noise figure.

4.2.4 Adaptive Array Antenna

The technical descriptions hereafter are following those given in the antenna manual [35] provided by the designer group from Laval University.

4.2.4.1 Antenna Components

Microstrip Antenna

The microstrip antenna was designed to operate at 29 GHz. Eight four-patch columns, each column being programmable in its amplitude and phase, constituted the antenna structure. It was mounted vertically on a pole attached to one of the measurement carts, thus creating a steerable horizontal beam through programming of phase shifters associated to individual columns. The antenna emitted a horizontally polarized radiowave.

Because of the existence of 0.97λ spacings between adjacent columns necessary for feed networks, where λ is the wavelength at 29 GHz, a grating lobe as far as 60° away from the main appeared in the antenna pattern for any absolute angle greater than 15° . For pointing angles in this range, the grating lobe could be as large as only 6 dB below the main lobe level. This issue brought about the need for special cares in dealing with data obtained for pointing angles with a grating lobe.

Phase Shifter

The phase shifters were controlled by the control module through varying the driving current for each shifter. Operating in 28–30 GHz band, each of them exhibited a 10- μ s response time. Phase control accuracy was somewhat lost due to the occurrence of hysteresis cycle in the phase shifter characteristic curve, leading to a maximum phase shift error of 10° .

Combiner

The element combined the received signals from all eight antenna columns. Coaxial cables were inserted before the combiner to compensate for different phase shifts due to different length among individual feed networks.

Control Module

The control module housed eight current sources for the phase shifters, each capable of supplying current in ± 500 mA span, digitally controlled by means of DAC888 8-bit converters via a DB25 connection. A full-scale step input yielded 63% full response after 9 μ s.

Personal Computer #2

The PC#2 incorporated a digital 24-bit I/O interface card to send control signals to the control module. The original C-written program for antenna control designed by the Laval team was modified to allow remote operation, with PC#1 on the other cart being the main controller and communicating with PC#2 over a serial communication cable.

Antenna Overall Characteristics

The antenna possessed a main beamwidth of 8° and an overall gain, inclusive of losses induced by the microstrip, phase shifters, combiners and connectors, of 11.25 dB. The antenna was positioned at a height of 1.88 m and able to sweep from -45° through $+45^\circ$. However, a large grating lobe occurred for main lobes pointing at absolute angles greater than 15° . Moreover, phase controller inaccuracy of $\pm 10^\circ$, making up a $\pm 1^\circ$ phase precision, should also be taken into account in analyses.

4.2.5 Data Acquisition and Storage

Power Meter HP437B and Power Sensor HP8482A

The power meter operated in a range of 100 kHz to 50 GHz with a 50-dB dynamic range. During its initialization, PC#1 remotely programmed it through the GPIB network to acquire the detected power in dB with 0.01 dB resolution, to apply a calibration factor of 97.5% as required by the sensor used, and to produce a 100 ms average reading. The averaging time was the shortest that could be programmed on the power meter, giving a minor limitation to the acquisition speed.

Oscilloscope HP54542A

The oscilloscope assumed two important roles; it displayed the transmitted and received sequences in real time, thus facilitating simple conduct of the ongoing acquisition, and performed

acquisition processes that involved sampling and data transfer to PC#1. It had a 500 MHz lowpass bandwidth and a programmable sampling rate of up to 2 GS/s. For the acquisition purpose, it was initialized to sample the incoming wave at 1 GS/s. Since the PN sequence was 1023 bits long and sent at 200 Mbps, it required 5115 samples of 1 ns spacing to collect one sequence period of data, with 5 samples in one bit duration.

A single-shot mode was programmed such that the real-time display and the acquisition ran in synchronism with the PN sequence waveform generation. This was done by employing triggering signals from the PWG for external trigger sources for the oscilloscope. The 16-bit/sample binary coded raw data were transferred to PC#1 over the GPIB bus for further storage and processing.

Since a synchronized work was desired for the PN sequence generation and the acquisition, an additional output port was made on the oscilloscope such that its internal 10 MHz source could be used to generate clocks for the PWG and for the IF generator block.

Personal Computer #1

PC#1 assumed the primary control over the entire process of measurement. It was connected via an GPIB network to the synthesizers (a1 and d1), the oscilloscope and the power meter #2. These buses carried control signals from and measurement data back to PC#1. The initialization of those apparatuses connected to GPIB network was invoked by a C program executed before the entire measurement started. Subroutines for control and data transfer were also available and called from time to time by the main acquisition program, also written in C language.

PC#1 as well communicated with PC#2 through a serial communication connection when a new pointing angle was required for the next acquisition. A protocol had been constructed that PC#1 would not start the acquisition steps until it received an acknowledgement signal from PC#2, assuring that no angle errors aside from those inherently induced were made.

Before the whole measurement process was embarked on, the Matlab-generated PN sequence was transferred to the PWG from PC#1 using its parallel printer port. After one sweep of acquisitions, PC#1 received the raw I and Q data for eleven angles from the oscilloscope and stored them in an optical storage to be described later. These data were stored in form of MAT

files facilitating easy processing with Matlab later on.

Acquisition Trigger Control Circuit

This circuit was the last part added to the system, controlled by PC#1 through an unused bit on the parallel port connecting PC#1 to the PWG. It was inserted between the PWG trigger lines and the oscilloscope external trigger input port to do its sole function, which was to stop the generation of triggering signals for acquisition while an acknowledgement was not yet received from PC#2 signaling that it had accomplished pointing angle programming as had been requested by PC#1.

Magneto Optical Disk Drive Subsystem Sony SMO-S501

The optical disk drive was used as permanent storage for raw data received from the oscilloscope. The double-sided diskette used with this equipment had a 300 MB capacity on each side.

4.2.6 Calibration

4.2.6.1 Calibration Setup

The setup for the calibration process barely differed from that for the actual measurements, the difference being in the back-to-back connection between the transmitter and receiver, thus removing the effects of antennas and radio link from the system, and in the insertion of variable attenuators at the transmitter. The configuration is described graphically by figure 4.3. The diagram is not completely shown, leaving only blocks that clarify its difference from that of the system for the actual measurement. The two new attenuator components have not been introduced in previous sections and hence will be explained in the following.

Programmable Step Attenuators HP33324 and HP33326 (i1)

These attenuators were cascaded and controlled through the use of the HP11713A Attenuator/Switch Driver. The HP33324 covered 0 to 11 dB range of attenuation with 1-dB steps, whereas the HP33326 provided a larger scale attenuation in the range of 0 to 90 dB with 10-dB steps. Therefore, the overall operation range was 0 to 101 dB with a 1-dB resolution. The

attenuators' operating frequency ranged from *DC to 40 GHz*. Their main function was to emulate the channel attenuation due to radio propagation loss in the absence of antennas. Throughout the entire calibration process, their value was constantly set to 32 dB.

Variable Attenuator HPR382A (i2)

This attenuator was aimed to allow flexible adjustment of the transmit power. It provided a 0–50 dB attenuation range and a *1 W* power capacity for frequencies between 26.5 and 40 GHz. Equipped with a rotary switch for its attenuation level adjustment, its resolution varied from 0.05 dB for 0–1 dB range, 0.1 dB for 1–10 dB, 0.2 dB for 10–20 dB, 0.5 dB for 20–40 dB, up to 1 dB for 40–50 dB attenuation range. However, for calibration purpose, a 0.1 dB resolution was assumed over 0–10 dB attenuation range, the others being the same as have been detailed above.

4.2.6.2 Calibration Curve

Figure 4.3: Block diagram of calibration setup.

The calibration was undertaken first to identify the linear operating range, upper-bounded by the compression point and lower-bounded by the noise floor, of the whole structure and thus to determine the appropriate transmit power taking into account the amount of channel fading the signal might encounter. Secondly, it was imperative to know the exact relation of the power meter reading and the actual received power level at, say, the output of receive antenna. From this relation, a power curve could be drawn and made use of to convert the acquired power values to actual absolute values or to values relative to a known reference.

From the calibration steps, it was found that the compression point was arrived at when the step attenuator was set to 32 dB and variable attenuator was kept at 0 dB. Keeping the step attenuation constant at 32 dB, the noise floor was approached when the variable attenuation shifted close to 25 dB. During the latter step, readings were taken from the display of power meter #1, with dB steps taking on values as described in previous section, resulting in a calibration power curve featured in Appendix C. It was revealed from this exercise that the 0 through -25 dB relative power range detected at the output of receive antenna was equivalent to the -5.43 through -29.9 dBm range of power meter reading.

Taking into account the known value of transmit power, which was 16 dBm, the compression point translated to -16 dBm ($= 16 \text{ dBm} - 32 \text{ dB}$) and noise floor to -41 dBm ($= 16 \text{ dBm} - 32 \text{ dB} - 25 \text{ dB}$). In subsequent chapters, however, the relative power range of 0 to -25 dB is used in analyses pertinent with power measurements. It is also necessary to point out from the above results that the dynamic range inherent in the measurement system was 25 dB. This dynamic range was expectedly able to compensate for power variations due to changes in measurement location and propagation distance.

4.3 Experimental Procedure

In every measurement session, the following steps were executed.

- a. The Matlab-generated maximum-length pseudonoise sequence was loaded into the PWG, which in turn produced analog periodic replicas of the sequence at 200 Mbps.
- b. After connecting the cables used for synchronization purposes between transmitter and receiver, both parts were turned on and left to warm up for about an hour. From this step on,

the 200-Mbps sequence readily biphase-modulated the 2.5 GHz IF carrier wave, which after upconversion was transmitted at 29.5 GHz. The phased array interfaced the radio wave to the receiver electronics, in which the RF signal was downconverted back to IF 2.5 GHz. Demodulation was performed at IF to yield two output waveforms, namely the I and Q signals, that constitute the complex envelope. These two signals were subsequently displayed on the oscilloscope screen, but their recording was not effected until a computer program that handled the sampling and recording processes was invoked (see step d).

- c. During this warmup time, location measurement and tabulation were carried out manually. Necessary adjustments of the transmitter and receiver positions were performed as well. These adjustments were intended to assure that the 0° angle of the receive array fell exactly on the line-of-sight path, making use of the fact that the 0° pointing angle should give the highest average received power among all reception angles.
- d. After warmup time, the acquisition program was executed from PC #1. This program asked the oscilloscope to sample the incoming I and Q signals for one sequence period, and to transfer these samples to PC #1 which in turn stored them on the optical disk in a Matlab's MAT-file format. It also sent a command to PC #2 that controlled the array antenna to change the pointing angle to that next in order. In this manner, a single MAT file stored acquisition data for one discrete-step sweep of eleven reception angles ranging from -30° through $+30^\circ$ with respect to the LOS path. The phrase "a single sweep of measurements" is used throughout this thesis to consistently denote the raw data or the results of their processing that belong to a single MAT file. It took approximately 3 seconds for the setup to achieve a single sweep measurement of complex envelopes, and another 5 seconds for power acquisitions. The acquisition flowchart is depicted in figure 4.4.
- e. For every arrangement of transmitter and receiver positions, five MAT files were collected, corresponding to five transmitter's positions 1 cm apart from each other. It was the transmitter, not the receiver, that was shifted because the transmit antenna had a wider main beam such that if some slight changes were detected in the received signals after the shift, it could be assumed safely that they occurred due to small changes in the multipath structure rather than changes in antenna direction. This step resulted in data reflecting the minor variations of

multipath structures in a local area influenced by the same reflecting and scattering objects. Also, the step permitted a computation of spatial power delay profile for each arrangement, and hence, allowed the calculation of channel rms delay spread.

The above steps were repeated for every combination of transmitter and receiver positions, with steps (a) and (b) being exceptions when it was possible to move the equipment without turning them off.

Figure 4.4: Data acquisition flowchart

4.4 Measurement Locations

As many as four rooms in several buildings within the campus of Carleton University were chosen for measurement locations. These rooms were selected to represent several types of environment, including laboratories, classrooms, computer rooms, and lobbies, in which a broadband wireless communication may take place. The following are the selected rooms and the environment type they represent.

- a. Minto 4050 : a laboratory of $9.6 \times 9.2 \times 2.6 \text{ m}^3$ dimension full of metal equipment arranged on metal shelves, wooden benches, and metal and wooden cabinets, with two sides covered by aluminium-framed glass windows. Six sets of 5 files of different Tx-Rx arrangements were collected in this room, with every file storing eleven angle-dependent acquisitions for one out of five transmitter micropositions, and every five-file group containing data from a specific Tx-Rx arrangement in a given location, Minto 4050 in this case.
- b. Minto 4038 : a $14.9 \times 9.2 \times 2.6 \text{ m}^3$ computer room bordered by glass windows on one side and filled with an array of plastic tables on which workstations are placed. As many as twelve five-file data sets were collected.
- c. Minto 4031 : one of the hallways on the 4th floor of Minto building, which measures $32.0 \times 2.5 \times 2.6 \text{ m}^3$. Here, ten data sets were recorded.
- d. Architecture 507 : one of two drafting rooms, $13.6 \times 13.3 \times 3.0 \text{ m}^3$ in dimension, with high wooden benches arranged in rows as in usual classrooms, and one wall covered by glass windows. Twelve sets of data were collected here.

Figure 4.5 sketches the layout of room 4038 of Minto building described above that typifies measurement locations chosen for this experiment. Points at which the transmitter and the receiver were positioned during the experiments are denoted respectively by small rectangles and circles, whereas the measured radio links in the room are given below the diagram. Floor plans of the other buildings are included in Appendix A. Based on the descriptions of measurement procedure and locations, a data collection structure results, as is sketched in figure 4.6. A total of 40 sets of 5 files were collected from measurement activities in the above-described four locations.

It is also important to note that a larger number of measurements, as can be seen from figure 4.5, were taken with the transmitter-receiver direct line parallel or almost parallel to the room length. The same condition also occurred for experiments in the other rooms.

4.5 Measurement System Limitations

There were several promising benefits of employing the array antenna in the measurement system. First, unlike in a multiple antenna system, acquisitions of received powers and impulse response estimates could be performed for various arbitrary reception angles at practically the same time with an antenna and a receiver. Data acquisitions for multiple angles could be achieved subsequently through antenna array programming and data recording.

Secondly, in cases where multiple-angle acquisitions for angular diversity measurement purposes were achieved by multiple antennas, a certain measure of space diversity was inevitable

Figure 4.5: Room 4038 of Minto Building

due to the physical spacings between antennas. The use of a single array antenna eliminated this problem such that information could be gathered on the characteristics of a system employing a “pure” angle-of-arrival diversity. Lastly, faster angular adjustment and, to some amount, higher pointing accuracy were the apparent advantage of the use of an electrically steered array antenna over that of a mechanically rotated one.

Nevertheless, there were limitations, some of which were fairly severe, imposed by the array antenna used in our system that eliminated some of the aforementioned benefits. First of all, albeit the antenna was able to point at directions between -45° and 45° , the usable range was made narrower by the appearance of a grating lobe for reception angles less than -15° or more than 15° .

Figure 4.6: Structure of measurement data collection

Unfortunately, the destructive effects of these lobes on the measurement data were realized only after all measurement activities had been accomplished and a major portion of the data had gone through processing. One way to avoid these effects was by using only data corresponding to pointing angles between -15° and 15° in angular correlation function computations. That left us with a limited amount of data for a narrower range of reception angles.

In addition, a relatively long time gap existed between the start of acquisition for a specific reception angle and that for the angle next in order. This interval, which was made up of times for *I-Q* sequences acquisition, power acquisition, data transfer and recording, and pointing angle remote programming, was in the order of hundreds of milliseconds and led to a total time for one sweep of acquisitions of about 8 seconds. Two consequences arose from this limitation. (a) It precluded the temporal variation measurement that normally required a sampling rate of at least twice the channel maximum Doppler shift. (b) Also, during some of these eight-second sessions of measurement, changes in the multipath environment owing to movement of people nearby were inevitable. As a result, while the entire ensemble of the collected data can be used for delay spread computations, only a subset of them may be used for angle-correlation analyses, the subset being those obtained for locations where people movement was restricted.

To mitigate the latter problem, measurements were undertaken with fixed transmitter and receiver, exclusive of people activities in the vicinity if possible. Meanwhile, so as to measure the spatial variation, a set of 5 acquisitions were collected for every transmitter-receiver position combination, as explained in section 4.3; each acquisition in the set corresponded to one of five different positions of transmitter separated 1 cm (or one wavelength) apart. Assuming unchanged environments, these sets of 5 data could be considered to represent the spatial measurement due to the transmitter motion. If the transmitter was assumed to move at 1 m/s, then these data would indicate the temporal change within a timespan of 50 ms with a sampling period of 10 ms. Each set of the data would as well be used later to obtain power delay profiles, one for each Tx-Rx arrangement and for each reception angle. Also, an additional set of measurement run to be described in more detail in later chapter were undertaken to confirm the channel stationarity.

Another problem was that, due to the presence of hysteresis in the phased-array antenna phase controller, the angle pointed at by the antenna was not necessarily the correct absolute value of the angle programmed on the controlling computer. However, it was believed that the step

taken between two discrete angles, the absolute values of which were not exactly known, was very close to what had been programmed on the computer. This inaccuracy in absolute value of pointing angle made futile any attempt to use a pointer for determining 0° reception angle. Instead, the 0° reference angle determination was at all times accomplished by adjusting the spatial and angular orientations of both transmit and receive antennas such that the highest received power considered to correspond with the reference angle was obtained. This problem further precluded any possibility to perform channel measurements under shadowed condition and thus limited the collected data to those for non-shadowed situations.

Chapter 5

Measurement Results, Analyses and Discussions

5.1 Data Processing

Two types of data were collected from measurement sessions, namely the 1 GHz-sampled received in-phase and quadrature sequences and the received power readings. A pair of one-period I and Q sampled sequences (a period being defined to comprise 5115 sample points of the received 1023-bit I or Q sequence) and a single power reading were obtained from a single acquisition performed for a given set of antenna pointing angle, local area, position in the local area and measurement location. Prior to further analyses, the following chain of processings were applied on the recorded complex envelope $I + jQ$.

- a. First processes were the conversion from binary coded sampled values to actual quantized voltage levels, and the removal of DC components from both real and imaginary parts. The conversion was performed separately for each of I and Q sequences as follows

$$y_I(k) = [(r_I(k) - ScYOrigin) \times ScYInc] + ScYRef \quad (5.1)$$

where $y_I(k)$ is the I sequence in its actual voltage levels and $r_I(k)$ is the I sequence raw samples. $ScYOrigin$, $ScYInc$ and $ScYRef$ are the Y-axis origin, incremental and reference values, respectively, automatically appended by the oscilloscope to each Matlab data file during its transfer to the optical disk. The same operation was executed on the Q sequence raw samples.

- b. The complex-valued outcome was afterward periodically crosscorrelated with a replica of the transmitted PN sequence by applying the following

$$\begin{aligned} R_{yx}(k) &= \sum_{n=0}^{5114} y(n)x^*((n-k))_{5115} \\ &= IDFT[DFT(y(k)) \cdot (DFT(x(k)))^*] \end{aligned} \quad (5.2)$$

where $R_{yx}(k)$ is the output-input crosscorrelation function, $x(k)$ and $y(k)$ are the 5115-point

sampled transmitted sequence and received complex envelope, respectively, $x((n-k))_{5115}$ represents circular shift operation of a 5115-point sequence $x(n)$, and $DFT(\bullet)$ and $IDFT(\bullet)$ represent the discrete Fourier transform operation and its inverse. Taking only values for positive k , this step produced a 5115-ns estimate of the baseband equivalent channel complex impulse response function for that particular location, local area, individual position in the local area, and antenna pointing angle.

- c. In accordance with findings of past works on indoor millimeter wave radio channel maximum multipath spread [25], [22], the impulse response estimate was next curtailed to merely 400 ns long starting from the point of the first-detected pulse. This cutback in duration was also justified by results obtained from some preliminary measurement tests that showed that the maximum spread of the mm-wave indoor radio channel impulse response never exceeded 400 ns. Also, the arrival time of the first-detected pulse peak was set to zero, such that the time axis represents the excess delay of the arriving component, i.e. the component arrival time relative to that of the first-detected pulse.
- d. Finally, the complex impulse response function estimate was normalized to possess a unit energy. It was found, however, that such normalization was unnecessary since, thanks to the exertion of the AGC, all the resultant impulse responses had identical energy.

Meanwhile, the received power level underwent value conversion based on the previously-established power calibration curve. Power readings that fell in between calibrated points were converted by means of interpolation. This resulted in the actual power value, relative to -16 dBm reference, detected at the receive array antenna port. This value was subsequently used in analyses that required power value as direct input, and also in scaling the earlier-computed impulse response estimate so that the estimate also conveyed the information on its power strength relative to the others. The complete listings of Matlab programs used in the entire processing is included in Appendix E.

In later analyses, the following notations are used. $P(L, l, p, \theta)$ denotes the power detected at the receive antenna output at location or room L , local area or Tx-Rx arrangement l and position p with receive pointing angle θ . Likewise, $\hat{h}(L, l, p, \theta, \tau)$ implies the complex value at τ ns excess delay of the baseband equivalent channel impulse response estimate at location L , local

area l and position p , with receive direction θ .

To avoid any confusion with the use of several angular quantities throughout the following discussions, figure 5.1 is provided to clarify their respective definitions. The terms *receive antenna pointing angle*, *reception angle*, and *misalignment angle* are all identically defined as the azimuthal angle toward which the receive antenna maximum gain is directed. These terms are therefore used interchangeably, and often referred to herein by a single word *angle*. These three quantities are expressed relative to the direction of the transmit antenna, which was invariably pointed toward the receiver. This reference direction is also referred to as the *LOS* or the *direct path*. Accordingly, a 0° reception angle represents the situation in which both antennas are aligned to each other. The *horizontal 3-dB beamwidth* of an antenna is defined as the width of the azimuthal angle range within which the antenna gain is at least 3 dB below its maximum. Similarly, the *vertical 3-dB beamwidth* is the elevation angle range over which the gain is at least 3 dB below the peak. Unless specifically stated otherwise, the terms *beamwidth* and *main beam* both refer to the antenna horizontal 3-dB beamwidth.

Figures 5.2 and 5.3 depict a set of typical results after processing some data collected for an arbitrary set of values of L (which happens to be room 4050 of Minto building in this

Figure 5.1: Top view of a transmit-receive antenna configuration with a pictorial description of angular and other parameters used in subsequent discussions.

Figure 5.2: Magnitudes of impulse response estimates normalized to have equal energy, with the associated reception angle and static rms delay spread values printed on each graph.

exemplary case), l and p . Figure 5.2 shows eleven equal energy impulse response estimates, each for a single discrete value of receiver pointing angle θ . Superimposed on each graph are the corresponding pointing angle θ and the rms delay spread for that particular angle. Figure 5.3 presents the received power for every direction of reception for the same set of L , l and p .

Some important points are revealed from these and subsequent processing results. First, in all impulse response estimates, including those collected for non-zero pointing angles, the first-detected pulse possesses a large magnitude relative to those of other pulses in the same function estimate. Since all measurements were performed under line-of-sight condition, this means that despite the high directivity of the receive array antenna, the LOS component is still one of the

Figure 5.3: Relative received power (dB) vs. reception angle ($^{\circ}$).

major contributors of the received power even when the antennas are misaligned, a phenomenon referred to as “spillover”, as has also been observed by Licht [12] using a similar experimental setup. In other words, for non-zero reception angles, the effects of transmit antenna gain and direct LOS path loss, which is the lowest among those of all possible paths, in most cases still outdo the effects of receive antenna gain and losses due to longer propagation path and reflections. Also, it should be noted that due to the limited number of discrete reception angles applied in the experiment, these receive antenna pointing angles do not necessarily match the angles of arrival, which are random in nature, of reflected signal components that actually exist. Hence, these individual reflected signal components may not undergo the receive antenna maximum gain.

Second, from power-angle profiles, it can be observed that the difference in dB of the received power at 0° and that at others are very high, a phenomenon naturally attributable to the narrow beam of the receive antenna, the 3-dB beamwidth of which is 8° . However, small

reductions often occur from the received power at 0° to those at $\pm 6^\circ$ due to the inevitably imperfect adjustment of the 0° pointing direction.

Third, it is as well noticable that the power-angle profile is not symmetrical around the 0° direction, showing the evidence of asymmetry in the multipath structure geometry.

Finally, it is observed that in almost all impulse response estimates displayed in figure 5.2, a common strong multipath component appears at an excess delay of around 70 ns and occasionally becomes the last-detected component. It is found later that this large multipath component may be related to the length of the room in which the impulse response estimates were collected since most of the measurement links were parallel or almost parallel to the room length. In the above case, for instance, the length of room 4050 in which the impulse response functions were acquired is approximately 32 ft, a distance traveled over by light in 32 ns or so. Thus the multipath component may be induced by a portion of energy that propagates over a distance equivalent to twice the room length. Since similar phenomena are also obtained from measurement results at other positions and in other rooms, a more thorough inspection of these phenomena is required and to be done later in this chapter.

5.2 Temporal Variation

Channel temporal variation becomes the first aspect to address of indoor radio channels under investigation since it is desired to understand the impact imposed by the lack of temporal measurements on the validity of results gained in the following analyses.

For instance, the calculation of diversity gain potentially obtained from the implementation of an angular diversity technique with selection combining requires several steps. These include the construction of cumulative distribution functions of the received signal envelopes for all available channels (i.e. channels of different reception angles) from continually sampled data of the time varying envelopes, the derivation of cumulative distribution function of the combined received envelopes if the diversity scheme with selection combining is employed, and the determination of reduction in required transmit power when switching from non-diversity to diversity system while still achieving the same performance measure such as 99% availability [18], [12]. This power reduction is the desired diversity gain.

The derivation of cumulative distribution functions as explained in [18] requires fading envelope data that result from continuous sampling of the received signal envelopes at a rate sufficiently high to capture deep fades. In conclusion, continually sampled data of the time varying received signal envelope are required to determine the experimental cumulative distribution function of the fading envelope and thus the diversity gain that can possibly be attained by employing several uncorrelated channels. The absence of such data from this experiment precludes the determination of potential gain offered by the use of angular diversity technique. However, if it turns out that the radio channel is so static that the temporal measurements are unnecessary, then the channel snapshots collected during this experiment can still be useful to obtain angular correlation functions and coefficients of correlation among powers detected at different angles.

Theoretically, assuming people or objects are moving indoors at a velocity of 1 m/sec, the maximum Doppler shift for radio frequency of 29.5 GHz is approximately 100 Hz, implying a sampling rate of at least 200 samples/sec so as to successfully capture the channel temporal variation. Ideally, a measurement configuration used for purposes of acquiring pointing angle-dependent time-varying impulse response estimates must be able to sample the channels of different reception angles at the same time at the above-specified rate. In case of acquisition with programmable receive array antenna, one angular sweep of measurements must be accomplished within the maximum sampling period, that is in 5 msec, in order for the system to be in time to start the next sweep while maintaining the minimum required sampling rate.

As described in the previous chapter, the considerably large inter-angle switching time, approximately 200 msec, required for data transfer from oscilloscope to personal computer and the hysteresis property of the array antenna phase shifters prevented the angular sweep to be performed accurately and fast enough to portray the temporal and angular variations of the channel response. If we insisted on sampling the channel under that situation, we would immediately face the fact that the channel variations were highly undersampled, not to mention the inaccuracy of the antenna pointing angle. These facts permitted only collections of individual snapshots of the radio channels instantaneous conditions, rather than a chain of continual snapshots.

To seek the evidence of the channel stationarity, additional measurements were performed

in room 4050 Minto CASE building. Transmitter-receiver separation was set to 3.2 m, with both antennas aligned and 0° reception angle becoming the reference angle. The radio channel was to be sampled periodically as fast as the equipment allowed. At each acquisition time, a complex impulse response estimate was collected, as well as the received power value. The limitations in data transfer speed and oscilloscope memory size permitted only a collection of eleven estimates at a time with inter-acquisition time of 200 msec, commensurate with an overall observation time of 2 seconds. During these eleven consecutive acquisitions, the receive array antenna was pointed at a fixed angle; two fixed angles were chosen for this purpose, namely -15° and 15° . A total of ten eleven-acquisition sessions were executed for each angle. During the entire session, people movements were intentionally made in the vicinity of the radio link, in such a manner that these motions never blocked the line-of-sight path.

After necessary processing as explained in the preceding section, a time-domain autocorrelation function was computed from a set of eleven impulse response estimates for a given pointing angle. The output of this computation was a function of excess delay and discrete time lag, which was a multiple of 200 msec. If $\hat{h}(nT_s, kT_a, a)$ or $\hat{h}_n(k, a)$ denotes the n^{th} component of the k^{th} estimate of presumably time-variant impulse response with sampling period T_s of 1 nsec and inter-acquisition period T_a of 200 msec, and $a = 1, \dots, 10$ represents the acquisition session index, then the autocorrelation function for data set a is

$$R_{\hat{h}\hat{h}}(n, m, a) = \frac{1}{11} \sum_{k=0}^{10} \hat{h}_n(k, a) \hat{h}_n^*((k-m), a) \quad (5.3)$$

Next, results for ten data sets were averaged to yield the average autocorrelation function for that particular reception angle, as expressed in the following

$$\begin{aligned} R_{\hat{h}\hat{h}}(n, m) &= \langle R_{\hat{h}\hat{h}}(n, m, a) \rangle_a \\ &= \frac{1}{10} \sum_{a=1}^{10} R_{\hat{h}\hat{h}}(n, m, a) \end{aligned} \quad (5.4)$$

It can be seen that for any arbitrary value of impulse response component index n :

- a. a constant level (time-invariant) impulse response component, i.e. $\hat{h}_n(k, a) = \hat{h}_n((k - m), a)$ in equation (5.3), will produce a triangular correlation function with a slope of -0.45/sec for positive m .
- b. $R_{\hat{h}\hat{h}}(n, 0)$ represents the average power of n^{th} component, relative to that of zero excess delay component.

The average autocorrelation function for the above-described single-room experiment in dB scale for 15° reception is depicted in figure 5.4, in which a point on the three-dimensional curve corresponding with excess delay τ and time lag Δt represents the correlation of impulse response values at the same excess delay τ but separated Δt seconds away in observation time. In other words, a profile obtained by slicing the three-dimensional curve perpendicular to the excess delay axis at a specific delay value τ describes the stationarity of the τ -delay component in the channel impulse response, or equivalently, the immobility of objects that caused it.

From Figure 5.4, it is observed that, for any excess delay, the correlation function exhibits relatively small variations along the correlation time lag axis. For example, at an excess delay of approximately 60 nsec, the function value drops only by 1 dB as observation time difference increases from 0 to 200 msec, and by 10 dB when the time lag extends to 2 seconds. This gradual reduction¹ along the time lag axis indicates that the radio channel environment was fairly stationary despite the presence of people activities nearby. A plausible physical explanation to this phenomenon arises from the fact that both ends of the radio link employed a directional antenna, whereas during the measurements, movements were made at positions alongside the radio link yet still outside the beamwidth of these antennas. Consequently, these movements did not induce any significant change in the channel impulse response since the radiation patterns of both antennas imposed high attenuations on components arriving from their directions. Since in the actual experiment performed for all specified discrete reception angles nearby people activities were avoided or restricted to regions relatively far from the direct LOS path vicinity, this result can also be safely extended to these situations.

Thus, for systems with directional antennas such as that used in our experiment, the indoor

---

1. The reduction is considered gradual when compared to what will result if the channel is perfectly constant, i.e. when the resultant correlation function is triangular in shape, with a slope of -0.45/sec for positive time lags.

Figure 5.4: Magnitude in dB of average time autocorrelation function of the channel impulse response estimate for 15° pointing angle.

radio channel is slowly varying as long as the changes in environment take place outside the beamwidth of the antennas and do not cause any blocking of the line-of-sight path. This finding justified the use of channel single snapshots, instead of incessantly sampled channel acquisitions, in subsequent analyses. In this manner, any averaging required in the analyses would be done over all locations and positions in all local areas.

5.3 Delay Spread Distribution

As has been found and rigorously studied in previous research efforts, radio channel delay spread, represented by its maximum or root-mean-square value, is an important broadband radio channel parameter that determines how significant the effects are of intersymbol interference due to channel dispersive nature on the performance of radio communication link [36], [37]. The delay spread of millimeter-wave channels therefore needs to be quantified from measurements before digital communication systems can be implemented on such channels while achieving a

satisfactory performance in terms of irreducible bit error rate.

While there have been a large number of indoor delay spread measurements reported for lower frequency ranges as summarized in [5], fewer empirical studies have also been undertaken to measure the delay spread values of mm-wave channels at various carrier frequencies ([22], [23], [24], [25]). Delay spread measurements reported in this thesis were aimed to enrich the knowledge on indoor mm-wave channel delay spread properties at 29.5 GHz frequency band by investigating its statistics for various rooms and reception angles.

Two parameters were considered here as the delay spread measure, namely the static maximum delay spread and the static root-mean-square (rms) delay spread. Both were derived from individual impulse response estimate by applying different values of threshold level ν . These values ranged from 5 dB through 30 dB below the impulse response maximum level with 5 dB spacings. In the subsequent discussion, the two parameters are simply referred to as the maximum and the rms delay spreads.

The above parameters were determined as follows. From a single impulse response estimate, a power delay profile $F(L, l, p, \theta, \tau)$ was obtained by squaring the magnitude of each component of the associated impulse response

$$F(L, l, p, \theta, \tau) = |h(L, l, p, \theta, \tau)|^2 \quad (5.5)$$

Next, threshold level ν was applied by setting to zero the components of $F(L, l, p, \theta, \tau)$ that lied more than ν dB below its peak, i.e. those having a decibel level lower than $10\log[F(L, l, p, \theta, \tau)] - \nu$. The resultant function is denoted hereafter by $F(L, l, p, \theta, \tau, \nu)$. From each function of this type, the maximum delay spread value $\tau_{max}(L, l, p, \theta, \nu)$ was obtained as the maximum timespan of the function, that is the entire interval time from the point at which the incoming signal becomes nonzero for the first time up to the point at which the signal falls back to zero for the last time. From the same power delay profile, the rms delay spread $\tau_{rms}(L, l, p, \theta, \nu)$ in nanoseconds was determined by applying the following equation

$$\tau_{rms}(L, l, p, \theta, v) = \sqrt{\frac{\sum_{\tau=1}^{400} \tau^2 F(L, l, p, \theta, \tau, v)}{\sum_{\tau=1}^{400} F(L, l, p, \theta, \tau, v)} - \left(\frac{\sum_{\tau=1}^{400} \tau F(L, l, p, \theta, \tau, v)}{\sum_{\tau=1}^{400} F(L, l, p, \theta, \tau, v)} \right)^2} \quad (5.6)$$

The maximum discrete delay value τ of 400 resulted from the impulse response estimate truncation to 400 ns long done previously as described in section 5.1.

Three types of distribution functions were established from the resulting ensemble of maximum and rms delay spreads. First, using all available values in each set of the maximum and rms delay spreads, a set of distribution functions were determined with threshold level as the varying parameter. The function thus represents the overall distribution of each parameter as a result of 29.5 GHz broadband transmission over indoor radio channels in all of the experiment locations for all propagation distances and reception angles. The second and third sets of functions were derived for different experiment locations and different reception angles, respectively.

It should be noted that due to the use of SS-based measurement system, of which the transmitted sequence was 1023 bit in length and the bit rate was 200 MHz, the equivalent pulse transmitted through the baseband equivalent of the whole system was triangular in shape (obtained when an autocorrelation process is operated on a replica of the transmitted pseudonoise sequence), with a peak level of 1023 times the magnitude of the “+1” level of the pseudonoise sequence and a time width of 10 ns. As a consequence, the received pulse would have an inherent spread with a maximum of 10 ns and, by applying equation (5.6), an rms value of 1.56 ns even if the system as a whole including the transmitter, the radio channel and the receiver was nondispersive in nature.

5.3.1 Delay Spread Distribution over All Measurements

Parts (a) and (b) of figure 5.5 display the complementary cumulative distribution functions (CCDFs) of $\tau_{max}(L, l, p, \theta, v)$ and $\tau_{rms}(L, l, p, \theta, v)$, respectively, with the threshold v being varied. The mean, standard deviation and maximum of each delay spread value set of different v are listed in table 5.1.

Figure 5.5: Complementary cumulative distribution functions of maximum and rms delay spreads for different threshold levels, computed over all measurements.

Table 5.1: Statistics of Channel Maximum and RMS Delay Spreads over All Measurements

Threshold ν (dB)	Maximum Delay Spread (ns)			RMS Delay Spread (ns)		
	Mean	Standard Deviation	99 th Percentile	Mean	Standard Deviation	99 th Percentile
5	9.2	21.2	98.0	2.8	8.8	42.8
10	19.1	34.4	149.0	5.0	11.2	51.0
15	37.4	49.8	219.0	8.1	13.1	54.9
20	68.9	65.1	224.0	11.7	14.3	56.9
25	112.5	74.2	328.0	15.2	14.6	58.5
30	172.5	86.6	393.0	18.3	14.5	59.5

Several observations made from both the table and the graph are as follows.

- a. As the threshold is increased, the means of both maximum and rms delay spreads increase. This can be explained by the fact that the lower the threshold is set, the larger number of received signal components that otherwise fall below the threshold level, are taken into account in the delay spread computations. A similar observation has also been reported for 1 GHz frequency band by Hashemi [38].
- b. An observation identical to that explained in (a) is also found for the 99th percentile of $\tau_{max}(L, l, p, \theta, \nu)$. However, the same does not hold for the case of 99th percentile of the rms delay spread $\tau_{rms}(L, l, p, \theta, \nu)$. It turns out from table 5.1 that the 99th percentile of the rms delay spread is not so sensitive to the threshold level used in its computation as is that of the maximum delay spread. For almost any level of threshold employed, the rms delay spread 99th percentile is found at all times to be approximately 50–60 ns.
- c. Although not shown in both table 5.1 and figure 5.5, it should be mentioned that a very small number of impulse response estimates (apparently less than 1% of the entire collection) exhibit rms delay spreads that may extend up to 130 ns.

5.3.2 Delay Spread Distribution for Various Experiment Locations

Figure 5.6 displays two different sets of curves. The first two graphs, presented by parts (a) and (b), depict the CCDFs of the maximum and rms delay spreads, respectively, for different rooms selected as measurement locations, with the threshold line ν being set to 20 dB below each power delay profile peak. Likewise, parts (c) and (d) show the CCDFs of the maximum and rms delay spreads, but with a threshold line positioned at 30 dB below the individual profile peak. To each curve in all graphs, a four-character code is tagged, indicating the room for which the curve represents the maximum or rms delay spread CCDF. Description of room dimensions and shapes is given previously in section 4.4 of this thesis, while floor plans are included in Appendix A.

Based on observations over these graphs, the following issues are found.

- a. Variation of the threshold level causes more pronounced changes in the distribution of the maximum delay spread than in that of the rms delay spread.

(a) $\tau_{max}(L, \nu)$, $\nu=20\text{dB}$ (b) $\tau_{rms}(L, \nu)$, $\nu=20\text{dB}$ (c) $\tau_{max}(L, \nu)$, $\nu=30\text{dB}$ (d) $\tau_{rms}(L, \nu)$, $\nu=30\text{dB}$

Figure 5.6: Complementary cumulative distribution functions of maximum and rms delay spreads for various experiment locations, with threshold levels at 20 and 30 dB below the power delay profile peak.

- b. It is interesting to note that for each room, large concentrations of maximum delay spreads exist at excess delays equivalent to even multiples of the length of the corresponding room. For instance, in room 4050, the length of which is 9.6 m or 31.5 ft, with 30 dB threshold, a large number of maximum delay spreads are concentrated around values of 70 ns, 135 ns and 200 ns, as is indicated by abrupt drops on the CCDF curve at these points in figure 5.6(c). Similar phenomena are found to occur in other rooms. In retrospect, experiments were done in all locations in most cases with the transmitter-receiver direct pathline being parallel or nearly parallel to the length of the respective rooms, as has been briefly described in section 4.4. Consequently, considering the narrow beams of the antennas, it can be surmised that one of the primary radio propagation modes, beside the direct LOS mechanism, is the multiple reflections off the walls behind the transmitter and the receiver, and that relatively large echoes can be expected to appear at excess delays commensurate with the even multiples of the room length.

A deeper insight into this relation of the channel maximum delay spread and the size of the room for which the delay spread measure was acquired was sought in the following manner. The set $\{\tau_{max}(L, \upsilon)\}$ of maximum delay spreads measured in room L with threshold level υ were normalized to the size $s(L)$ of room L . For this purpose, in accordance with the above-stated conjecture the room length in feet was chosen to represent $s(L)$.

A family of new quantities was subsequently obtained, each of which is denoted by $\tau_{norm}(L, \upsilon)$ and derived as follows

$$\tau_{norm}(L, \upsilon) = \frac{\tau_{max}(L, \upsilon)}{s(L)} \quad (5.7)$$

Recalling that the speed of light is approximately 1 ft/ns, $\tau_{max}(L, \upsilon)$ may represent the excess propagation distance in feet of the last-detected multipath component, where the excess distance of a component is simply the difference in propagation length between that and the first-detected component. Since $\tau_{max}(L, \upsilon)$ is expressed in nanoseconds and $s(L)$ in feet, $\tau_{norm}(L, \upsilon)$ therefore measures the maximum excess propagation distance relative to the room length.

The CCDF curves of $\tau_{norm}(L, \upsilon)$ for all different rooms at a given threshold υ are altogether plotted in the same graph. Three graphs of this type for 10, 20 and 30 dB threshold are

shown in figure 5.7.

- a. Comparisons between figures 5.6(a) and 5.7(b), and between 5.6(c) and 5.7(c) indicate that, except for the hallway case (room 4031), the CCDF curves of $\tau_{max}(L, v)$ for various rooms that lie somewhat sparsely are shifted nearer to one another after normalization.
- b. The normalized maximum delay spread CCDF curves of all rooms show fast drops around excess propagation distances of even multiples of the room length. This observation is in favor of the previous supposition that multiple reflections off the walls behind the transmit and receive antennas exhibit a relatively significant power contribution. Their power contribution is sufficiently high that very often they become the last component detected above the applied threshold despite their lengthy propagation path. This observation, however, needs a theoretical support, which is performed in a manner described in the next point.
- c. A simple verification step is performed to confirm the observation in point (b). The verification is done by calculating the relative power of the pulse associated to the second-order reflection off the endwalls with respect to the first detected pulse, and by comparing the calculation result to that shown by the measured impulse response estimate (those collected for 0° reception are selected for this comparison). In the computation process, it is assumed that the reflection coefficient of the endwalls is of unit magnitude, that each of the first pulse and the pulse detected at a delay of second-order reflection consists only of a single ray (i.e. no vectorial addition of other, unresolved rays), that the LOS ray and the multiply reflected rays experience the same antenna gain, and that the measurement link is exactly parallel to the room length. The relative power of the second-order reflection can then be computed by adopting a propagation exponent of two for each ray as follows

$$P_{2nd, dB} = 20 \log \left(\frac{d}{d + 2s} \right) \quad (5.8)$$

where d and s represent the propagation distance of the direct ray and the room length, respectively, in meters. The result shows that the calculated relative power of the pulse arriving from a doubly-reflected path may be less or larger than the measured, with a maximum difference of around 5 dB. While the small differences between the prediction and the measurement, which might have been created by the aforesaid assumptions, tend to

Figure 5.7: Complementary cumulative distribution functions of maximum delay spread normalized to the corresponding room size, with threshold levels at 10, 20 and 30 dB below the power delay profile peak.

support our conjecture, this result does not completely confirm our conjecture. Another set of measurements in other rectangular rooms may hence be needed to check whether the same phenomenon happens consistently in rooms of such kind.

- d. When the threshold is shifted to 10 dB, there are practically only two types of resultant impulse response estimates, the ones with a single, LOS component, and the others with multiple components that include the first-detected, LOS component and the last-detected component at a delay equivalent to that of a second order reflection along the room length. Very few impulse response estimates exhibit the presence of multipath components beyond the latter path at 10 dB threshold. In fact, from figure 5.7(a) it can be seen that most of impulse response estimates collected in the hallway at 10 dB threshold are single-pulse functions. It follows that in many cases the multipath components have a magnitude of 10 dB or more below the first-detected, direct path component.
- e. In general, it can be observed from the figures that the larger in length is the room, the higher the concentrations of its normalized maximum spread at smaller values. This is especially true when we compare the CCDFs of these rooms for the 20 dB threshold case.

5.3.3 Delay Spread Distribution for Different Reception Angles

To gain knowledge of the impact of variation in reception angle on the channel delay spread, CCDFs of the delay spread parameters are plotted in figure 5.8. Parts (a) and (b) of the figure present the CCDFs of the maximum and rms delay spreads, respectively, for various receive antenna pointing angles, with a 20 dB threshold below power delay profile peak. CCDFs plotted in parts (c) and (d) are computed in a similar way to those shown in (a) and (b), the only difference being in the threshold level which is set to 30 dB below the peak.

Name tags are only given to curves of CCDFs of important pointing angles, such as 0° , and to those showing interesting behaviors, such as that exhibiting a tendency somewhat different from others. This is done since, as can be observed from the figures, most of the curves exhibit similar properties. These curves are lying so close to each other that identifying any one out of them is very difficult. This fact, however, reveals that the maximum or rms delay spreads for different reception angles, with a few exceptions, show insignificant differences in their statistics

Figure 5.8: Complementary cumulative distribution functions of the maximum and rms delay spreads for various reception angles, with threshold levels being set at 20 and 30 dB below the power delay profile peak.

and distributions. The following are points of great interest found from observing the plots in figure 5.8.

- a. Despite the high directivity of both the transmit and the receive antennas, the case of 0° reception angle, for which both antennas are aligned, is not necessarily the one that gives the smallest values of maximum and rms delay spreads. This is probably related to the presence in any impulse response estimate at 0° reception of a large, long-delayed component multiply-reflected off the room endwalls, as has been stated in section 5.3.2, particularly when the measurement link is almost parallel to the sidewalls (or perpendicular to the endwalls) such that the even order reflection enters the receive antenna through its main lobe.
- b. A Kolmogorov-Smirnov test on a 200-sample data set shows that the uncertainty range of the distribution function computed from the data set is ± 0.12 when a 99% confidence level is desired [16]. This uncertainty range is then applied to each of the CCDFs in figure 5.8. This exercise reveals that for the maximum delay spread at 30 dB threshold, except for reception angles of -30° , -18° , 12° , 18° , and 24° , the CCDF for any reception angle resides in the uncertainty region of that for any other angle, as is shown by part (c) of figure 5.8. This is also true for CCDFs of the rms delay spread at 30 dB threshold, as can be seen in part (d) of the same figure. However, this does not hold when a 20 dB threshold is applied on each impulse response estimate (see parts (a) and (b) of the figure).
- c. The -18° and $+18^\circ$ cases tend to yield the smallest values of rms and maximum delay spreads. Many plausible causes can be thought of as the explanation of such results. Those among others are the presence of a grating lobe for the above angles and the physical geometry of each experimental link with regard to the room layout. To gain more understanding of what could have caused such outcomes, the computation of experimental CDFs of the maximum delay spreads was attempted for any combination of room and reception angle despite the limited data available. The individual results for rooms 4050, 4038, A507 and 4031 at 30 dB threshold, depicted respectively in figure 5.9(a), (b), (c) and (d), do demonstrate that in every room the same trend occurs that the $\pm 18^\circ$ receptions give rise to the smallest of maximum and rms delay spreads. Since all rooms, in spite of their similarity in size and shape, are diverse in volume, positions and types of objects and furniture inside them, the above

(a) $\tau_{max}(\theta, \nu)$, $\nu=30\text{dB}$, room 4050 (b) $\tau_{max}(\theta, \nu)$, $\nu=30\text{dB}$, room 4038 (c) $\tau_{max}(\theta, \nu)$, $\nu=30\text{dB}$, room A507 (d) $\tau_{max}(\theta, \nu)$, $\nu=30\text{dB}$, room 4031

Figure 5.9: Complementary cumulative distribution functions of the maximum delay spread measured in individual experimental rooms, with the threshold level being set at 30 dB below the power delay profile peak.

phenomenon is perhaps more related to an unknown, inherent nature of the receive array antenna. A more thorough investigation, however, is not attempted here.

- d. Another angle-dependent phenomenon also happens for -30° and 24° angles, which tend to give the largest values of rms and maximum delay spreads over the entire measurements and over measurements in each room.

5.4 Angle-Dependent Received Power

In an indoor setting where a wireless network is deployed with base station and user terminals employing directional antennas, the situation most expected to give the highest performance of a specific user-to-base line-of-sight radio link is when antennas at both ends of the link are precisely aligned. As such, the base antenna will be very likely to receive maximum power from the user, as long as nothing intervenes in between. However, signals from another active user terminal making an angle α with the LOS pathline of the former antennas may still act as a potential interference although they might have been somewhat attenuated due to the base antenna directivity. This situation is illustrated in figure 5.10, in which user terminal A is the desired and B the interfering terminal. It is thus desired to quantify the amount of interference a directional-to-directional antenna system will suffer from.

Figure 5.10: A possible setting of network of base station and user terminals with directional antennas where interference is caused by signals with different angle of arrival.

To do so, from every data file containing a set of angle-dependent power values, another set of parameters was created that contained the ratio of power detected at pointing angle θ to that at zero pointing angle, i.e. $P(L, l, p, \theta)/P(L, l, p, 0^\circ)$. This quantity, the angle-dependent normalized power, had a value between 0 and 1 since the power received at zero angle was always higher than that at any other angle. A total of 200 sets of this quantity were obtained, each from a single measurement data file. Next, the mean $\mu_R(\theta) = \langle P(L, l, p, \theta)/P(L, l, p, 0^\circ) \rangle_{L, l, p}$ and standard deviation $\sigma_R(\theta)$ of the normalized power were computed over all locations, local areas and positions in them. The results are depicted in figure 5.11. Using the above notations, interference coming from reception angle θ at the same distance as that of the desired 0° signal is quantified by $\mu_R(\theta)$.

Great care should be taken in interpreting this quantity and thus a clearer picture of what actually happened during the experiment is needed in the analysis. Sketches drawn in figure 5.12 display two possible settings of the transmit and receive antennas. The first sketch depicts the

Figure 5.11: Average θ -to- 0° received power ratio $\mu_R(\theta)$, with superimposed vertical lines showing range of mean $\pm$ standard deviation $\sigma_R(\theta)$.

situation in which the receive array main beam makes a 10° angle with the imaginary direct line linking both antennas, with the transmit antenna being pointed at the receiver. In the second sketch, the receive antenna main beam is shown to be directed toward the 20° azimuth with respect to the transmit-receive direct pathline, for which condition a relatively high grating lobe starts to appear at approximately -40° azimuthal angle. Patterns shown for the receive antenna are obtained from measurement [35] and are drawn in dB scale, while those for the transmit horn are shown to have an ideal 30° beamwidth. Although these exemplary conditions assume reception angles that were not used in the actual experiments due merely to the radiation pattern data unavailability for the experimental pointing angles, physical understanding is expected to be gained about signal reception scenarios with or without the grating lobe appearance by observing these two kinds of typical situations. The following are discussions on significant points observed from the experimental results and the accompanying sketches of figure 5.12.

Figure 5.12: Two possible transmit and receive antennas settings, (a) without and (b) with the presence of a grating lobe on the receive antenna horizontal pattern.

First, recalling the discussion in section 5.1, for any reception angle the major portion of received power comes from the direct LOS component. At some occasions, however, this domination is not so significant. In a situation like this, the power received from a transmitter at direction θ is also determined by the power conveyed by each resolvable multipath component that may come from a very different direction from θ . Translating this back to the above-described indoor setting, the interference signal power does not necessarily all come from angle θ . Rather, its angles of arrival depend on the multipath structure of the indoor environment being considered. In other words, different portions of the interference signal power that reach the receive antenna may have before undergone reflections off different objects located at various positions relative to the receiver's position in the room.

In relation with reasoning in the preceding paragraph and by considering the two probable scenarios described by figure 5.12, it is realized that the power received from a nonaligned transmit antenna may be the aggregate of (i) portions of power propagating over the direct LOS pathline and thus received by the array antenna through its 0° sidelobe, (ii) portions that encounter reflections and enter the antenna right through its main lobe, (iii) portions that reach the array antenna through its grating lobe (for cases where $|\theta| > 15^\circ$), and (iv) other portions that arrive from any other direction after some reflections or scatterings and encounter the receive antenna sidelobe. Similarly for the zero reception angle, the received power consists of contributions by pulses coming from (i) the direct path through the receive antenna main lobe, (ii) the even order reflections off the walls behind the transmit and receive antennas, if such mechanism is feasible, that lead to reception through the receive array main lobe and (iii) other paths that lead to reception by the array antenna sidelobes.

Figure 5.11 reveals that, apart from $\pm 6^\circ$ receptions, the received signal level at a nonzero reception angle, even at any of those angles whose radiation patterns exhibit a large grating lobe, is on average 10 dB or more below that at zero angle. Recalling that in any measured impulse response estimate it has been found that the component arriving through the direct path is predominant in power contribution (e.g. as has been demonstrated by figure 5.7(a) from which it is noticeable that only a small percentage of the impulse response estimates have multipath components that differ in level from the peak by 10 dB or less), a safe conclusion can be drawn that in most cases for nonzero reception angles, the total absolute power contribution of the

multipath components, i.e. those arriving by mechanisms (ii) to (iv) described in the preceding paragraph, is on average more than 10 dB below the received power at 0° reception. Also, for nonzero reception angles, it should always be noticed that a strong reflection that lies exactly on the path experiencing the receive antenna main lobe may not exist all the times.

On the basis of the aforesaid points, it can be safely stated that a large power contribution of reflected waves cannot be relied upon if one wants to implement angle diversity scheme with a number of directional antennas pointing to different directions. For example, if the direct path is blocked by some object or person causing a drop in the received power on the receive antenna branch most aligned to the transmitter, one must rely upon the received power on any other branch which on average is 10 dB lower than that of the aligned antenna branch. In fact, Licht in [12] found that due to the “spillover” phenomenon and the physical closeness of these antenna branches, when the direct path of the aligned antenna branch is blocked, the power on any other branch is also reduced, indicating the suppression of the direct path component. Hence in this more severe yet plausible case, if an angle diversity receiver is desired, the receiver must be able to somehow capture the reflected energy, which is very small if existent, under a severely shadowed condition that makes the signal on the most aligned antenna branch unusable.

As for the $\pm 6^\circ$ reception angles, at which the received power on average shows a relatively small degradation from that at zero angle, with a large variation as indicated by the large power standard deviation for each of those angles, an explanation for the results is as follows. Since the receive array antenna has a 3 dB beamwidth of 8° , when it is pointed to either of the $\pm 6^\circ$ angles the 0° direction falls on one of the sloping edges of the narrow main beam. This may explain the relatively small average reduction in the received power at $\pm 6^\circ$ from that at 0° , compared to results for other nonzero angles. Moreover, because from previous results it prevails that for any pointing angle, including any of the $\pm 6^\circ$ angles, the direct path component arriving from 0° direction is the major power contributor, a slight variation in the array antenna physical orientation, which shifts the 0° pathline up or down the edge of the antenna main beam, gives rise to a large variation in the received power at $\pm 6^\circ$ angles.

The asymmetry observed in the angle-dependent received power around 0° angle is most

possibly caused by the asymmetry in physical geometry of the environments chosen as experimental sites.

Finally the above result shows that:

- a. An interfering signal of transmitted EIRP and distance identical to those of the desired signal but arriving from a direction other than that of the direct path causes as low an interference power as 10 dB below the desired signal power, as long as it does not originate from a source inside or close to the region covered by the receive antenna pattern main lobe.
- b. The quantity $\mu_R(\theta)$ also indicates the average loss in power when the directional transmit and receive antennas are misaligned by θ . Such a high loss, approximately 6 dB for 6° misalignment, is expected when high directivity antennas are in use.

5.5 Angular Correlation Function

In this section, computations and analyses of angular correlation functions are discussed. The discussion is important to quantify the correlations among the radio channel responses for different values of the receiver pointing angle. The measurement results, while they could not be used to determine the potential angular diversity gain as explained in section 5.2, are still significant to measure the correlation between two different angular diversity branches, and thus to determine the most appropriate pointing angles to be taken as these diversity branches.

Methods similar to that used in section 5.2 for time correlation function computations were employed. Equations similar in form to (5.3) and (5.4) were applied to every set of eleven impulse response estimates measured in the same location and position (the same values of L , l and p) but with different reception angle θ . If $\hat{h}(L, l, p, k\theta_s, nT_s)$ or $\hat{h}_n(L, l, p, k)$ represents the n^{th} sampling point of one of the estimates belonging into a given set of equal L , l and p , where $\theta_s = 6^\circ$, $k = -5, -4, \dots, 4, 5$, and $T_s = 1$ ns, then the angular autocorrelation function for the set is

$$R_{\hat{h}\hat{h}}(n, m, L, l, p) = \frac{1}{11} \sum_{k=-5}^5 \hat{h}_n(k, L, l, p) \hat{h}_n^*((k-m), L, l, p) \quad (5.9)$$

The final form of the autocorrelation function to be used in the analysis was obtained by averaging

$R_{\hat{h}\hat{h}}(n, m, L, l, p)$ over the entire measurements, that is

$$R_{\hat{h}\hat{h}}(n, m) = \langle R_{\hat{h}\hat{h}}(n, m, L, l, p) \rangle_{L, l, p} \quad (5.10)$$

Due to the presence of a grating lobe for absolute pointing angles greater than 15° , two correlation functions were determined. The first involved all pointing angles used in the experiments, and the other took into account only those within the $\pm 12^\circ$ range. The first thus would have 21 points of angular differences ranging from -60° to 60° with 6° steps, while the second spanned from -24° to 24° . Both functions expressed in dB relative to their values at $n=m=0$ are depicted in figure 5.13; part (a) illustrates the 21-point correlation function, whereas (b) plots the 9-point function. Like the time correlation functions, it can be seen that pointing angle-invariant impulse response estimates, i.e. $\hat{h}_n(k, L, l, p) = \hat{h}_n((k-m), L, l, p)$ in equation (5.9), will lead to a triangular correlation function. Such a function will have a slope of $-1/(30^\circ)$ if pointing angles only in the $\pm 12^\circ$ range are considered, or $-1/(66^\circ)$ if all angles in the $\pm 30^\circ$ range are taken into account in the computation, for positive m . The value of each function at excess delay τ and angular difference $\Delta\theta$ represents the average correlation of the τ -excess delay components of two impulse response estimates obtained for two different reception angles separated by $\Delta\theta$. Since the effects of differences in multipath structures of various experimental environments are averaged by the application of equation (5.10), the resultant functions reflect only the influences of the antenna patterns.

From both figures we may note that for any excess delay τ at non-zero differential angles, the values of the correlation functions are lower than those at 0° by more than 10 dB. In general, the greater the angular difference, the smaller the correlation, with few exceptions. In figure 5.13(a) the function rises above -20 dB at 48° difference before plunging back to -30 dB at 54° and moving up again to -25 dB at 60° difference. The rise in correlation at 60° difference may be explained by the fact that to obtain the function value at that difference, only two impulse responses were involved, i.e. those associated with $\pm 30^\circ$ reception angles, for which a large grating lobe appears at approximately $\mp 30^\circ$ such that both reception angles essentially produce almost identical antenna radiation patterns. This similarity in receive antenna pattern leads to a

(a)

(b)

Figure 5.13: Angular autocorrelation functions averaged over all measurements, (a) involving pointing angles within the $\pm 30^\circ$ range and (b) involving only those in the $\pm 12^\circ$ range.

high correlation of impulse responses of those two reception angles. The other spike at 48° , however, is inexplicable. The same also holds for the small rise at 18° in figure 5.13(b).

The above observation indicates that using array antennas of the same type as ours, sufficiently small correlation (less than 0.1) between two channels of different reception angles can be achieved by pointing the antennas at two angles that differ by as small as 6° . This result may be related to the narrow main lobe of the receive antenna, the 3-dB half width of which is 4° . Smaller correlations can certainly be obtained by making use of two pointing angles with a larger difference, as demonstrated by the decreasing value of correlation when the angular difference is increased. However, this result applies under static conditions only as it results from measurements in various rooms with no people movement nearby (as has also been explained in section 5.2), and hence can not be used to determine the required angle separation for an angular diversity system employed under fading conditions.

5.6 Power-Distance Relation

Relation between the received power and the indoor radio propagation distance is of great interest because complete knowledge of how the received signal power behaves as a receiving terminal moves away from the base radio transmitter is useful in the determination of indoor cell size given the transmitted power level. In addition, the power-distance relation in a line-of-sight situation in which both transmitter and receiver reside in the same room but with antennas being nonaligned, to our knowledge, has never before been investigated. Any result from this effort is invaluable for performance prediction pertinent with power reception in indoor wireless systems that employ single or multiple directional antennas at either user or base terminal.

The power-distance relation model most commonly adopted by researchers is the inverse n^{th} power model [5]. In this model, the received power on average is assumed to vary as the inverse n^{th} power of the propagation distance, with the distance exponent n being dependent on the type of environment. If the power is expressed in dBm, or alternatively dB relative to a known reference level, and the distance in logarithmic scale of meter, the curve transforms to a straight line. The relation, accordingly, can be expressed in a form of linear equation

$$\bar{P}(d) = P(1) - 10n \log_{10} d \quad (5.11)$$

where d is the distance in meter between the transmit and receive antennas, n is the exponent of distance, $\bar{P}(d)$ is the average relative received power in dB at distance d , and $P(1)$ is the relative received power in dB for 1 m separation between transmitter and receiver in free space. The value of $P(1)$ is dependent on the antenna patterns, the radio frequency used and the transmitted power. The exponent n determines the downward slope of the straight line formed by equation (5.11).

If on average the relative received power in dB does have linear relationship with the distance in logarithmic scale, deviations from the linear fit observed in experiments are attributed to the presence of uncontrolled variables whose effects on the received power are not taken into account in equation (5.11) [15]. In our case, these uncontrollable variables include the variations in multipath environment and partial shadowing [20] because of changes in the environment of different rooms and different transmitter-receiver arrangements taken in performing the experiments. These variations induce differences between the measured power values and the average values obtained from equation (5.11). For each pair of measured data $\{d_i, P(d_i)\}$, where i is data index, the difference ε_i in dB is found to equal $P(d_i) - \bar{P}(d_i)$.

The distance exponent n is the parameter that characterizes the environment of the radio communication system. Hence, dissimilar types of indoor environment possess possibly different values of n [5]. To determine the distance exponent for a given type of environment, a common method is by performing a linear curve-fitting over pairs of distance and measured power obtained in that environment to acquire the values of $\hat{P}(1)$ and $\hat{n}$ that estimate their actual values in equation (5.11). The curve-fitting is performed to achieve the least sum of squared error between

the linear fit and the actual measured data, that is to minimize $\sum_{i=1}^N \varepsilon_i^2$, where N is the data size.

The above estimates can be obtained by applying the following pair of equations derived from equations (2.20) and (2.21)

$$\hat{n} = -\frac{1}{10} \left[\frac{N \sum_{i=1}^N (\log_{10} d_i) P(d_i) - \left(\sum_{i=1}^N \log_{10} d_i \right) \left(\sum_{i=1}^N P(d_i) \right)}{N \sum_{i=1}^N (\log_{10} d_i)^2 - \left(\sum_{i=1}^N \log_{10} d_i \right)^2} \right] \quad (5.12)$$

$$\hat{P}(1) = \frac{1}{N} \sum_{i=1}^N P(d_i) + 10 \hat{n} \frac{1}{N} \sum_{i=1}^N \log_{10} d_i \quad (5.13)$$

It is apparent from the preceding paragraphs that the set of errors $\{\epsilon_i\}$ can be used to quantify the effects of the third, uncontrolled random variable, i.e. the location- and arrangement-varying multipath environment, on the set of measured powers $\{P(d_i)\}$. However, to be able to quantify their effects, an additional assumption is required that random errors (in dB) for different distances are independent with identical zero-mean, σ_ϵ^2 -variance normal distributions [15].² The desired measure of their varying effects can in turn be obtained by computing the unbiased estimate of variance $\hat{\sigma}_\epsilon^2$ of the zero-mean random variable ϵ , for which $\{\epsilon_i\}$ are the experimental samples, using the following formula

$$\hat{\sigma}_\epsilon^2 = \frac{1}{N-2} \sum_{i=1}^N \epsilon_i^2 \quad (5.14)$$

In the case of the data pool acquired from our experiments, to each of the above-mentioned variables except the propagation distance, another independent variable, the reception angle θ , was inserted. Aside from that, prior to linear curve-fitting execution on the experimental data, the following data preprocessing steps were taken. First, a collection of transmitter-receiver distance data were constructed from location measurements tabulated during experimental sessions, as explained in section 4.3. This step ended up with an ensemble of values of $\{d(L, l, p)\}$, which denotes the distance d measured at location L , for transmitter-receiver arrangement l and position p . Tx-Rx distances in the resulting set ranged from 3 m to 10 m. Next, each entry of this collection was paired up with its associated set of eleven angle-dependent average received

2. This becomes the only reason for the selection of a lognormal distribution as the hypothesized distribution to describe the received power spatial variation later on, since there is no known physical background to such hypothesis for results of measurements in various rooms for different reception angles under an invariably line-of-sight condition.

powers $P(L, l, p, \theta)$.

Only a single collection of distances versus angle-dependent powers would result from accumulation of all measurements performed in all locations, with all measurements in the hallway (room 4031) being an exception; data pairs for this environment were excluded from the processing. The reason was that this room has a physically distinct multipath structure compared to the other, normal rooms. Hallways, favorite target locations for many indoor mm-wave measurements ([18], [24], [17], [19]), have been reported to exhibit wave guiding properties induced by the parallel floor and ceiling and the two sidewalls. Hence, the results of the following analysis were expected to show the average characteristic of power-distance for rooms of size and contents similar to those of the remainder.

Following the preceding discussion, the set on which curve fitting would be operated could be written as $\{d_i, P(d_i, \theta)\}$. The power-distance linear curve fitting was performed for each reception angle, producing for each angle a pair of $\hat{P}(1, \theta)$ and $\hat{n}(\theta)$ values as are listed in table 5.2.

From each of these pairs, a linear curve is plotted for the associated reception angle as depicted in figure 5.14. On the same plots, points are drawn to represent the measured data. In figure 5.15, linear curves computed for all individual receive antenna pointing angles are presented in the same plot, thereby allowing direct comparisons of either $\hat{P}(1, \theta)$ or $\hat{n}(\theta)$ among all curves for different angles to be conveniently done. From the measured data sets and the resultant linear curves, samples of errors $\{\epsilon_i(\theta)\}$ were computed, while the squared error estimate $\hat{\sigma}_{\epsilon(\theta)}^2$ for each reception angle was obtained by applying equation (5.14). The standard deviations of random variable $\epsilon(\theta)$, square root of the squared error estimates, are also given for all angles in table 5.2.

To explore more deeply the impact of the varying environment on the angle-dependent received power level, from the sample set $\{\epsilon_i(\theta)\}$ a cumulative distribution function was constructed for every θ . Their plots are presented in figure 5.16. On each of these curves, a zero-mean gaussian distribution function having variance of $\hat{\sigma}_{\epsilon(\theta)}^2$ is superimposed to facilitate comparison between the empirical error distribution function and that of the corresponding

gaussian random variable.

In relation with the plots in figure 5.16, Kolmogorov-Smirnov goodness-of-fit test [15] was performed with a zero-mean $\hat{\sigma}_{\varepsilon(\theta)}^2$ -variance normal distribution being conjectured as the null hypothesis. Maximum absolute difference $D_{max}(\theta)$ between the values of the experimental and its associated theoretical normal CDF is tabulated for each reception angle, along with the data size $N(\theta)$, in table 5.2. The previously-stated null hypothesis for any angle cannot be rejected with a confidence level of 0.99 if the difference between the experimental and the hypothesized CDFs for that angle is in the range of $\pm 1.63/(\sqrt{N(\theta)})$ [16], or ± 0.133 for data size of 150 (clustered in sets of five values, with each of the 30 sets corresponding to a particular propagation distance).

Figure 5.14: Scatter plots of relative received power versus distance for various values of reception angles, on each of which the associated linear curve fit is superimposed.

Figure 5.15: Linearly fitted curves of relative received power (dB) vs. propagation distance (m) in logarithmic scale for all reception angles ($^{\circ}$).

The following are points of interest resulting from observations over the results presented in the preceding figures and table.

- a. The relative average received power in dB has been found to be a linearly decaying function of distance in logarithmic scale for any reception angle, or in other words, the inverse power of distance model well describes the power-distance relation regardless of the receive antenna pointing direction for situations in which the transmitter and the receiver are situated line-of-sight in the same room.
- b. The distance exponent for various angles ranges from 0.31 to 1.41. This means that in the selected environments, for any pointing angle of the experimental receive antenna, the median received power decreases more gradually with distance than it does in free space.
- c. For the aligned antenna case ($\theta=0^{\circ}$), the propagation exponent is 0.93, very small relative to the free space index of two normally expected considering the high directivities of the antennas used in the experiment. Conjectured causes are described at the end of this section,

together with the results of simulations for verification purposes.

- d. For large reception angles, i.e. those having a magnitude greater than 6° , the received power is very small and sometimes falls below the measurement system's noise floor, especially for measurement links of large distances. In such cases, the level of the noise floor is taken as the measured received power. Hence, this phenomenon is probably the cause of the small values of propagation index n for large pointing angles.
- e. The average received power at 1 m distance is the highest for 0° angle, i.e. when the transmit and receive antennas are aligned. Over all results, $\hat{P}(1, \theta)$ ranges from -15.58 to -0.08 dB.

Figure 5.16: Plots of experimental cumulative distribution functions of deviations in dB of the measured received powers from the associated linear-fit means for different reception angles, along with the theoretical normal CDFs of the same means and variances.

Table 5.2: List of $\hat{n}(\theta)$, $\hat{P}(1, \theta)$, $\hat{\sigma}_{\varepsilon}(\theta)$, $D_{max}(\theta)$, $N(\theta)$, and the 1st percentiles of experimental and theoretical Gaussian CDFs of the random error

Reception angle ($^{\circ}$)	$\hat{n}(\theta)$	$\hat{P}(1, \theta)$	$\hat{\sigma}_{\varepsilon}(\theta)$	$D_{max}(\theta)$	$N(\theta)$	1 st percentile (exp.)	1 st percentile (normal)
-30	0.71	-15.58	2.41	0.10	150	-3.84	-5.62
-24	1.14	-12.81	1.99	0.06	150	-3.90	-4.65
-18	0.86	-11.75	1.32	0.05	150	-3.73	-3.07
-12	1.11	-8.19	2.29	0.15	150	-6.37	N/A
-6	1.41	-2.35	3.03	0.20	150	-8.14	N/A
0	0.93	-0.08	0.78	0.07	150	-2.50	-1.81
6	0.31	-11.77	3.49	0.10	150	-6.77	-8.14
12	0.86	-12.34	1.33	0.05	150	-4.10	-3.09
18	0.99	-11.76	1.71	0.08	150	-3.92	-3.98
24	1.24	-9.19	2.53	0.16	150	-7.99	N/A
30	0.88	-12.63	2.21	0.09	150	-4.79	-5.16

N/A = not computed since the lognormal hypothesis is rejected.

When converted to the actual received power using the result of section 4.2.6, the span becomes approximately -32 to -16 dBm. When $\theta=0^{\circ}$, $\hat{P}(1, \theta)$ is equal to -0.08 dB prior to conversion, and is approximately -16 dBm after conversion. For this 0° case, the received power calculation at 1 m distance is tried using the Friis free space equation (despite the possibility that 1 meter distance lies in the near-field region of the transmit or receive antenna) as follows [7]

$$P(1) = P_T + G_T + G_R + 20\log\left(\frac{\lambda}{4\pi}\right) \quad (5.15)$$

where $P(1)$ is in dBm, P_T is the 18 dBm transmitted power, G_T and G_R respectively represent the 15 dB transmit and 11 dB receive antenna gains, and λ denotes the radio wavelength approximately equal to 1 cm for the 29.5 GHz frequency. From this exercise, $P(1)$ is found to

equal -18 dBm, which differs only by 2 dB from the -16 dBm result of the regression analysis.

- f. The standard deviation estimates of power variation over various locations and pointing angles range between 0.78 and 3.49 dB.³ The smallest of them being at 0° angle has been expected since at 0° reception the main beams of the transmit and receive antennas point to one another such that the majority of the received power is delivered to the receive antenna through the direct path which is merely affected by the propagation distance but not by the multipath structure of the experimental sites. The largest standard deviations occur for $\pm 6^\circ$ angles because, in addition to the spatial variation, the other, uncontrolled independent variables also include the small variation in the receive antenna pointing accuracy that induces a great variation in the received power at these reception angles, as has also been pointed out in section 5.4.
- g. While for several non-zero reception angles the hypothesis that the random error $\varepsilon(\theta)$ in dB is normally distributed with zero mean and $\hat{\sigma}_{\varepsilon(\theta)}^2$ variance, i.e. $\varepsilon(\theta) \sim N(0, \hat{\sigma}_{\varepsilon(\theta)})$, is rejected, the same hypothesis could not be rejected for $\theta=0^\circ$. This implies that the power fluctuation induced by spatially random multipath and partial shadowing in an indoor line-of-sight setting is lognormally distributed for the case of aligned antennas. The same statement, however, does not always apply for indoor links with misaligned antennas.
- h. The power margin to account for the spatial variations [20] is found experimentally to span from 2.50 through 8.14 dB to yield 99% reliability for the specified pointing angles, as referred to in table 5.2 as the 1st percentiles of the experimental error distributions.
- i. A power margin for spatial variations of local mean can as well be found for a specified reliability from the theoretical lognormal distribution [39] at pointing angles for which the lognormal hypothesis is not rejected. Assuming target reliability of 99%, the power margin is simply the absolute value of the 1st percentile of a zero-mean $\hat{\sigma}_{\varepsilon(\theta)}^2$ -variance Gaussian distribution. The power margin $PM(\theta)$ for reception angle θ is hence found to be

---

3. These results, as well as later results on the power margin to cover the spatial variation, are specific to measurements in which a directional-to-directional antenna configuration was realized in any case under a line-of-sight condition.

$$PM(\theta) = 2.33 \hat{\sigma}_\varepsilon(\theta) \quad (5.16)$$

Applying the above expression to the result for each pointing angle, the power margin is discovered to range from 1.81 dB to 8.14 dB. This result is given in table 5.2.

- j. As seen in figure 5.15, the results of linear curve fitting do not seem to show any symmetry around zero pointing angle. The asymmetry can be explained by the fact that the experimental environments are not symmetrical around the direct LOS path.

Conjectures on Causes of the Low Propagation Exponent and Their Verifications

One plausible cause of the phenomenon described in point (c) above is explained as follows. In retrospect, many of the measurement links were set up 190 cm above the floor, fairly close to such structures as the 260 cm high room ceiling and workstations on desks which are 130 cm tall (see, for example, figure 4.5 that demonstrates such a situation). Considering the vertical 3 dB-beamwidth of the antennas (30° for the horn and 15° for the array) and the grazing angle of signal incidence on the structures' surfaces, especially for longer distances, these structures may serve as good reflectors, producing strong multiple paths which add up vectorially at the receiver. As has also been surmised by Xia *et al* in [40], a propagation exponent of less than two can be expected in this situation due to the increasing magnitude of reflection coefficient with the increasing distance, such that at larger distances the vectorial summation will give a greater impact on the received power. In comparison, a two-ray model simulation and its empirical verification reported in [40] yield a propagation exponent for various types of outdoor line-of-sight environments in the 0.8-1.6 range at distances closer than a break point, which corresponds to the distance at which the first Fresnel zone touches the ground. For our case, assuming the worst-case condition (i.e. the closest structure, which is a workstation set up on a desk, lies 60 cm from the LOS line at around the halfway from the transmitter to the receiver), the break distance is found to be 144 meters, far larger than ranges taken in our measurements, using the following expression [40]

$$\begin{aligned}
 d_b &= \frac{1}{\lambda} \sqrt{(\Sigma^2 - \Delta^2)^2 - 2(\Sigma^2 + \Delta^2)\left(\frac{\lambda}{2}\right)^2 + \left(\frac{\lambda}{2}\right)^4} \\
 \Sigma &= h_1 + h_2 \\
 \Delta &= h_1 - h_2
 \end{aligned} \tag{5.17}$$

where d_b is the break distance in meters, h_1 and h_2 are the transmit and receive antenna heights, respectively, which are both 0.6 meter in this case, and λ is the radio frequency wavelength, which is approximately 0.01 meter for the 29.5 GHz frequency. For higher antennas or more distant reflecting structures, larger break distances will result. To verify this suggested explanation, two types of computer simulations have been performed, one for deterministic cases and one other for cases with random antenna height. The results for the deterministic cases follow.

- a. As a start, a simulation is performed based on the very basic, two-ray model, similar to that used in reference [40], with an isotropic antenna at both sides installed at a fixed height of 1.9 m from the floor. The resultant pattern with its two-segment regression curve is very similar to that presented in [40] and is depicted in figure 5.17(a) with the values of predicted break point (d_b , or BP in the figure), and near- and far-region propagation exponents (n_1 and n_2) printed above the box. Since both antennas are horizontally polarized, the reflection coefficient is computed using [40]

$$\Gamma(\phi) = \frac{\cos(90^\circ - \phi) - a\sqrt{\epsilon_r - \sin(90^\circ - \phi)^2}}{\cos(90^\circ - \phi) + a\sqrt{\epsilon_r - \sin(90^\circ - \phi)^2}} \tag{5.18}$$

where ϕ is the angle of wave incidence on the reflector surface, ϵ_r is the relative permittivity of the reflector (which is taken to be 6.5 in the simulation, based on the measured refractive index of concrete at 57.5 GHz [41]), and a is 1 for a horizontal reflector or $1/\epsilon_r$ for a vertical reflector.

- b. Three-ray model is used with a fixed floor at the same position as above and a fixed ceiling at 0.7 m above the antenna height. Three segments are identified, separated in between by two break points corresponding to the ceiling and floor reflections. The result is shown in figure 5.17(b).

Figure 5.17: Results of simulations for deterministic settings

- c. Three-ray model is assumed as before, with the antennas set to resemble the experimental antennas. The result is shown in figure 5.17(c).
- d. Five-ray model is assumed by adding two side walls located at 4 meter distance to the right and left of the radio link. Figure 5.17(d) demonstrates the result.

From these simulations we arrive at some conclusions as follows. First, the average received

Figure 5.18: Histograms of the simulation results for the case of randomized floor and walls.

power does show a distance exponent smaller than two for short distances. Second, the experimental antenna pattern (which, for the receive antenna, has two side lobes on the elevation plane pointing toward the floor and the ceiling, see figure D.2) results in reduced exponent. Third, a larger number of multiple reflections taken into account decreases the propagation exponent.

It should be noted, however, that in the experiment, measurements were taken at various positions and in various rooms. This leads to the need for a simulation with the positions of the "floor" and the walls being randomized. The setting is essentially the same as that for the five-ray model described in point (d) above, except for the randomizing. The "floor" distance from the antenna level ranges from 0.6 to 1.9 meters, whereas the walls are located between 3 and 6 meters beside the radio link. Additionally, the propagation distances, the actual number of which is 30, are made random in the range of the actual experimental distances (3–10 meters). Two numbers are assumed, 30 and 100, of the data points (i.e. distances) for each run, while the total number of simulation runs is 10,000 for each of the above-mentioned number of points. The results are shown in the form of histograms in figure 5.18. It can be concluded that a smaller number of data

Figure 5.19: Simulated received power for the distance range of 3–10 meters, with randomly varying positions of floor and walls.

points produces a larger variation of the propagation exponent. For a 30-point linear curve fitting, an exponent value of one (as can be seen in figure 5.18(a)) is not impossible yet is not very likely to occur either.

Finally a conjecture can be made that the low measured exponent may have been the result of the manner in which the actual measurements were done. During the experiments, the 0° reference direction was determined by adjusting the position and orientation of the transmitter cart. This may have unintentionally placed the receiver at the location of one of the crests illustrated in figure 5.17(a) on any occasion. To check this possibility, a plot shown in figure 5.19 is made so as to provide a visual inspection of the slope of the dashed, straight line connecting the peaks on the received power curve within the range of 3–10 meters. It can be seen from comparing the values on this straight line at 3 and 6 meters, that a twofold increase in distance results in a decrease in received power peak of approximately 4 dB. This means that if the measured powers are located at these peaks, the resultant propagation exponent is $4/(10 \times \log 2) = 1.33$, which is far below two.

5.7 Link Budget Analysis

This section seeks to exemplify link budget analysis by making use of some parameters found from previous discussions. Several attempts have also been made in the past to perform such calculation for mm-wave band [12], [20]. Therefore, the result sought in this section is aimed to provide a comparison to them, pertinent especially to practical application of results specific to circumstances of our experiments. Values of parameters required in the following calculation but missing from our previous results, such as the temporal fade margin, will be adopted from reference [12] for it made use of the same frequency band, binary symbol rate and indoor environments.

The link budget applies for one-room indoor line-of-sight communication systems that utilize directional antennas of the type used in the experiment at both ends of every base-user link. Several assumptions are used in the calculation. An input noise temperature of 300 K, a data rate of 200 Mbps, a receiver noise figure of 6 dB, an antenna temperature of 300 K, a maximum error rate requirement of 10^{-3} for differentially-detected binary phase shift keying, and a maximum coverage range of 10 m are assumed. The effective noise temperature T_r of the receiver (assumed to incorporate the transmission line and the chain of receiving equipment up to the point before the detector) can be obtained from its noise figure F_r of 6 dB as follows [42]

$$\begin{aligned} T_r &= (F_r - 1)T \\ &= 894 \text{ K} \end{aligned}$$

where T is the reference noise temperature of 300 K. Summing the temperatures of the antenna and the receiver [42], the overall system effective temperature is found to equal 1194 K. As a result, the overall system noise figure is $F_s = 1 + T_s/T = 1 + 1194/300 = 5$, or equivalently 7 dB.

Since there is always a possibility of antenna misalignment for any directional-to-directional antenna setting, particularly because the base station is fixed, a power margin is included to account for losses due to such misalignment. Assuming a maximum antenna pointing error of $\pm 6^\circ$, the power margin of 6 dB is taken from the largest possible loss, which as seen in figure 5.11 occurs at 6° pointing angle. A 13 dB short-term fading power margin [12] is also included to protect the link from time-varying multipath and short-time blocking due to people

movement in the room under consideration. Also, as has been pointed out in section 5.6, another power margin is required for protection against spatial variation of received power. For this parameter, an 8 dB margin has been chosen on the basis of the 1st percentile of the experimental CDF associated with the 6° reception as can be seen in table 5.2. This guarantees the protection even when as large an antenna mispointing as 6° happens.

Finally, a distance exponent of 0.93 is assumed in the propagation path loss calculation, which excludes the effects of transmit and receive antenna gains. This entry is obtained by computing $(\lambda/(4\pi))^2 d^{-n}$ [20], where λ denotes the radio wavelength, approximately 1 cm at 29.5 GHz, d denotes the reception range, 10 meters being the maximum, and $n=0.93$ is the distance exponent. The total gain provided by the directional antennas of the type employed in our experiment is 26 dB (G_T = transmit antenna gain = 15 dBi, and G_R = receive antenna gain = 11 dBi). This value is thus used in the following calculation. The link budget calculation is detailed in table 5.3 given below.

Table 5.3: Link Budget Calculation

Noise power spectral density N_o ($k=1.38 \times 10^{-23}$ J/K, $T=300$ K)	-174	dBm/Hz
Required E_b/N_o (2-DPSK, $P_e=10^{-4}$)	9	dB
Symbol rate $R=200$ MHz	83	dBHz
Receiver noise figure (including the antenna temperature, see text)	7	dB
Required received power	-75	dBm
Path loss ($d=10$ m, $n=0.93$, excluding the antenna gains)	71	dB
Directional antenna gains ($G_T G_R$)	-26	dBi
Short-term fading [12]	13	dB
Spatial variation (see section 5.6)	8	dB
Antenna misalignment (see section 5.4)	6	dB
Total power margin	27	dB
Required transmitted power	-3	dBm

As a comparison to the result given in the table, the same calculation that uses a propagation exponent n of 2 is performed and results in a required transmit power of 7 dBm to achieve the same performance.

Chapter 6

Conclusions

6.1 Thesis Summary

6.1.1 Summary of Experiment

The work reported in this thesis sought to investigate the characteristics of broadband millimeter-wave indoor radio channels when at both the transmitter and the receiver a directional antenna is used. In addition, the research was also intended to study the effect of the variation in reception angle on the channel characteristics.

In attempting to reach these goals, experimental activities were conducted in several buildings in the campus of Carleton University. The experiments were performed by repetitively generating a 1023-bit pseudonoise sequence at 200 Mbps rate, and transmitting it over a 29.5 GHz carrierwave through a biphase modulation process. The signal was transmitted from a 30°-beamwidth horn antenna, and was received by a programmable phased-array receive antenna of 8° beamwidth. Following the *I-Q* demodulation process at the receiver, the resultant complex envelope was sampled and stored. Subsequently, the sampled complex envelope was crosscorrelated with a replica of the transmitted sequence, resulting in an estimate of the lowpass equivalent of the radio channel complex impulse response.

Field measurements were accomplished by acquiring the complex envelope samples and the received power for each of eleven reception angles programmed on the receive antenna controller, that range from -30° to 30° with 6° spacings in between. As many as four rooms were selected as experimental sites, one of which was a hallway while the rest were of rectangular shape with similar size.

The following are the results of analyses on the acquired collection of radio channel impulse response estimates. These results, in general, may be extended to other measurement systems and environments of the type similar to those used in our work. The numerical results, however, are unique to the systems and indoor environments with precisely the same parameters as those of ours.

6.1.2 Summary of Results

a. Measurement Results

A total of 40 sets of complex envelope and received power data per reception angle were collected from the entire measurement activities in all of the selected rooms. After some calibration and crosscorrelation procedures, the desired impulse response estimates resulted. From observations over these response estimates and received powers, several points were discovered. First, for any reception angle, the signal component arriving from the line-of-sight path is the highest in power contribution among all distinct components detected by the receiver. Also, components that arrive at the receiver at excess delays equivalent to even multiples of the room dimension along which the measurement link is stretched, referred to as the dominant dimension, commonly appear in the measured impulse responses of various reception directions. Lastly, the received power in any case is the highest when the receive antenna is pointed exactly toward the transmit antenna.

b. Delay Spread Distribution

When the maximum and rms delay spread cumulative distribution functions were concerned, several observations were made. Over all measurements it was found that the 99th percentiles of the maximum and rms delay spreads are respectively 393 ns and 60 ns. This implies that an observation window of 400 ns long is adequate in order to capture the entire duration of the radio channel impulse response of interest in 99% of all cases.

When a separate cumulative distribution function was computed for each room, it was found that a large proportion of the entire maximum delay spread collection are located near values equivalent to even multiples of the room dimension approximately parallel to the measurement link. It can be inferred from this finding that propagation over even-order reflection paths along the room dimension parallel or approximately so to the experimental link (i.e. the previously-defined dominant dimension) is one of the dominant modes of indoor radio propagation. A very simplified calculation of the theoretical magnitude of such even-order reflections tends to be in favor of the conjecture. It was also observed that the larger is this dimension of the room, the more concentrated is the maximum delay spread distribution to smaller values relative to the dimension.

An observation over the distribution of the maximum and rms delay spreads at different directions of reception shows that while the delay spread parameters vary from one reception angle to another when results for any individual experimental position are concerned, in overall the distribution curves for all individual angles are very similar to each other. This has as well been confirmed through a Kolmogorov-Smirnov test on the data measured with a 30 dB threshold applied on each impulse response estimate. Another significant finding is that the 0° reception does not necessarily produce the shortest delay spreads, which perhaps arises from the presence of strong even-order reflections that arrive through the receive antenna pattern main lobe.

c. Angle-Dependent Received Power

From an observation on the average ratio of received power at a certain angle to that at the zero reception angle, it was found that on average an interfering signal of the same distance and power as the desired signal is suppressed by 10 dB below the desired so long as the interfering source is located far outside the region illuminated by the receive antenna main lobe, e.g. making an azimuthal angle of 12° or more with the zero angle line from which the desired signal arrives in our case of 8° beamwidth receive antenna. It was also found, however, that using the same directional antenna, a power loss of as high as 6 dB may occur when a 6° misalignment between the transmit and the receive antennas exists.

The above result also shows that if an angle diversity is to be implemented using multiple antennas of the same type as above, each of which is directed to a different direction, there will be a large difference between the average powers received by different antenna branches, which may reduce the effectiveness of the angle diversity technique in combating fading, as has been pointed out by Licht [12].

d. Angle Correlation Function

When the quantification of variations of the radio channel impulse response due to changes in direction of reception in various experimental rooms was attempted through the autocorrelation functions of the response estimates, and when the results for all rooms were averaged, it was found that signal components received at the same excess delay on two different antennas separated by an angle of 6° or more show a correlation of 0.1 at most. This result

suggests that using 8° -beamwidth antennas, like the one used in our system, as angle diversity branches, a separation among branches of as small as 6° in pointing angle is sufficient to obtain an effective use of the angle diversity technique. This possibility is, however, made less useful by the large difference between the received power at zero angle and that at any other angle, as has been mentioned previously.

e. Power-Distance Relation

The regression analysis operated on all the results of each reception angle demonstrated that the propagation exponent is invariably less than two. For large nonzero angles the low exponent may be the result of low received powers falling below the noise floor of the measurement system, for which the noise floor level is taken as the detected power level. The zero reception angle that produces the highest received powers among all angles, however, results in an exponent of 0.93, much lower than the expected near-freespace exponent. Conjectured causes of this phenomenon are the very short measurement distances, the too small number of measured distances, and the manner of measurement that may have captured only the peaks of the distance-varying received power. These conjectures have been verified through simulations to possibly be the real causes.

It was as well found from the subsequent distribution-fit analysis that for some of the reception angles, including 0° , the variation of received power around its distance-dependent value estimated from the regression is lognormally distributed with a mean of zero and a variance equal to the sample variance of the measurements. This variation is attributed to changes in the multipath structure as the measurement link changes in position. For 0° , the power margin required to protect the link against this spatial variation is 2.5 dB. However, in the subsequent link budget calculation, a power margin of 8 dB is used to account for the spatial variation considering the possibility of a misalignment of 6° between the two antennas.

6.2 Recommendations for Future Work

a. Measurement System Improvement

In future work that involves the measurement system used in this research, it should

always be kept in mind the presence of a large grating lobe in the receive array antenna radiation pattern for pointing angles larger than 15° in magnitude. However, an antenna replacement that exhibits no grating lobe is recommended, since it eliminates any uncertainty on the measurement results that may arise for the above pointing angles.

The present system is limited in its storage capability such that it is only able to collect measurements of eleven reception angles at a time. A larger storage capacity is therefore needed when more reception angles are considered for the future experiment. The use of a better apparatus for power measurement, which imposes lower noise floor, should also be considered in the future.

b. Further Study

Learning from the measurement results that one of the dominant propagation modes inside a rectangular room is the even-order reflections approximately parallel to the radio line-of-sight link, it is perhaps useful to undertake a similar measurement with two antennas, one of which is in approximately back-to-back position relative to the other that is aligned to the transmit antenna. With the first antenna facing the wall behind the latter, aligned antenna, the first antenna can be expected to receive stronger, odd-order reflections, which may result in a more effective use of the angle diversity technique. Also, since the experiments were performed only in rooms of rectangular shape, it may be desirable to learn about the channel characteristics in rooms of other shapes, such as lobbies which usually are irregularly shaped.

In addition, a further study is perhaps needed to clarify two points left unclear from the analyses in this work. The first is the cause of smallest delay spreads that occur for reception angles $\pm 18^\circ$. It should be made certain whether this phenomenon is induced by the antenna characteristic, as has been conjectured, or not. Also, it is probably interesting to further pursue the explanation of the low propagation exponent found in this research. A second set of measurements using the same system in the same rooms are perhaps required to verify the above result.

List of References

- [1] E. Ayanoglu, K.Y. Eng and M.J. Karol, "Wireless ATM: Limits, Challenges and Proposals", *IEEE Personal Communications Magazine*, pp. 18-35, August 1996.
- [2] R.O. LaMaire, A. Krishna, P. Bhagwat and J. Panian, "Wireless LANs and Mobile Networking: Standards and Future Directions", *IEEE Communications Magazine*, pp. 86-94, August 1996.
- [3] D. Falconer, "A System Architecture for Broadband Millimeter-Wave Access to an ATM LAN", *IEEE Personal Communications Magazine*, pp. 36-41, August 1996.
- [4] D. Buchholz, P. Odlyzko, M. Taylor and R. White, "Wireless In-Building Network Architecture and Protocols", *IEEE Network Magazine*, pp. 31-38, November 1991.
- [5] H. Hashemi, "The Indoor Radio Propagation Channel", *Proceedings of the IEEE*, vol. 81, no. 7, pp. 943-968, July 1993.
- [6] J.G. Proakis, *Digital Communications*, Third Edition, McGraw-Hill, 1995.
- [7] T.S. Rappaport, *Wireless Communications Principles and Practice*, Prentice-Hall, 1996.
- [8] W.C. Jakes, *Microwave Mobile Communications*, Wiley, 1974.
- [9] D.G. Brennan, "Linear Diversity Combining Techniques", *Proceedings of the IRE*, pp. 1075-1102, June 1959.
- [10] W.C.-Y. Lee, "Effects on Correlation Between Two Mobile Radio Base-Station Antennas", *IEEE Transactions on Communications*, vol. COM-21, no. 11, pp. 1214-1224, November 1973.
- [11] K. Pahlavan and A. Levesque, *Wireless Information Networks*, Wiley-Interscience, 1995.
- [12] U. Licht, *Characterization of 29 GHz Indoor Radio Channels Using a 2-Channel Angle*

Diversity System, MEng Thesis, Department of Systems and Computer Engineering, Carleton University, 1996.

- [13] D. Tholl, M. Fattouche, R.J.C. Bultitude, P. Melançon and H. Zaghloul, "A Comparison of Two Radio Propagation Channel Impulse Response Determination Techniques", *IEEE Transactions on Antennas and Propagation*, vol.41, no. 4, pp. 515-517, April 1993.
- [14] J.G. Proakis and D.G. Manolakis, *Digital Signal Processing: Principles, Algorithms, and Applications*, Prentice-Hall, 1996.
- [15] I. Miller and J.E. Freund, *Probability and Statistics for Engineers*, Second Edition, Prentice-Hall, 1977.
- [16] W.J. Dixon and F.J. Massey, Jr., *Introduction to Statistical Analysis*, 3rd edition, McGraw-Hill, 1969.
- [17] G.A. Kalivas, M.El-Tanany and S.A. Mahmoud, "Millimeter-Wave Channel Measurements for Indoor Wireless Communications", *42nd IEEE Vehicular Technology Conference*, VTC'92, pp. 609-612, May 1992.
- [18] G.A. Kalivas, M. El-Tanany and S. Mahmoud, "Millimeter-Wave Channel Measurements with Space Diversity for Indoor Wireless Communications", *IEEE Transactions on Vehicular Technology*, vol. 44, no. 3, pp. 494-505, August 1995.
- [19] A.R. Tharek and J.P. McGeehan, "Indoor Propagation and Bit Error Rate Measurements at 60 GHz Using Phase-Locked Oscillators", *38th IEEE Vehicular Technology Conference*, VTC'88, pp.127-133, May 1988.
- [20] R.J.C. Bultitude, R.F. Hahn and R.J. Davies, "Propagation Considerations for the Design of an Indoor Broadband Communications System at EHF", *Proceedings of Wireless '95*, Calgary, pp. 637-656, July 1995.
- [21] P.F.M. Smulders and A.G. Wagemans, "Millimeter-Wave Biconical Horn Antennas for Near Uniform Coverage in Indoor Picocells", *Electronic Letters*, vol. 28, no. 7, pp. 679-680,

March 1992.

- [22] P.F.M. Smulders and A.G. Wagemans, "Frequency-Domain Measurement of the Millimeter Wave Indoor Radio Channel", *IEEE Transactions on Instrumentation and Measurement*, vol. 44, no. 6, pp. 1017-1022, December 1995.
- [23] P.F.M. Smulders and A.G. Wagemans, "Wideband Indoor Radio Propagation Measurements at 58 GHz", *Electronic Letters*, vol. 28, no. 13, pp. 1270-1271, June 1992.
- [24] L. Talbi and G.Y. Delisle, "Experimental Characterization of EHF Multipath Indoor Radio Channels", *IEEE Journal on Selected Areas in Communications*, vol. 14, no. 3, pp. 431-440, April 1996.
- [25] R. Davies, M. Bensebti, M.A. Beach and J.P. McGeehan, "Wireless Propagation Measurements in Indoor Multipath Environments at 1.7GHz and 60GHz for Small Cell Systems", *41st IEEE Vehicular Technology Conference, VTC'91*, St. Louis, pp. 589-593, May 1991.
- [26] A. Kajiwar, "Effects of Polarization, Antenna Directivity, and Room Size on Delay Spread in LOS Indoor Radio Channel", *IEEE Transactions on Vehicular Technology*, vol. 46, no. 1, pp. 169-175, February 1997.
- [27] P.F.M. Smulders and G.J.A.P. Vervuurt, "Influence of Antenna Radiation Patterns on Mm-Wave Indoor Radio Channels", *International Conference on Universal and Personal Communications, ICUPC'93*, 1993.
- [28] T. Manabe, Y. Miura and T. Ihara, "Effects of Antenna Directivity and Polarization on Indoor Multipath Propagation Characteristics at 60 GHz", *IEEE Journal on Selected Areas in Communications*, vol. 14, no. 3, pp. 441-448, April 1996.
- [29] S. Guerin, "Indoor Wideband and Narrowband Propagation Measurements Around 60.5 GHz in An Empty and Furnished Room", *IEEE 46th Vehicular Technology Conference*, Atlanta, Georgia, pp. 160-164, April 28-May 1, 1996.

- [30] P.A. Matthews and A.H. Abubakar, "Direction of Arrival of Radio Signals Within Buildings", *8th International Conference on Antennas and Propagation*, March 1993.
- [31] J.-G. Wang, A.S. Mohan and T.A. Aubrey, "Angles-of-Arrival of Multipath Signals in Indoor Environments", *IEEE 46th Vehicular Technology Conference*, Atlanta, Georgia, pp. 155-159, April 28-May 1, 1996.
- [32] S.Y. Seidel and H.W. Arnold, "Propagation Measurements at 28 GHz to Investigate the Performance of Local Multipoint Distribution Service (LMDS)", *Globecom'95*, Singapore, pp. 754-757, November 1995.
- [33] T. Manabe, K. Sato, H. Masuzawa, K. Taira, T. Ihara, Y. Kasashima and K. Yamaki, "Polarization Dependence of Multipath Propagation and High Speed Transmission Characteristics of Indoor Millimeter-Wave Channel at 60 GHz", *IEEE Transactions on Vehicular Technology*, vol. 44, no.2, pp. 268-273, May 1995.
- [34] Y. Roy, *Measurement Setup for Complex Envelope Measurements at mm-Wave*, laboratory notes.
- [35] S. Damphousse, M. Lecours, M. Pelletier, N. Hassaine and G.Y. Delisle, *Technical Report: A 29 GHz Discrete Element Adaptive Phased Array Antenna System*, Radio Communication and Signal Processing Laboratory, Laval University, 1995.
- [36] B. Glance and L.J. Greenstein, "Frequency-Selective Fading Effects in Digital Mobile Radio with Diversity Combining", *IEEE Transactions on Communications*, vol. COM-31, no. 9, pp. 1085-1094, September 1983.
- [37] J.C.-I. Chuang, "The Effects of Time Delay Spread on Portable Radio Communications Channels with Digital Modulation", *IEEE Journal on Selected Areas in Communications*, vol. SAC-5, no. 5, pp. 879-889, June 1987.
- [38] H. Hashemi, "Impulse Response Modeling of Indoor Radio Propagation Channels", *IEEE Journal on Selected Areas in Communications*, vol 11, no. 7, pp. 967-977, September 1993.

- [39] R. Steele, *Mobile Radio Communications*, Pentech Press, London, 1992.
- [40] H.H. Xia, H.L. Bertoni, L.R. Maciel, A. Lindsey-Stewart and R. Rowe, "Radio Propagation Characteristics for Line-of-Sight Microcellular and Personal Communications", *IEEE Transactions on Antennas and Propagation*, vol. 41, no. 10, pp. 1439-1447, October 1993.
- [41] K. Sato, T. Manabe, J. Polivka, T. Ihara, Y. Kasashima and K. Yamaki, "Measurement of the Complex Refractive Index of Concrete at 57.5 GHz", *IEEE Transactions on Antennas and Propagation*, vol. 44, no. 1, pp. 35-40, January 1996.
- [42] B. Sklar, *Digital Communications Fundamentals and Applications*, Prentice Hall, 1988.

Appendix A

Floor Plans of Measurement Locations

Figure A.1: Floor plan of the 4th floor of Carleton University's Minto CASE Building, with shaded areas indicating experimental sites.

Figure A.2: Floor plan of the 4th floor of Carleton University's Architecture Building, with the shaded area indicating the room chosen as the experimental site.

Appendix B

Photographs of Measurement Setup

Figure B.1: Transmitter cart that carries the PC that controls the whole measurement process, the transmitting subsystems (including the modulator and the upconverter), the receiver's demodulator, the data acquisition subsystems, and the transmit horn antenna.

Figure B.2: Receiver cart that carries the receive phased-array antenna, the PC that controls the antenna phase-shifters, and the receiver's downconverter.

Appendix C

Calibration Curve

Figure C.1: Calibration curve for power measurement. The vertical axis shows the reading from power meter #2 (see figure 4.2(b)), whereas the horizontal shows the associated actual power level detected at the output of the receive antenna (the input to the receiver's RF stage) relative to a reference level of -16 dBm.

Appendix D

Antenna Radiation Patterns

(a)

(b)

Figure D.1: E plane of the receive array antenna radiation pattern, at pointing angles of (a) 0° , (b) 10° , and (c) 20° .

Figure D.1: E plane of the receive array antenna radiation pattern, at pointing angles of (a) 0° , (b) 10° , and (c) 20° (Continued).

Figure D.2: H plane of the receive array antenna radiation pattern for pointing angles of 0° and 30° .

Figure D.3: E plane of the transmit horn antenna radiation pattern [12].

Figure D.4: H plane of the transmit horn radiation pattern [12].

Appendix E

Data Analysis Program Listings

E.1 Introduction

In the following, programs are listed that were used in the processing and analytical steps of the experimental data. There are mainly six types of these programs:

- a. group of programs that treat the raw data and convert them into impulse response estimates and actual received powers.
- b. program used in the temporal variation analysis.
- c. group of programs invoked to compute the delay spread measures (the rms and maximum values) and to construct their experimental cumulative distribution functions.
- d. program that computes the angle-dependent received power ratio as is used in the analysis of different-angle interference.
- e. group of programs that produce the angle correlation functions.
- f. group of programs that perform regression analyses for power-distance relation.

E.2 Programs for Data Processing

a. prodisk.m

```
%PRODISK.M
%
%   is used to process MAT files containing oscilloscope raw data.
%   First, the program reads a MAT file from floppy disk.
%   It computes the actual received power (ActualPower) and 401-point (ns)
%   complex cross-correlation between the received complex envelope and
%   a replica of the transmitted sequence (Crosscorr) which is the channel
%   impulse response estimate. Finally, both new functions are saved back
%   in a file in the network with the same name as before, together with
%   two other variables, namely Angls and filename, ready for subsequent
%   processing.
%   (c)1996 Gamantyo Hendrantoro.

clear;
load caltable;
load pn1023;
filedatabase = ['h4050a';'h4050b';'h4050c';'h4050d';'h4050e';'h4050f';...
               'h4038a';'h4038b';'h4038c';'h4038d';'h4038e';'h4038f';...
               'h4038g';'h4038h';'h4038i';'h4038j';'h4038k';'h4038l';...
               'ha507a';'ha507b';'ha507c';'ha507d';'ha507e';'ha507f';...
               'ha507g';'ha507h';'ha507i';'ha507j';'ha507k';'ha507l';...]
```

```

'h4031a','h4031b','h4031c','h4031d','h4031e','h4031f';...
'h4031g','h4031h','h4031i','h4031j'];

for floppy=1:40,

    %%%%%%%%%%% IMPORTING ALL 5 FILES FROM ONE FLOPPY
    xx = input(['Press return when ',filedatabase(floppy,:), ' is ready '], 's');

    for p=1:5,
        do(['!mcopy a:',filedatabase(floppy,:),int2str(p),...
            '.mat /home/gamantyo/matlab/MAT-files']);
    end

    %%%%%%%%%%% PROCESSING 5 MAT FILES ONE BY ONE
    for p=1:5,
        processedfile = [filedatabase(floppy,:),int2str(p),'.mat'];
        disp(['LOADING ',processedfile,'...']);
        eval(['load ',processedfile]);
        Angls = PtrToAngls - 30;
        for m=1:NmbrOfSweeps*NmbrOfStps,
            if (Power(m)<-30) | (Power(m)>-6),
                disp(['Power value at ',int2str(6*m-36),' invalid']);
                disp(['It is ',num2str(Power(m))]);
            end
            for n=2:160,
                if Power(m)>=PowerReading2(n),
                    ActualPower(m) = ReceivedPower(n-1) +...
                        (Power(m)-PowerReading2(n-1))/...
                        (PowerReading2(n)-PowerReading2(n-1)) *...
                        (ReceivedPower(n)-ReceivedPower(n-1));
                    break;
                else
                    ActualPower(m) = -25;
                end
            end
        end
    end

    %%%%%%%%%%% START LOOPING FOR EACH ANGLE VALUE
    for k=1:NmbrOfSweeps*NmbrOfStps,
        disp(['ANGLE : ',int2str(Angls(k))]);
        disp('Converting data to real voltage values...');
        index1 = ((2*k-2)*ScAcqLen+1:(2*k-1)*ScAcqLen);
        index2 = ((2*k-1)*ScAcqLen+1:2*k*ScAcqLen);
        index = ((k-1)*ScAcqLen+1:k*ScAcqLen);
        indexkor = index;

        % REAL PART
        Volt(index)=(ScCh(index1)-ones(ScAcqLen,1)*ScYOrigin(1,1)).*...
            (ones(ScAcqLen,1)*ScYInc(1,1))+...
            ones(ScAcqLen,1)*ScYRef(1,1);
        m=mean(Volt(index));
        Volt0Mean(index)=Volt(index)-ones(1,ScAcqLen)*m;

        % IMAGINARY PART
        Volt(index)=(ScCh(index2)-ones(ScAcqLen,1)*ScYOrigin(1,2)).*...
            (ones(ScAcqLen,1)*ScYInc(1,2))+...
            ones(ScAcqLen,1)*ScYRef(1,2);
        m=mean(Volt(index));
        Volt0Mean2(index)=Volt(index)-ones(1,ScAcqLen)*m;
        clear m;

        Volt0Mean(index) = Volt0Mean(index) + i*Volt0Mean2(index);
        clear Volt0Mean2;

```

```

        disp('Preprocessing is done, now doing the correlation...');

        Voltdummy = Volt0Mean(index)';
        Crosscorr(index) = kor(Input0Mean, Voltdummy);
        disp('Done.');
```

end

```

%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%% END OF LOOP

dummy = [];
MidStep = (NmbrOfStps+1)/2;
[xmax ymax]=max(abs(Crosscorr((MidStep-1)*5115+1:MidStep*5115)));
lower = ymax-50;
upper = lower+430;

for m=1:NmbrOfSweeps*NmbrOfStps,
    dummy = [dummy Crosscorr((m-1)*5115+lower+1:(m-1)*5115+upper+1)];
end;

Crosscorr = dummy;
disp(['Saving ',processedfile,'...']);
eval(['save ',processedfile,' ActualPower Crosscorr Angls']);
end

disp('Done.')
```

do('!eject');

end

E.3 Program for Temporal Variation Analysis

a. *timecorr.m*

```

%TIMECORR.M
%   Produces the time correlation function to be used in temporal variation analysis
%   (c)1996 Gamantyo Hendrantoro.

clear;
fileno = [1 2 3 4 5; 6 7 8 9 10];
angstr = 'mp';

for k=1:2,
    avtimecorr = zeros(11,371);
    for m=1:2,
        ente = input(['Insert floppy that contains g',angstr(k),'15_',...
            int2str(fileno(m,1)),' and press enter'],'s');
        do('!volcheck');
        for n=1:5,
            GoOn=1;
            disp(' ');
            filename = ['g',angstr(k),'15_',int2str(fileno(m,n)),'.mat'];
            do(['!mcopy a:',filename,' /home/gamantyo/matlab/MAT-files']);
            command = ['load ',filename];
            disp(command); eval(command);

            if any(ActualPower == -35), % Received power at noise floor
                GoOn = 0;
                disp(['No correlation function is to be computed for ',filename]);
            else
                disp('Computing correlation function for each excess delay...');
                for b=1:11,
                    index = (b-1)*401-30:b*401-1;
                    complexIR(b,1:371) = Crosscorr(index)/max(abs(Crosscorr(:)));
                end
            end
        end
    end
end

```

```

end

for exdelay=1:371,    %%% M=11, 2*M-1=21
    %%% USE ONLY POSITIVE LAGS (including ZERO LAG)
    timecorr(1:21,exdelay) = xcorr(complexIR(:,exdelay),'biased');
    CmplxTimeCorr(1:11,exdelay) = timecorr(11:21,exdelay);
end
CmplxTimeCorr = CmplxTimeCorr/max(abs(CmplxTimeCorr(1,:)));
avtimecorr = avtimecorr + CmplxTimeCorr;
end
command = ['save ',filename,' ActualPower Angls Crosscorr TimeRef ',...
    'Timer CmplxTimeCorr'];
disp(command); eval(command);
do(['!mdel a:',filename]);
do(['!mcopy /home/gamantyo/matlab/MAT-files/',filename,' a:',filename]);
do(['!rm /home/gamantyo/matlab/MAT-files/',filename]);
end
if m==2, % Save a PDPfile for same angle
    AvTimeCorr = avtimecorr / max(abs(avtimecorr(1,:)));
    AvTimeCorrDel = ( sum(AvTimeCorr') / sum(ceil(abs(AvTimeCorr(1,:)))) )';
    pdpfile = ['pdp_',angstr(k),'15.mat'];
    command = ['save ',pdpfile,' AvTimeCorr AvTimeCorrDel'];
    disp(command); eval(command);
    do(['!mcopy /home/gamantyo/matlab/MAT-files/',pdpfile,' a:',pdpfile]);
    do(['!rm /home/gamantyo/matlab/MAT-files/',pdpfile]);
end
do('!eject');
end
end
end

```

E.4 Program for Delay Spread Distribution Computation

a. *spreadcomp.m*

```

%SPREADCOMP.M
%   Computes rms and maximum delay spreads from individual responses.
%   (c)1996 Gamantyo Hendrantoro.

clear;
filedatabase = ['h4050a','h4050b','h4050c','h4050d','h4050e','h4050f';...
    'h4038a','h4038b','h4038c','h4038d','h4038e','h4038f';...
    'h4038g','h4038h','h4038i','h4038j','h4038k','h4038l';...
    'ha507a','ha507b','ha507c','ha507d','ha507e','ha507f';...
    'ha507g','ha507h','ha507i','ha507j','ha507k','ha507l';...
    'h4031a','h4031b','h4031c','h4031d','h4031e','h4031f';...
    'h4031g','h4031h','h4031i','h4031j'];

%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%% Initialization of Static Delay Spread variables
StatRmsSpread5 = []; StatRmsSpread10 = []; StatRmsSpread15 = [];
StatRmsSpread20 = []; StatRmsSpread25 = []; StatRmsSpread30 = [];
StatRmsSpread35 = []; StatRmsSpread40 = [];
StatSpread99 = []; StatSpread90 = [];
StatMaxSpread5 = []; StatMaxSpread10 = []; StatMaxSpread15 = [];
StatMaxSpread20 = []; StatMaxSpread25 = []; StatMaxSpread30 = [];
StatMaxSpread35 = []; StatMaxSpread40 = [];

for floppy = 1:40, %%%%%%%%%%%%% PROCESSING 40 FILES ONE BY ONE
    filename = filedatabase(floppy,:);

    %%%%%%%%%%%%%%% LOADING DATA FROM 5 FILES

```

```

for k=1:5,
    command = ['load ',filename,int2str(k)];
    disp(command); eval(command);
    complexIR(k,:) = Crosscorr(1,:) / sqrt(sum((abs(Crosscorr(1,:)).^2)/11));
    RxPower(k,:) = ActualPower(1,:);

    for ang=1:11,
        rowindexPDP = (ang-1)*431+30:ang*431-1; % 401 points out of 431
        statpdp1 = (abs(complexIR(k,rowindexPDP))).^2;
        for th=5:5:40,
            statpdp1 = statpdp1/max(statpdp1);
            aboveth = (statpdp1>10^(-th/10));
            %%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%% RMS Delay Spread computation
            sump = sum(statpdp1.*aboveth);
            sumptau = sum(statpdp1.*aboveth.*tau);
            sumptau2 = sum(statpdp1.*aboveth.*tau.^2);
            rms = sqrt(sumptau2/sump - (sumptau/sump)^2);
            command = ['StatRmsSpread',int2str(th),...
                ' = [StatRmsSpread',int2str(th),' rms];'];
            eval(command);
            cump = cumsum(statpdp1.*aboveth);

            %%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%% Maximum Delay Spread computation
            pdp2 = statpdp1.*aboveth;
            for nstart = 1:401,
                if pdp2(nstart) > 0
                    break;
                end
            end
            for nend = 402:-1:2,
                if pdp2(nend-1) > 0,
                    break;
                end
            end
            maxspread = nend - nstart;
            command = ['StatMaxSpread',int2str(th),...
                ' = [StatMaxSpread',int2str(th),' maxspread];'];
            eval(command);
        end;
    end
end

command = ['save HSPREAD.mat StatRmsSpread5 StatRmsSpread10 StatRmsSpread15 ',...
    'StatRmsSpread20 StatRmsSpread25 StatRmsSpread30 StatRmsSpread35 ',...
    'StatRmsSpread40 ',...
    'StatMaxSpread5 StatMaxSpread10 StatMaxSpread15 StatMaxSpread20 ',...
    'StatMaxSpread25 StatMaxSpread30 StatMaxSpread35 StatMaxSpread40 '];
disp(command);
eval(command);

```

b. cdfspread1.m

```

%spreadcdf1.m
%   Computes delay spread complementary cumulative distribution functions
%   (c)1996 Gamantyo Hendrantoro.

close all
%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%% Computing Overall CDFs
clear
plotline=['- ' ;'-.' ;'- - ' ;'- ' ;'-.' ;'- - ' ;'- ' ;'-.' ;'- - '];
resol = 1;
load HSPREAD

```



```

disp('Computing CDFs for Rms Delay Spreads...');
disp('Th      Static Max      Static Rms');
disp('dB  mean  std  max  mean  std  max');
for thr = 5:5:30, % or use HIST function followed by CUMSUM
    for spr = 0:resol:400,
        command = ['cdfStatMax',int2str(thr),' = [cdfStatMax',int2str(thr),...
            ' sum(cut(StatMaxSpread',int2str(thr),' ,spr))/2090];']; %2090=#floppies*55
        eval(command)
        command = ['cdfStatRms',int2str(thr),' = [cdfStatRms',int2str(thr),...
            ' sum(cut(StatRmsSpread',int2str(thr),' ,spr))/2090];'];
        eval(command)
    end
    command = ['aveStatMax = mean(StatMaxSpread',int2str(thr),' ');']; eval(command);
    command = ['aveStatRms = mean(StatRmsSpread',int2str(thr),' ');']; eval(command);
    command = ['stdStatMax = std(StatMaxSpread',int2str(thr),' ');']; eval(command);
    command = ['stdStatRms = std(StatRmsSpread',int2str(thr),' ');']; eval(command);
    command = ['dumb = sort(StatMaxSpread',int2str(thr),' ');']; eval(command);
    maxStatMax = dumb(ceil(.99*length(dumb)));
    command = ['dumb = sort(StatRmsSpread',int2str(thr),' ');']; eval(command);
    maxStatRms = dumb(ceil(.99*length(dumb)));
    disp([sprintf('%2.0f',thr),sprintf('%6.1f',aveStatMax),...
        sprintf('%6.1f',stdStatMax),sprintf('%6.1f',maxStatMax),...
        sprintf('%7.1f',aveStatRms),sprintf('%6.1f',stdStatRms),...
        sprintf('%6.1f',maxStatRms)]);
end

disp('Plotting CDF graphs...');
for thr = 5:5:30,
    figure(1); hold on
    command = ['plot(0:resol:400,cdfStatMax',int2str(thr),' ,plotline(thr/5,:))'];
    eval(command)
    figure(2); hold on
    command = ['plot(0:resol:400,cdfStatRms',int2str(thr),' ,plotline(thr/5,:))'];
    eval(command)
end

figure(1); hold off; axis([0 400 0 1]); grid
ylabel('P(Max. Spread > abscissa)'); xlabel('Maximum Spread (ns)');
title('CDF of Static Maximum Delay Spread');
legend('-5dB',' -10dB',' -15dB',' -20dB',' -25dB',' -30dB');

figure(2); hold off; axis([0 100 0 1]); grid
ylabel('P(Rms Spread > abscissa)'); xlabel('Rms Spread (ns)');
title('CDF of Static Rms Delay Spread');
legend('-5dB',' -10dB',' -15dB',' -20dB',' -25dB',' -30dB');

```

c. *cdfspread2.m*

```

%CDFSPREAD2.M
%   Computes CDFs for individual rooms
%   (c)1996 Gamantyo Hendrantoro.

clear
plotline=['- ' ;'-.' ;'--' ;': ' ;'- ' ;'-.' ;'--' ;': ' ;'- ' ;'-.' ;'--' ;': '];
resol = 1;
load HSPREAD
NoPosInRoom = [0 4 12 11 10]; NoPosInRoom = cumsum(NoPosInRoom);
RoomName=['4050';'4038';'A507';'4031'];

for a=1:6, figure(a); clf; end
disp('Room Th      Maximum Spread      RMS Spread')
disp('      mean  std  max  mean  std  max')

```

```

for room=1:4,
    for thr = 10:10:30,
        eval(['clear cdfStatMax',int2str(thr)]);
        eval(['clear cdfStatRms',int2str(thr)]);
    end
    for thr = 10:10:30,
        command = ['dataMax = StatMaxSpread',int2str(thr),...
                    '(NoPosInRoom(room)*55+1:NoPosInRoom(room+1)*55);'];
        eval(command)
        command = ['dataRms = StatRmsSpread',int2str(thr),...
                    '(NoPosInRoom(room)*55+1:NoPosInRoom(room+1)*55);'];
        eval(command)
        aveMax = mean(dataMax); stdMax = std(dataMax); maxMax = max(dataMax);
        aveRms = mean(dataRms); stdRms = std(dataRms); maxRms = max(dataRms);
        disp([RoomName(room,:),sprintf('%4.0f',thr),...
              sprintf('%7.1f',aveMax),sprintf('%6.1f',stdMax),sprintf('%6.1f',maxMax),...
              sprintf('%7.1f',aveRms),sprintf('%6.1f',stdRms),sprintf('%6.1f',maxRms)]);

        for spr = 0:resol:400,
            command = ['cdfStatMax',int2str(thr),' = [cdfStatMax',int2str(thr),...
                        ' sum(cut(StatMaxSpread',int2str(thr),...
                        '(NoPosInRoom(room)*55+1:NoPosInRoom(room+1)*55),spr))/',...
                        ' ((NoPosInRoom(room+1)-NoPosInRoom(room))*55)];'];
            eval(command)
            command = ['cdfStatRms',int2str(thr),' = [cdfStatRms',int2str(thr),...
                        ' sum(cut(StatRmsSpread',int2str(thr),...
                        '(NoPosInRoom(room)*55+1:NoPosInRoom(room+1)*55),spr))/',...
                        ' ((NoPosInRoom(room+1)-NoPosInRoom(room))*55)];'];
            eval(command)
        end
    end
end

disp(' ');
disp('Plotting CDF graphs...');
for thr = 20:10:30,
    figure(2*room+5); hold on
    command = ['plot(0:resol:400,cdfStatMax',int2str(thr),' ,plotline(thr/5,:))'];
    eval(command)
    figure(2*room+6); hold on
    command = ['plot(0:resol:400,cdfStatRms',int2str(thr),' ,plotline(thr/5,:))'];
    eval(command)
end

figure(2*room+5); hold off; axis([0 400 0 1]);
ylabel('P(Max. Spread > abscissa)'); xlabel('Maximum Spread (ns)');
title(['CDF of Static Maximum Delay Spread, Room:',RoomName(room,:)]);
legend('-5dB', '-10dB', '-15dB', '-20dB', '-25dB', '-30dB', '-35dB', '-40dB');

figure(2*room+6); hold off; axis([0 100 0 1]);
[x y]=max(cdfStatRms40); y = ceil(y/10)*10; axis([0 y 0 1]);
ylabel('P(Rms Spread > abscissa)'); xlabel('Rms Spread (ns)');
title(['CDF of Static Rms Delay Spread, Room:',RoomName(room,:)]);
legend('-5dB', '-10dB', '-15dB', '-20dB', '-25dB', '-30dB', '-35dB', '-40dB');

figure(1); plot(0:resol:400,cdfStatMax30,plotline(room,:)); hold on;
figure(2); plot(0:resol:400,cdfStatRms30,plotline(room,:)); hold on;
figure(3); plot(0:resol:400,cdfStatMax20,plotline(room,:)); hold on;
figure(4); plot(0:resol:400,cdfStatRms20,plotline(room,:)); hold on;
figure(5); plot(0:resol:400,cdfStatMax10,plotline(room,:)); hold on;
figure(6); plot(0:resol:400,cdfStatRms10,plotline(room,:)); hold on;
end

figure(1); hold off; title('CDF of Static Maximum Delay Spread, threshold:-30dB');
axis([0 410 0 1]);

```

```

xlabel('Maximum Spread (ns)'); ylabel('P(Maximum Spread > abscissa)');
legend('4050','4038','A507','4031');

figure(2); hold off; title('CDF of Static Rms Delay Spread, threshold:-30dB')
axis([0 140 0 1]);
xlabel('Rms Spread (ns)'); ylabel('P(Rms Spread > abscissa)');
legend('4050','4038','A507','4031');

figure(3); hold off; title('CDF of Static Maximum Delay Spread, threshold:-20dB')
axis([0 410 0 1]);
xlabel('Maximum Spread (ns)'); ylabel('P(Maximum Spread > abscissa)');
legend('4050','4038','A507','4031');

figure(4); hold off; title('CDF of Static Rms Delay Spread, threshold:-20dB')
axis([0 140 0 1]);
xlabel('Rms Spread (ns)'); ylabel('P(Rms Spread > abscissa)');
legend('4050','4038','A507','4031');

figure(5); hold off; title('CDF of Static Maximum Delay Spread, threshold:-10dB')
axis([0 410 0 1]);
xlabel('Maximum Spread (ns)'); ylabel('P(Maximum Spread > abscissa)');
legend('4050','4038','A507','4031');

figure(6); hold off; title('CDF of Static Rms Delay Spread, threshold:-10dB')
axis([0 140 0 1]);
xlabel('Rms Spread (ns)'); ylabel('P(Rms Spread > abscissa)');
legend('4050','4038','A507','4031');

```

d. spreadcdf3.m

```

%SPREADCDF3.M
%      Computes CDFs for individual angles
%      (c)1996 Gamantyc Hendrantoro.

clear
plotline=['- ','-.-','--','- ','-.-','--','- ','-.-','--','- ','-.-','--']; resol = 1;
load HSPREAD

for fig=1:26, figure(fig); clf; end
disp('Angle Th      Maximum Spread      RMS Spread')
disp('      mean    std    99%    mean    std    99%')

for angle=1:11,
    for thr = 10:10:20,
        eval(['clear cdfStatMax',int2str(thr)]);
        eval(['clear cdfStatRms',int2str(thr)]);
    end
    for thr = 10:10:20,
        command=['dataMax=StatMaxSpread',int2str(thr),...
            '(angle:11:2090);']; %2090=#floppies*55
        eval(command)
        command = ['dataRms = StatRmsSpread',int2str(thr), '(angle:11:2090);'];
        eval(command)
        aveMax = mean(dataMax); stdMax = std(dataMax);
        dumb = sort(dataMax); maxMax = dumb(ceil(.99*length(dumb)));
        aveRms = mean(dataRms); stdRms = std(dataRms);
        dumb = sort(dataRms); maxRms = dumb(ceil(.99*length(dumb)));
        disp([sprintf('%4.0f',angle*6-36),sprintf('%4.0f',thr),...
            sprintf('%7.1f',aveMax),sprintf('%6.1f',stdMax),sprintf('%6.1f',maxMax),...
            sprintf('%7.1f',aveRms),sprintf('%6.1f',stdRms),sprintf('%6.1f',maxRms)]);
        for spr = 0:resol:400,
            command = ['cdfStatMax',int2str(thr), ' = [cdfStatMax',int2str(thr),...
                ' sum(cut(StatMaxSpread',int2str(thr),...
                '(angle:11:2090),spr))/190];'];

```

```

        eval(command)
        command = ['cdfStatRms',int2str(thr),' = [cdfStatRms',int2str(thr),...
                    ' sum(cut(StatRmsSpread',int2str(thr),...
                    '(angle:11:2090),spr))/190];'];
        eval(command)
    end
end

figure(2*angle+3); clf;
figure(2*angle+4); clf;
for thr = 10:10:20,
    figure(2*angle+3); hold on
    command = ['plot(0:resol:400,cdfStatMax',int2str(thr),'plotline(thr/5,:))'];
    eval(command)
    figure(2*angle+4); hold on
    command = ['plot(0:resol:400,cdfStatRms',int2str(thr),'plotline(thr/5,:))'];
    eval(command)
end

figure(2*angle+3); hold off; axis([0 400 0 1]);
ylabel('P(Max. Spread > abscissa)'); xlabel('Maximum Spread (ns)');
title(['CDF of Static Maximum Delay Spread, Angle:',int2str(angle*6-36)]);
legend('-5dB','-10dB','-15dB','-20db','-25dB','-30dB','-35dB','-40dB');

figure(2*angle+4); hold off; axis([0 100 0 1]);
[x y]=max(cdfStatRms40); y = ceil(y/10)*10; axis([0 y 0 1]);
ylabel('P(Rms Spread > abscissa)'); xlabel('Rms Spread (ns)');
title(['CDF of Static Rms Delay Spread, Angle:',int2str(angle*6-36)]);
legend('-5dB','-10dB','-15dB','-20db','-25dB','-30dB','-35dB','-40dB');

figure(1); plot(0:resol:400,cdfStatMax10,plotline(angle,:)); hold on;
figure(2); plot(0:resol:400,cdfStatRms10,plotline(angle,:)); hold on;
figure(3); plot(0:resol:400,cdfStatMax20,plotline(angle,:)); hold on;
figure(4); plot(0:resol:400,cdfStatRms20,plotline(angle,:)); hold on;
end

figure(1); hold off;
title('CDF of Static Maximum Delay Spread, threshold:-10dB'); axis([0 410 0 1]);
xlabel('Maximum Spread (ns)'); ylabel('P(Maximum Spread > abscissa)');
legend('-30','-24','-18','-12','-6','0','+6','+12','+18','+24','-30');

figure(2); hold off;
title('CDF of Static Rms Delay Spread, threshold:-10dB'); axis([0 100 0 1]);
xlabel('Rms Spread (ns)'); ylabel('P(Rms Spread > abscissa)');
legend('-30','-24','-18','-12','-6','0','+6','+12','+18','+24','+30');

figure(3); hold off;
title('CDF of Static Maximum Delay Spread, threshold:-20dB'); axis([0 410 0 1]);
xlabel('Maximum Spread (ns)'); ylabel('P(Maximum Spread > abscissa)');
legend('-30','-24','-18','-12','-6','0','+6','+12','+18','+24','+30');

figure(4); hold off;
title('CDF of Static Rms Delay Spread, threshold:-20dB'); axis([0 100 0 1]);
xlabel('Rms Spread (ns)'); ylabel('P(Rms Spread > abscissa)');
legend('-30','-24','-18','-12','-6','0','+6','+12','+18','+24','+30');

```

e. spreadcdf4.m

```

%spreadcdf4.m
%       Computes CDFs for individual rooms, normalized to room dimension (length)
%       (c)1996 Gamantyo Hendrantoro.

clear
close all

```



```

legend('4050','4038','A507','4031');

figure(2); hold off; title('CDF of Normalized Rms Delay Spread, threshold:-30dB');
axis([0 1.5 0 1]);
xlabel('Rms Spread (ns)'); ylabel('P(Rms Spread > abscissa)');
legend('4050','4038','A507','4031');

figure(3); hold off; title('CDF of Normalized Maximum Delay Spread, threshold:-20dB');
axis([0 10 0 1]);
xlabel('Maximum Spread (ns)'); ylabel('P(Maximum Spread > abscissa)');
legend('4050','4038','A507','4031');

figure(4); hold off; title('CDF of Normalized Rms Delay Spread, threshold:-20dB');
axis([0 1.5 0 1]);
xlabel('Rms Spread (ns)'); ylabel('P(Rms Spread > abscissa)');
legend('4050','4038','A507','4031');

figure(5); hold off; title('CDF of Normalized Maximum Delay Spread, threshold:-10dB');
axis([0 10 0 1]);
xlabel('Maximum Spread (ns)'); ylabel('P(Maximum Spread > abscissa)');
legend('4050','4038','A507','4031');

figure(6); hold off; title('CDF of Normalized Rms Delay Spread, threshold:-10dB');
axis([0 1.5 0 1]);
xlabel('Rms Spread (ns)'); ylabel('P(Rms Spread > abscissa)');
legend('4050','4038','A507','4031');

```

E.5 Programs for Different-Angle Interference Analysis

a. PxPo.m

```

%PxPo.m
%      - computes ensemble average of Px/Po for any angle x
%      - computes correlation coefficient of Px and Po for any angle x
%      (c)1996 Gamantyo Hendrantoro

clear
filedatabase = ['h4050a','h4050b','h4050c','h4050d','h4050e','h4050f';...
                'h4038a','h4038b','h4038c','h4038d','h4038e','h4038f';...
                'h4038g','h4038h','h4038i','h4038j','h4038k','h4038l';...
                'ha507a','ha507b','ha507c','ha507d','ha507e','ha507f';...
                'ha507g','ha507h','ha507i','ha507j','ha507k','ha507l';...
                'h4031a','h4031b','h4031c','h4031d','h4031e','h4031f';...
                'h4031g','h4031h','h4031i','h4031j'];

PxPo = zeros(1,11);
setPxPo = [];
P = [];

for floppy = 1:40,
    for fileno = 1:5,
        filename = [filedatabase(floppy,:),int2str(fileno)];
        command = ['load ',filename];
        disp(command); eval(command);

        setPxPo = [setPxPo; 10.^((ActualPower-ActualPower(6))/10)];
        PxPo = PxPo + 10.^((ActualPower-ActualPower(6))/10);
        P = [P; 10.^(ActualPower/10)];
    end
end

PxPo = PxPo/PxPo(6); stdPxPo = std(setPxPo);

```

```

figure; plot(-30:6:30,PxPo,'--'); hold on;
for ang=1:11
    if ang~=6
        plot([Angls(ang) Angls(ang)],[PxPo(ang)+stdPxPo(ang) PxPo(ang)-stdPxPo(ang)],'-')
    end
    cc = corrcoef(P(:,[ang 6]));
    ccPxPo(ang) = cc(1,2);
end
axis([-30 30 0 1]); title('Ratio of Received Powers at Angles X and 0 deg.')
ylabel('Power Ratio (linear scale)'); xlabel('Pointing angle (X deg.)')

figure; plot(-30:6:30,ccPxPo,'--'); axis([-30 30 0 1])
title('Correlation coefficient of Received Powers at Angles X and 0 deg.')
ylabel('Correlation Coefficient (linear scale)'); xlabel('Pointing angle (X deg.)')

command = ['save hpowerangle.mat PxPo stdPxPo ccPxPo'];
disp(command); eval(command)

```

E.6 Programs for Angle Correlation Function Computation

a. *ac30.m*

```

%ac30.m
%      Creates complex angular correlation functions of IRs for -30 to 30 degrees
%      one for each file in a 5-file set.
%      (c)1996 Gamantyo Hendrantoro

clear;
%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%% SETTING SOME INITIAL VALUES
room = ['4050','4038','a507','4031'];
filedatabase = ['h4050a','h4050b','h4050c','h4050d','h4050e','h4050f';...
               'h4038a','h4038b','h4038c','h4038d','h4038e','h4038f';...
               'h4038g','h4038h','h4038i','h4038j','h4038k','h4038l';...
               'ha507a','ha507b','ha507c','ha507d','ha507e','ha507f';...
               'ha507g','ha507h','ha507i','ha507j','ha507k','ha507l';...
               'h4031a','h4031b','h4031c','h4031d','h4031e','h4031f';...
               'h4031g','h4031h','h4031i','h4031j'];

%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%% SUMMING CORRELATION FUNCTIONS FROM FLOPPIES
for fileno = 1:38,
    filename = filedatabase(fileno,:);
    for q=1:5,
        GoOn = 1;
        processedfile = [filename,int2str(q),'.mat'];

        command = ['load ',processedfile];
        disp(command); eval(command);

        if any(ActualPower(4:8) == -35), % Received power at noise floor
            GoOn = 0;
            disp(['No correlation function is to be computed for ',processedfile]);
        else
            disp('Computing correlation function for each excess delay...');
            for k=1:11, % only 5 angles without grating lobe are used (-12 to 12)
                m = k;
                index = (m-1)*431+30:m*431-1;
                complexIR(k,1:401) = Crosscorr(index)/max(abs(Crosscorr(:)));
                complexIR(k,1:401) = complexIR(k,1:401) .* ...
                    (abs(complexIR(k,1:401)) > 0.0316);
                complexIR(k,1:401) = complexIR(k,1:401) * 10^(ActualPower(m)/20);
            end
        end
    end
end

```

```

end

for n=1:401,%%% M=11, 2*M-1=21
    %%% USE ONLY POSITIVE LAGS (including ZERO LAG)
    if all(complexIR(:,n) == 0),
        CmplxAngCorr(1:11,n) = zeros(11,1);
    else
        anglecorr(1:21,n) = xcorr(complexIR(:,n),'biased');
        CmplxAngCorr(1:11,n) = anglecorr(11:21,n);
    end
end
CmplxAngCorr = CmplxAngCorr/max(max(abs(CmplxAngCorr)));
end

disp(' ');
if GoOn & exist('CmplxAngCorr2'),
    command = ...
        ['save ',processedfile,' ActualPower Crosscorr Angls CmplxAngCorr ',...
        'CmplxAngCorr2'];
elseif GoOn,
    command = ...
        ['save ',processedfile,' ActualPower Crosscorr Angls CmplxAngCorr'];
elseif exist('CmplxAngCorr2'),
    command = ...
        ['save ',processedfile,' ActualPower Crosscorr Angls CmplxAngCorr2'];
else
    command = ['save ',processedfile,' ActualPower Crosscorr Angls'];
end
disp(command); eval(command);
end
end

```

b. aac30.m

```

%aac30.m
%      Creates angular correlation functions averaged over local area
%      (c)1996 Gamantyo Hendrantoro

clear;
%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%% SETTING SOME INITIAL VALUES
room = ['4050','4038','a507','4031'];
filedatabase = ['h4050a','h4050b','h4050c','h4050d','h4050e','h4050f';...
                'h4038a','h4038b','h4038c','h4038d','h4038e','h4038f';...
                'h4038g','h4038h','h4038i','h4038j','h4038k','h4038l';...
                'ha507a','ha507b','ha507c','ha507d','ha507e','ha507f';...
                'ha507g','ha507h','ha507i','ha507j','ha507k','ha507l';...
                'h4031a','h4031b','h4031c','h4031d','h4031e','h4031f';...
                'h4031g','h4031h','h4031i','h4031j'];

%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%% SUMMING CORRELATION FUNCTIONS FROM FLOPPIES
for fileno = 1:38,
    filename = filedatabase(fileno,:);
    flags = zeros(1,401);
    templat = zeros(11,401);

    for k=1:5, %%% Process all 5 files one by one
        GoOn = 1;
        processedfile = [filename,int2str(k),'.mat'];

        command = ['load ',processedfile];
        disp(command);
        eval(command);

        if ~exist('CmplxAngCorr'), % The file has no angular correlation function 2

```



```

        GoOn = 0;
        disp(['No correlation function #1 is found in ',processedfile]);
    else
        templat = templat + CmplxAngCorr;
        flags = flags - sign(abs(CmplxAngCorr(1,:)));
    end
    clear CmplxAngCorr;
end

dummyflags = flags; %% required because there's another variable of the same
                    %% name from the pdpfile to be loaded
pdpfile = ['PDP',filename(2:6),'.mat'];

clear AverAngCorr2 AverAngCorr;
command = ['load ',pdpfile];
disp(command);
eval(command);

clear AverAngCorr
disp('Computing the average angular correlation function ...');
for k=1:401,
    if ~dummyflags(k), % No correlation functions to be averaged
        AverAngCorr(1:11,k) = zeros(11,1);
    else
        AverAngCorr(1:11,k) = templat(1:11,k)/dummyflags(k);
    end
end
flags = dummyflags;
if ~exist('AverAngCorr2'),
    command = ['save ',pdpfile,' MaxSpread10 MaxSpread20 MaxSpread30 PDP',...
        ' RmsSpread10 RmsSpread20 RmsSpread30 Spread90 Spread99',...
        ' AverAngCorr flags flags2 ;'];
else
    command = ['save ',pdpfile,' MaxSpread10 MaxSpread20 MaxSpread30 PDP',...
        ' RmsSpread10 RmsSpread20 RmsSpread30 Spread90 Spread99',...
        ' AverAngCorr2 AverAngCorr ',...
        ' flags flags2 ;'];
end
disp(command);
eval(command);
end

```

c. *aaca30.m*

```

%aaca30.m
%   Computes the angle correlation function averaged over all local areas.
%   (c)1996 Gamantyo Hendrantoro

clear;

%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%% SETTING SOME INITIAL VALUES
room = ['405C','4038','a507','4031'];

for m=1:4, %% There are 4 rooms
    command = ['Corr',room(m,:), ' = zeros(11,401);'];
    disp(command); eval(command);
    command = ['flag',room(m,:), ' = zeros(1,401);'];
    disp(command); eval(command);
end

filedatabase = ['h4050i','h4050j','h4050k','h4050l';...
    'h4038a','h4038b','h4038c','h4038d','h4038e','h4038f';...
    'h4038g','h4038h','h4038i','h4038j','h4038k','h4038l';...
    'ha507a','ha507b','ha507c','ha507d','ha507e','ha507f';...

```

[illegible]

```
command = ['save correla Corral1 Corr4050 Corr4038 Corr4031 Corra507'];
disp(command); eval(command);
```

E.7 Programs for Power-Distance Relation Analysis

a. *pdstep1.m*

```
%PDSTEP1.M creates a MAT-file (hpowerdis.mat) that contains angle-distance matrix
%           with received powers as its entries
%           (c)1997 Gamantyo Hendrantoro

clear;

%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%% SETTING SOME INITIAL VALUES
filedatabase = ['h4050g','h4050h','h4050i','h4050j','h4050k','h4050l';...
               'h4038a','h4038b','h4038c','h4038d','h4038e','h4038f';...
               'h4038g','h4038h','h4038i','h4038j','h4038k','h4038l';...
               'ha507a','ha507b','ha507c','ha507d','ha507e','ha507f';...
               'ha507g','ha507h','ha507i','ha507j','ha507k','ha507l'];

pda = [];

for floppy = 1:30,
    for fileno = 1:5,
        filename = [filedatabase(floppy,:),int2str(fileno),'.mat'];
        command = ['load ',filename];
        disp(command); eval(command);
        pdanewline = zeros(1,21);
        actpow = ActualPower; % Use actpow as a dummy variable
        [x y] = max(actpow);
        indexshift = y - 6;
        actpow = abs(sign(actpow)).*sign(actpow+35).*actpow;
        pdanewline(6-indexshift:16-indexshift) = actpow;
        pda = [pda; pdanewline];
    end
    disp(' ');
end

distance = [3.2 4.6 5.2 4.7 3.5 5.0];
distance = [distance 10.4 10.2 7.2 7.5 6.5 6.8 6.5 3.7 3.3 6.2 6.1 3.8];
distance = [distance 9.8 7.8 6.2 4.1 3.0 3.6 4.6 5.5 7.3 8.8 7.7 4.8];
dist=[distance; distance; distance; distance; distance]; dist=dist(:);
PDA = [dist pda(:,6:16)];
readme = ['PDA stores unsorted distances on first column and corresponding',...
         ' received powers for all individual angles (-30:6:30) on others.'];
command = ['save powerdis.mat PDA readme'];

for ang=1:11,
    XY = [];

    %%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%% USE THESE 5 LINES (no averaging)
    for fileno=1:150, % = #floppies * 5
        if PDA(fileno,ang+1),
            XY = [XY; [PDA(fileno,1) PDA(fileno,ang+1)]];
        end
    end
    command = ['PD',int2str(ang),' = XY;'];
    disp(command); eval(command);
end

command = ['save hpowerdis.mat PDA readme PD1 PD2 PD3 PD4 PD5 PD6 PD7 PD8 ',...
         'PD9 PD10 PD11'];
disp(command); eval(command);
```

b. pdstep2.m

```

%PDSTEP2.M operates on powerdis.mat to yield the linear curve fit of power in
%      dB and distance in logarithmic scale for each pointing angle.
%      It computes one-segment linear regression analyses.
%      (c)1997 Gamantyo Hendrantoro

close all
clear
load hpowerdis

%%% Sorting the distances in ascending order
for k=1:11,
    strk = int2str(k);
    command = ['pd',strk,' = PD',strk,',';']; eval(command);
    command = ['[sorted,index] = sort(pd',strk,','(:,1));']; eval(command);
    command = ['pd',strk,' = [sorted pd',strk,','(index,2)];']; eval(command);
end;

%%% Printing the scatter plots
figure(1)
for k=1:11,
    subplot(4,3,k),
    command = ['pair = pd',int2str(k),',';'];
    eval(command)
    semilogx((pair(:,1)),pair(:,2),'.');
    axis([3 10 -35 0])
    text(8,-5,sprintf('%3.0f',k*6-36))
    hold on
end

for k=1:11,
    %%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%% Computing coefficients
    strk = int2str(k);
    command = ['ab = polyfit(10*log10(pd',strk,','(:,1)),pd',strk,','(:,2),1);'];
    eval(command);
    AB(k,1:2) = ab;

    %%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%% Computing error
    command = ['er',strk,' = pd',strk,...
        '(:,2) - AB(k,2) - AB(k,1)*10*log10(pd',strk,','(:,1));'];
    eval(command);
    command = ['AB(k,3) = sqrt(sum(er',strk,','.^2)/(length(er',strk,')-2));'];
    eval(command);

    %%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%% Computing correlation coefficient
    command = ['cc = corrcoef(10*log10(pd',strk,','(:,1)),pd',strk,','(:,2));'];
    eval(command);
    AB(k,4) = cc(1,2);

    figure(1)
    subplot(4,3,k),
    semilogx([3:10],AB(k,2)+AB(k,1)*10*log10([3:10]),'-')
end

figure(1)
subplot(4,3,2); title('Relative Rx power(dB) - Distance(log scale)')
subplot(4,3,10); xlabel('Distance(m)')
subplot(4,3,11); xlabel('Distance(m)')
subplot(4,3,12); xlabel('Distance(m)')

figure(2)
semilogx([1:10],AB(:,2)*ones(1,length([1:10]))-AB(:,1)*10*log10([1:10]),'y')

```

```

axis([1 10 -35 5])
legend('-30','-24','-18','-12','-6','0','6','12','18','24','30')
title('Relative Rx Power (dB) vs Distance (log)')
ylabel('Relative Rx Power (dB)'); xlabel('Distance (m)')

figure(3)
semilogx([3:10],AB([6 1:3],2)*ones(1,length([3:10]))+...
    AB([6 1:3],1)*10*log10([3:10]),'y')
axis([3 10 -30 -5]); legend('0','-30','-24','-18')
title('Relative Rx Power (dB) vs Distance (log)')
ylabel('Relative Rx Power (dB)'); xlabel('Distance (m)')

figure(4)
semilogx([3:10],AB([6 4 5 7 8],2)*ones(1,length([3:10]))+...
    AB([6 4 5 7 8],1)*10*log10([3:10]),'y')
axis([3 10 -30 -5]); legend('0','-12','-6','6','12')
title('Relative Rx Power (dB) vs Distance (log)')
ylabel('Relative Rx Power (dB)')
xlabel('Distance (m)')

figure(5)
semilogx([3:10],AB([6 9:11],2)*ones(1,length([3:10]))+AB([6
9:11],1)*10*log10([3:10]),'y')
axis([3 10 -30 -5])
legend('0','18','24','30')
title('Relative Rx Power (dB) vs Distance (log)')
ylabel('Relative Rx Power (dB)')
xlabel('Distance (m)')

disp('angle      n      P(1)    standard dev.    corr.coeff.')
for k=1:11,
    disp([sprintf('%5.0f',k*6-36) sprintf('%10.3f',AB(k,1)) ...
        sprintf('%10.3f',AB(k,2)) sprintf('%15.2f',AB(k,3)) ...
        sprintf('%10.3f',AB(k,4))]);
end

readme = ['PDA stores sorted distances on first column and corresponding',...
    ' received powers for all individual angles (-30:6:30) on others.'];
command = ['save hpowerdis.mat PDA readme PD1 PD2 PD3 PD4 PD5 PD6 PD7 PD8 ',...
    'PD9 PD10 PD11 er1 er2 er3 er4 er5 er6 er7 er8 er9 er10 er11'];
disp(command); eval(command);

```

c. *pdstep3.m*

```

%PDSTEP3.M
% Builds ccdfs of er's in powerdis.mat
% Prints them and their normal fits
% (c)1997 Gamantyo Hendrantoro

close all; clear
load hpowerdis

%%% Sorting the distances in ascending order
for k=1:11,
    strk = int2str(k);
    command = ['pd',strk,' = PD',strk,',';]; eval(command);
    command = ['[sorted,index] = sort(pd',strk,'(:,1));']; eval(command);
    command = ['pd',strk,' = [sorted pd',strk,'(index,2)];']; eval(command);
end;

for k=1:11,
    %%%%%%%%%%%%%%%%%%%%%%%%%%%%%%% Computing coefficients
    strk = int2str(k);
    command = ['ab = polyfit(10*log10(pd',strk,'(:,1)),pd',strk,'(:,2),1);'];

```

```

eval(command); AB(k,1:2) = ab;

%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%% Computing error
command = ['er',strk,' = pd',strk,...
          '(:,2) - AB(k,2) - AB(k,1)*10*log10(pd',strk,'(:,1));']; eval(command);
command = ['AB(k,3) = sum(er',strk,'.^2)/(length(er',strk,')-2);']; eval(command);
end

figure

%%% Building experimental CDF for every angle
for k=1:11,
    strk = int2str(k);
    command = ['cdf',strk,' = sum(er',strk,'*ones(1,200)<=ones(length(er',strk,...
          '),1)*linspace(-4*sqrt(AB(k,3)),4*sqrt(AB(k,...
          ',3)),200)); cdf',strk,' = cdf',strk,'/length(er',strk,');'];
    eval(command);
    subplot(4,3,k)
    command = ['plot(linspace(-4*sqrt(AB(k,3)),4*sqrt(AB(k',...
          ',3)),200),cdf',strk,'','-'')'];
    eval(command);
    axis([-10 10 0 1]); hold on
end

%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%% Building CDF of zero-mean normal R.V. for every angle
disp('Data for Kolmogorov-Smirnov goodness-of-fit test');
disp('Angle Dmax      N      99%      99%');
disp('                exp      norm');
for k=1:11,
    index = linspace(-4*sqrt(AB(k,3)),4*sqrt(AB(k,3)),200);
    cdfnorm = [];
    for a=index,
        zeroa = linspace(0,a,round(abs(a)*100));
        integrand = 1/sqrt(2*pi)/sqrt(AB(k,3)) * exp(-zeroa.^2/2/AB(k,3));
        cdfnorm = [cdfnorm .5+trapz(zeroa,integrand)];
    end
    subplot(4,3,k); plot(index,cdfnorm,'-'); hold off
    ang = sprintf('%3.0f',k*6-36); text(6,.1,ang)
    command = ['D = max(abs(cdfnorm - cdf',int2str(k),''))']; eval(command);
    command = ['CDF = pd',int2str(k),'']; eval(command);
    command = ['er = sort(er',int2str(k),'')']; eval(command);
    e99 = er(ceil(.01*length(er)));
    display = [sprintf('%5.0f',...
          k*6-36),sprintf('%5.2f',D),sprintf('%5.0f',length(CDF)),...
          sprintf('%7.2f',e99),sprintf('%7.2f',-2.33*sqrt(AB(k,3)))];
    disp(display);
end

subplot(4,3,4); ylabel('P(Error(dB) < abscissa)');
subplot(4,3,11); xlabel('Error (dB)')

```

d. *pdsim1.m*

```

% PDSIM1.M for power-distance simulation
% 5-ray model, random floor at 1.3-1.9m, fixed ceiling at 0.7m
% two random walls at 1-6 and 2-6 meters
% Tx antenna: 30deg elevation beamwidth
% Rx antenna: 15deg elevation beamwidth, with 2 sidelobes at +/-23deg
% (c)1997 Gamantyo Hendrantoro

clear

D = 10000;

```

```

r0 = 1:0.01:D;
N = length(r0);

h1 = rand(1,N)*1.3 + 0.6;      % distance of reflector
h2 = ones(1,N)*0.7;
h3 = rand(1,N)*1 + 5;
h4 = rand(1,N)*2 + 4;

r1 = sqrt(r0.^2 + (2*h1).^2); % path length through reflection
r2 = sqrt(r0.^2 + (2*h2).^2);
r3 = sqrt(r0.^2 + (2*h3).^2);
r4 = sqrt(r0.^2 + (2*h4).^2);

Q1 = atan(r0./(2*h1)); % incidence angle on reflector surface
Q2 = atan(r0./(2*h2));
Q3 = atan(r0./(2*h3));
Q4 = atan(r0./(2*h4));

% antenna gain
rad = pi/2 - Q1;
G1 = cos(rad).^10 .* ...
    (abs(cos(rad*pi/8)).^0.25 .* ( (2/pi*rad+1).*(rad<0) + ...
    (-2/pi*rad+1).*(rad>=0) ));
rad = pi/2 - Q2;
G2 = cos(rad).^10 .* ...
    (abs(cos(rad*pi/8)).^0.25 .* ( (2/pi*rad+1).*(rad<0) - ...
    (-2/pi*rad+1).*(rad>=0) ));
G3 = cos(pi/2 - Q3).^(10+60);
G4 = cos(pi/2 - Q4).^(10+60);

E1 = 6.5; % relative permittivity
E2 = 6.5;
E3 = 6.5;
E4 = 6.5;

% reflection coefficient
R1 = (cos(Q1) - sqrt(E1-sin(Q1).^2))./(cos(Q1) + sqrt(E1-sin(Q1).^2));
R2 = (cos(Q2) - sqrt(E2-sin(Q2).^2))./(cos(Q2) + sqrt(E2-sin(Q2).^2));
R3 = (cos(Q3) - sqrt(E3-sin(Q3).^2))./E3./(cos(Q3) + ...
    sqrt(E3-sin(Q3).^2))./E3;
R4 = (cos(Q4) - sqrt(E4-sin(Q4).^2))./E4./(cos(Q4) + ...
    sqrt(E4-sin(Q4).^2))./E4;

freq = 29.5e9;

lambda = (3e8./freq);
k = 2*pi./(3e8./freq);

Pr = abs(((3e8./freq)/4/pi * ones(1,N)) .* (...
    exp(-j*k*r0)./r0 +...
    (G1.*R1).*exp(-j*k*r1)./r1 +...
    (G2.*R2).*exp(-j*k*r2)./r2 + ...
    (G3.*R3).*exp(-j*k*r3)./r3 + ...
    (G4.*R4).*exp(-j*k*r4)./r4 ...
    ));
PdB = 10*log10(Pr/max(Pr));

figure
semilogx(r0,PdB,':'); hold on

ylabel('Relative received power (dB)')
xlabel('Propagation distance (m)')
axis([3 D min(PdB) 0])

```

```

BP1 = (4*h1(1).^2 - lambda.^2/4)/lambda;
BP2 = (4*h2(1).^2 - lambda.^2/4)/lambda;

[x ind1] = max(r0>BP1);
[x ind2] = max(r0>BP2);

coef = polyfit(10*log10(r0(1:ind2)),PdB(1:ind2),1);
n1 = coef(1);
semilogx(r0(1:ind2-1:ind2),coef(2)+10*n1*log10(r0(1:ind2-1:ind2)));

coef = polyfit(10*log10(r0(ind2:ind1)),PdB(ind2:ind1),1);
n2 = coef(1);
semilogx(r0(ind2:ind1-ind2-1:ind1),...
         coef(2)+10*n2*log10(r0(ind2:ind1-ind2-1:ind1)));

coef = polyfit(10*log10(r0(ind1:N)),PdB(ind1:N),1);
n3 = coef(1);
semilogx(r0(ind1:N-ind1-1:N),coef(2)+10*n3*log10(r0(ind1:N-ind1-1:N)));

title(sprintf('BP1 = %8.1f m; ',BP2),...
       sprintf('n1 = %4.2f',-n1));

```

e. pdsim2.m

```

%PDSIM2.M to compute distribution function, mean and standard deviation of
%   propagation exponent n.
% 0 = LOS
% 1 = ground
% 2 = ceiling
% 3 = wall1
% 4 = wall2
clear

NS = 10000;

for X = 1:NS, % NS = number of runs

%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%

    N = 100; % N = number of points

    r0 = rand(1,N)*7.4 + 3; % LOS distance

    h1 = rand(1,N)*1.3 + 0.6; % distance of reflector
    h2 = ones(1,N)*0.7;
    h3 = rand(1,N)*1 + 5;
    h4 = rand(1,N)*1 + 5;

    r1 = sqrt(r0.^2 + (2*h1).^2); % path length through reflection
    r2 = sqrt(r0.^2 + (2*h2).^2);
    r3 = sqrt(r0.^2 + (2*h3).^2);
    r4 = sqrt(r0.^2 + (2*h4).^2);

    Q1 = atan(r0./(2*h1)); % incidence angle on reflector surface
    Q2 = atan(r0./(2*h2));
    Q3 = atan(r0./(2*h3));
    Q4 = atan(r0./(2*h4));

    rad = pi/2 - Q1;
    G1 = cos(rad).^10 .* ...
         (abs(cos(rad*8)).^0.25 .* ( (2/pi*rad+1).*(rad<0) + ...
         (-2/pi*rad+1).*(rad>=0) ));
    rad = pi/2 - Q2;

```



```

G2 = cos(rad).^10 .* ...
    (abs(cos(rad*8))).^0.25 .* ( (2/pi*rad+1).*(rad<0) - ...
    (-2/pi*rad+1).*(rad>=0) );
G3 = cos(pi/2 - Q3).^(10+60);
G4 = cos(pi/2 - Q4).^(10+60);

E1 = ones(1,N)*6.5;    % relative permittivity
E2 = ones(1,N)*6.5;
E3 = ones(1,N)*6.5;
E4 = ones(1,N)*6.5;

                                % reflection coefficient
R1 = (cos(Q1) - sqrt(E1-sin(Q1).^2))./(cos(Q1) + sqrt(E1-sin(Q1).^2));
R2 = (cos(Q2) - sqrt(E2-sin(Q2).^2))./(cos(Q2) + sqrt(E2-sin(Q2).^2));
R3 = (cos(Q3) - sqrt(E3-sin(Q3).^2))./E3./(cos(Q3) + ...
    sqrt(E3-sin(Q3).^2))./E3;
R4 = (cos(Q4) - sqrt(E4-sin(Q4).^2))./E4./(cos(Q4) + ...
    sqrt(E4-sin(Q4).^2))./E4;

freq = 29.5e9;

lambda = (3e8./freq);
k = 2*pi./(3e8./freq);

Pr = abs(((3e8./freq)/4/pi * ones(1,N)) .* ...
    (exp(-j*k*r0)./(ones(nf,1) * r0) + ...
    ones(nf,1)*(G1.*R1).*exp(-j*k*r1)./(ones(nf,1)*r1) + ...
    ones(nf,1)*(G2.*R2).*exp(-j*k*r2)./(ones(nf,1)*r2) + ...
    ones(nf,1)*(G3.*R3).*exp(-j*k*r3)./(ones(nf,1)*r3) + ...
    ones(nf,1)*(G4.*R4).*exp(-j*k*r4)./(ones(nf,1)*r4))) ...
    .^2;

PdB = 10*log10(Pr);

coef = polyfit(10*log10(r0),PdB,1);
n = coef(1);
nn = [nn n];

%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%

end

m = mean(-nn)
s = std(-nn)
figure
hist(-nn,50)
title(['No. of runs=',int2str(NS),...
    '; No. of data per run=',int2str(N),...
    sprintf('; Average=%4.2f',m),...
    sprintf('; Std dev=%4.2f',s)])
ylabel('Frequency of occurrence')
xlabel('Propagation exponent, n')

```