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EMPIRICAL EVALUATION OF DIVERSITY TECHNIQUES FOR INDOOR DIGITAL RADIO AT 1.75 GHz

by

Stephen Ross Todd, B.A.Sc.

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

Masters of Engineering

**Ottawa-Carleton Institute for Electrical Engineering
Faculty of Engineering
Department of Systems and Computer Engineering
Carleton University
December 15, 1992**

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

**“Empirical Evaluation of Diversity Techniques
for Indoor Digital Radio at 1.75 GHz”,**

submitted by Stephen R. Todd in partial fulfillment of the requirements
for the degree of Masters of Engineering.

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Abstract

The expanding international market for portable and mobile communications has created a great interest in indoor wireless communications. Major impairments to signal quality occur in buildings due to the severe multipath fading and the dynamics of a changing radio channel. For low cost portable telephones some practical micro-diversity techniques are required. While considerable theory has been developed on these techniques there are few published results of indoor diversity measurements, particularly in the 1.6–2.0 GHz frequency band. This thesis evaluates three diversity techniques at 1.75 GHz based on narrowband measurements of envelope fading due to both temporal changes in the channel and the motion of a four-branch diversity receiver. The experiments were conducted in four university buildings having differing layouts, wall construction and traffic.

Analysis results provide a direct comparison of two-branch space, two-branch frequency and four-branch space/frequency (hybrid) diversity. The average diversity gain of two-branch space diversity was relatively insensitive to antenna spacing from one full wavelength down to one-quarter wavelength. The average diversity gain at 99% time availability was 10 dB. For frequency diversity the minimum carrier separation for best performance was between 5 and 10 MHz which provided similar performance to the space diversity. The four-branch hybrid diversity results provided gains of 15 dB. The effects of various buildings and path topologies on the average output fading statistics was only 1 to 2 dB at 99% suggesting that results may be useful for other similar areas. There were however significant variations in envelope cross-correlation and diversity output fading between individual one-minute measurements. These variations in statistics were reduced for higher order diversity but had significant effects on the worst case bit error rate of differential phase shift keying. Simulations of two practical switched combining techniques showed that the switch and stay strategy performance is highly dependent on variable fading statistics and switch delays in excess of 10 milliseconds, while the switched diversity based on a direct comparison of branches in alternating timeslots was less sensitive to these impairments. These results would be useful to predict communication reliability and required fade margins in similar buildings.

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Table of Contents

Abstract	iii
Acknowledgments	iv
List of Tables	viii
List of Figures	ix
List of Plates	xi
List of Acronyms and Abbreviations	xii
List of Mathematical Symbols and Notation	xiv
Chapter 1	INTRODUCTION 1
1.1	Background 1
1.2	Thesis organization 4
1.3	Thesis contributions 5
Chapter 2	BACKGROUND ON INDOOR WIRELESS COMMUNICATIONS, RADIO PROPAGATION AND DIVERSITY TECHNIQUES 8
2.1	Portable communication systems 8
2.2	Multipath radio wave propagation 11
2.3	Diversity techniques for digital communications 16
2.3.1	Micro-scopic diversity types 16
2.3.2	Combining techniques 17
2.3.3	Practical issues and impairments 25
Chapter 3	CURRENT RESEARCH AND THESIS MOTIVATION 29
3.1	Current research on indoor radio propagation and diversity performance estimation 29
3.1.1	Indoor radio channel measurements and models 29
3.1.2	Diversity performance estimates based on theoretical channel models and simulation 31

3.1.3	Measurements to evaluate diversity performance in buildings	32
3.2	Analysis of current research	34
3.3	Thesis problem statement	35
Chapter 4	EXPERIMENTAL PROCEDURE AND APPARATUS	38
4.1	Introduction	38
4.2	Indoor environment	38
4.2.1	Primary measurement area	38
4.2.2	Secondary measurement areas	39
4.3	Measurement Apparatus	41
4.4	Procedure to measure fading	43
Chapter 5	ANALYSIS AND RESULTS	52
5.1	Introduction	52
5.2	Envelope fading statistics	53
5.3	Envelope cross-correlation	53
5.4	Selection combined fading statistics	58
5.4.1	Effects of antenna and frequency separation on average diversity gain	61
5.5	Diversity performance in four buildings.	66
5.6	Variation of diversity performance	70
5.7	Practical selection combining techniques	74
5.7.1	Switch and stay combining	75
5.7.2	Switched diversity with a TDM format	81
5.7.3	Discussion	85
5.8	Bit error rate computation for DPSK	85
5.9	Application of diversity results to a link budget	91
5.10	Summary of Analysis	94

Chapter 6	CONCLUSION	98
6.1	Thesis summary	98
6.1.1	Choice of parameters for space, frequency and hybrid diversity	99
6.1.2	Direct comparison of three diversity techniques and the effects of different environments	100
6.1.3	Variations in correlation, fading statistics and diversity performance	101
6.1.4	Practical combining techniques	101
6.1.5	Application to digital communications and systems design	102
6.2	Recommendations for further research	103
6.2.1	Further analysis with this data	103
6.2.2	Further measurements with this apparatus	104
Bibliography		106
Appendix A	Measurement environments	110
Appendix B	Photographs of measurement equipment	115
Appendix C	Data acquisition hardware and software	119
Appendix D	Antenna verification	121
Appendix E	Program listings of analysis software	126
E.1	Conversion of data to power values in dBm	126
E.2	Computation of CDF with various combining	127
E.3	Switch and stay combining	132
E.4	Switched diversity with TDM format	134
E.5	Computation of average BER from envelope fading CDF	136
E.6	Computation of BER for DPSK	137

List of Tables

Table 2.1	Cordless telephone system parameters	10
Table 2.2	Diversity gains in dB at 99% availability with Rayleigh fading . .	22
Table 5.1	Average SC diversity gains in dB versus ΔF and ΔS for frequency diversity (a), for space diversity (b) and four-branch hybrid diversity (c).	65
Table 5.2	Summary of path types in each building.	67
Table 5.3	Summary of average CDF values in dB at 99 % availability and corresponding diversity gain using SC. Results are for all measured paths.	68
Table 5.4	Summary of average CDF values in dB at 99 % availability and corresponding diversity gain using SC. Results are separated into OBS and LOS paths.	69
Table 5.5	The effect of switching delay on SAS combining for typical results of path 101 and unusual results of path 108.	80
Table 5.6	Link budget for cordless telephone system based on DECT standard using measured diversity results.	95

List of Figures

Figure 2.1	CDF of Rayleigh distributed envelope power with SC diversity assuming two and four branches with independent fading. . . .	20
Figure 2.2	Block diagram of switched diversity technique	23
Figure 2.3	Block diagram of feedback diversity technique	26
Figure 4.1	Primary measurement area with transmitter locations (X) and receiver locations (R).	40
Figure 4.2	1.75 GHz Dual Frequency Transmitter.	44
Figure 4.3	1.75 GHz Frequency and Space Diversity Receiver.	45
Figure 4.4	Receiver calibration file with ADC value versus RF input power. .	46
Figure 4.5	Top view of antenna movement around guide for antenna 1 (A1) and antenna 2 (A2).	48
Figure 4.6	Received envelope from antenna moving in a circle at 0.5 m/s. .	50
Figure 4.7	Effect of sampling rate on the CDF of envelope fading.	51
Figure 5.1	CDF of received signal fading on four branches from one file. . .	54
Figure 5.2	CDF of single branch envelope fading from 868 OBS paths. . . .	55
Figure 5.3	Histograms of envelope correlation coefficients.	57
Figure 5.4	Received envelope from one measurement for one-, two- and four-branch diversity with SC using $\Delta F = 10$ MHz and $\Delta S = 1\lambda$. . .	59
Figure 5.5	CDF of selection combined envelope for one path using four-branch hybrid diversity with $\Delta F = 10$ MHz and $\Delta S = \lambda$. . .	60
Figure 5.6	Average CDF for selection combined envelopes from 22 OBS paths using $\Delta F = 10$ MHz and $\Delta S = \lambda/4$	63
Figure 5.7	Average CDF for selection combined envelopes from 22 OBS paths using $\Delta F = 2$ MHz and $\Delta S = \lambda/4$	64
Figure 5.8	Average and worst case envelope fading statistics using two-branch space diversity and four-branch hybrid diversity with $\Delta F = 10$ MHz and $\Delta S = 1\lambda$	72
Figure 5.9	Distribution of one-, two- and four-branch CDF values at 99% for all OBS and LOS paths in all buildings.	73

Figure 5.10	Typical result for CDF of envelope fading on branch 1 (BR1) and branch 3 (BR3) using SAS with threshold levels in dB. From path 101, first space diversity pair.	76
Figure 5.11	Typical results for SAS from path 101 with knee of CDF shifting down for increasing delay. Delays are in msec.	78
Figure 5.12	Example of SAS output fading with bad input fading from path 108. Delays are in msec.	79
Figure 5.13	Simulated TDM frame format	82
Figure 5.14	Typical results for switched TDM technique from path 101. Timeslot durations are in msec.	83
Figure 5.15	Switched TDM technique CDFs from unusual fading on path 108. Timeslot durations are in msec.	84
Figure 5.16	BER of DPSK with average and worst case statistics from 35 OBS paths for $\Delta F = 10$ MHz and $\Delta S = 1\lambda$ compared to ideal SC with Rayleigh fading shown in dotted lines.	90
Figure A.1	Secondary measurement area in the engineering annex with transmitter locations (T) and receiver locations (R).	110
Figure A.2	Measurement locations on the second floor of Administration building	111
Figure A.3	Measurement locations on the third floor of Administration building	112
Figure A.4	Measurement locations on second level of library	113
Figure A.5	Measurement locations on top floor of Mackenzie building	114
Figure D.1	E-field azimuth cut for single monopole antenna	122
Figure D.2	E-field azimuth cut with second antenna at 0° and $\Delta S = \lambda/4$	123
Figure D.3	E-field elevation cut for single monopole antenna	124
Figure D.4	E-field elevation cut with second antenna at 90° and $\Delta S = \lambda/4$	125

List of Plates

Plate 1	1.75 GHz dual frequency transmitter (left) and four-branch diversity receiver (right).	116
Plate 2	Top view of receive antennas and guiding frame	117
Plate 3	Side view of receive antennas on foam supports separated by one half wavelength (8.5 cm)	118

List of Acronyms and Abbreviations

ADC — analog to digital converter
AGC — automatic gain control
Amp — amplifier
BER — bit error rate
BR# — branch number
CCI — co-channel interference
CDF — cumulative distribution function
CT-1 — Cordless Telephone system, first generation
CT-2 — Cordless Telephone system, second generation
ct/sqkm — circuits per square kilometer
CW — continuous wave
dB — decibel
dBm — decibels reference to one milliwatt of power
DCA — dynamic channel allocation
DECT — Digital European Cordless Telecommunications
DPSK — differential phase shift keying
EGC — equal gain combining
FDM — frequency division multiplexing
FSK — frequency shift keying
FSL — free space loss
GHz — gigahertz = one billion cycles per second
GMSK — gaussian minimum shift keying
HF — high frequency
Hz — hertz (cycles per second)
IF — intermediate frequency
ISDN — integrated digital services network
ISI — intersymbol interference
Kbps — kilo-bits per second
KHz — kilohertz = one thousand cycles per second

LCR — level crossing rate
LOS — line of sight
MHz — megahertz = one million cycles per second
MRC — maximal ratio combining
m — metre
m/s — metres per second
msec — milliseconds = 10^{-3} seconds
mW — milliwatt = 10^{-3} watts
OBS — obstructed
PBX — private branch exchange
PC — personal computer
PCN — personal communications network
PCS — personal communication services
PIFA — planar inverted-F antenna
pdf — probability density function
pmf — probability mass function
RF — radio frequency
rms — root mean square
Rx — receiver
RSSI — received signal strength indicator
s — second
SAE — switch and examine (combining)
SAS — switch and stay (combining)
SC — selection combining
SNR — signal to noise ratio
TDD — time division duplex
TDM — time division multiplexing
TDMA — time division multiple access
Tx — transmitter
wrt — with respect to

List of Mathematical Symbols and Notation

- a_k — multiplier coefficient on the k^{th} branch
 A — switching threshold voltage
 B_d — Doppler spread
 $c(\tau;t)$ — impulse response of time varying multipath channel
 $C(f;t)$ — multipath channel transfer function
 C — covariance matrix of data matrix
 e — 2.71828183
 d/dx — derivative with respect to x
 D — straight line path distance in metres
 $E[\bullet]$ — expected value operator
 E_b — energy per bit
 f_c — carrier frequency
 f_d — Doppler frequency variable
 $f_{d_{\max}}$ — maximum Doppler frequency
 I_0 — zeroth order modified bessel function
 k — Boltzman constant = 1.37×10^{-23} Joule/degree
 k^{th} — branch indicator
 K — ratio of diffuse to dominant components for a Rician distribution
 L — number of diversity branches
 m^2 — square metres
 n_k — noise voltage on the k^{th} branch
 $\overline{n_k^2}$ — mean noise power on the k^{th} branch
 N — distance/power law exponent
 N_0 — noise power density per Hz
 $p_{\gamma,2}(x)$ — pdf of SNR of two-branch combiner
 $p_{\gamma,4}(x)$ — pdf of SNR of four-branch combiner
 $p_\alpha(x)$ — pdf of envelope fading
 $p_k(x)$ — pmf with respect to x
 $P(\alpha \leq x)$ — CDF of envelope fading

P_2 — probability of bit error
 P_{2_2} — probability of bit error with two-branch combiner
 P_{2_4} — probability of bit error with four-branch combiner
 $r(t)$ — equivalent lowpass received signal
 $r_k(t)$ — signal plus noise on the k^{th} branch
 R — data matrix (60,000 $\times$ 4)
 R_s — optimum symbol rate
 $s_k(t)$ — signal envelope on the k^{th} branch
 T — time duration for local averages (less than a symbol duration)
 T_1 — time duration for short term averages (in this work, 60 seconds)
 T_2 — time duration for long term averages
 T_{amb} — ambient temperature
 T_q — average fade duration
 T_m — multipath spread
 T_s — symbol interval
 T_{sd} — switching time delay
 V — velocity
 W — message bandwidth
 x — dummy variable
 α — radio channel attenuation factor
 $\alpha_n(t)$ — attenuation factor for the n^{th} path
 γ_b — SNR per information bit
 γ_c — SNR at combiner output
 γ_k — local SNR on the k^{th} branch
 $(\Delta f)_c$ — coherence bandwidth
 ΔF — frequency diversity carrier separation
 ΔS — space diversity antenna separation
 $(\Delta t)_c$ — coherence time
 Γ — average SNR over period T_1
 Γ_b — average SNR per information bit

Γ_c — average SNR at combiner output
 θ_n — phase shift for the n^{th} path
 λ — one free space wavelength at the carrier frequency
 π — 3.14159265
 ρ_n — envelope cross-correlation coefficient
 $\rho(i,j)$ — cross-correlation coefficient between the i^{th} and j^{th} branch
 σ^2 — variance
 $\tau_n(t)$ — propagation delay for the n^{th} path
 τ_q — average fade duration
 $\phi(t)$ — phase of the equivalent lowpass channel
 $\phi_c(\tau;\Delta t)$ — auto-correlation function of channel impulse response
 $\phi_C(\Delta f;\Delta t)$ — auto-correlation function of channel transfer function
 $\sum_a^b$ — summation from a to b
 $\int_a^b$ — integration from a to b
 $\prod_a^b$ — product from a to b

Chapter 1

INTRODUCTION

1.1. Background

Indoor radio communications

There has been great interest in wireless communications that would allow personal and portable communications to anyone, anywhere. This tetherless communication system concept is often referred to as universal portable communications [1] or personal communication services (PCS). Communication systems addressing the needs of mobile users, such as cellular radio, have become increasingly popular. There has also been a huge demand for cordless telephones in residential applications [2]. Recently there have been standards to address a growing demand for in-building wireless communications to support commercial use. These are typically in the form of a wireless private branch exchange (PBX) to serve multiple users in an office. Radio systems addressing this need however, must contend with many impairments in the indoor communications channel.

For narrow band communication services a dominant impairment is the reduction of the received signal level. Severe signal fading from multipath propagation is a major impairment to signal quality. These fluctuations typically occur within short distances less than a wavelength for a moving receiver. Receive signal power can decrease by a factor of 1000 or 30 dB. Temporal variations occur due to the movement of personnel or equipment. Larger scale and longer time variations may occur throughout buildings due to shadowing from walls or large equipment.

The large area and long time variations are typically mitigated through the use of a macro-diversity scheme where a portable will communicate with one of several base stations whichever has the best signal. The small scale, short term and temporal variations of the channel require a micro-diversity scheme to combat the fading. Micro-diversity can be considered to be the first line of intelligent defence against fading compared with the brute force approach of increasing transmit power which requires more battery power and weight for the portable and may increase interference between users. Other techniques

such as coding and interleaving are applied to reduce the effect of the fading on the demodulated output in contrast to diversity which, in the case of pre-detection combining, modifies the fading statistics at the input to the demodulator to reduce the likelihood of deep fading and resulting bit errors. Micro-diversity, which can be implemented at the base station or portable receiver, selects from multiple replicas of the transmitted signal on multiple "branches" in the receiver. These branches may take the form of two or more antennas or multiple receivers tuned to different frequencies. In this thesis we evaluate micro-diversity techniques in a context where the large scale variations have been reduced through the use of macro-diversity.

Related research

Micro-diversity has been examined in detail for the application to a variety of propagation media and there has been in depth treatment of combining techniques [3, 4, 5], simulated performance estimates [6], and applications to system link design [2] within the context of a macroscopic diversity system to mitigate shadow fading [7]. Despite these achievements, there are few results that indicate the best operating parameters for indoor radio diversity systems, the expected variations in diversity performance or the amount of envelope correlation between diversity branches. In many proposed systems the required fade margin is based on average diversity performance estimates from computer simulations or based on a Rayleigh distribution. Indoor radio channel models have been developed to predict the fading and time dispersive nature of the channel [8, 9], however, there have been few published results of diversity performance over this channel based on measured data. Some experiments have been conducted in the 800 and 900 MHz bands using two-branch diversity [10, 11, 12, 13] where different types were compared based on independent experiments and average fading statistics. Recently there has been interest in operating at twice these frequencies in the 1.6 – 2.0 GHz band [14, 15] where more spectrum is being allocated to PCS. Some systems propose the use of four-branch diversity for robust in-building communications [15]. In the literature there is also a lack of direct comparisons between two-branch diversity schemes.

Thesis motivation

Empirical estimates of two-branch and four-branch diversity performance in the 1.6 – 2.0 GHz band, addressing the above issues, would be useful for the design of indoor portable terminals and systems as envisaged in [1]. It is important to know the best parameters to use, the trade-offs between complexity and performance, variations in performance, and the effect of various environments and path topologies. The implementation of diversity can have a large impact on over all transmission system design. Space diversity antenna separation has a substantial impact on the portable design while frequency diversity implementation is subject to allocated bandwidths for a service.

Thesis objective

Our objective is to evaluate diversity performance in buildings in the 1.6 – 2.0 GHz frequency band. To do this we use an empirical approach. The objective of this thesis is to present a review of the experiment and an analysis of realistic and practical diversity techniques for indoor wireless communications based on measured data. Specifically, we consider the case of a moving portable receiver and a stationary transmitter.

Empirical approach

The approach consisted of three stages which were: the design of the experimental apparatus, the measurements and then the data analysis.

The equipment designed consisted of a transmitter and receiver. The transmitter emitted two continuous wave (CW) carriers on different frequencies which were subject to fading of the radio channel which was measured using a four-branch, two-antenna, dual frequency receiver.

Measurements were conducted in four different buildings in various sized rooms with typical levels of personnel activity for each area and under conditions that approximate the use of a portable telephone. At some areas multiple measurements were made with various antenna spacings and frequency separations. A centre frequency of 1.75 GHz was chosen to allow comparisons with previous temporal fading measurements [9].

Data analysis involved the grouping of similar measurements and the writing of software to extract specific information useful in the design of systems employing diversity techniques. The simultaneous measurement of fading on four diversity branches allowed for a direct comparison between space diversity, frequency diversity and space/frequency (hybrid) diversity. Simple selection combining techniques were implemented in software and measurements were applied to a digital modulation technique and a system link budget.

1.2. Thesis organization

In Chapter 2 we provide a review of relevant background information which consists of three parts. A description of portable communications systems is presented first. This is intended to familiarize the reader with current and future portable systems and it establishes a context for our research. Second, the nature of multipath radio propagation is reviewed and relevant terms are defined. The third section describes diversity techniques and details some specific types which we consider in this thesis.

Chapter 3 presents a literature review of applicable research to our study, an analysis of this research and a statement of the objectives of this thesis. The first section presenting the review is comprised of three parts. We describe results from indoor radio measurements, theoretical and simulated diversity performance, and experiments to measure diversity operation in buildings. In the second section the analysis of the research highlights relevant topics that lay the ground work for our study. The final section presents the thesis problem statement which outlines our research objectives.

Chapter 4 contains a description of the experimental procedure and apparatus. The beginning of this chapter describes the data storage requirements. Next, the indoor environments are described with detail of the layout and construction type. Then we describe the design and specifications of the equipment. The chapter concludes with a detailed description of the procedure used to measure the fading, the data acquisition and a sample of the measured fading.

In Chapter 5 we provide a detailed analysis of the collected data to meet the thesis objectives. Observations and results are presented. There are 10 sections including an introductory section on the computing requirements and a summary section which

highlights the main results of the chapter. The second and third section present an analysis of the envelope fading statistics and the cross-correlation results. In the fourth section we analyze the fading statistics of selection combined envelopes and examine the effects of antenna and frequency separation on diversity performance. The fifth section extends this analysis based on measurements in four different buildings and section 6 considers the variations of diversity performance. In section 7 we simulate two practical combining techniques with the measured data and in the eighth section we examine the performance of differential phase shift keying (DPSK) with the measured fading statistics. The second last section presents the application of the diversity results in a link budget design.

Conclusions are presented in the last Chapter 6. The main results and contributions of the thesis are summarized in the first section. In the second section we make recommendations for further research based on our results.

A bibliography follows and appendices present supporting information on the experimental procedure and analysis.

1.3. Thesis contributions

This thesis presents an empirical study of the performance of space and frequency micro-diversity techniques for the indoor radio channel at 1.75 GHz with applications to digital communications. An experimental apparatus has been developed to allow simultaneous four-branch measurements of envelope fading. Analysis software and batch programs have been designed for the evaluation of diversity techniques from large groups of measured data. Based on a literature review presented in Chapter 3, the author knows of no such previous investigation. General contributions from our analysis are as follows:

1. An empirical evaluation of three diversity techniques at 1.75 GHz to mitigate narrowband envelope fading. The antenna and frequency separations are determined for good diversity performance.
2. A direct comparison of two-branch space diversity and two-branch frequency diversity with four-branch space/frequency (hybrid) diversity based on simultaneous four-branch measurements.

3. An analysis of diversity performance differences in four buildings with different characteristics of open layout, soft partitions, plasterboard divisions and concrete walls.
4. The variations in envelope cross-correlation, fading statistics and diversity performance are studied. These results allow for a more accurate prediction of communication reliability.
5. Practical switched diversity combining techniques have been evaluated based on measured data. Simple diversity combiner implementations are required for low cost portable terminals.
6. The effect of the measured diversity performance on digital communications systems has been presented. The bit error rate (BER) is evaluated for DPSK with average and worst case fading statistics.

A detailed summary of contributions is presented in the final chapter, section 6.1.

Context of this study

The context and specific parameters under which this study was undertaken were as follows:

1. Micro-diversity was examined with the assumption that in a practical system a macro-diversity system would be used to mitigate the shadow fading.
2. Measurement areas included offices divided by concrete walls, plasterboard walls and soft partitions and open areas.
3. Antennas were quarter wavelength vertically polarized monopole antennas with ground plane and having omni-directional radiation patterns.
4. The antennas were moving at a velocity 0.5 metres per second.
5. Antenna spacings used were from 0.25 to 1.0 wavelengths.
6. Frequency diversity spacing was 1, 2, 5, 10 and 20 MHz.
7. Statistics of the envelope fading were computed based on a 60 second measurement period to approximate the average duration of a telephone call.

We restricted our study to CW envelope measurements for the following reasons:

1. For the narrowband communication systems under consideration, the envelope fading is the dominant effect on the channel.
2. For practical diversity combining systems such as selection combining and switch and stay, the branch selection is based on received signal strength indicators or envelope power.
3. Phase variations in the channel may be significant for coherent demodulation systems, however less so with differential detection which we are considering.
4. CW measurements provided significant information with a relatively simple apparatus.

Chapter 2

BACKGROUND ON INDOOR WIRELESS COMMUNICATIONS, RADIO PROPAGATION AND DIVERSITY TECHNIQUES

2.1. Portable communication systems

In this section we describe some new and future portable communication systems in which the existence of signal fading suggests the use of diversity techniques. In the experimental procedure and analysis in this thesis we consider diversity performance in the context of such systems.

Portable communication systems and their associated technologies are emerging rapidly in the wake of the success of cellular mobile phone systems. The business and market for these systems has grown fastest in Europe where the second generation digital Cordless Telephone system (CT-2) [16, 17] is now operational in the United Kingdom (as of 1989) and the Digital European Cordless Telecommunications system (DECT) is being planned for operation in 1992 [14].

The basic system concept and vision of universal portable communications is described in [1, 2]. Current systems are described as second generation following the first generation cordless telephone and telepoint in the UK. There are now standards and specifications for these systems which are discussed in [18, 16, 14, 17, 15]. Both CT-2 and DECT systems are aimed at three markets: residential, public (Telepoint), and business (PBX). Under each configuration the radio link design is different and the relative importance of the different channel impairments varies. In a public location, for example, the range of coverage will be large so signal levels must be maintained under fading and shadowing. In a high user density area such as a business office the effect of interference will likely predominate. System standards are chosen to reflect these requirements with the constraint that the user must have the equivalent quality of communications as a wire-line service. The literature suggests that a target for portable communication reliability is 99 percent, although this parameter may vary between service providers.

CT-2 System

This is a second generation system initiated after the first generation, CT-1, where spectrum was limited and roaming was not permitted. CT-2 permits inbound and outbound calls at Telepoint facilities and is less expensive than cellular telephone to the user. It is operational in U.K. as of 1989 and North America and Europe are adopting it as a standard for the first systems to address business office needs. Refer to table 2.1 for capacity, frequency, channels, and access scheme. CT-2 has been designed to support a 32 Kbps voice channel and has a total of 40 duplex channels. The design approach to obtain the goal of high capacity is similar to mobile cellular. The frequency re-use distance is made large enough to avoid significant co-channel interference (CCI). For initial frequency assignment, the channel access is based on a measure for CCI at call setup only. Standards and information on this system are found in [16, 18, 17].

DECT System

This is a third generation digital European cordless telecommunication standard. It is ISDN compatible and provides for a wide range of data and voice services. DECT should feature very high capacity and should provide "seamless" hand over between micro-cells. Refer to table 2.1 for a summary of capacity, frequency, and total number of channels. This system design is based on a micro-cellular matrix of frequency and time slots using time division multiple access (TDMA). Dynamic channel allocation (DCA) for variable data rates using many time slots could be supported. Papers on DECT implementation suggest that system should be designed for optimum CCI performance. Standards and information on this system can be found in [14, 19].

Table 2.1: Cordless telephone system parameters

Item	Units	CT-2	DECT
Implementation Date		1989	1992
Capacity Objective	ct/sqkm	5,000	50,000
Range	metres	< 200	< 200
Telephone service areas		Telepoint, residential, office, PBX	
Data Services (ISDN)		32 Kbps possible	variable rate
Frequency band	MHz	864 - 868	1880 - 1900
Max. no. of carriers/cell		40	11
Channel type		Time Division Duplex (TDD)	
Channel Access		FDMA	TDMA
TDD channels / carrier		1 at 32 Kbps	12 at 32 Kbps
Max. no. channels/cell		40	132
Channel allocation		Trunking	Dynamic
TDD frame rate	Hz	500	100
No. time slots/carrier		2	24
Portable avg. TX power	mW	10	10
Portable peak power	mW	20	240
RF channel bandwidth	KHz	100	1728
Modulation type		FSK 2-level	GMSK
Data rate/carrier	Kbps	72	1152

Belcore framework

Belcore in Red Bank, New Jersey has prepared a generic framework for portable communications systems [15] to create a dialogue between services providers and telephone equipment suppliers. Though this is not a standard it may have a significant impact on determining the shape of third generation PCS in North-America. The framework considers applications where tetherless access would be desired. This includes indoor and outdoor applications. Base stations may be positioned on telephone poles to allow radio access to homes or they may be placed inside large buildings. The system concept suggests time division multiplexing (TDM) for the base to portable link and TDMA on the reverse link. Proposed transmission rates are 225 kilo-symbols per second, with two bits per symbol. The frequency band for this type of system depends on government regulation in the United States, but, currently frequencies slightly below 2 GHz are being considered.

2.2. Multipath radio wave propagation

The variabilities inherent in communication channels can be usefully divided into two groups of "short term" and "long term" fading or into two other groups, "small scale" or "large scale" variations. Short term and small scale fading directly affect the received signal characteristics and hence affect the distributions of errors in a digital link. Large scale and long term variations determine the mean values of the small scale and short term variations respectively and they essentially control the long term or large area availability of the communication channel.

Small scale variations in propagation characteristics can, for the case of a moving radio terminal, create short term signal variations as can also be found in radio channels with internal disturbances. This can be modelled as a stochastic process and can be represented as a time varying filter as described in [20, 21] provided that the variations are slow compared to the symbol time of the digital modulation. Following the model described in [21] the impulse response of the time varying multipath channel can be

described as:

$$c(\tau; t) = \sum_n \alpha_n(t) e^{-j2\pi f_c \tau_n(t)} \delta[\tau - \tau_n(t)]$$

where :

$\alpha_n(t)$ is the attenuation factor for the n^{th} path

$\tau_n(t)$ is the propagation delay for the n^{th} path

f_c is the carrier frequency

(1)

Multiple paths change in electrical length and in signal strength as scatterers move or internal effects in the channel vary. The varying phases ($2\pi f_c \tau_n$) of all received components result in constructive or destructive vector addition and they are very sensitive to propagation delays, τ_n , because the carrier frequency is usually large.

The multipath channel transfer function is the Fourier transform of $c(\tau; t)$ and is denoted by $C(f; t)$

$$C(f; t) = \int_{-\infty}^{\infty} c(\tau; t) e^{-j2\pi f \tau} d\tau$$
(2)

A received multipath signal $r(t)$ can be considered a vector sum of the signals from all paths with varying phase over 0 to 2π . For the case of an unmodulated carrier without noise, the equivalent lowpass received signal can be expressed as

$$\begin{aligned} r(t) &= \sum_n \alpha_n(t) e^{-j2\pi f_c \tau_n(t)} \\ &= \sum_n \alpha_n(t) e^{-j\theta_n(t)} \end{aligned}$$
(3)

where θ_n is the phase shift due to the product of the delay and the carrier frequency.

A derivation based on the central limit theory concludes that the received signal, in the absence of noise, looks like a narrow band of gaussian noise that can be represented by quadrature gaussian components. The classical derivation shows that the impulse response envelope $|c(\tau, t)|$ is distributed according to a Rayleigh distribution and that the phase is uniformly distributed over 0 to 2π . This "Rayleigh fading" fits many measured results closely and allows relatively simple analysis of multipath fading. However, not all short term fading fits this distribution such as the case where one of the signal paths

carries a much stronger and dominant component. This case has been shown to fit well to a Rician distribution. Other cases of more severe fading can be modelled with a Nakagami distribution.

Rayleigh fading envelope distribution

The probability density function (pdf) for a Rayleigh distribution is derived from the central chi-square distribution [21] and has the form

$$p_{\alpha}(x) = x/\sigma^2 e^{-x^2/2\sigma^2} \quad x \geq 0 \quad (4)$$

where the random variable α is the square root of the sum of the squares of two gaussian random variables with zero mean and variance of σ^2 . The cumulative distribution function (CDF) is:

$$P(\alpha \leq x) = 1 - e^{-x^2/2\sigma^2} \quad x \geq 0 \quad (5)$$

Frequency selective fading

Correlation functions and power spectral densities described in [21] provide useful descriptions of channel characteristics that determine the type of distortion on a digital signal. Assuming uncorrelated scattering, the auto-correlation function of $c(\tau;t)$ can be written as $\phi_c(\tau;\Delta t)$. If $\Delta t = 0$, $\phi_c(\tau)$ is the average power output of the channel as a function of the time delay and is known as the multipath intensity profile. A multipath spread can be defined as the range of time delay over which $\phi_c(\tau)$ is non-zero and will be written as T_m . In channels where path lengths differ greatly, T_m can be very large causing inter-symbol interference (ISI). A similar auto-correlation in the frequency domain of the transfer function $C(f;t)$ is expressed as $\phi_C(\Delta f;\Delta t)$. When $\Delta t = 0$, $\phi_C(\Delta f)$, is the spaced frequency correlation function and when $\Delta f = 0$, $\phi_C(\Delta t)$, is the spaced time correlation function. The values of Δf over which $\phi_C(\Delta f)$ is non-zero is loosely termed the "coherence bandwidth" of the channel, $(\Delta f)_c$. Due to the inverse Fourier transform relationship with $\phi_c(\tau)$, the following approximation is valid

$$(\Delta f)_c \approx 1/T_m \quad (6)$$

Similarly, a “coherence time” of the channel, $(\Delta t)_c$, is defined as the range of Δt over which $\phi_C(\Delta t)$ is greater than zero. If we assume that a digital signaling waveform has a period T_s , and a corresponding message bandwidth W ; the fading is described as frequency selective if $W \gg (\Delta f)_c$ and the signal is “wideband”. If the converse is true, $W \ll (\Delta f)_c$, then the signal will experience frequency non-selective, or “flat” fading and can be called a “narrowband” signal. Considering the case of an unmodulated carrier with a flat fading channel, the received signal can be represented by

$$\begin{aligned} r(t) &= C^*(0;t) \\ &= \alpha(t)e^{-j\phi(t)} \end{aligned} \quad (7)$$

where $\alpha(t)$ represents the envelope and $\phi(t)$ is the phase of the equivalent lowpass channel. By the same approach described to model the impulse response, the complex valued narrowband transfer function $C(0;t)$ can be modeled as a zero mean gaussian process in which the envelope $\alpha(t)$ is Rayleigh distributed and $\phi(t)$ is uniformly distributed over 0 to 2π .

Doppler spread

Time variations of the channel parameters and of the auto-correlation function, $\phi_C(\Delta f; \Delta t)$, can be expressed by the Doppler power spectrum of the channel $S_c(f_d)$ which is the Fourier transform of $\phi_C(\Delta f; \Delta t)$. The values of Doppler frequency (f_d) over which $S_c(f_d)$ is non-zero is the Doppler spread (B_d) which is double sided and is inversely related to the coherence time, $(\Delta t)_c$. Radio channels associated with fast moving terminals change quickly and have a large Doppler spread. For a moving terminal at a velocity, V , the maximum Doppler frequency shift is:

$$f_{d_{max}} = V/\lambda \quad (8)$$

where λ is the wavelength of the carrier frequency.

Frequency non-selective slowly fading channels

For a digital communication with symbol period T_s , the channel is “slowly fading” if $T_s \ll (\Delta t)_c$, ie: the channel does not vary during the time of one symbol interval. If

a further constraint is added such that $T_m \ll T_s \ll (\Delta t)_c$, the transmission experiences no ISI due to delay spread and no distortion due to channel variation during one symbol period. This is known as a frequency non-selective slowly fading channel and it is the simplest of fading channels to analyze. The constraint for this can be summarized, for the case of $W \approx 1/T_s$, by the relation

$$T_m B_T \ll 1 \quad (9)$$

and is also known as an under spread channel. Signal processing in a receiver is simplified when operating over an underspread channel.

Clarification of time scales used for averaging

The notion of short-term and long-term variations in the propagation channels is useful and at this point we elaborate on time scales for future reference. Following [3], three time scales can be considered. Short intervals of length T are intervals that are short in comparison to the length of time it takes the channel to change significantly but are long compared to a symbol interval. In the context of fading, the parameters that are averaged over this interval are described by the adjective “local”. Intermediate intervals, T_1 , are long in comparison to the time it takes for the channel to vary significantly. It is variations within the period of time, T_1 that directly affect the distribution of errors over a digital link and whose envelope fluctuations can often be modelled by the Rayleigh, Rician or Nakagami distributions. The mean values in these distributions are averaged over T_1 . In the context of moving terminals, the period T_1 should be long in comparison to a Doppler cycle (the inverse of the maximum Doppler frequency, $f_{d_{max}}$). For example, at a velocity of 0.5 m/s and carrier frequency of 1.75 GHz we have $f_{d_{max}} = 2.9 \text{ Hz}$. A period of $T_1 = 60$ seconds would represent 175 Doppler cycles and, for the case of Rayleigh fading, there would be 30 deep fades at a level 20 dB below the average. Variations of T_1 averages are observed over a longer period T_2 which would be long enough for the surroundings and the channel to change, for example, a few 10's of minutes.

2.3. Diversity techniques for digital communications

These techniques all provide some form of signal redundancy to overcome deep signal fades. The approaches can be divided into two categories: micro-scopic diversity to combat small scale and short term fading and macro-scopic diversity to mitigate the variations in large area signal strength. This review deals with the first technique of micro-diversity usually under the assumption that large scale variations in the mean signal are either not significant or that the variations have been reduced by using some form of macro-diversity. Discussion is restricted to diversity operation over frequency non-selective slowly fading channels.

2.3.1 Micro-scopic diversity types

Diversity operation relies on the fact that two or more independently fading signals will rarely be in a deep fade at the same time. Thus it requires more than one transmission path with relatively uncorrelated fading. By simply selecting the best signal or by adding the signals at the receiver, the availability can be increased dramatically. There are several methods to provide multiple independently fading replicas of the transmitted signal and these are listed here. These approaches are all useful for a frequency non-selective slowly fading channel.

Frequency Diversity

This technique is commonly used on high frequency (HF) and terrestrial microwave with some use in the land mobile channel. Two or more paths can be uncorrelated providing the frequencies are separated by more than the coherence bandwidth of the medium. Recent portable communications standards and proposals [14, 15] which use time division multiplexing (TDM) and frequency division multiplexing (FDM) have structures which allow the use of frequency diversity.

Space diversity

This is used on HF, mobile, indoor, and terrestrial links. It is very popular since it requires no extra bandwidth, power or time delay. Limitations are usually based on

the physical size of the antennas. In general the antennas must be spaced enough to have low a correlation of envelope fading (< 0.7) between signals. The required spacing is a function of the proximity of the nearby scatterers and the spread of arrival angles of the various signal paths. With nearby scatterers the required separation is usually smaller [4, 22].

Polarization diversity

Generally this approach requires a separate transmitter on each polarization and has been shown effective in mobile communications [22]. However, when significant depolarization of the signal occurs during transmission, as shown for the indoor channel, a single transmitter can be used with a dual polarized antenna at the receiver.

Angle diversity

This method relies on a directive antenna that allows reception of multipath rays arriving from a limited number of directions only as shown by [4]. At any one time it is unlikely that all signals in all directions will be in a deep fade.

2.3.2 Combining techniques

In this section the methods of combining the multiple diversity paths are reviewed. These methods all have the same objective, that is, to improve the availability or signal to noise ratio (SNR) of the signal. The various techniques differ in the trade off between performance and complexity.

General form of linear combiner

The general linear combiner can be expressed in the form:

$$r(t) = a_1 r_1(t) + a_2 r_2(t) + \dots + a_L r_L(t) = \sum_{k=1}^L a_k r_k(t)$$

where :

$r(t)$ is the combined received signal at time t

a_k is the multiplying coefficient on the k^{th} branch

$r_k(t)$ is the signal on the k^{th} branch at time t

L is the number of diversity branches

(10)

In the above equation the received signal on each branch is composed of a signal term and a noise term such that:

$$r_k(t) = s_k(t) + n_k$$

where

$s_k(t)$ is the signal envelope

n_k is the noise voltage

(11)

$s_k^2/2$ is the local mean signal power per branch

$\overline{n_k^2}$ is the mean noise power per branch

$\gamma_k = \frac{s_k^2}{\overline{n_k^2}}$ is the local SNR on the k^{th} branch

Different linear combining techniques can be realized by adjustment of the complex valued multipliers on each branch. Three well known techniques are selection combining (SC), equal gain combining (EGC) and maximal ratio combining (MRC). SC is the simplest of the three techniques and MRC is the most complex. MRC is also the best performing technique of these three. With EGC the weighting coefficients are all equal to unity. The signals must be co-phased which is usually done with a phase locked loop. The MRC technique weights each coefficient proportionally to the SNR on that branch and also requires co-phasing of the signals.

Selection combining

With reference to equation 10 the selection combining (SC) technique chooses the n^{th} branch such that:

$$r_n \geq r_k \quad \text{for } k = 1, 2, \dots, L$$

$$\text{Therefore: } a_k = \begin{cases} 1 & \text{for } k = n \\ 0 & \text{for } k \neq n \end{cases} \quad (12)$$

The selecting of the best branch is typically done based on the value of a received signal strength indicator (RSSI) on each branch which may be available from an automatic gain control (AGC) system. These methods actually choose the branch with the largest r_k (signal plus noise) instead of the preferred SNR. An alternative approach is to select a branch based on an error monitoring scheme which is indirectly a measure of SNR.

Theoretical fading distribution of envelope power with SC diversity

The CDF of the power of a Rayleigh fading envelope is useful for comparisons in later sections. Let γ_k represent the SNR on the k^{th} branch of a selection combiner with Rayleigh envelope fading described by equation 4. Since the signal power is proportional to the square of envelope it follows that the cumulative distribution function of γ_k on the k^{th} branch is

$$P(\gamma_k \leq x) = \int_0^x 1/\Gamma \cdot e^{-x'/\Gamma} dx' = 1 - e^{-x/\Gamma} \quad (13)$$

where Γ is the average SNR on one branch

and for the selection combined SNR of L branches (γ_c) the CDF is

$$P(\gamma_c \leq x) = P[\gamma_1, \gamma_2, \dots, \gamma_L \leq x] = \prod_{k=1}^L P(\gamma_k \leq x) \quad (14)$$

$$= [1 - e^{-x/\Gamma}]^L$$

These CDFs are plotted in Fig. 2.1 for one-, two- and four-branch diversity assuming independent fading on each branch.

The average SNR of an L -branch SC output can be computed by the standard approach used to find the expected value of a random variable, that is, integrate the product of the pdf and SNR over all values of SNR. This has been shown in [3] to be:

$$\Gamma_c = \Gamma \sum_{k=1}^L 1/k \quad (15)$$

where Γ_c is the average combined SNR

Figure 2.1: CDF of Rayleigh distributed envelope power with SC diversity assuming two and four branches with independent fading.

Time availability

A useful statistical measure of radio communication system performance over fading channels is the outage rate or the time availability. The availability is the percentage of time or physical area in which the received signal level is above a threshold for good communications. In our discussion we refer to the received signal level referenced to its mean value during the measurement duration. For a Rayleigh distributed envelope as shown in Fig. 2.1 the signal power level will be above a value which is 20 dB below its mean level about 99% of the time, that is, the fade depth is -20 dB at 99% availability. This corresponds to an outage rate of 10^{-2} which is often specified for indoor communications [15].

Diversity gain

If there is a specified system outage rate or availability requirement, it is useful to evaluate the performance of diversity on this basis. Usually this is quantified by measuring the amount we can reduce the transmitter power and achieve the same availability compared to a system with no diversity. This is commonly referred to as a "diversity gain" or more precisely as a "local reliability diversity gain". This can be read directly from the CDF of the SNR or envelope fading distributions as shown in Fig. 2.1 for SC.

The diversity gains for SC, EGC and MRC are summarized in table 2.2 for Rayleigh fading at an availability of 99% [3]. It is evident that the extra gain achieved by adding a branch decreases as the number of branches is increased. There is also a clear performance advantage with the higher complexity techniques. Despite these advantages the two techniques of EGC and MRC are excluded from further analysis in our study because of the need for co-phasing the signals which would add significantly to complexity. Since our concern in this thesis is with practical diversity techniques that could easily be implemented on a hand held telephone, we consider only selection combining and various forms thereof.

Switched diversity

Sub-optimal selection combining techniques are used for their simplicity. A few desirable qualities of these techniques for use in portable radio communications are:

1. practicality of implementation
2. combine signals in the front end of receiver
3. small decision delay
4. small amounts of processing required
5. select based on average power or some other average metric such as mean BER

A practical and simple implementation of selection combining is switched diversity in which multiple diversity antennas are connected via a switch to just a single receiver. This technique can also reduce the rate at which we change between branches as compared with pure SC. A generic two-branch switched scheme is illustrated in Fig. 2.2.

Instead of directly comparing the signals on each branch, the circuitry switches branches when the current signal level falls below a certain threshold. There are various algorithms that can be implemented for this structure. Switch and stay selection combining (SAS) is a type of switched diversity in which a branch is switched out whenever the signal level or signal quality indicator drops below a pre-defined threshold. That is, it switches on negative threshold crossings of the envelope to another branch which is likely above threshold. If, however, it switches to a branch which is already

Table 2.2: Diversity gains in dB at 99% availability with Rayleigh fading

L (# of branches)	Selection	Maximal Ratio	Equal Gain
2	10	12	11
3	14	16.5	16
4	16	19.5	18.5
6	18	22.5	21.5
8	19	24.5	23.5

Figure 2.2: Block diagram of switched diversity technique

below threshold then it stays. The advantage in staying is to reduce the number of switching transients and, in a two-branch combiner, there would be little gained by switching back to the original branch. This strategy is analyzed in [4] for the case of independent Rayleigh fading and equal local mean signals on each branch. In conditions of fading where the mean of the signals (over period T_1) is constant, the threshold may be fixed to a value slightly above the operating threshold of the receiver. In cases where there is a large scale variation present the threshold should be changed based on the most recent estimate of the mean signal level. A small variation on this strategy is switch and examine (SAE) in which the circuit switches branches again if it switches to a branch below threshold. Its advantage is that it will perform slightly better than SAS in terms of probability of being in a fade, but, it will switch more often.

A switched diversity scheme that can be used in a time division multiplexed (TDM) system [15] is described here. In this two-branch technique the first branch is switched in during one timeslot and the second branch during the next timeslot. An average quality metric is computed and compared for the two branches which results in a selection of the branch to use for the third timeslot. Performance of such a scheme would depend on the length of the timeslot and the rate of fading.

The distribution of the envelope α_c at the output of the SAS combiner has been derived for the case of Rayleigh fading [4] and is of the form:

$$P(\alpha_c \leq x) = \begin{cases} (1+q)P(\alpha \leq x) - q & x > A \\ qP(\alpha \leq x) & x \leq A \end{cases}$$

where

$$A \text{ is the switching threshold} \quad (16)$$

$$q = P(\alpha \leq A)$$

α is the Rayleigh distributed envelope on each branch

The CDF of the switched combined output is equal to that of SC in equation 14 only at the threshold "A". Below this value it has the slope of the Rayleigh CDF and above A it degrades to a single branch Rayleigh CDF.

Due to the two way nature of the communications a further reduction in complexity in the portable can be achieved by using "feedback diversity" as illustrated in Fig. 2.3.

The antenna switching function can be situated at the base station. The portable terminal only requires circuitry to decide if the received signal meets a predefined quality. It then relays this information back to the base station where the transmitter switches to another antenna. This can yield effective diversity reception for the outbound link. A key parameter controlling the performance of this system is the time delay between when a low signal is detected at the portable and when the signal from the new transmit antenna is received. A delay may be due to propagation, averaging period, or a frame length in a time division duplex (TDD) system.

2.3.3 Practical issues and impairments

Correlated signals on branches

In general, an increase in the envelope cross-correlation coefficient (ρ_{α}) between diversity branches will decrease the absolute performance of all diversity combining schemes. However the performance on an outage basis decreases slowly with an increase in ρ_{α} . Among others, Jakes [4] has shown that significant diversity gains using SC can be achieved with $\rho_{\alpha} = 0.8$. Assuming Rayleigh fading at a required availability of 99% the SC diversity gains are equal to 10.5, 8.5, and 7.0 dB for values of $\rho_{\alpha} = 0, 0.6$, and 0.8 respectively. The CDF for SC with envelope correlation coefficient ρ_{α} is presented in [22] as follows:

$$P(\gamma_r \leq x) = \frac{x^2}{\Gamma^2(1 - \rho_{\alpha})} \quad \text{for } x \ll 1 \quad (17)$$

Unequal local mean signal powers on each branch

Given L-branch SC diversity, if M signals are significantly below the local mean signal on the other branches, then the outage performance is equivalent to that of an (L-M)-branch combiner [22]. For two-branch SC there is considerable degradation from the ideal case of equal mean carrier to noise ratios on each branch ($\Gamma_1 = \Gamma_2$) where the diversity gain for Rayleigh fading is 10.5 dB. For an availability of 99% the degradation

Figure 2.3: Block diagram of feedback diversity technique

is as follows: with $\Gamma_1 = 2 \times \Gamma_2$ the diversity gain is reduced to about 9 dB and with $\Gamma_1 = 10 \times \Gamma_2$ it is only 5.5 dB.

For a two-branch SC with unequal power on each branch a modified CDF of the combined SNR has the form:

$$P(\gamma \leq x) = 1 - e^{-(x/\Gamma_1)} Q(a, \sqrt{\rho_v} b) - e^{-(x/\Gamma_2)} [1 - Q(\sqrt{\rho_v} a, b)]$$

where

$$a = \sqrt{\frac{2x}{\Gamma_2(1 + \rho_v)}}$$

$$b = \sqrt{\frac{2x}{\Gamma_1(1 - \rho_v)}} \quad (18)$$

$$Q(a, b) = \int_b^{\infty} e^{-(0.5)(a^2 + x^2)} I_0(ax) x \, dx$$

ρ_v is the envelope correlation coefficient

I_0 is the zeroth order modified Bessel function

Given this analytical result based on the different values of the mean signal on each branch, the practical issue is to determine what this difference is for a given diversity configuration. Space diversity systems are likely to exhibit unequal mean signals due to antenna pattern variation. As shown in [22] these variations can be up to 5 dB for omnidirectional base station antennas. However, these effects are not so great for mobile terminals where the antenna patterns are affected by local scatterers and mounting methods. To accurately predict the diversity performance degradation due to this effect, we require statistics of the mean signal levels.

Switching effects and impairments in SC

The effect of switching delays has been investigated in the context of feedback diversity from a mobile to the base station in [4]. The switching delay arises from the time it takes the mobile/portable unit to communicate the need to switch back to the base station where the diversity transmit antennas are located. In a land mobile system this may be dictated by the propagation delay time whereas in an indoor radio communications system this may be set in the protocol for a time division duplex (TDD) scheme or a

TDMA scheme where delays may range from 2 msec to 10 msec. The tolerable delay for a switched diversity system depends on the switching threshold (A) and the average fade duration (τ_q). For Rayleigh fading, the fade duration is:

$$\tau_q = \frac{\sigma}{\sqrt{\pi}} \times \frac{e^{A^2/2\sigma^2} - 1}{Af_{d_{max}}} \quad (19)$$

where : $f_{d_{max}}$ is the maximum doppler frequency
 σ^2 is the variance

If the switching delay (T_{sd}) is too long then the signal at the receiver may drop too far (below detection threshold) before the transmit antenna is switched. From [4] it is observed that, for Rayleigh fading, there may still be some diversity advantages up to a delay of

$$T_{sd} < 0.29 \times \tau_q \quad (20)$$

and that for $T_{sd} < 0.14 \times \tau_q$ there is no effect on the diversity gain.

Switching transient noise has been studied in [4]. Both amplitude and phase transients are reported. The phase transients can be assumed uniformly distributed over $-\pi$ to $+\pi$. The switching rate of the transient noise is dependent on the average level crossing rate (LCR) and the transient duration is inversely proportional to the intermediate frequency (IF) filter bandwidth.

Chapter 3

CURRENT RESEARCH AND THESIS MOTIVATION

3.1. Current research on indoor radio propagation and diversity performance estimation

3.1.1 Indoor radio channel measurements and models

Indoor radio propagation suffers from severe attenuation of the signal through walls and obstructions causing shadowing and also from multipath induced small scale variations in the channel due to movement of people or movement of the terminals. Typical transmission distances vary from 5 m to 200 m. Large scale path loss with distance is significant and is usually expressed in terms of a distance/power law where attenuation is modelled as decaying with distance to an exponent “N”. The small scale or short term channel variations are measured over a period T_1 of a few tens of seconds for moving terminals to a few minutes for stationary terminals. For a terminal moving at a velocity of 1.5 m/s, with frequencies of 0.9 and 1.7 GHz, the maximum Doppler frequencies are 4.5 Hz and 9 Hz respectively.

Alexander's work in the early 1980's [23, 24] studied radiowave propagation in buildings at 900 MHz. He reported on signal strength distributions and on average signal strength decay with distance (distance/power law). Using a random motion throughout steel walled buildings he found many signal strength distributions had a fairly close fit to a Rayleigh distribution. He found a gradient for this distance/power law of $N=5.7$ when varying the y-intercept and the slope to obtain a best fit line which had a fit within ± 10 dB. His 1983 paper reported on the gradients values for various buildings which varied between $N=6.5$ and $N=1.2$.

Bultitude [25] reported pathloss measurements at 900 MHz in hallways where the mean power varied with distance to the power of -1.8 , close to free space loss. For stationary terminals, temporal qualities were measured. Fades lasted 20 to 40 seconds due to movement of personnel and fade depths were down to -30 dB from median level. The CDF of envelope fading was fitted to a Rician distribution with K ratios of -6.8 dB or

-11 dB depending on the building. The value K defines the ratio of the dominant signal to diffuse ones. The worst Rician case is Rayleigh fading with $K = -\infty$ dB. It was suggested that the movement of people perturbs the spatial distribution of the signal level so that it varies in time and that the signal received during a quiescent period represents the dominant signal in a Rician distribution and the signals received during movement of people are the diffuse components. Based on measured time correlation functions the value of the coherence time was approximately 4 seconds, or equivalently, the Doppler spread for a stationary terminal is 0.25 Hz.

Owen and Pudney conducted measurements of pathloss indoors at 1700 MHz and 900 MHz [26] to determine the effect of walls and floors on the pathloss. Using a distance/power law with a correction factor for wall losses they found that both frequencies had a pathloss gradient of 3.5 but the 1 metre intercept points were 29 dB and 23 dB for the high and low frequency respectively. They found that propagation loss at 1700 MHz through walls was 3 to 4 dB for double plasterboard and 7 to 9 dB for breeze block or brick. Note that they allowed the 1 metre intercept point to vary to obtain a best fit. Since the intercept points are not the same as those used by Alexander, we cannot make a direct comparison of the gradient between the two results.

Rappaport and McGillem [27] conducted narrow band measurements in factories at 900 MHz. In this work they constrained the intercept point to be the value of free space loss at 10 wavelengths. They reported that path loss is dependent on local surroundings and is log-normally distributed. Obstructed paths had loss variation with distance to the exponent $N = 2.4$ to 2.8 . Small scale spatial fades of down to -30 to -35 dB were measured over a few wavelengths. Temporal fading was Rician distributed and small-scale signal fluctuations due to receiver motion were Rayleigh distributed except in line of sight (LOS) cases where they were Rician with an equivalent K ratio of -2 dB. In heavy clutter areas, the statistics of the envelope fading fit well to ratios of $-7 < K < -4$ dB. They found that there is little correlation between transmitter and receiver separation and small-scale fading distributions because of the proximity of the scatterers.

Saleh and Valenzuela [8] measured delay profiles using a radar like transmitter. They proposed a multipath ray model for the temporal channel and concluded that the 'door

channel is quasi-static or very slowly time varying due to people's movements as was determined in [25]. They also noted that the nature and the statistics of the channel's impulse response is virtually independent of the states of polarization of the transmitting and receiving antennas if no LOS path exists. In other words there was considerable cross coupling of signals due to the scattering. The distance/power law varied from $N = 3$ to 4. The maximum observed delay spread was 100 to 200 ns in rooms and up to 300 ns in hallways, and the measured rms delay spread within rooms had a median value of 25 ns and a maximum of 50 ns. The model proposed that the rays of the received signal, from reflections off of objects, arrive in clusters and that the received ray amplitudes are independent Rayleigh random variables with variances that decay exponentially with cluster delay as well as with ray delay within a cluster. Phase angles are independent uniform random variables over 0 to 2π .

A paper by Keenan and Motley [28] provided another analysis of large scale pathloss in buildings with correction factors for wall loss and floor loss.

Blanco [29] proposed a multirician fading model for mobile radio communication systems which is applicable to indoor radio. Based on the number of major scatterers in the model, the received signal consists of a finite number of specular components in addition to a diffuse component. It was shown that this may create fading deeper than Rayleigh fading. It is a general model that includes Rayleigh and Rician as special cases.

3.1.2 Diversity performance estimates based on theoretical channel models and simulation

There have been considerable efforts in the analysis of diversity techniques for general fading channels and simulations of diversity systems for mobile radio communications. In a classic paper on linear diversity combining [3] Brennan derived the performance of various combining techniques assuming a Rayleigh fading channel. This paper published in 1959 is quite general and may apply to different radio channels but refers most often to the high frequency (HF) radio channel. He also considered some effects of non-Rayleigh fading distributions and correlated fading on diversity performance. Relevant points from this were discussed in Chapter 2. Another classic paper published in 1960 by

Stein [30] considered the problem of estimating diversity performance for tropospheric scatter propagation. This has many useful discussions on diversity in general. A book by Jakes published in 1974 [4] analyzes the performance of diversity for the mobile radio channel and provides a detailed analysis of diversity techniques based on analytical work, and some measurements. The analysis in this book is useful for estimating diversity performance for the indoor radio channel because of its detail and because there are some similarities between the mobile and indoor channel.

Analytical and simulation results for indoor radio diversity systems are few in number. A 1983 paper by Cox [6] examined antenna diversity performance with random handset orientation for the indoor/outdoor environments using computer simulation. The multipath propagation model used here is almost the same as that used by Jakes [4] for the mobile radio environment. Winters [31] studied the capacity of an indoor wireless communication system using diversity and assuming a Rayleigh fading channel using Monte Carlo computer simulations. Beaulieu considered the performance of switched diversity systems on Nakagami fading channels [32].

There are a number of other works which consider the use of diversity coupled with an adaptive algorithm, equalization or coding for indoor radio systems. Three of these are [33, 34, 35].

3.1.3 Measurements to evaluate diversity performance in buildings

Motley and Alexander published a paper in 1983 [10] describing the results of an experiment to determine the performance of space diversity applied at a base station for 900 MHz cordless telephones indoors. Measurements were made with a portable receiver moving randomly in various rooms in one building. They computed signal strength distributions and estimated mean BER for the system in question. The average diversity gain was approximately 10 dB at 99% availability for an antenna separation of $\lambda/4$.

Bergmann and Arnold published a short paper in 1986 [11] reporting on measurements of polarization diversity performance in buildings using a microstrip patch antenna with horizontal and vertical polarizations. They analyzed signal strength for line of sight (LOS) and non-LOS cases and computed the correlation coefficients between two

branches. Results indicated that correlation coefficients were low enough to allow good diversity operation. They also plotted histograms of these correlations which showed a large variation from -0.4 to $+0.8$.

Arnold, Cox and Murray published a paper in 1988 [10] in which they investigate the performance of macro-scopic diversity to mitigate shadow fading in buildings. Their results based on many previous measurements indicated that three branch macro-scopic diversity can reduce the link margin required for 99% area coverage by 10 dB.

Tsunekawa published a paper in 1989 [36] that describes a very compact pair of planar inverted-F antennas (PIFA) to provide diversity on a hand-held portable unit. The application was for mobile (outdoors) communications, however, due to the small size such an approach may be useful for indoor communications. The analysis of the diversity performance consisted of average correlation calculations. (This did not consider the effect of variable mean signal levels between the two branches).

In a masters thesis by Lemieux in 1990 [12] the performance improvement of space, frequency and polarization diversity was measured and analyzed. This work was conducted for the 900 MHz band in one engineering building in three sets of measurements for each diversity type. The objective was to evaluate diversity performance based on narrowband CW fading signal strength by computing the cumulative distribution function (CDF) of the signal envelopes. Parameters for space and frequency diversity were varied to determine the difference in performance. Results based on averaged distributions indicated that frequency diversity required frequency separation (ΔF) of 10 MHz to obtain maximum improvement. For the antenna spacings between 1 and 2 wavelengths the average space diversity performance was close to that predicted by a Rayleigh fading distribution, that is, an improvement of 10 dB at a signal availability of 99%. Polarization diversity was measured with two monopole antennas, one horizontal and the other vertical. This showed poorer performance than the previous types.

In a subsequent paper by Lemieux et al [13] the performance of the three diversity schemes are compared based on the previous measurements. It is concluded that space diversity operates well for $\Delta S > 3/4 \lambda$ and frequency diversity operates well for $\Delta F = 5$ MHz and no further improvement is found for $\Delta F > 10$ MHz.

Recent papers published by Todd, El-Tanany and Mahmoud [37, 38, 39, 40] have investigated the performance of diversity techniques at 1.75 GHz for indoor digital radio communications.

3.2. Analysis of current research

In general the previous indoor radio channel modelling has been aimed at developing a channel model however this has not been directly aimed at predicting diversity performance. Theoretical models and simulations have made significant progress in estimating *average* diversity performance [3, 4, 5]. There has also been some extensive simulated results [6], however, they have not addressed the issue of what diversity parameters to use for the indoor radio channel at 1.7 GHz, nor the issue of variability of fading with and without diversity techniques. There is also a lack of direct comparisons between two-branch diversity schemes.

With respect to previous diversity measurements we summarize a few key points:

1. There have been measurements to determine operating parameters for space and frequency diversity at 900 MHz [12] but not in the 1.7 GHz band.
2. These previous measurements [10, 11, 12, 13] have not considered the worst case diversity performance and the effect of this worst case on the BER of digital modulation techniques.
3. There have been no previous four-branch measurements.
4. There have been no previous direct comparison between space and frequency diversity or between two- and four-branch diversity. If measurements of space and frequency diversity are made at different times as in [12] then the channel will be different and one cannot guarantee that there will be the same temporal changes in the channel unless this is planned. Thus simultaneous measurements are the best way to compare the diversity types.
5. The antenna spacing measured has previously been too large for a portable terminal. Previous work by Lemieux et al [13] used a minimum value of $3/4 \lambda$ which at 900 MHz is 25 cm. Work by Motley and Alexander [10] used a $\lambda/4$ spacing at 900 MHz which is 8.3 cm and is also fairly large for a

portable telephone. Work presented in [36] used small antenna spacings yet these results were for mobile radio and considered only average correlation results.

3.3. Thesis problem statement

Required information for incorporating diversity into system design

For a portable communication system design employing diversity, the following information is required:

1. An estimate of average diversity performance with fading statistics over durations of interest ie: 1 to 2 minutes.
2. The parameters to obtain this performance.
3. A performance comparison between diversity types.
4. Variations of diversity performance.
5. Trade-off complexity versus performance (two-branch versus four-branch and switched diversity versus SC).
6. Estimates of how these results effect the BER of digital modulation techniques.

The problem

In this thesis we address the problem of estimating realistic diversity performance for portable indoor radio communications at 1.75 GHz. To obtain "realistic" estimates of diversity performance we rely on measured data from multiple diversity branches rather than on radio channel models. We are specifically concerned with diversity performance for moving radio terminals. The problem encompasses the design of radio channel measurement equipment, the collection of channel fading data and most importantly the data analysis to extract relevant information on diversity performance.

Approach to solution

The solution to the problem is comprised of three steps:

1. **Experimental design:** This step comprises design and testing of radio measurement equipment, design of calibration software and data file storage method and the development of the experimental procedure to measure fading.
2. **Measurements:** This involves data collection using a diversity transmitter and receiver in over 135 locations in five different office areas. Measurements are made during times of typical activity, in various rooms and in four different buildings under conditions that approximate the use of a portable telephone.
3. **Data analysis:** Software is written to analyze data from similar groups of measurements. The data analysis was as follows. Two-branch space diversity is directly compared to two-branch frequency diversity and to four-branch hybrid diversity. We perform simple selection combining (SC) in software after data collection. Envelope cross-correlations are calculated and envelope fading of diversity is measured for various frequency separations and antenna spacings which would be applicable given the physical size of portable telephones and data terminals. The distributions of correlation, diversity gain and fading statistics at different locations are also investigated. The diversity output statistics are applied to a digital modulation technique and various sub-optimal but practical selection combining strategies were simulated. The analysis is concluded with an application to system link design.

Benefits of solving problem

It is important to solve this problem for the following reasons:

1. Due to importance of 1.6 – 2.0 GHz band to next generation wireless systems.
2. To have direct comparisons of three diversity types space, frequency and combined space/frequency diversity.

3. The diversity parameters have a significant impact on system design:
 - a. Frequency separation has direct effect on frequency coordination and total system bandwidth.
 - b. Antenna spacing has major impact on portable terminal design, on antenna design and overall approach to implementing micro-diversity, i.e.: do we use two antennas at just the base station or also on the portable?
4. For high reliability and quality communications (equivalent to wired telephones) we should know the worst case performance, *not just the average*.

There are secondary benefits for solving this problem:

1. It is important to consider practical combining techniques that have low complexity and require low amounts of processing.
2. It is useful to gain some insight into the envelope correlation between diversity channels for work on more advanced combining techniques.
3. It is useful to know how this impacts on the system link budget.

Chapter 4

EXPERIMENTAL PROCEDURE AND APPARATUS

4.1. Introduction

The indoor environments are described, the equipment design is summarized and measurement procedure is explained. The apparatus and procedure described here were used to collect data of simultaneous envelope fading on four diversity branches. The results reported here are for the case of moving receiver antennas for both obstructed (OBS) and line of sight (LOS) paths.

There were in total 428 one minute measurements in various buildings and locations. The first 328 measurements were conducted in the primary measurement area for a detailed analysis of diversity with varying parameters. There were multiple measurements made over each transmission path. These were collected from February 18 to April 4, 1991 and represented 150 megabytes of data stored in a binary format. The secondary group of measurements were conducted in four separate buildings with a fixed set of parameters. There was one measurement made over each of the 100 paths. The data was collected over a duration of two weeks in July 1992. The data occupied approximately 50 megabytes of memory in binary format. In total the data occupied 200 megabytes in a binary format and over 800 megabytes when converted to real number format suitable for analysis.

4.2. Indoor environment

4.2.1 Primary measurement area

This experiment was conducted on one 30×45 square metre (m^2) floor of an engineering building at Carleton University which is illustrated in Fig. 4.1. This is named the engineering annex. Measurements were made in 10 rooms and 3 hallways of different shapes and sizes. The room sizes varied from $13 \times 13 m^2$ to $3.3 \times 3.5 m^2$. The rooms and hallways differed in the number of doors and windows, ceiling height, number of people, activity level and furnishings. The construction type of the floor was common

to other office buildings, employing double plasterboard dividing walls, cement floors with linoleum tiles and an acoustic tile drop ceiling below a corrugated steel roof. It also had some interior windows reinforced with 1/2 inch steel mesh, steel doors and cement exterior walls. Furnishings in the various rooms included wooden desks, book shelves, computer workstations, high wooden tables, metal filing cabinets, tall metal racks and laboratory instruments. Of the measured transmission paths there were 5 LOS and 29 OBS paths used in our analysis.

The large scale pathloss at 1.75 GHz versus distance has been computed for this area [39]. Assuming free space loss at 1 metre, the pathloss increases by an amount proportional to the distance raised to the power $N = 2.85$. A slightly more accurate description of the large scale propagation environment was made by assuming a free space loss ($N=2.0$) plus an attenuation factor or wall loss of 3.0 dB per wall.

4.2.2 Secondary measurement areas

A group of five office environments were chosen in four different buildings on the Carleton University campus to allow examination of building dependent differences in diversity performance. Internal layouts of these areas are shown in appendix A. The areas are described here:

1. The engineering annex: The same area where the primary measurements were taken. The exact measurement locations differed from those of the primary measurements to allow more LOS paths. There were a total of 38 measurements made here.
2. Second floor administration building: this area was chosen as representative of a small but open office environment with soft partitions at chest level. There were a total of 9 work areas which were staffed to half capacity at the time of the measurements. The floor was carpeted and the walls containing the area were plasterboard.
3. Third floor administration building: this office was relatively cluttered with soft partitions above head level and with some ceiling height walls containing individual offices. The area was busy and some of the corridors between

Figure 4.1: Primary measurement area with transmitter locations (X) and receiver locations (R).

partitions were only 1.5 metres wide. The L-shaped area was surrounded by brick exterior walls with windows and interior walls of brick and plasterboard. There were a total of 25 measurements made in the two areas in the administration building.

4. Second floor of library: this large open area was furnished with small wooden cubicle desks 1.5 m in height and it was surrounded with large 3 m high bookshelves. The central area was similar to a large open office concept. There were no interior walls in the area. The exterior walls of this new section of the library were brick with tall windows. There were 23 measured paths from this building.
5. Third floor of Mackenzie engineering building: this unique building is constructed with brick exterior walls and cement block interior walls. The measured area was the top floor of the third quadrant of the building which consisted of an outside ring of offices and classrooms with tall narrow windows, all interconnected by a square hallway. On the interior of the hallways there were two large laboratories separated by a plasterboard wall. There were steel meshed windows between these labs and the hallway. At the time of the measurements the building was undergoing renovations and the suspended ceiling tiles had been removed in the hallways. Data was collected for 14 transmission paths in one quadrant of this building.

4.3. Measurement Apparatus

The equipment consisted of a portable transmitter and receiver on wooden carts. The key parts are described below. Photographs of the equipment are shown in Appendix B.

Antennas

Quarter wavelength monopole antennas with four wire radials for a ground plane were built from brass and mounted on wood and foam supports. They were each tested (in a large anechoic chamber) for a good impedance match at 1.75 GHz and for omnidirectional radiation patterns. The effect of the proximity of the two diversity receive

antennas was tested and it was observed that the only reduction in gain occurred in the direction of the second antenna. These variations over an azimuth of about 90 degrees were 5, 2 and 1 dB for antenna spacings of $\lambda/4$, $\lambda/2$ and λ respectively. Antenna measurements are detailed in appendix D. The antennas were mounted vertically on wooden masts at a height of 1.6 metres. A low dielectric rigid foam was used directly under the antenna to minimize disturbances to the radiation pattern. In this configuration the antennas received the vertical component of the electric field.

Transmitter

The transmitter as shown in Fig. 4.2 was designed to transmit two continuous wave (CW) carriers separated by ΔF and centred at 1.75 GHz. The frequency separation was adjustable from 1 to 20 MHz and the transmit power could vary from a maximum of +30 dBm per carrier to a minimum of -80 dBm in 1 dB increments.

Receiver

A receiver illustrated in Fig. 4.3 was designed to simultaneously detect the signal envelope from four diversity branches. Each antenna was connected via a flexible coaxial cable to a front end amplifier and then a separate RF downconversion and detection was implemented for each branch. Two frequency synthesizers were adjusted to allow detection of frequency diversity signals spaced from 1 to 20 MHz. The dynamic range of 50 dB was adequate to reliably detect deep signal fades more than -30 dB. Two oscilloscopes were used to visually monitor the received envelope.

Data Acquisition

After the detectors each received branch was amplified to match the maximum input level of -40 dBm to the full scale value of 10 volts. This precise adjustment maximized the possible dynamic range of the measurement. For data acquisition we used a 4 input 12 bit analog to digital converter (ADC) which was located inside a personal computer with a maximum acquisition speed of 8 KHz per input. Data was written directly to a hard disk and subsequently stored on a mass storage optical disk for the following analysis.

Calibration

The transmitter power was calibrated using a microwave spectrum analyzer to within ± 0.5 dB of its true value. The transmitter was then used as the source to calibrate all four receiver branches over the full dynamic range. Calibration files were created to allow accurate conversion from ADC integer output to power values in dBm and to account for the non-linearity of the detectors. The calibration file for the primary measurements is shown in Fig. 4.4.

4.4. Procedure to measure fading

The procedure to measure deep fades was to adjust the transmit power level such that the average received level was approximately 8 to 10 dB below the maximum input level of -40 dBm. After some initial test runs a reference voltage corresponding to this level was noted and signal levels were verified by visual inspection of the oscilloscope display. With this procedure the variations of the received envelope were always within the receiver's 50 dB dynamic range. The exact transmitter power was recorded for each measurement to allow calculation of the absolute signal attenuation over the path.

The transmitter on a wooden cart was located in areas in the engineering annex where people were working or walking by. One receive location was in a radio laboratory and the other in a hallway near the centre of the floor; see Fig. 4.1. Measurements were taken during times of typical activity in each of the areas. In classrooms there were between 20 and 30 students present. In laboratories or hallways the number of people varied between 2 to 10. In total there were 35 transmission paths for the results reported here. Since our receiver was difficult to move, it was placed at representative transmit locations and the transmitter was moved to typical receiver locations. Due to the close proximity of scatterers in both transmit and receive locations, it was assumed that the diversity performance would be the same for both cases so that similar results would be obtained if the receiver had been moved to many locations.

Measurements in the secondary areas were also conducted during the day time with normal levels of activity. In these areas and for the second set of measurements for the

Figure 4.2: 1.75 GHz Dual Frequency Transmitter.

Figure 4.3: 1.75 GHz Frequency and Space Diversity Receiver.

Figure 4.4: Receiver calibration file with ADC value versus RF input power.

annex the transmitter was normally located in just one area and the receiver moved to many locations. The exception to this procedure was on the third floor of the administration building which was too crowded to allow the large receiver cart complete access so the transmitter was moved to the various locations.

To simulate motion of the portable, the two receive antennas were mounted on a bracket to allow varying separations and were moved by hand around a 1 metre circumference (diameter of 1.8λ) circular guide cut from thin wood; see Fig. 4.5. This size of circle was chosen to allow the antennas to travel through several wavelengths yet stay in the same local area. A circular path was selected rather than a random one because each measurement could then be made with a relatively repeatable motion and velocity. The antenna bracket was moved at 1/2 revolution per second (0.5 m/s) which is a slow walking speed. Since the bracket was moved by hand, the antennas were always near the person holding the bracket, a situation similar to the location of a cordless telephone.

Up to 15 diversity configurations were tested for each path yielding a total of 132 measurements for each configuration. The transmitter carrier frequency separation, ΔF , was set to 1, 2, 5, 10 and 20 MHz. The receive antenna spacing, ΔS , was set to $\lambda/4$, $\lambda/2$, and λ (4.3 cm, 8.6 cm and 17.1 cm). The smallest of these spacings is significant since it could be roughly the width of a hand held portable.

Sampling rate and measurement duration

Two key parameters in measuring the fading signal envelope are the sampling rate and the duration of the measurement. Since the fading envelope at a moving receiver is characterized by deep and fast nulls, the interval between samples should be small enough so that fades of a certain depth are in fact sampled.

The minimum acceptable sampling rate was chosen to reliably measure fades that were -20 dB below the mean level. We conducted a test in which the sample rate was set to 8000 Hz per input channel for a reference (the maximum that hardware permitted) and the receive antenna was moved in a circular motion at 0.5 m/s. The cumulative distribution function (CDF) of the received envelope power was computed from this data and stored for reference. The data sequence was decimated by factors of two and the

Figure 4.5: Top view of antenna movement around guide for antenna 1 (A1) and antenna 2 (A2).

computation repeated until there was a noticeable error between the CDF of the decimated data and that of the reference at a fade value of -20 dB. At an effective sampling rate of 1000 Hz the error in the CDF at -25 dB was less than $\pm 2 \times 10^{-4}$. Fig. 4.6 shows the received envelope fading from one moving antenna and Fig. 4.7 shows the effect of sampling rate on the CDF of the envelope fading.

The chosen measurement duration was based on the following factors:

1. It should be long enough to collect sufficient number of samples for statistical validity.
2. It should be long enough to capture typical variations in the channel due to movement of people.
3. It should approximate the duration of a telephone call
4. It should not be so long as to render the data manipulation extremely tedious.

A measurement duration for each experiment of 60 seconds was chosen which provided $60,000$ samples per branch which allowed for 600 samples at the outage rate of 10^{-2} . This duration was also sufficient to capture typical variations in the channel due to the movement of people. This duration is approximately that of a telephone call. With these parameters, the number of samples per wavelength travelled was:

$$\text{samples}/\lambda = \frac{60,000 \text{ samples}}{0.5 \text{ m/s} \times 60 \text{ s} \times 5.83\lambda/\text{m}} = 313 \quad (21)$$

Average received power

The average received power of the signal sequences was computed using linear units of milliwatts and then converting the result to logarithmic values of dBm. This average was used to normalize the received envelopes such that the 0 dB value represents the one minute averaged power.

Figure 4.6: Received envelope from antenna moving in a circle at 0.5 m/s.

Figure 4.7: Effect of sampling rate on the CDF of envelope fading.

Chapter 5

ANALYSIS AND RESULTS

5.1. Introduction

In this chapter we describe the analysis of the collected data. Results from measurements in the engineering Annex are presented for envelope fading statistics and envelope cross-correlations. We apply ideal selection combining in software to the collected data and compare its performance with theory. The effect of various diversity parameters such as antenna spacing, ΔS , and frequency separation, ΔF , are examined on two-branch space and frequency diversity and on four-branch hybrid diversity. The data collected from the secondary measurement areas in four buildings is used to determine the effect of different environments on diversity performance. The variations of fading with diversity is examined to determine the worst case performance and the distribution of output statistics with both LOS and OBS transmission paths. We then consider two practical selection combining techniques which we implement in software and simulate using selected data files. We use results of envelope fading to predict the BER performance with DPSK. To put some of the results into context we apply four-branch hybrid diversity to a radio link budget and specify the system reliability. Finally we summarize the results of this section.

Analysis programs were written in a high level math language called "Matlab" operating on Sun workstations under the Unix environment. Important routines are listed in Appendix E. The programs are also portable to a PC based platform. Due to the very large size of the data files and the number of files the use of a fast workstation such as a Sparc-2 with 16 Mega-bytes of memory was important. This allowed for quick development of routines and where necessary some slow running portions were written as subroutines in the C language. Access to a large file server of approximately 300 Mega-bytes capacity was important to allow batch programs to be run through large groups of data files. Data acquisition programs and data analysis programs were written to be general enough to allow use in other experiments. They have been used with two other projects, one at millimeter wavelengths and another for cellular radio.

5.2. Envelope fading statistics

CDF of envelope fading

Measurements with the moving receive antennas showed numerous signal fades down to -30 dB from the mean as shown in Fig. 4.6 in Chapter 4. For each one minute measurement the average received power was calculated and used to normalize the data to power values expressed as a ratio in dB relative to the mean power. Histograms of the fading were computed for each branch by dividing the received power into 51 one dB bins. This was normalized by the total number of samples in the file (60,000) to provide a discrete pdf which we term the probability mass function (pmf). This pmf was cumulatively summed for all bins to provide a discrete cumulative distribution function (CDF) of the normalized envelope. This was computed for each branch. The programs for this procedure are documented in Appendix E. The CDF of the envelope fading on four single branches from one measurement is shown plotted on a logarithmic scale in Fig. 5.1.

Global maximums, averages and minimums were calculated to show the range of CDF values over all measurements. Results from 868 one minute measurements taken in the Annex of single branch envelope fading are shown in Fig. 5.2. These indicate that the average value of the CDF's are close to Rayleigh as expected; the outage rate, given a fade depth of 20 dB, is approximately 10^{-2} , or equivalently, the availability is 99%. There is, however, considerable spread between the best case and worst case statistics. At 99% availability, the worst case CDF is more than 6 dB lower than average. The difference between the best case and the worst case shown here for OBS paths is 18 dB.

5.3. Envelope cross-correlation

Many theoretical analysis of diversity combining systems have assumed that the fading statistics on each branch are independent and hence uncorrelated [3]. However, this is often not the case, as shown in [4, 22] for land mobile radio channel. An increase in the envelope cross-correlation coefficient between diversity branches will decrease the absolute performance of all diversity combining schemes. The degradation to selection

Figure 5.1: CDF of received signal fading on four branches from one file.

Figure 5.2: CDF of single branch envelope fading from 868 OBS paths.

combining performance, however, is relatively small for correlation values less than 0.5 [4]. The cross-correlation was computed as follows: let $r_j(k)$ represent the received signal power in watts on the j^{th} branch at sample k . It is normalized by the local mean power of the branch. These samples are stored in an $n \times 4$ matrix ($n=60,000$) where each column corresponds to a branch and each row represents a sample. We denote this matrix as R . The covariance matrix of R is C where

$$C = \frac{1}{(n-1)} \times R^T \times R \quad (22)$$

The cross-correlation coefficient at the zeroth lag between the i^{th} and j^{th} branch is

$$\rho(i, j) = \frac{C(i, j)}{\sqrt{C(i, i) \times C(j, j)}} \quad (23)$$

This correlation was calculated for all pairs of space and frequency diversity branches with moving antennas and normal levels of activity on the floor. Results are shown as histograms in Fig. 5.3 which are based on 132 one minute measurements. The mean value of correlations for $\Delta F = 1$ and 20 MHz, not shown on the histograms, were 0.946 and 0.115 respectively. The results suggest that frequency diversity at 1.75 GHz would be effective for $\Delta F \geq 5$ MHz and marginally better for $\Delta F = 10$ MHz where 91 % of the values were less than +0.5. The distribution of space diversity correlation values showed only minor changes for the three antenna spacings and in general suggests good diversity performance at $\Delta s = \lambda, \lambda/2$ and $\lambda/4$. The majority of measurements were made when there was some level of activity in the rooms or hallways. Under these circumstances the correlation values for space diversity and for frequency diversity with $\Delta F > 5$ MHz were usually close to zero. In the few cases when there was little or no movement on the floor there tended to be more cases of correlations at the extreme values around ± 0.5 . This might be explained by the existence of standing wave patterns under static channel conditions.

Figure 5.3: Histograms of envelope correlation coefficients.

5.4. Selection combined fading statistics

Diversity operation relies on the fact that two or more independently fading signals will rarely be in a deep fade at the same time. Thus it requires more than one transmission path over which the fading is relatively uncorrelated. By simply selecting the best signal or by adding the signals at the receiver, the availability can be increased dramatically.

The four branches of data were divided into two pairs of frequency diversity branches (branch 1 and 2, and branch 3 and 4), two pairs of space diversity branches (branch 1 and 3, and branch 2 and 4) and one group of four branches for combined space and frequency diversity which we call "hybrid diversity". Each one minute measurement included two frequency diversity measurements, one on each antenna and two space diversity measurements, one on each frequency.

Selection Combining

Ideal selection combining selects the largest signal from L branches where in our measurements, $L=2$ or $L=4$. The performance of selection combining is well documented [4]. SC was implemented in software using normalized signal levels. Fig. 5.4 shows a sample of selection combined received signal envelopes for one-, two- and four-branch diversity. The four-branch space and frequency hybrid diversity shows no deep fades in this 4 second interval.

The CDF of the combined signals was computed in the same manner done for the single branch case. Fig. 5.5 shows the CDF of single branch versus four-branch hybrid diversity for one 60 second measurement. The CDF's of the four branches were multiplied to form a product CDF that represents the theoretical joint distribution of four-branch selection combining if the signals on all branches were independent and hence uncorrelated. As shown in Fig. 5.5 the measured selection combined CDF in this particular case has better statistics than this theoretical curve.

Figure 5.4: Received envelope from one measurement for one-, two- and four-branch diversity with SC using $\Delta F = 10$ MHz and $\Delta S \approx 1\lambda$.

Figure 5.5: CDF of selection combined envelope for one path using four-branch hybrid diversity with $\Delta F = 10$ MHz and $\Delta S = \lambda$.

5.4.1 Effects of antenna and frequency separation on average diversity gain

Fading statistics for groups of measurements

The following procedure was used to analyze the fading statistics from individual measurements and from groups of measurements:

1. Select group of measurements (For example, all OBS paths in Annex).
2. Specify files based on diversity parameters ΔF and ΔS .
3. Compute mean envelope power in dBm from four branches and normalize the received envelope sequences, or, to exclude the effect of varying mean levels, normalize with respect to the mean for each branch.
4. Apply SC algorithm to all normalized envelope sequences.
5. Compute histograms of envelope power in 1 dB power bins. Do this for one-, two- and four-branch diversity for each measurement.
6. Normalize histograms by the total number of samples to create a discrete pdf (pmf). Then compute the cumulative sum to obtain a discrete CDF defined in 1 dB steps from -40 dB to +10 dB. Store results for each data file.
7. Compute average CDFs for the group.
8. Compute maximum CDFs for the group.
9. Store results for the group.

The software for this procedure is detailed in Appendix E.2.

All CDF's with the same values of ΔF and ΔS were averaged over 22 transmission paths. There were 88, 44 and 22 CDF's averaged for the one-, two- and four-branch diversity respectively. Some typical results are shown in figures 5.6 and 5.7. As an absolute reference only, the Rayleigh distribution is plotted for one-, two- and four-branch selection combining. From Fig. 5.6 where $\Delta F = 10$ MHz and $\Delta S = \lambda/4$, the advantage offered by two- and four-branch diversity is clear. As might be expected by the envelope correlation results, the average combined statistics for space and hybrid diversity did not vary much with antenna spacings of $\lambda/4$, $\lambda/2$ and λ . However, the

average statistics for frequency and hybrid diversity show degradation with a smaller frequency separation of $\Delta F = 2$ MHz as shown in Fig. 5.7.

Average diversity gain

The values of diversity gain in dB as described in section 2.3.2 were calculated for each two- and four-branch combination for each of the measurement paths. The diversity gains calculated here were based on normalized mean signal levels for each branch and thus exclude variations due to unequal means. The variations of the local mean signal levels varied by ± 3 dB as reported in [40], however, this had negligible effects on the average gain values reported here. These relative gains were measured with respect to the local average single branch CDF from each path. Global average gains were then calculated by averaging over all gain measurements with the same values of ΔF and ΔS for each path.

Two branch diversity gain

Results of two-branch diversity gain calculations are shown in table 5.1 (a) and 5.1 (b) based on 22 paths (values for $\Delta F = 1$ and 20 MHz were calculated from 6 paths). These indicate that the average two-branch space diversity gain at 99% availability is relatively constant and greater than 10 dB for $\Delta S = \lambda/4$, $\lambda/2$ and λ . This independence was also reported in [10] for a 900 MHz receiver moving in a random fashion throughout a room. This measured gain compares favorably to the theoretical performance of 10 dB advantage at 99% availability for two-branch selection combining with Rayleigh fading reported in [3, 4]. For 99 % availability, the averaged two-branch frequency diversity gain for $\Delta F = 10$ and 20 MHz was 10.4 dB and 10.6 dB respectively. There appears to be little advantage in choosing the wider frequency separation. These average results are comparable with space diversity and to two-branch diversity performance found for the 900 MHz band [10, 13]. Average diversity gain at $\Delta F = 5$ MHz was 9.6 dB which is more than 2 dB better than that reported at 900 MHz in the same building [12]. The poorer diversity gain at the lower spacings of 1 and 2 MHz can be explained by the high average envelope correlation values reported earlier in this section and the effect of this correlation on the CDF as described by equation 17.

Figure 5.6: Average CDF for selection combined envelopes from 22 OBS paths using $\Delta F = 10$ MHz and $\Delta S = \lambda/4$.

Figure 5.7: Average CDF for selection combined envelopes from 22 OBS paths using $\Delta F = 2$ MHz and $\Delta S = \lambda/4$.

Table 5.1: Average SC diversity gains in dB versus ΔF and ΔS for frequency diversity (a), for space diversity (b) and four-branch hybrid diversity (c).

ΔF (MHz)	Availability		
	90.0 %	99.0 %	99.9 %
1.0	1.5	4.0	9.2
2.0	2.8	6.7	12.6
5.0	4.6	9.6	15.4
10.0	5.6	10.4	15.8
20.0	4.8	10.6	16.0

(a) Two-branch frequency diversity

ΔS (λ)	ΔS (cm)	Availability		
		90.0%	99.0%	99.9%
1/4	4.3	5.2	10.4	16.9
1/2	8.6	5.3	10.0	15.3
1	17.1	5.7	10.8	16.4

(b) Two-branch space diversity

ΔF (MHz)	Antenna spacing (ΔS)		
	$\lambda/4$	$\lambda/2$	λ
2.0	12.8	13.4	13.8
5.0	14.9	15.0	15.4
10.0	15.5	15.7	16.1

(c) Four-branch hybrid diversity at 99% availability

Four branch hybrid diversity

A four-branch hybrid space and frequency diversity receiver was evaluated by selecting the largest signal from the four receiver branches. Average diversity gains over 22 paths are shown in table 5.1 (c) for 99% availability. The effect of varying ΔS was minor as it was for space diversity, but again there were noticeable effects of varying ΔF as with two-branch frequency diversity. The theoretical predicted four-branch diversity gain given Rayleigh fading and selection combining is 16 dB [3]. This value was achieved for a frequency separation of 10 MHz and an antenna spacing of 1 wavelength.

5.5. Diversity performance in four buildings.

We measured diversity performance on over 100 transmission paths in four buildings whose environments were described in chapter 4 with floor layouts illustrated in appendix A. The antenna motion and sampling rate were kept the same as in the primary measurements in the Annex. The following parameters were used:

1. $\Delta F = 5$ MHz
2. $\Delta S = \lambda/2$
3. Ideal sample by sample selection combining
4. Receive antenna velocity = 0.5 metres per second

The transmitter was placed in a representative transmit location and the receiver was moved to various receive locations. These locations were selected to obtain a mix of LOS and OBS transmission paths as opposed to the primary measurements which concentrated on OBS paths. The number of different paths is indicated in table 5.2.

Average values of CDF results at 99% availability

Average values of the CDF for the single branch input fading statistics are shown with the diversity combined (output) CDF in table 5.3. Input and output CDF's are averaged from all paths in the specified building. The diversity gain values here are simply differences between the average input and output fading CDF at 99% availability. In these tables it is the value of the output CDF that is important. This value dictates

how much fade margin would be required in a radio communication link. The single branch average fading is not as severe as our first measurements from the Annex likely because this average result includes LOS cases which may have less severe fading.

Results in table 5.3 indicate that there are relatively small variations in the output CDF's between the different buildings despite larger variations in the input statistics. Space diversity in the Library had the best output fading statistics of -7.6 dB compared with the worst in the Mackenzie building which had -8.9 dB. The variation of 1.3 dB is not that large which is good for extending results to other buildings. The frequency diversity averages had a similar variation but had poorer absolute values by 1.5 to 2.0 dB. This may be due to higher envelope cross-correlation as noticed in section 5.3. A larger frequency separation, that is, $\Delta F > 5$ MHz, is likely needed to maximize diversity performance for these different buildings. Four-branch hybrid diversity had a difference of 1.6 dB between the best and worst average performance.

LOS versus OBS paths

The effect of numerous LOS paths was investigated. Results are presented in table 5.4. The average output CDF values for OBS paths have small variations of only 0.7 dB but the LOS paths have large variations in average values between buildings of 3.2 dB (for frequency diversity). These larger variations for LOS cases may not be significant since in general the outputs statistics are always better for LOS cases than the OBS cases. This suggests that we may use the OBS results for a link design.

Table 5.2: Summary of path types in each building.

Building	Number of path types		
	OBS	LOS	Total
Annex	28	10	38
Admin	13	12	25
Library	14	9	23
Mackenzie	11	3	14
All buildings	66	34	100

Table 5.3: Summary of average CDF values in dB at 99 % availability and corresponding diversity gain using SC. Results are for all measured paths.

Building	All measured paths		
	input CDF	output CDF	gain
Annex	- 15.6	- 8.3	7.3
Admin	- 16.0	- 8.7	7.3
Library	- 13.2	- 7.6	5.6
Mackenzie	- 16.7	- 8.9	7.8
All buildings	- 15.1	- 8.3	6.8

(a) Two-branch space diversity

Building	All measured paths		
	input CDF	output CDF	gain
Annex	- 15.6	- 9.9	5.7
Admin	- 16.0	- 10.8	5.2
Library	- 13.2	- 9.3	3.9
Mackenzie	- 16.7	- 10.4	6.3
All buildings	- 15.1	- 10.0	5.1

(b) Two-branch frequency diversity

Building	All measured paths		
	input CDF	output CDF	gain
Annex	- 15.6	- 4.8	10.8
Admin	- 16.0	- 6.4	9.6
Library	- 13.2	- 5.0	8.2
Mackenzie	- 16.7	- 5.8	10.9
All buildings	- 15.1	- 5.3	9.8

(c) Four-branch hybrid diversity

Table 5.4: Summary of average CDF values in dB at 99 % availability and corresponding diversity gain using SC. Results are separated into OBS and LOS paths.

Building	Obstructed paths			Line of sight paths		
	input CDF	output CDF	gain	input CDF	output CDF	gain
Annex	- 18.8	- 9.0	9.8	- 11.7	- 6.8	4.9
Admin	- 16.4	- 8.8	7.6	- 15.5	- 8.5	7.0
Library	- 18.4	- 8.3	10.1	- 10.0	- 6.6	3.4
Mackenzie	- 16.9	- 8.9	8.0	- 16.1	- 8.6	7.5
All buildings	- 17.8	- 8.8	9.0	- 12.3	- 7.4	4.9

(a) Two-branch space diversity

Building	Obstructed paths			Line of sight paths		
	input CDF	output CDF	gain	input CDF	output CDF	gain
Annex	- 18.8	- 10.1	8.7	- 11.7	- 9.5	2.2
Admin	- 16.4	- 10.8	5.6	- 15.5	- 10.9	4.6
Library	- 18.4	- 10.8	7.6	- 10.0	- 7.7	2.3
Mackenzie	- 16.9	- 10.3	6.6	- 16.1	- 10.4	5.7
All buildings	- 17.8	- 10.4	7.4	- 12.3	- 9.4	2.9

(b) Two-branch frequency diversity

Building	Obstructed paths			Line of sight paths		
	input CDF	output CDF	gain	input CDF	output CDF	gain
Annex	- 18.8	- 4.7	14.1	- 11.7	- 5.2	6.5
Admin	- 16.4	- 6.3	10.1	- 15.5	- 6.4	9.1
Library	- 18.4	- 5.0	13.4	- 10.0	- 5.2	4.8
Mackenzie	- 16.9	- 5.7	11.2	- 16.1	- 6.2	9.9
All buildings	- 17.8	- 5.2	12.6	- 12.3	- 5.7	6.6

(c) Four-branch hybrid diversity

The frequency diversity gains for LOS measurements were consistently lower than those for space diversity. This result was also noticed by observation of the received envelope fading on the oscilloscopes during the measurements. That is, there appeared to be significant correlation of the frequency diversity envelopes for LOS measurements.

Effect of interchanging transmitter and receiver positions

It is of interest to compare the average gain values measured for the annex in the primary and secondary measurements. The measurement locations for the OBS paths were similar however the positioning of the transmitter and receiver was reversed for the secondary experiment. Primary measurements were conducted in February during times of high classroom activity while the secondary experiment was conducted in July when the labs were still used but the classrooms were empty. Comparing the result from table 5.1 (a) for space diversity with $\Delta S = \lambda/2$, a gain of 10.0 dB, to that from table 5.4 (a), a value of 9.8 dB, we see there is very little difference. This appears to justify the assumption made in section 4.4 that similar results would be observed if the transmitter and receiver were interchanged. If we compare parts (b) of the same two tables for frequency diversity with $\Delta F = 5$ MHz it is apparent that the gain values are slightly different. The values are 9.6 dB and 8.7 dB for the primary and secondary measurements respectively. The hybrid diversity values are 15.0 dB and 14.1 dB. This suggests that frequency diversity may be slightly more sensitive to changes in local scattering and that the assumption may not be as valid.

5.6. Variation of diversity performance

Worst case CDF from OBS paths

CDFs of combined envelopes were computed as described in section 5.4.1. These statistics were stored in a database and then used to compute both average and worst case statistics for each set of antenna and frequency separation. The results illustrated in Fig. 5.8 are based on measurements from 35 paths with $\Delta F = 10$ MHz and $\Delta S = 1\lambda$. There were four single branch CDFs, two two-branch CDFs and one four-branch CDF per path

which provided a data base of 140, 70 and 35 CDFs respectively. Five of the 35 paths were line of sight (LOS) and results included the effect of local mean variations of the received signal. While these mean variations had little effect on average results, they did increase the difference between average and worst case CDF results.

Two factors contribute to the variation of the diversity combined CDFs: different single branch CDFs as shown in Fig. 5.2 and the variations in envelope cross-correlations. The latter has an effect on the diversity improvement [4] and the correlation values have been shown to vary from -0.4 to $+0.9$ in the indoor environment [37].

As with the larger group of 868 measurements in Fig. 5.2, the average single branch CDF for this group is close to Rayleigh distributed and the maximum value (worst case) is 6 dB lower at an outage rate of 10^{-2} . The benefit of diversity is clear for the two branch and four-branch combining where the improvement in the average statistics at 99% availability is 10.3 and 16.0 dB respectively. The CDF for two-branch frequency diversity with $\Delta F = 10$ MHz (not shown here) were very close to that for the space diversity. These results also show that the 6 dB difference between average and worst case CDF decreases to 5.1 and 2.9 dB at 99% availability for the two-branch and four branch cases, respectively. Thus the diversity improves the average statistics and also decreases the variability in statistics over all paths. Alternatively, the improvement of the worst case CDFs is greater than the improvements between the average CDFs.

Variations of CDF at 99% availability in four buildings

We consider here the variation in the CDF at 99% between individual measurements throughout the four buildings. The average results were shown in section 5.5. The distributions of the CDF values are illustrated as histograms in Fig. 5.9 for single branch (input) fading and for the three diversity types. These distributions include values for both OBS and LOS paths in the four different buildings. There is a significant spread in these values for the single branch fading with a minimum of -26.6 dB to a maximum of -2.3 dB. Thus the required fade margin for 99% availability without diversity varies between paths by 24.3 dB. This variation is larger than the 18 dB value reported in section 5.2 and is likely due to the inclusion of LOS paths. The benefit of diversity

Figure 5.8: Average and worst case envelope fading statistics using two-branch space diversity and four-branch hybrid diversity with $\Delta F = 10$ MHz and $\Delta S = 1\lambda$.

Figure 5.9: Distribution of one-, two- and four-branch CDF-values at 99% for all OBS and LOS paths in all buildings.

is evident for the two- and four-branch diversity CDF values. Not only has the mean value improved as indicated in table 5.3, but the variations in values have been reduced. The spread of values is 15.3 dB, 13.1 dB and 7.8 dB for frequency, space and hybrid diversity respectively. If fade margins at 99% availability were to be estimated from these results, we could report that 95% (for example) of cases fall above -23.4 dB, -14.6 dB, -12.8 dB and -9.1 dB for single branch fading, frequency diversity, space diversity and hybrid diversity respectively. These results illustrate that diversity is effective in reducing variations of fading statistics caused by different paths in different buildings. Further insight into the variability of the fading statistics could be obtained by plotting histograms from specific sub-groups of this data such as for the various buildings with LOS or OBS paths.

5.7. Practical selection combining techniques

In this section we simulate two sub-optimal forms of selection combining and evaluate their performance based on selected measurements from the primary measurement area in the annex. Our approach in this analysis is to compare these practical techniques to ideal selection combining for both typical and atypical fading for OBS paths. It was noticed in section 5.2 and Fig. 5.2 that there is considerable variation in envelope fading statistics for OBS paths however most of the fading distributions were within 2 dB of a Rayleigh distribution at 99% availability. A comparison of the variation between the fading on two diversity branches from each measurement showed a typical difference of 1 to 3 dB at 99%. In some unusual cases the fading statistics on two branches differed by as much as 8 dB. We present simulation results for the switched combining techniques described in section 2.3.2. Instead of presenting averaged results we present two cases which were representative. A measurement result from path 101 was representative of "typical" cases where the fading differs by 1 to 3 dB between branches and path 108 was the chosen example of cases where fading between branches varied by up to 8 dB. For both paths the receiver was in the radio laboratory. For path 101 the transmitter was in a large classroom with one person present and people walking by in the hallway. For

path 108 the transmitter was in a small computer room with 2 to 3 people present but with a few people entering and leaving via two steel doors.

We constrained the analysis to measurements with the parameters $\Delta F = 10$ MHz and $\Delta S = 1\lambda$. For these parameters there were 25 other OBS paths with typical fading similar to path 101 and 3 paths with unusual fading like that found for path 108. The 5 LOS paths had unusual and largely varying fading statistics also, however, this fading was less severe and was not considered for this investigation.

5.7.1 Switch and stay combining

A description of switch and stay (SAS) combining is presented in section 2.3.2. Figure 2.2 illustrates the generic block diagram of a two-branch switched antenna scheme. The SAS technique switches branches on the negative threshold crossings of the received signal. In this simulation we varied the switching threshold level and the switching delay to determine effects on the output fading statistics. Detailed software listings for this simulation are provided in Appendix E.

Effect of switching threshold

The threshold is chosen to optimize performance at a desired outage rate. For example, with Rayleigh fading as shown in [4], the best threshold for a required outage of 10^{-2} is -10 dB below the mean signal level. Analytical results for the ideal case predicted from equation 16 show that SAS performance matches that of SC at the threshold point which looks like a “knee” in the CDF.

The effect of different switching threshold levels on the measured statistics at path 101 is shown in Fig. 5.10. As the threshold is lowered the knee of the CDF curve is lowered. At lower outage rates, the CDFs for the cases of lower thresholds are superior. This result was similar for the other 25 OBS files with similar fading and shows some agreement with the theoretical results noted above. However, for the measured results there was always a degradation of 1 to 2 dB from SC at the knee point. This may be due to the differences of 1 to 3 dB in fading statistics between the two branches.

Figure 5.10: Typical result for CDF of envelope fading on branch 1 (BR1) and branch 3 (BR3) using SAS with threshold levels in dB. From path 101, first space diversity pair.

Effect of switch delay

Switching delay in a SAS scheme may occur for a few reasons. In an implementation as illustrated in Fig. 2.2 there may be delay in measuring the signal level due to a narrow band filter or a required averaging period. The scheme of feedback diversity illustrated in Fig. 2.3 suffers from delay due to propagation time or the timing of a TDD protocol. Due to the reduction in complexity at the portable the feedback scheme is attractive for indoor wireless systems. When switch delay is introduced the CDF degrades as shown in Fig. 5.11 (a) for path 101. Above 10 milliseconds the benefit of diversity is minimal. Such a delay would be experienced in the Digital European Cordless Telephone system based on the frame rate [14].

A similar effect to varying the threshold is noticed⁴ for increasing delay where at the lower outage rates the diversity performance is actually better. For small delays, less than half the duration of a fade, the switch delay has the same effect as reducing the threshold level at which switching takes place. Hence, the knee of the CDF curve shifts down and the diversity performance at lower outage rates improves. However, this improvement is at outage rates below that of interest. This effect is noticeable for the result shown in Fig. 5.11 (a) and even more apparent in another measurement result shown in Fig. 5.11 (b).

Unusual and severe fading

In some unusual cases of input statistics worse than Rayleigh fading, for example path 108, the knee of the CDF shifts up and the diversity gain at 99% availability improves with *increasing* delay as illustrated in Fig. 5.12 (a) or in row 5 of table 5.5. In such a case the threshold may need adjustment to optimize performance. It was also noticed that in some cases where the statistics of the two branches differ greatly that the best case single branch was superior to some combined CDFs with large delay as shown in Fig. 5.12 (b) or in row 7 of table 5.5. Such an effect indicates that the SAS performance cannot be accurately predicted from the theory which assumes identical statistics on each branch and independent fading.

Results of diversity gain at 99% availability versus delay are presented in table 5.5 for the four pairs of branches at each of the paths 101 and 108. Results from Fig. 5.11(a),

(a) Space diversity pair branch 1 (BR1) and branch 3 (BR3)

(b) CDF for frequency diversity from branch 1 and 2

Figure 5.11: Typical results for SAS from path 101 with knee of CDF shifting down for increasing delay. Delays are in msec.

(a) Example with bad fading. Space diversity

(b) Example with different CDFs. Frequency diversity

Figure 5.12: Example of SAS output fading with bad input fading from path 108. Delays are in msec.

2 OF 2

PM-1 3 1/2" x 4" PHOTOGRAPHIC MICROCOPY TARGET
NBS 1010a ANSI/ISO #2 EQUIVALENT

Table 5.5: The effect of switching delay on SAS combining
for typical results of path 101 and unusual results of path 108.

path #	type	Worst case branch CDF 99%	Diversity gain in dB at 99% (wrt worst case branch)							
			Best one- branch	Switching delay in milliseconds						Ideal SC
				20	16	12	8	4	0	
101	space	-20.0	2.3	3.4	5.5	6.8	7.9	9.0	10.1	11.2
101	space	-21.3	1.5	2.6	4.2	6.2	8.3	9.9	11.4	12.9
101	freq.	-20.0	0.0	1.6	3.5	5.0	6.8	6.7	8.8	10.5
101	freq.	-21.4	3.7	4.6	5.7	7.0	7.7	7.9	7.9	11.6
108	space	-24.4	1.1	4.9	4.8	4.8	3.8	3.1	2.4	9.6
108	space	-15.6	0.4	2.0	2.5	3.0	3.9	5.0	6.0	7.0
108	freq.	-24.4	8.4	7.9	8.1	8.4	8.4	8.4	8.4	15.3
108	freq.	-23.1	7.7	8.6	8.7	9.2	9.2	8.7	8.1	10.9

Fig. 5.11 (b), Fig. 5.12 (a) and Fig. 5.12 (b) are summarized in the first, third, fifth and seventh rows respectively.

For the purpose of comparison, the maximum tolerable delay was calculated for the cases of Rayleigh fading based on equations 19 and 20 from section 2.3.2 with a velocity of 0.5 m/s. This calculation indicates that a delay of 9 msec would be the maximum tolerable delay before significant losses in diversity gain. Our measurement based simulation results are close to this but show that the switching strategy is slightly more sensitive to delay with these statistics.

5.7.2 Switching diversity with a TDM format

The time division multiplex (TDM) format used for this simulation is shown in Fig. 5.13. This represents the transmission mode suggested in [15] for outbound communications, that is, the base station to portable link. Portable number "n" receives communications from the base station once every frame for a duration equal to one timeslot. Preliminary proposals for these values were for a timeslot equal to 2 msec and a frame duration of 16 msec though these values are subject change. The transmission rate was 225 kilo symbols per second with 2 bits per symbol.

We simulated a diversity technique assuming two branches on the portable receiver where the portable measures the average signal power on each branch during the preceding two timeslots and then chooses the best for reception. This assumes perfect synchronization of timeslots. The signal level measurement is either one or two timeslots old by the time the portable receives its information timeslot. An equivalent switching delay is thus 1.5 times the timeslot duration. Our objective in simulating this technique is to determine its sensitivity to timeslot duration and to compare it with the SAS technique. We selected the same representative data files for simulation to provide a direct comparison with SAS combining. Timeslot durations were varied from 2 to 40 msec. Detailed software listings for this simulation are provided in Appendix E.

Results shown in the four CDF plots are from the same data files as SAS results. That is, figures 5.14 (a), 5.14 (b), 5.15 (a) and 5.15 (b) can be directly compared to the

Figure 5.13: Simulated TDM frame format

(a) Space diversity pair branch 1 (BR1) and branch 3 (BR3)

(b) CDF for frequency diversity from branch 1 and 2

Figure 5.14: Typical results for switched TDM technique from path 101. Timeslot durations are in msec.

(a) Space diversity pair branch 1 (BR1) and branch 3 (BR3)

(b) CDF for frequency diversity from branch 1 and 2

Figure 5.15: Switched TDM technique CDFs from unusual fading on path 108. Timeslot durations are in msec.

SAS CDFs in figures 5.11 (a), 5.11 (b), 5.12 (a) and 5.12 (b). By comparing the results for the typical fading at path 101 to the SAS results, it is apparent that there is no knee to the CDF. This technique is superior to SAS at both low delays and large delays. The performance for a timeslot of 2 msec is also better than SAS with no delay.

The two results from path 108 in Fig. 5.12 indicate that this switched combining yields superior output statistics to that of SAS when the fading is considerably different on the two branches.

5.7.3 Discussion

The simulated results show a distinct performance advantage for the switched-TDM technique compared to SAS. Both techniques perform poorer than pure SC. The switched-TDM technique is inherently different than SAS in that it is continuously comparing the two branches against each other as opposed to SAS where the signal is compared to a threshold. Both techniques have a similar complexity so this is considered to be a fair performance comparison. However, for feedback diversity, SAS has considerably less complexity with one antenna at the portable and may therefore still be an attractive alternative for cordless telephone systems. Also, the second technique relies on a TDM outbound channel which is not an option for current systems such as CT-2 and DECT. Other switched techniques such as SAE described in section 2.3.2 may provide a better trade-off between outage performance and complexity. Further insight into the merits of these techniques could be gained by an analysis of other performance parameters such as the switching rate and the fade duration.

5.8. Bit error rate computation for DPSK

In this section we determine the BER performance of differential phase shift keying (DPSK) modulation using diversity with both Rayleigh fading and measured fading results. A useful technique to predict average BER knowing the first order statistics of the fading envelope and the receiver probability of error for a given SNR is to use a quasi-static approximation. This is valid when the rate of fading is slow enough that the channel can be considered static over a number of symbol intervals [21]. For a binary

modulation scheme under non-fading conditions, with no inter-symbol interference the probability of error (P_2) can be determined from the local SNR per information bit (γ_b) which can be expressed as $P_2(\gamma_b)$ where $\gamma_b = \alpha^2 E_b/N_0$ with α representing the channel gain. For a fading channel, α is a random variable and so is γ_b , therefore $P_2(\gamma_b)$ is a conditional probability conditioned on γ_b . The distribution of γ_b is proportional to that of α^2 and is dependent on the average SNR per bit, Γ_b given by:

$$\Gamma_b = E_b/N_0 E[\alpha^2] \quad (24)$$

and $E[\cdot]$ is the expected value operator. This conditioning can be removed by averaging $P_2(\gamma_b)$ over the pdf for γ_b . The pdf of the instantaneous value of γ_b is dependent on the mean SNR (Γ_b) and is written as $p(\gamma_b | \Gamma_b)$. The average BER is thus a function of the mean SNR and may therefore be obtained, for the case of binary signalling, by:

$$P_2(\Gamma_b) = \int_0^{\infty} P_2(\gamma_b) p(\gamma_b | \Gamma_b) d\gamma_b. \quad (25)$$

This relationship may be evaluated directly in the case when α is Rayleigh distributed or numerically when the statistics of the fading are determined empirically. We first evaluate this analytically for the case of Rayleigh fading and then evaluate the average BER numerically based on the measured fading statistics.

We use DPSK for this analysis since a reasonable estimate of performance can be made with information of the signal envelope only. We assume that there are no significant phase variations within the duration of two consecutive bit periods, (valid for a slowly fading channel and bit rates of interest) [21].

BER for Rayleigh fading

Here we determine the diversity improvement in average BER for DPSK using two-branch and four-branch selection combining operating over a flat, slowly fading Rayleigh channel.

(a) Determine mean BER for single branch.

The distribution of γ_b is proportional to that of α^2 which has a chi-squared distribution, and thus is of the form:

$$p(\gamma_b | \Gamma_b) = \frac{1}{\Gamma_b} e^{-\gamma_b/\Gamma_b} \quad \gamma_b \geq 0 \quad (26)$$

Given the static BER performance of DPSK :

$$P_2(\gamma) = 1/2 e^{-\gamma} \quad (27)$$

and the pdf from equation 26, we average over all γ to obtain the average BER :

$$P_2 = \int_0^{\infty} P_2(\gamma) p(\gamma) d\gamma = 1/2\Gamma \int_0^{\infty} e^{-\gamma} e^{-\gamma/\Gamma} d\gamma = 1/(2(1 + \Gamma)) \quad (28)$$

where we let $\Gamma = \Gamma_b$ to simplify the notation.

(b) Determine mean BER for two-branch and four-branch case.

Assume that mean SNR's on each branch are equal to Γ .

From equation 13, the cumulative distribution function of γ_k on the k^{th} branch is :

$$P(\gamma_k \leq x) = 1 - e^{-x/\Gamma} \quad (29)$$

and from equation 14 for the selection combined SNR of L branches, (γ_c), the CDF is :

$$P(\gamma_c \leq x) = \left[1 - e^{-x/\Gamma} \right]^L \quad (30)$$

Next, we differentiate with respect to x to obtain the modified pdf $p_{\gamma_{c2}}(x)$ for the two-branch SC output :

$$p_{\gamma_{c2}}(x) = d/dx \left\{ \left[1 - e^{-x/\Gamma} \right]^2 \right\} = 2/\Gamma \left[e^{-x/\Gamma} - e^{-2x/\Gamma} \right] \quad (31)$$

and for four branches:

$$p_{\gamma_{c4}}(x) = d/dx \left\{ \left[1 - e^{-x/\Gamma} \right]^4 \right\} = 4/\Gamma \left[e^{-x/\Gamma} - 3e^{-2x/\Gamma} + 3e^{-3x/\Gamma} - e^{-4x/\Gamma} \right] \quad (32)$$

The average BER can now be found for two-branch SC:

$$\begin{aligned}
 P_{2i} &= \int_0^{\infty} P_2(\gamma) p_{\gamma,2}(\gamma) d\gamma = 1/\Gamma \int_0^{\infty} e^{-(1+1/\Gamma)\gamma} - e^{-(1+2/\Gamma)\gamma} d\gamma \\
 &= 1/(\Gamma + 1) - 1/(\Gamma + 2) \\
 &= 1/((\Gamma + 1)(\Gamma + 2))
 \end{aligned} \tag{33}$$

and similarly for four-branch SC:

$$\begin{aligned}
 P_{24} &= \int_0^{\infty} P_2(\gamma) p_{\gamma,4}(\gamma) d\gamma \\
 &= 2/\Gamma \int_0^{\infty} e^{-(1+1/\Gamma)\gamma} - 3e^{-(1+2/\Gamma)\gamma} + 3e^{-(1+3/\Gamma)\gamma} - e^{-(1+4/\Gamma)\gamma} d\gamma \\
 &= 2 [1/(\Gamma + 1) - 3/(\Gamma + 2) + 3/(\Gamma + 3) - 1/(\Gamma + 4)]
 \end{aligned} \tag{34}$$

This result indicates that the average BER for L-branch selection combining varies with $(1/\Gamma)^L$ as it does for maximal ratio combining. The average BER performance of DPSK with selection combining derived in equations 28, 33, and 34 are plotted as dotted lines in Fig. 5.16.

BER for measured fading

Here we estimate the average BER of DPSK based on the measured diversity results. For this we use the measured discrete CDFs. The worst case and average CDFs are evaluated from 35 OBS paths measured in the engineering annex from which there were 140, 70 and 35 one minute measurements for four-branch, two-branch and single branch diversity respectively. The CDF results were shown earlier in Fig. 5.8 of section 5.6 which are for the diversity parameters $\Delta F=10$ MHz and $\Delta S=1$ λ with ideal SC and include the variations due to unequal mean signal levels between branches.

The average BER was computed as follows:

1. Compute average and maximum CDF for group of measurements as described in section 5.4.1 under the heading, "Fading statistics for groups of measurements". The software for this procedure is detailed in Appendix E.2.

2. Convert the stored CDF results using a discrete differentiation to obtain the conditional pmfs $p_k(\gamma_b | \Gamma_b)$ conditioned on the average SNR. This is done by taking the difference between consecutive CDF values in each power bin.
3. For a range of average SNR, $0dB \leq \Gamma_b \leq 40dB$, we compute the following:
 - a. Following equation 25, the average BER is the summation (discrete integration) of the product of the pmf and the static BER formula as shown here:

$$P_2(\Gamma_b) = \frac{1}{2} \sum_{k=1}^N e^{-\gamma_b} p_k(\gamma_b | \Gamma_b). \quad (35)$$

- b. For each Γ_b , equation 35 is evaluated for $k = 1$ to 51 representing the power bins from -40 dB to $+10$ dB. The local SNR takes on the values $-40 + \Gamma_b \leq \gamma_b \leq 10 + \Gamma_b$.

The software for this procedure is detailed in Appendix E.5 and E.6. The result of these computations using the worst case and average CDFs from one-, two- and four-branch combining is shown in Fig. 5.16.

Discussion

The average BER results based on the average CDF's are fairly close to the Rayleigh fading references shown in dotted lines. The relative performance to the Rayleigh case can be explained by the statistics on which these results are based as shown earlier in Fig. 5.8. The improvements compared to single branch fading at a BER of 10^{-3} were 12.3 dB and 18.4 dB for two-branch and four-branch diversity respectively.

The worst case (maximum) results differ considerably from the averages. This can also be explained by the earlier CDF curves. The benefit of higher order diversity is noticed here also as it was in section 5.6. The variations between the average case and the worst case at a BER of 10^{-3} were 6.7 dB, 5.5 dB and 2.3 dB for one-, two- and four-branch diversity respectively.

Figure 5.16: BER of DPSK with average and worst case statistics from 35 OBS paths for $\Delta F = 10$ MHz and $\Delta S = 1\lambda$ compared to ideal SC with Rayleigh fading shown in dotted lines.

5.9. Application of diversity results to a link budget

Assumptions

In this section we present an application of the diversity results to a link budget based on parameters similar to the Digital European Cordless Telecommunications system (DECT). We analyze the DECT system for the following reasons:

1. DECT is an emerging international standard for which there are a sufficient number of known parameters to allow a detailed analysis. The Bellcore framework is not yet solidified.
2. Our measurements have been conducted in buildings which is the target area of application for DECT. The Bellcore framework considers outdoor applications.
3. DECT hardware specifications which allow for soft handoff would permit the use of both space and frequency diversity while no such arrangement exists for CT-2.

We consider a base to portable transmission for a single cell on one floor of an office. The link budget format follows an approach similar to the one in [2]. The total link gain in dB should exceed the total link attenuation, the difference being the link implementation margin. The complete link budget is shown in table 5.6 with the following sections describing the entries.

Receiver sensitivity

The receiver sensitivity is based on hardware specifications for DECT from [14, 19]. It is assumed that the noise bandwidth of the receiver is equal to the physical channel bandwidth of 1728 KHz. The noise figure of 3 dB is based on [2] and the SNR threshold is based on discriminator detection of gaussian minimum shift keying (GMSK) modulation for a BER of 10^{-3} [19].

Total link gain

Antenna gains for the portable are estimated to be lower than that for the base station [2]. The peak transmit power is based on 10 milliwatts average power but with a duty cycle of 1 in 24 based on information in table 2.1. Thus the peak power is 240 milliwatts or +24 dBm.

Total link attenuation

The total attenuation is the summation of three components, the large scale pathloss versus distance, a log-normal distributed variable representing shadowing and a variable to represent the short term fading which is often estimated as Rayleigh distributed. Macro-diversity and micro-diversity are applied to reduce two fading components.

Large scale pathloss

Results of large scale pathloss from our measurements of the engineering annex presented in [39] and by others in various buildings [25, 26, 9, 27, 28, 41] predict the variation of the mean signal level versus distance. For example, using results from [39] the mean pathloss is calculated as follows:

$$PL = FSL(1m) + 10 \times N \times \log_{10}(D) \quad (36)$$

Where FSL(1m) is the free space loss at one metre at a carrier frequency of 1.75 GHz (37.3 dB), $N = 2.85$ and D is the distance in metres. For a cell radius of 200 m the estimate of mean pathloss is 102.9 dB.

Shadowing margin based on service probability and macro-diversity

The service probability defines the percentage of physical areas that will have a signal level above a particular value and thus dictates the extra margin required to overcome the log-normally distributed variations due to shadowing. When macro-diversity is employed, the shadowing margin can be reduced. For example it has been shown that the required shadowing margin using four-branch macro-diversity is 10 dB for 99% service probability [2, 7] and that without diversity a margin of 25 dB would be required.

Diversity type and parameters

For high quality portable communications we consider applying a four-branch micro-diversity system with two antennas with a separation of $\Delta S = \lambda/4$ on the portable terminal. The other two branches could be realized by using frequency diversity which could be supported within the DECT protocol. Since a total bandwidth of 20 MHz has been allocated to the service, a frequency separation ΔF of 5 MHz would be practical. In this example we use measured results from the four buildings in section 5.5 for four-branch hybrid diversity with $\Delta F \approx 5$ MHz and $\Delta S = \lambda/2$. (Based on the primary measurements, there was a negligible difference between $\Delta S = \lambda/2$ and $\Delta S = \lambda/4$).

Fading margin based on time availability and micro-diversity

Time availability, or signal availability as previously described, specifies the percentage of time that the signal will be above a particular value for a given time period. When micro-diversity is employed, the fade margin can be reduced. For example assuming the Rayleigh fading model without diversity, a fade margin of 20 dB would be required for 99% time availability. With four-branch SC diversity, the fade margin could be reduced to 4 dB as shown in Fig. 2.1 and table 2.2.

Adjustment to required fading margin based on measured results

Based on our measured results of the envelope CDF at 99% time availability shown in Fig. 5.9 in section 5.5 it is apparent that a value of 4 dB for the fade margin is optimistic. If each measurement period represented a telephone call then a significant number of the calls would not meet the desired availability specification. Our results for four-branch hybrid diversity indicated that based on 100 measurements in four buildings a fade margin of 10 dB would be required to meet the 99% availability specification at all 100 of the locations. Since 95 of the 100 measurements have a value of between -2.5 to -9.1 dB we could specify that 95% of the calls will meet the time availability specification of 99% if we apply four-branch diversity and allot a fade margin of 9.1 dB.

Discussion

The link parameters in table 5.6 has a positive implementation margin which is required for reliable operation. Without the benefit of diversity to combat fading and shadowing the total link attenuation would increase significantly. The link specifications are for 99% of locations, with 99% time availability during a one minute call for 95% of the telephone calls.

5.10. Summary of Analysis

In section 5.2 we reported on the envelope fading which showed deep, -30 dB fades. When fading statistics for OBS paths are measured over the duration of one minute there is a significant variation of 18 dB at 99% availability. The average CDF was close to Rayleigh distributed however, the worst case CDF was 6 dB lower than this at 99% availability.

In section 5.3 we presented results of envelope cross-correlation from the primary measurement area. For space diversity the mean values were less than 0.01. There was only a very small increase in mean values from 0.003 to 0.009 for $\Delta S = \lambda$ to $\Delta S = \lambda/4$. The result that there was such a low value of correlation for $\Delta S = \lambda/4$ might be attributed to temporal variations in the channel. There were also significant variations of correlation in individual measurements and there were an almost equal number of negative correlations to positive correlations. It is the maximum correlation that yields the worst diversity performance. Similar negative correlations were observed by Bergmann and Arnold [11] for polarization diversity. The frequency diversity results showed higher mean correlation values from 0.80 to 0.16 for $\Delta F = 2$ and 10 MHz. These significant correlations for the small frequency separation suggests that this separation is approaching the coherence bandwidth of the channel. There were also large variations between different measurements however few negative values.

In section 5.4 results show clearly the benefit of selection combining to reduce the number of fades. In some cases the SC envelope fading is better than the joint CDF. These cases may be due to the negative correlation noted above. In sub-section 5.4.1 the

Table 5.6: Link budget for cordless telephone system
based on DECT standard using measured diversity results.

Link parameters	Value	Units
Thermal noise at 25 Celcius (kT_{amb})	-174	dBm/Hz
Noise bandwidth	+62	dB-Hz
Noise figure of receiver	+3	dB
SNR threshold for BER = 0.001	+11	dB
RECEIVER SENSITIVITY	-98	dBm
Receiver sensitivity (subtract)	-(-98)	dBm
Transmit antenna gain	+8	dBi
Receive antenna gain	-4	dBi
Peak transmit power	+24	dBm
TOTAL LINK GAIN	+126	dB
Large scale pathloss ($F=1.75$ GHz, $N=2.85$, $D=200$ m)	102.9	dB
Fading margin (4-branch hybrid diversity for 99% time availability in 95% of one minute calls)	9.1	dB
Shadowing margin (4-branch macro-diversity at 99%)	10	dB
TOTAL ATTENUATION	122	dB
IMPLEMENTATION MARGIN	+4	dB

CDF's from many measurements in the engineering Annex were averaged and compared. A comparison of the average CDF's shows that the *average* results are close to those predicted by a Rayleigh distribution. The space diversity results are slightly better than those for large frequency diversity separation. For smaller frequency separation of 2 MHz there was a 5 dB difference at 99% availability. Four-branch hybrid diversity average statistics were typically better than the four-branch Rayleigh distribution case. A comparison of average diversity gains at 99% shows that for space diversity a value of approximately 10 dB is attained at all three antenna spacings while for frequency diversity we require a minimum value of ΔF between 5 and 10 MHz for similar results. Four-branch hybrid diversity had an average gain of 15 dB for $\Delta F > 5$ MHz and $\Delta S \geq \sqrt{2}$.

In section 5.5 the results of the secondary measurements in four different buildings were presented. Two thirds of the paths were OBS and the remaining were LOS. A significant result from these measurements is that there was a relatively small variation in the output statistics of only 1.3 dB between the average results in the four buildings. This suggests that the detailed results from the primary area in the Annex may be useful in predicting performance in other buildings. The LOS paths showed considerable variation in the fading however the OBS paths typically had more severe fading.

In section 5.6 the variations in the selection combined fading statistics were studied. The single branch fading case from the Annex, which had a worst case fading that was 6 dB lower than the average at 99% availability, was compared to the selection combined outputs. For the case of $\Delta F = 10$ MHz and $\Delta S = 1\lambda$, the space diversity output had a worst case fading 5.1 dB worse than average and the four-branch hybrid scheme had a worst case fading 2.9 dB worse than average. The diversity showed a trend to not only reduce fading but to reduce the variability in the fading as the number of branches was increased. Based on measurements in all four buildings there were large variations of the CDF results at 99% availability as shown by the spread of the histograms. Fortunately there was a considerable reduction in the variation of fading for higher order combiners. The variations of the CDF at 99% were for single branch, frequency, space and hybrid diversity 24.3 dB, 15.1 dB, 13.1 dB and 7.8 dB respectively.

In section 5.7 two practical implementations of selection combining were compared.

The SAS combining showed effects of threshold variations similar to that predicted by theory. The effect of switch delay on this technique was significant. Delays longer than 10 msec severely reduced the diversity improvements for the typical results. In some unusual fading cases the SAS output performs poorly and in an unpredictable fashion. Despite these drawbacks the feedback diversity technique using only one antenna may be an attractive trade-off between performance and complexity. A switched technique based on a TDM frame format was also simulated. It used the average signal level on the two antennas in the two preceding timeslots as a basis for selecting the best branch. The results showed that this technique is less sensitive to time delays than SAS and that in general the results were within 1 dB of SC for a timeslot of 2 msec. This approach also had well behaved output statistics in unusual cases of varying input fading statistics.

In section 5.8 the average BER of DPSK was evaluated based on a quasi-static approximation. This was done for the average and worst case statistics. The average statistics yielded a BER close to the derived BER assuming Rayleigh fading. The gain in average SNR was 12.3 dB and 18.4 dB at a BER of 10^{-3} for two- and four-branch diversity. The worst case BER was significantly poorer than average by 6.7 dB, 5.5 dB and 2.3 dB for one-, two- and four-branch diversity respectively.

In section 5.9 we applied micro-diversity in a link budget for a cordless telephone system. The four-branch diversity performance was used to reduce the required fade margin to 9.1 dB. Based on the variability in performance we specified that the link would have an excess implementation margin of 4 dB for 99% of locations with 99% time availability. Due to the variation of diversity output statistics this specification was further defined to be valid for 95% of the telephone calls.

Chapter 6

CONCLUSION

6.1. Thesis summary

This research was motivated by the expanding market for indoor wireless communication systems. In order to provide high quality communications to users, a suitable technique is required to mitigate the severe signal fading that occurs in buildings. Diversity techniques which make use of multiple replicas of the transmitted signal have previously been proven effective in the mobile radio and HF radio environment. Diversity realizations can be relatively simple and are thus an attractive means to reducing deep signal fading for narrow band indoor radio. For these techniques to be useful, a good understanding of their performance characteristics is necessary. The best compromise between implementation complexity, and system level planning depends on the choice of the diversity parameters. For the three types of diversity considered in this thesis space diversity, frequency diversity and combined space/frequency (hybrid) diversity, it is the antenna and carrier frequency separations that are important. Trade-offs between different combining techniques also effect the complexity and performance balance.

After a review of portable communication systems, multipath propagation, and diversity techniques, presented in Chapter 2, a literature review was summarized in Chapter 3 covering current research in indoor radio propagation and diversity measurements. The current literature covers models of indoor radio propagation in detail however these studies did not address issues that are important to designing practical diversity systems in the 1.6–2.0 GHz band with realistic performance estimations. To this end a measurement apparatus was designed and experiments conducted in five university environments as described in Chapter 4. The primary measurement area was in an area referred to as the engineering Annex which had many rooms of various sizes divided by plasterboard walls. In the Annex there were multiple one-minute measurements per path to compare different antenna and frequency separations for mostly obstructed (OBS) paths. The secondary measurements were conducted in four buildings including open office layouts and highly isolated office areas and included many line of sight (LOS) paths. The analysis

in Chapter 5 was aimed at answering key points important to the implementation and performance prediction of space and frequency diversity techniques. These contributions are summarized in the following subsections.

6.1.1 Choice of parameters for space, frequency and hybrid diversity

The comparison of diversity performance with different parameters of ΔF and ΔS was based on extensive measurements in the primary measurement area, the engineering Annex. The diversity results were based on measurements with moving antennas, mostly OBS paths and typical amounts of movement of people. Data was analyzed for both envelope cross-correlation and diversity gain at a time availability of 99% using ideal selection combining (SC).

Results of average envelope cross-correlation coefficients were less than 0.1 for space diversity with $\Delta S = \lambda$, $\lambda/2$, and $\lambda/4$. The smallest of these spacings is only 4.3 cm which would allow use on a portable telephone. Average correlations less than 0.5 were obtained for frequency diversity with $\Delta F = 5$, 10 and 20 MHz. The effect of antenna separation on diversity gains was insignificant. For $\Delta S = \lambda$, $\lambda/2$, and $\lambda/4$ the average diversity gains were 10 dB. This is a useful result considering the small size of portable telephones. It should be emphasized that these results were for moving terminals and people.

Average diversity gain using selection combining was at least 10 dB for frequency diversity with $\Delta F > 5$ MHz. Measurements in the Annex with OBS paths indicated that with $\Delta F = 5$ MHz there was very little reduction in diversity gain as compared to $\Delta F = 10$ MHz and to space diversity performance. For $\Delta F = 20$ MHz there was no more improvement. With more LOS measurements in the other buildings there was a degradation of 1 to 2 dB for frequency diversity with $\Delta F = 5$ MHz as compared with space diversity. This suggests that for maximum frequency diversity benefit we may choose to use a value between 5 and 10 MHz.

Four-branch hybrid diversity with $\Delta F \geq 5$ MHz and $\Delta S \geq \lambda/4$, provided an average diversity gain ≥ 15 dB. The performance of this four-branch technique showed a dependence on ΔF similar to frequency diversity.

6.1.2 Direct comparison of three diversity techniques and the effects of different environments

A comparison of the improvement in envelope fading statistics between diversity types was based on measurements from 135 locations of one minute duration in the primary and secondary measurement areas which included five office areas in four buildings. These results included both OBS and LOS paths and were based on ideal SC. A result from fading statistics that is useful when input statistics vary considerably is the value of the cumulative distribution function (CDF) at 99% availability. This describes the necessary fade margin for a link budget design.

The difference between two- and four-branch diversity performance indicates that there are significant gains in using the greater number of branches. For OBS paths measured in the engineering Annex these average gains were approximately 5 to 6 dB above two-branch diversity which is comparable to theory based on Rayleigh fading. In the same area two-branch space and two-branch frequency diversity provided average gains of approximately 10 dB provided that ΔF was greater than 5 MHz.

In the secondary measurements areas the LOS paths had single branch fading statistics that were less severe, however, the diversity output statistics were comparable to those of the OBS paths. The LOS diversity gains were lower but more importantly the diversity output fading statistics for LOS and OBS cases did not differ drastically.

The effect of various buildings on the average output fading statistics were relatively small. In the open area of the library the two-branch space diversity CDF at 99% was -7.6 dB compared with -8.9 dB for the Mackenzie building with concrete walls. For frequency diversity with $\Delta F = 5$ MHz the best case was again the library with -9.3 dB and the worst case was the Administration building with -10.8 dB. For four-branch hybrid diversity which uses both space and frequency diversity, the best case was in the Annex with -4.8 dB and the worst case in Administration which had output fading of -6.4 dB at 99% time availability. The relatively small differences in average fading statistics suggests that results may be useful for predicting diversity performance in other buildings similar to these.

6.1.3 Variations in correlation, fading statistics and diversity performance

The average results summarized above allow a reasonable comparison between diversity types for different antenna and frequency separations. There were, however, significant variations in the envelope correlations and single branch fading between the individual one-minute measurements. These variations resulted in differences in diversity gain and diversity output fading statistics.

The distribution of correlation values for space diversity were centred about zero with some extreme values as low as -0.6 and up to 0.8 . The spread of correlation values for frequency diversity were dependent on ΔF . For $\Delta F = 2$ MHz the values were between 0.5 and $+1.0$ whereas for $\Delta F = 10$ MHz the spread was from -0.3 to $+0.85$. These variations are significant since the envelope correlation has a direct effect on diversity gain.

The input fading statistics between one-minute measurements over the one minute duration varied considerably. In the Annex for 868 OBS measurements there was a variation in the CDFs of 18 dB at 99% availability between the worst and best case. The average CDF was close to Rayleigh distributed but the worst case CDF was 6 dB lower than this. For the four buildings with LOS paths the difference between best and worst case CDFs was 24.3 dB.

Results from two-branch space and four-branch hybrid diversity with ideal SC for OBS paths showed that the worst case CDFs were 5.0 dB and 2.5 dB worse than their respective averages at 99% availability. There was a tendency for this difference to decrease with increasing diversity order. This effect was even more noticable in the results from the four buildings which included LOS paths. The values of CDF at 99% varied by 24.3 dB, 15.3 dB, 13.1 dB and 7.8 dB for single branch fading, frequency diversity, space diversity and hybrid diversity respectively. The diversity therefore reduced the variations in output statistics between one minute measurements.

6.1.4 Practical combining techniques

Simple implementations of diversity combining are required for portable terminals. Two sub-optimal techniques of selection combining (SC) were evaluated based on selected

measurements which were chosen to be representative both typical fading and unusual severe fading.

The switch and stay (SAS) technique which has the advantage of requiring only one receiver was compared with SC. The effect of varying the threshold agreed well with theory but the best performance of SAS was typically 1 to 2 dB poorer than SC. An even simpler implementation using feedback diversity was considered. This technique was sensitive to switching delays. For a delay longer than 10 msec the diversity gain was significantly reduced. This should be considered if applying feedback diversity in systems where the delay due to the time division duplex rate is near 10 msec as in DECT. In unusual and severe fading cases the performance of SAS was poor and unpredictable.

A switched technique based on a time division multiplex (TDM) frame format was simulated. It used the average signal level on the two antennas in the two preceding timeslots as a basis for selecting the best branch. The results showed that this technique is less sensitive to time delays than SAS and that in general the results were within 1 dB of SC for a timeslot of 2 msec. This approach also had well behaved output statistics in unusual cases of varying input fading statistics.

6.1.5 Application to digital communications and systems design

The mean BER of DPSK was evaluated for the average and worst case statistics from OBS path measurements in the Annex. This computation assumed that the fading was quasi-static. This was done for single branch fading, space diversity and four-branch hybrid diversity. The results of this showed that BER based on average single branch measured statistics was comparable to an analytic result using Rayleigh fading. The performance degradation using worst case statistics was 6.7 dB of SNR at BER of 10^{-3} . For space diversity there was a significant improvement in BER and there was a 5.5 dB degradation using the worst case statistics. For hybrid diversity the difference between the required SNR for average and worst case fading was reduced to 2.3 dB.

The diversity results were applied to a link budget for indoor wireless communications. The parameters were based on values suitable for DECT. To accurately predict the reliability of an indoor communication system we should account for the variabil-

ity of diversity output fading statistics between telephone calls. A link design may use the average performance, the worst case performance or a value representing a specified percentage of the calls. By choosing the best 95% of results, the diversity fade margin results from the four buildings at 99% time availability were -23.4 dB, -14.6 dB, -12.8 dB and -9.1 dB for single branch fading, frequency diversity, space diversity and hybrid diversity respectively. The example link budget was specified to be accurate for 95% of the telephone calls for the desired availability. These results suggest the use of extra fade margin to account for the variability in fading statistics.

6.2. Recommendations for further research

Our findings in this thesis suggest investigation into the following topics.

6.2.1 Further analysis with this data

Location specific effects on diversity performance

Further insight into diversity operation could be obtained by averaging CDF results from specific groups. The software is organized and the data is documented to allow division of measured data into relevant groups. For example, the effect of the proximity of people, the proximity of scatterers or the effect of distance could be studied. A study of the characteristics of the environment for the worse than average results may be useful.

Effect of temporal channel variations on envelope correlation

It was noticed that envelope correlation values were related to the number of moving people in the channel. This could be investigated further with this measured data. Such an analysis may provide insight into the cases where diversity performance may vary.

Model fading distributions

We may consider fitting this envelope fading data to Nakagami or multirician fading distributions which model fading worse than Rayleigh. We might then use an analytical approach to predict diversity performance.

Simulation with other practical SC techniques

Further study into the performance of switched diversity schemes is recommended. It would be useful to simulate the performance of these techniques for all the measured data. This would however require significant computing time. Further sub-optimal switched diversity schemes could be evaluated. For example a SAS algorithm with a variable threshold level may be useful. We may also consider a switch and examine strategy (SAE) confined to switch only at frame intervals and using a word error rate for the switch decision.

Impact of diversity on fade durations

The length of time between fades can be important in the context of TDD and TDM access protocols. In the event of burst errors due to a fade some protocols will re-transmit a word or frame. If the frame length was much longer than the average time between fades then the likelihood of good transmission would decrease. Analysis could be conducted on this measured data to determine the effect of diversity on the fade durations and the non-fade durations.

6.2.2 Further measurements with this apparatus

Various antenna positions and motion

Using the equipment described in chapter 4, further measurements could provide insight into the effect of various antenna positions such as near the ceiling and near the head with various orientations. One could consider random motion and different velocities. In particular it would be useful to collect data from stationary diversity antennas for application to wireless local area networks. This may be significant for space diversity performance. It would be useful to determine the smallest antenna spacing usable under static channel conditions.

Extend measurements to more buildings

It would be useful to conduct measurements in other buildings such as in a gymnasium, a shopping mall, factories and residential buildings. These different environments may have an impact on the required diversity parameters and the variability of the envelope fading statistics.

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Appendix A

Measurement environments

Figure A.1: Secondary measurement area in the engineering annex with transmitter locations (T) and receiver locations (R).

Figure A.2: Measurement locations on the second floor of Administration building

Figure A.3: Measurement locations on the third floor of Administration building

Figure A.4: Measurement locations on second level of library

Figure A.5: Measurement locations on top floor of Mackenzie building

Appendix B

Photographs of measurement equipment

Plate 1 1.75 GHz dual frequency transmitter (left) and four-branched diversity receiver (right).

Plate 2 Top view of receive antennas and guiding frame

Plate 3. Side view of receive antennas of our ship. It's separated by one half wavelength (8.6 cm).

Appendix C

Data acquisition hardware and software

The data acquisition employed the following hardware and software:

1. PC 286 computer with 40 Mbyte hard disk and Ethernet card.
2. Data acquisition card by Data Translation. Model DT2827 with the following parameters:
 - a. Input voltage 0 to +10 volts.
 - b. 12 bit, 4096 discrete voltage levels
 - c. Simultaneous sampling on four channels.
 - d. Sample rate of 1000 Hz.
3. Controlling software was a custom "C" program named "ad.exe" written by Mr. B. Sullivan at Carleton University. This wrote data to buffers and directly to the hard disk drive. The maximum data acquisition speed of 32,000 samples per second was limited by the speed of writing to disk. With four input channels this provided a maximum rate of 8,000 samples per second per channel.
4. Read/write optical drive by Corel with 650 Mbyte capacity per disk.

The hard disk was segmented to allow 32 Mbyte of collected data before transfer to the optical disk. For the one minute files collected in these measurements the integer format files were each 500 Kbytes allowing for over 60 files to be stored on the hard disk. Data was transferred from storage to file servers on the Sun Unix network via Ethernet. A custom program entitled "admat.exe" was used to convert the integer format data to Matlab readable files.

Data file format

The following variables are contained in each data file:

- Ch — matrix with four column vectors containing data (60,000x4)
- ChSamplesPerSec — sample rate per channel

CollectForSeconds — duration of measurement

CollectionDate — date of measurement

CollectionTime — time of measurement

DataFile — name of data file

NumOfChannels — number of channels is 4

Appendix D

Antenna verification

The omni-directional antennas were tested in a large anechoic chamber to verify performance. The measured parameters are as follows:

1. Return loss at 1.75 GHz was better than 22 dB indicating a good impedance match at 50 ohms.
2. The bandwidth for 20 dB return loss was 100 MHz.
3. The antenna gain relative to an isotropic radiator was 1.8 dBi.

Radiation patterns

1. Fig. D.1 shows the electric field (E-field) azimuth cut for a single antenna.
2. Fig. D.2 shows the effect on E-field azimuth cut when the second antenna is placed at a spacing of $\Delta S = \lambda/4$ at 0 degrees azimuth. There is approximately 5 dB reduction in gain in this direction. The gain reduction for other spacings was 2 dB and 1 dB for $\lambda/2$ and λ respectively.
3. Fig. D.3 shows the E-Field elevation cut for a single antenna.
4. Fig. D.4 shows the effect on E-field elevation cut when second antenna is placed at $\Delta S = \lambda/4$ at 90 degrees (horizontal) elevation.

Figure D.1: E-field azimuth cut for single monopole antenna

Figure D.2: E-field azimuth cut with second antenna spaced at $\Delta S = \lambda/4$ at 0 degrees

Figure D.3: E-field elevation cut for single monopole antenna

Figure L $\vec{E}$ -field elevation cut with second antenna spaced at $\Delta S = \lambda/4$ at 90 degrees

Appendix E

Program listings of analysis software

E.1. Conversion of data to power values in dBm

File: ADCtoDBM2.m

```
% Matlab program ADCtoDBM2 (sub program)
%
% Purpose
Convert ADC values into received power values in dBm. Revision 2 uses
a simple look up table approach using a new calibration matrix "newcal".
This is considerably faster than version 1. For explanation of the
variables, see the introductory text in file cdfbat.m.
%
% Author Stephen Todd
% Version # 1,2,3 % Date 1991 May 17, October 1, 1992 Feb.7

% filename math/ADCtoDBM2.m
% input files cal/fastcalnf10d.mat, and data vectors
% output files result variables created in workspace
% called by cdf8.m %

% A DATA FILE MUST BE PRESENT IN THE WORKSPACE (example= d131135.mat)
% This is obtained using admat.exe in DOS and transferring file to SUN.
%
% Load in the calibration file.
load cal/fastcalnf10d
ChSize = size(Ch);
numsamp = ChSize(1);
numchan = NumOfChannels;

% Calibration info. is stored in 4 column vectors in the matrix
"newcal". Each entry is the received power that corresponds to a
particular ADC % value. ADC value can be determined by subtracting
one from the row index. Thus the received power is simply the value of
newcal evaluated at the value of the data vector Ch plus 1. For faster
execution, initialize large matrices to zero.

rxdbm = zeros(numsamp,numchan); %
rxdbm(:,1) = newcal( Ch(:,1)+1, 1); % 1 1
rxdbm(:,2) = newcal( Ch(:,2)+1, 2); % 1 2
rxdbm(:,3) = newcal( Ch(:,3)+1, 3); % 2 1
rxdbm(:,4) = newcal( Ch(:,4)+1, 4); % 2 2
```

```

clear Ch; % clear raw data matrix

% Convert raw data into milliwatts. %
rxmw = zeros(numsamp,numchan);
for i = 1:4 rxmw(:,i) = (10).^(rxdbm(:,i)./10);
end;

% Calculate the average power per channel and the total average.
temppower = 0;
meanmw1 = mean(rxmw(:,1)); avg(1,1) = 10*log10(meanmw1); temppower = temp-
power + meanmw1;
meanmw2 = mean(rxmw(:,2)); avg(1,2) = 10*log10(meanmw2); temppower = temp-
power + meanmw2;
meanmw3 = mean(rxmw(:,3)); avg(2,1) = 10*log10(meanmw3); temppower = temp-
power + meanmw3;
meanmw4 = mean(rxmw(:,4)); avg(2,2) = 10*log10(meanmw4); temppower = temp-
power + meanmw4;
avgmw = temppower/numchan; avgdbm = 10*log10(avgmw);

% Calculate the maximum and minimum values of received power in rxdbm.
maxchdbm = max(rxdbm); minchdbm = min(rxdbm);
maxdbm = max(maxchdbm); mindbm = min(minchdbm);

% end of ADCToDBM2.m %

```

E.2. Computation of CDF with various combining

File: cdf9.m

```

% Matlab program CDF9 (sub program)
% Purpose
% Given data from file this calculates the
% normalized data vector, the average power
% the histogram, the PMF, and the CDF for
% the data vector and for combinations of
% the data vectors using the maximum value.
% Also calculates combinations of cdf's
% such as products, and calculates the
% diversity improvements.
% Author Stephen Todd
% Version # Date %
% 8 1991 Nov. 1
% 9 1992 Apr. 23
% 9 1992 June 23 add hyst.
% 9 1992 Aug. 28 % 9 1992 Sept 21 add SWTDM
% 9 1992 Sept 24 add some defaults
% filename cdf9.m

```

```

% input files cal fastcalnf10d.mat
% dat DkkkFmSn.mat
% output files result variables in workspace
% programs called ADCtoDBM2.m, sasd4.m, sasd4.mex
% sc_hyst.m, sc_hyst.mex
% calling programs cdfbatch2a.m, sasbat_auto.m
% sc_hyst_auto.m
% associated programs cdfavg.m, divimpmeas.m,
% corrhist.m, plot___
% test programs run 1st simcal2.m, randist2.m, simdet2.m
% version2 has: column oriented channel data stored in a n by 4 matrix,
% channels 1 2 3 4 and power from -90 to -40 dBm
% version3 calculates the value of numsamp from workspace variables
% version4 calculates correlation between received power on channels
% version5 uses matrix cdf(numchan,numbin) and "pwrbin" for power bins
% and it initializes large matrices to improve execution time.
% version6 has simplified calculation of milliwatt vectors and also
% calls cdfcombine.m and DivImp.m to combine cdf's and to
% calculate diversity improvements. See these programs.
% version7 also calculates max and min of rxdm to ensure in range
% Version 8 new features:
% Faster conversion program (ADCtoDBM2.m) in cdf8.m.
% Provides results for envelopes normalized by the
% average of 1 channel (as it was) or by mean of all 4.
% This is specified by the variable "env_norm_by # ch".
% Version 9 new features:
% This includes variations on the type of selection
% combining. Denoted by the variable "SC_type".
% = 'Ideal' Ideal combining, choose max value
% = 'SAS_D' Switch and Stay with Delay
% ref. Jakes pp 399-411
% = 'Hyst.' Combining with hysteresis
% eg. Bell Cellular receivers % = 'SWTDM' Switched diversity based on
TDM outbound
% eg. Bellcore protocol
%***** must load cal/fastcalnf10d *****
%***** must load a data file from dat *****
%***** must enter a value for env_norm_by_ch *****
%***** must enter a value for SC_type *****

if exist('Ch') == 1
ChSize = size(Ch);
numsamp = ChSize(1);
end;
numchan = NumOfChannels;
numbin = 51;
NumDivType = 7;
pwrbin=[ -40:1:10 ]; % Specify power bin boundaries for histogram

% Defaults -----

```

```

if exist('SC_type') == 0
SC_type = 'Ideal'; % default to ideal selection combining
end;
if exist('env_norm_by_ch') == 0
env_norm_by_ch = 1; % default to normalizing by one channel
end;
if exist('do_corr') == 0
do_corr = 0; % default to no correlation computation
end;
if exist('do_cdf') == 0
do_cdf = 1; % default to doing CDF computation
end;

% -----
% Convert raw data to 4 columnwise vectors of power in dBm stored in
% the matrix (numsamp X 4) called "rxdbm".
% Also, compute the average powers of each channel "avg(i,j)"
% and of all channels combined "avgdbm".
if exist('Ch') == 1
ADCToDBM2;
end;
% Calculate correlation coefficient between channels.
if do_corr == 1
chancorr = corrcoef(rxmw);
end;
clear rxmw;

% Normalize the received power.
% To include effects of variable local mean signals, we normalize by
% the mean power of all 4 channels combined.
% To exclude the effects of variable mean signal levels on each branch
% we normalize each branch by its mean power. This provides results that
% assume equal means ( as done for original analysis ).
rxdb = zeros(numsamp,numchan);
if env_norm_by_ch == 4
rxdb(:,1) = rxdbm(:,1) - avgdbm;
rxdb(:,2) = rxdbm(:,2) - avgdbm;
rxdb(:,3) = rxdbm(:,3) - avgdbm;
rxdb(:,4) = rxdbm(:,4) - avgdbm;
end;
if env_norm_by_ch == 1
rxdb(:,1) = rxdbm(:,1) - avg(1,1);
rxdb(:,2) = rxdbm(:,2) - avg(1,2);
rxdb(:,3) = rxdbm(:,3) - avg(2,1);
rxdb(:,4) = rxdbm(:,4) - avg(2,2);
end;
clear rxdbm;

% Create histogram of normalized receive power.
% To plot: bar( pwrbin,rxhist11 )

```

```

if do_cdf==1
[ rxhist11 ] = hist( rxdb(:,1), pwrbin );
[ rxhist12 ] = hist( rxdb(:,2), pwrbin );
[ rxhist21 ] = hist( rxdb(:,3), pwrbin );
[ rxhist22 ] = hist( rxdb(:,4), pwrbin );

    % Create PMF of normalized receive power.
% To plot: bar( pwrbin,pmf11 )
pmf11 = rxhist11/numamp;
pmf12 = rxhist12/numamp;
pmf21 = rxhist21/numamp;
pmf22 = rxhist22/numamp;

    % Calculate cdf of normalized receive power per channel.
% To plot: plot( pwrbin,cdf(1,:) )
% Version 9 modification uses matlab function "cumsum"
cdf(1,:) = cumsum(pmf11);
cdf(2,:) = cumsum(pmf12);
cdf(3,:) = cumsum(pmf21);
cdf(4,:) = cumsum(pmf22);
end; % of if

    % Combine the received envelope using the chosen combining type.
sldb = zeros(numamp,NumDivType);
% Compute Ideal SC normalized signals.
if SC_type == 'Ideal'

% Diversity type depends on channels combined. Diversity AF & AF
% -----
sldb(:,1) = max( rxdb(:,1) , rxdb(:,3) ); %space 11 21
sldb(:,2) = max( rxdb(:,2) , rxdb(:,4) ); %space 12 22
sldb(:,3) = max( rxdb(:,1) , rxdb(:,2) ); %frequency 11 12
sldb(:,4) = max( rxdb(:,3) , rxdb(:,4) ); %frequency 21 22
sldb(:,5) = max( rxdb(:,1) , rxdb(:,4) ); %hybrid 11 22
sldb(:,6) = max( rxdb(:,2) , rxdb(:,3) ); %hybrid 12 21
% For four channel hybrid diversity...
sldb(:,7) = max( sldb(:,5) , sldb(:,6) ); end;

    % Compute Switch and Stay combined envelopes with switching delay.
if SC_type == 'SAS_D'
% Check and set defaults.
if exist('th') == 0 | exist('delay') == 0
th = -10;
delay = 50; % in milliseconds
end;
dy = round(delay/1000* ChSamplesPerSec); % delay in samples
sldb(:,1) = sasd4(rxdb(:,1) , rxdb(:,3),th,dy,numamp);

```

```

sldb(:,2) = sasd4(rxdb(:,2) , rxdb(:,4),th,dy,numsamp);
sldb(:,3) = sasd4(rxdb(:,1) , rxdb(:,2),th,dy,numsamp);
sldb(:,4) = sasd4(rxdb(:,3) , rxdb(:,4),th,dy,numsamp);
sldb(:,5) = sasd4(rxdb(:,1) , rxdb(:,4),th,dy,numsamp);
sldb(:,6) = sasd4(rxdb(:,2) , rxdb(:,3),th,dy,numsamp);
% For four channel hybrid diversity (apply SAS to SAS output) (not opt.)
sldb(:,7) = sasd4(sldb(:,5) , sldb(:,6),th,dy,numsamp);
end;

```

```

% Compute combined envelopes with hysteresis.
if SC_type == 'Hyst.'
% Check and set defaults.
if exist('th') == 0 | exist('delay') == 0
th = -3;
delay = 10; % in milliseconds
end;
dy = round(delay/1000* ChSamplesPerSec); % delay in samples
sldb(:,1) = sc_hyst2(rxdb(:,1) , rxdb(:,3),th,dy,numsamp);
sldb(:,2) = sc_hyst2(rxdb(:,2) , rxdb(:,4),th,dy,numsamp);
sldb(:,3) = sc_hyst2(rxdb(:,1) , rxdb(:,2),th,dy,numsamp);
sldb(:,4) = sc_hyst2(rxdb(:,3) , rxdb(:,4),th,dy,numsamp);
sldb(:,5) = sc_hyst2(rxdb(:,1) , rxdb(:,4),th,dy,numsamp);
sldb(:,6) = sc_hyst2(rxdb(:,2) , rxdb(:,3),th,dy,numsamp);
% For four channel hybrid diversity (apply sc_hyst to output) (not opt.)
sldb(:,7) = sc_hyst2(sldb(:,5) , sldb(:,6),th,dy,numsamp);
end;

```

```

% Compute combined envelopes with Bellcore's switched TDM protocol.
if SC_type == 'SwTDM'
% Check and set defaults.
if exist('timeslot') == 0 | exist('frame') == 0
timeslot = 2; % in milliseconds
frame = 16; % in milliseconds
end;
if exist('method') == 0
method = 2;
end;
ts_samp = round(timeslot/1000* ChSamplesPerSec); % in samples
fr_samp = round(frame/1000* ChSamplesPerSec); % in samples
sldb(:,1) = sw_tdm(rxdb(:,1) , rxdb(:,3),ts_samp,fr_samp,numsamp,method);
sldb(:,2) = sw_tdm(rxdb(:,2) , rxdb(:,4),ts_samp,fr_samp,numsamp,method);
sldb(:,3) = sw_tdm(rxdb(:,1) , rxdb(:,2),ts_samp,fr_samp,numsamp,method);
sldb(:,4) = sw_tdm(rxdb(:,3) , rxdb(:,4),ts_samp,fr_samp,numsamp,method);
sldb(:,5) = sw_tdm(rxdb(:,1) , rxdb(:,4),ts_samp,fr_samp,numsamp,method);
sldb(:,6) = sw_tdm(rxdb(:,2) , rxdb(:,3),ts_samp,fr_samp,numsamp,method);
sldb(:,7) = zeros(numsamp,1); % no four branch version of algorithm
end; % of if

```

```

clear rxdb;

```

```

    % For each combined envelope, compute the CDF.
    if do_cdf ==1
    for j = 1:NumDivType
    % Calculate histograms of selection combined receive vectors
    % To plot: bar( pwrbin,sldbhist(j,:) )
    [ sldbhist(j,:) ] = hist( sldb(:,j) , pwrbin );
    % Create pmf of selection combined signals from histogram
    % To plot: bar( pwrbin,sldbpmf(j,:) )
    sldbpmf(j,:) = sldbhist(j,:) / numsamp;
    % Calculate cdf of selection combined signals from histogram
    % To plot: plot( pwrbin,sldbpdf(j,:) )
    sldbpdf(j,:) = cumsum(sldbpmf(j,:));
    end

    % Call to subprogram "cdfcombine" to compute averages and products of
    cdf's
    cdfcombine;
    end; % of if
    % Call to subprogram "DivImp" to calculate diversity improvement vectors.
    DivImp;
    % end of CDF9 %

```

E.3. Switch and stay combining

File: sasd4.c

This C program implements the same function as the Matlab function in sasd4.m. It was written in C and linked to Matlab with the hope that it would run much faster than the Matlab equivalent. Initial Edit Date: Apr. 23, 1992. Last Edit Date: Apr. 23, 1992. Last Edited File: sasd4.c. Written By: Steve Todd. Compiler: cc (Sun 4 c). Compilation & link commands: "cmex -O4 sasd4.c"

```

function sas_out=sasd4(r1,r2,th,dy,numsamp) */

#include <math.h> #include <cmex.h> int irint(); #define null 0L
/* Input Arguments */
#define r1In 0 #define r2In prhs[1]
#define thIn prhs[2] #define dyIn prhs[3]
#define numsampIn prhs[4]

/* Output Arguments */
#define sas_outOut plhs[0]

static SASD4(sas_out, r1, r2, th, dy,numsamp)
double sas_out[], r1[], r2[], th; int dy,numsamp;
/* This is the computation of the output combined envelope * (

```



```

    unsigned int k, N, numsw, next_sw, branch, last_branch; N=numsamp -
    1; sas_out[0]=r1[0]; last_branch=1; numsw=0; next_sw=0;

```

```

    for (k=1; k<N; k++) {

        if ( last_branch == 1 )
        { if (r1[k-1] == th || r1[k] == th)
          { next_sw=k+dy; }
          if (k==next_sw)
          { branch=2;
            sas_out[k]=r2[k];
            numsw=numsw+1; }
          else
          { branch=1;
            sas_out[k]=r1[k]; }
          }
        else
        { if (r2[k-1] == th || r2[k] == th)
          { next_sw=k+dy; }
          if (k==next_sw)
          { branch=1;
            sas_out[k]=r1[k];
            numsw=numsw+1; }
          else
          { branch=2;
            sas_out[k]=r2[k]; } }
        last_branch = branch; } }

```

***** end of SASD4 *****

```

    user_fcn(nlhs, plhs, nrhs, prhs)
    int nlhs, nrhs;
    Matrix *plhs[], *prhs[];

```

```

    * This is the entry point to the Matlab linkable routine.  Matlab
    * * calls this function when the .MEX file is invoked.
    * * Inputs:
    * * 1) int nlhs: Number of left-hand-side arguments
    * * 2) int nrhs; Number of right-hand-side arguments
    * * 3) Matrix *plhs[]: An array of pointers to the lhs arguments.
    * * It contains nlhs pointers.
    * * 4) Matrix *prhs[]: An array of pointers to the rhs arguments.
    * * It contains nrhs pointers.
    * {

```

```

    unsigned int col;
    double *numsamp, *dy;
    double *sas_out;
    double *r1, *r2, *th;

```

```

    * Check for proper number of input & output argument *
    if (nrhs != 5)
        mex_error("Function SASD4 requires 5 input arguments.");
    if (nlhs != 1)
        mex_error("Function SASD4 requires one output argument.");

    * Assign pointers to the various parameters *
    r1 = (*prhs[1]).pr;
    r2 = (*prhs[2]).pr;
    th = (*prhs[3]).pr;
    dy = (*prhs[4]).pr;
    numsamp = (*prhs[5]).pr;

    * Check input argument ranges *
    if ((*numsamp) > 65535)
        mex_error("Function SASD4 requires (numsamp) <= 65535.");

    * Create a matrix (numsamp rows by 1 column) for the return argument *
    col = *numsamp;
    sas_outOut = create_matrix(col,1,REAL);
    sas_out = (*sas_outOut).pr;

    * Do the actual computations in a subroutine *
    SASD4(sas_out,r1,r2,*th,rint(*dy),rint(*numsamp));

    /* end user_fcn */

```

E.4. Switched diversity with TDM format

File: sw_tdm.m

```

function sw_tdm = sw_tdm(r1,r2,ts,fr,numsamp,method);
% Matlab program sw_tdm.m %
% Purpose %
% Switched combining algorithm for Time Division Multiplex (TDM) trans-
mission. This models the outbound channel when a TDM format is used based
on Bellcore's proposals. The two preceding frames are analyzed, once for
antenna 1 and once for antenna 2. The antenna with the highest average
power is selected. (for 2 branch only)
% Reference: FA-TSY -001013 issue 1, March 1990. pg 45.
% (later we may use hysteresis with a threshold) %
% Author Stephen Todd %
% Version Date % 1 1992 Sept. 21 %
% filename sw_tdm.m
% calling programs cdf9.m %

```

```

% INPUT PARAMETERS
% r1, r2 input signals in dB from each branch
% ts = time slot length in samples (eg 2)
% ts = time slot length in samples (eg 2)
% num_samp = number of samples in file (eg 30,000)
% num_ts = for loop, ts = matrix factor

% COMPUTE PARAMETERS
% per frame = ts * % number of timeslots per frame (eg 3)
num_ts = num_samp / ts; % number of timeslots (eg 60000 / 2 = 30000)
num_ts = num_samp / ts; % number of frames (eg 60000 / 16 = 3750)

% CREATE TWO MATRICES WITH num_ts ROWS AND ts COLUMNS
% Each row is one timeslot with the column entries being the
% samples in the timeslot. This is used for averaging.
R1=zeros(ts,num_ts);
% define as the transpose R1' = r1;
% pg 2-35 in matlab manual: reshaping R1=R1';
% transpose to obtain desired form
R1=zeros(ts,num_ts);
% define as the transpose R2' = r2;
% pg 2-35 in matlab manual: reshaping R2=R2';
% transpose to obtain desired form

% AVERAGE COLUMNS IN MATRICES AND PUT RESULT IN LAST COLUMN
% Since values are in dB we use a subprogram called "meandb".

R1(:,3)=meandb_rows(R1);
% compute mean of each row and put in col. 3 R2(:,3)=meandb_rows(R2);
% compute mean of each row and put in col. 3

% IMPLEMENT SELECTION COMBINING ROUTINE FOR EACH TIMESLOT
% Starting with timeslot 3 we look at the average signal of the
% two previous timeslots. For the largest signal branch, we assign the
% samples in that branch to the current output timeslot.

if method == 1
% The for loop will iterate num_ts times minus 2. (eg 2 to 30,000)
branch = zeros(num_ts,1);
r_out=zeros(num_samp,1);
branch(1)=1; % arbitrary decision
r_out(1:2)=r1(1:2);
branch(2)=1;
r_out(3:4)=r1(3:4);
for ts_count = 3:num_ts
if R1(ts_count-2,3) > R2(ts_count-1,3)
branch(ts_count)=1;
r_out(ts_count*ts-1:ts_count*ts)=r1(ts_count*ts-1:ts_count*ts);
else
branch(ts_count)=2;
r_out(ts_count*ts-1:ts_count*ts)=r2(ts_count*ts-1:ts_count*ts);
end
end

```

```

end; % of if
end; % of ts_count

    else % use faster vector method

        % Shift R1 down by 2 ts and R2 down by 1 ts. Then compare them.
        RP1=zeros(num_ts+2,1);
        RP1(3:num_ts+2)=R1(1:3);
        RP2=zeros(num_ts+2,1);
        RP2(3:num_ts+1)=R2(1:3);
        RPdif=RP1-RP2; % difference vector
        RRsing=sign(RPdif); % +1 = br1, -1 = br2
        RR10=ceil((RRsing+1)/2); % +1 = br1, 0 = br2

        brlon=ones(num_ts,1); % branch indicator by time-slot
        brlon(3:num_ts) = RR10(3:num_ts); % branch indicator by time-slot
        %Create vector indicating branch selection by sample-. %
        brlon_samp=ones(numsamp,1);
        % predetermine the size
        for k_ts = 1:ts
            brlon_samp(k_ts : ts : numsamp-ts+k_ts) = brlon;
        end; % of k_ts

        %Use branch selection vector to create final output vector. %
        r_out = r1 .* brlon_samp + r2 .* (1-brlon_samp);

    end; % of if
    sw_tdm=r_out;
% end of sw_tdm.m %

```

E.5. Computation of average BER from envelope fading CDF

File: avg_ber_from_cdf.m

```

function y = avg_ber_from_cdf(cdf,pwrbins,gamma,mod_type,range);
% % Matlab program avg_ber_from_cdf.m %
% Purpose
% Compute average BER versus average Eb/No,
% given the CDF of fading envelope, the desired range
% of Eb/No (Gamma) values and the modulation type.
% Author Stephen Todd %
% Version # 1, 2
% Date 1992 February 14, Apr. 22 %
% filename ~todd math/avg_ber_from_cdf.m

```

```

% input file: vector of  $\gamma$ , Gamma, pwrbin
% output file: vector of average BER
% calling programs interactive and others %

    env_pmf = cdf_to_pmf(cdf);
    numG = max(size(gamma)); numbin = max(size(pwrbin));
    if mod_type == 'dpsk' % DPSK
        Pe = ber_dpsk(pwrbin', gamma, numbin, numG);
    else
        if mod_type == 'dqpsk' % DQPSK
            Pe = ber_dqpsk(pwrbin', gamma, numbin, numG, bug);
        else
            echo on
            % Invalid modulation type.
            echo off
        end;
    end;

    PeAvg = Pe' * env_pmf';
    y = PeAvg;

```

E.6. Computation of BER for DPSK

File: ber_dpsk.m

```

function y = ber_dpsk(env, gamma, samples, num_gamma);
% Matlab program BER_DPSK.m (function) %
% Purpose % % Compute BER of DPSK
% operated over the slow fading channel as described
% by the 1, 2 & 4 branch envelope fading measurements
% that are stored as vectors in the data files. %
% Author Stephen Todd %
% Version # 1 % Date 1992 February 11 %
% filename ~todd/math/ber_dpsk.m
% input files vectors rxdb and Gamma in the workspace
% output files matrix of ber
% calling programs ber.m %
% env is the normalized envelope power in dB
% Gamma is the average  $E_b/N_0$  at which to evaluate the BER %

    y = zeros(samples, num_gamma);
    for i = 1:num_gamma
        y(:,i) = 0.5 .* exp(-10 .* ((env + gamma(i))./10));
    end;

    % end of BER_DPSK.m %

```

END

15-1 2-93

FIN