1

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

National Library
of Canada

Bibliothèque nationale
du Canada

Acquisitions and
Bibliographic Services Branch

Direction des acquisitions et
des services bibliographiques

395 Wellington Street
Ottawa, Ontario
K1A 0N4

395, rue Wellington
Ottawa (Ontario)
K1A 0N4

Your file Votre référence

Our file Notre référence

NOTICE

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

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

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

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

AVIS

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

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

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

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

Canada

Filter Structures for Equalization and Diversity Combining

by

Shivakumar Venkataraman, B. Tech.

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

Master of Engineering

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

May 12, 1994

© copyright

1994, Shivakumar Venkataraman

National Library
of Canada

Acquisitions and
Bibliographic Services Branch

395 Wellington Street
Ottawa, Ontario
K1A 0N4

Bibliothèque nationale
du Canada

Direction des acquisitions et
des services bibliographiques

395, rue Wellington
Ottawa (Ontario)
K1A 0N4

Your file *Votre référence*

Our file *Notre référence*

**THE AUTHOR HAS GRANTED AN
IRREVOCABLE NON-EXCLUSIVE
LICENCE ALLOWING THE NATIONAL
LIBRARY OF CANADA TO
REPRODUCE, LOAN, DISTRIBUTE OR
SELL COPIES OF HIS/HER THESIS BY
ANY MEANS AND IN ANY FORM OR
FORMAT, MAKING THIS THESIS
AVAILABLE TO INTERESTED
PERSONS.**

**L'AUTEUR A ACCORDE UNE LICENCE
IRREVOCABLE ET NON EXCLUSIVE
PERMETTANT A LA BIBLIOTHEQUE
NATIONALE DU CANADA DE
REPRODUIRE, PRETER, DISTRIBUER
OU VENDRE DES COPIES DE SA
THESE DE QUELQUE MANIERE ET
SOUS QUELQUE FORME QUE CE SOIT
POUR METTRE DES EXEMPLAIRES DE
CETTE THESE A LA DISPOSITION DES
PERSONNE INTERESSEES.**

**THE AUTHOR RETAINS OWNERSHIP
OF THE COPYRIGHT IN HIS/HER
THESIS. NEITHER THE THESIS NOR
SUBSTANTIAL EXTRACTS FROM IT
MAY BE PRINTED OR OTHERWISE
REPRODUCED WITHOUT HIS/HER
PERMISSION.**

**L'AUTEUR CONSERVE LA PROPRIETE
DU DROIT D'AUTEUR QUI PROTEGE
SA THESE. NI LA THESE NI DES
EXTRAITS SUBSTANTIELS DE CELLE-
CI NE DOIVENT ETRE IMPRIMES OU
AUTREMENT REPRODUITS SANS SON
AUTORISATION.**

ISBN 0-315-98589-5

Canada

Name SHIVAKUMAR VENKATARAMAN

Dissertation Abstracts International is arranged by broad, general subject categories. Please select the one subject which most nearly describes the content of your dissertation. Enter the corresponding four-digit code in the spaces provided.

ENGINEERING-ELECTRONICS & ELECTRICAL

SUBJECT TERM

0544 U.M.I.
SUBJECT CODE

Subject Categories

THE HUMANITIES AND SOCIAL SCIENCES

COMMUNICATIONS AND THE ARTS

Architecture	0729
Art History	0377
Cinema	0900
Dance	0378
Fine Arts	0357
Information Science	0723
Journalism	0391
Library Science	0399
Mass Communications	0708
Music	0413
Speech Communication	0459
Theater	0465

EDUCATION

General	0515
Administration	0514
Adult and Continuing	0516
Agricultural	0517
Art	0273
Bilingual and Multicultural	0282
Business	0688
Community College	0275
Curriculum and Instruction	0727
Early Childhood	0518
Elementary	0524
Finance	0277
Guidance and Counseling	0519
Health	0680
Higher	0745
History of	0520
Home Economics	0278
Industrial	0521
Language and Literature	0279
Mathematics	0280
Music	0522
Philosophy of	0998
Physical	0523

Psychology	0525
Reading	0535
Religious	0527
Sciences	0714
Secondary	0533
Social Sciences	0534
Sociology of	0340
Special	0529
Teacher Training	0530
Technology	0710
Tests and Measurements	0288
Vocational	0747

LANGUAGE, LITERATURE AND LINGUISTICS

Language	
General	0679
Ancient	0289
Linguistics	0290
Modern	0291
Literature	
General	0401
Classical	0294
Comparative	0295
Medieval	0297
Modern	0298
African	0316
American	0591
Asian	0305
Canadian (English)	0352
Canadian (French)	0355
English	0593
Germanic	0311
Latin American	0312
Middle Eastern	0315
Romance	0313
Slavic and East European	0314

PHILOSOPHY, RELIGION AND THEOLOGY

Philosophy	0422
Religion	
General	0318
Biblical Studies	0321
Clergy	0319
History of	0320
Philosophy of	0322
Theology	0469

SOCIAL SCIENCES

American Studies	0323
Anthropology	
Archaeology	0324
Cultural	0326
Physical	0327
Business Administration	
General	0310
Accounting	0272
Banking	0770
Management	0454
Marketing	0338
Canadian Studies	0385
Economics	
General	0501
Agricultural	0503
Commerce-Business	0505
Finance	0508
History	0509
Labor	0510
Theory	0511
Folklore	0358
Geography	0366
Gerontology	0351
History	
General	0578

Ancient	0579
Medieval	0581
Modern	0582
Black	0328
African	0331
Asia, Australia and Oceania	0332
Canadian	0334
European	0335
Latin American	0336
Middle Eastern	0333
United States	0337
History of Science	0585
Law	0398
Political Science	
General	0615
International Law and Relations	0616
Public Administration	0617
Recreation	0814
Social Work	0452
Sociology	
General	0626
Criminology and Penology	0627
Demography	0938
Ethnic and Racial Studies	0631
Individual and Family Studies	0628
Industrial and Labor Relations	0629
Public and Social Welfare	0630
Social Structure and Development	0700
Theory and Methods	0344
Transportation	0709
Urban and Regional Planning	0999
Women's Studies	0453

THE SCIENCES AND ENGINEERING

BIOLOGICAL SCIENCES

Agriculture	
General	0473
Agronomy	0285
Animal Culture and Nutrition	0475
Animal Pathology	0476
Food Science and Technology	0359
Forestry and Wildlife	0478
Plant Culture	0479
Plant Pathology	0480
Plant Physiology	0817
Range Management	0777
Wood Technology	0746
Biology	
General	0306
Anatomy	0287
Biostatistics	0308
Botany	0309
Cell	0379
Ecology	0329
Entomology	0353
Genetics	0369
Limnology	0793
Microbiology	0410
Molecular	0307
Neuroscience	0317
Oceanography	0416
Physiology	0433
Radiation	0821
Veterinary Science	0778
Zoology	0472
Biophysics	
General	0786
Medical	0760

EARTH SCIENCES

Biogeochemistry	0425
Geochemistry	0996

Geodesy	0370
Geology	0372
Geophysics	0373
Hydrology	0388
Mineralogy	0411
Paleobotany	0345
Paleoecology	0426
Paleontology	0418
Paleozoology	0985
Palynology	0427
Physical Geography	0368
Physical Oceanography	0415

HEALTH AND ENVIRONMENTAL SCIENCES

Environmental Sciences	0768
Health Sciences	
General	0566
Audiology	0300
Chemotherapy	0992
Dentistry	0567
Education	0350
Hospital Management	0769
Human Development	0758
Immunology	0982
Medicine and Surgery	0564
Mental Health	0347
Nursing	0569
Nutrition	0570
Obstetrics and Gynecology	0380
Occupational Health and Therapy	0354
Ophthalmology	0381
Pathology	0571
Pharmacology	0419
Pharmacy	0572
Physical Therapy	0382
Public Health	0573
Radiology	0574
Recreation	0575

Speech Pathology	0460
Toxicology	0383
Home Economics	0386

PHYSICAL SCIENCES

Pure Sciences	
Chemistry	
General	0485
Agricultural	0749
Analytical	0486
Biochemistry	0487
Inorganic	0488
Nuclear	0738
Organic	0490
Pharmaceutical	0491
Physical	0494
Polymer	0495
Radiation	0754
Mathematics	0405
Physics	
General	0605
Acoustics	0986
Astronomy and Astrophysics	0606
Atmospheric Science	0608
Atomic	0748
Electronics and Electricity	0607
Elementary Particles and High Energy	0798
Fluid and Plasma	0759
Molecular	0609
Nuclear	0610
Optics	0752
Radiation	0756
Solid State	0611
Statistics	0463

Applied Sciences

Applied Mechanics	0346
Computer Science	0984

Engineering	
General	0537
Aerospace	0538
Agricultural	0539
Automotive	0540
Biomedical	0541
Chemical	0542
Civil	0543
Electronics and Electrical	0544
Heat and Thermodynamics	0348
Hydraulic	0545
Industrial	0546
Marine	0547
Materials Science	0794
Mechanical	0548
Metallurgy	0743
Mining	0551
Nuclear	0552
Packaging	0549
Petroleum	0765
Sanitary and Municipal	0554
System Science	0790
Geotechnology	0428
Operations Research	0796
Plastics Technology	0795
Textile Technology	0994

PSYCHOLOGY

General	0621
Behavioral	0384
Clinical	0622
Developmental	0620
Experimental	0623
Industrial	0624
Personality	0625
Physiological	0989
Psychobiology	0349
Psychometrics	0632
Social	0451

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

"FILTER STRUCTURES FOR EQUALIZATION AND DIVERSITY COMBINING"

submitted by Shivakumar Venkataraman, B. Tech.

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

Thesis supervisor

Thesis supervisor

Chair, Department of Systems and Computer Engineering

Carleton University

July 12, 1994.

Abstract

The past few years have seen considerable growth in the areas of wireless communications. This trend is expected to continue with the emphasis shifting from variable speed, low quality voice communication services to high speed, high quality services incorporating voice, video and data. The major impediment which has to be overcome is the severe channel characteristics that cause multipath fading and intersymbol interference causing irreducible bit error rates. To meet the grade of service required for data transmission, the link has to be made very reliable using a combination of sophisticated signal processing and channel coding techniques. An important component of future systems is thus antenna diversity combining and equalization. Base stations can be expected to have long equalizers (to overcome and exploit long delay spreads) along with a significant order of antenna diversity combining (both microscopic and macroscopic). While offering considerable gains, equalization and antenna diversity combining do not use up the frequency spectrum and do not involve a huge increase in complexity.

This work is aimed at studying the issues involved in systems that use both diversity combining and equalization. We study several filter structures performing equalization and diversity combining in terms of the impact of several variables on the bit error rate performances. The variables include training length, signal to noise ratios, length of equalizer, tracking properties at different Doppler frequencies and presence of co-channel interference. The results indicate that filter structures differ significantly with respect to each of the above parameters and choice of the filter structure should be made with these effects taken into consideration.

A number of new filter structures are introduced which offer the same performance as that of well known structures but with significant reduction in the computational complexity when used in a slow fading channel with long equalizers (i. e. with a large number of taps) and sufficiently long training sequences. The RLS algorithm is used for adaptation. Some of the new filter structures are shown to offer significant enhancements in performance at a lower complexity when short training sequences are used.

Acknowledgments

I am deeply indebted to my thesis supervisors, Dr. Rafik A. Goubran and Dr. Samy A. Mahmoud, for their constant encouragement and guidance throughout the program. Their suggestions and help have played a crucial role in bringing the research to fruition. I also wish to express my gratitude to my friends and the departmental staff for helping me with the numerous problems I faced during the course of the work and as a new graduate student.

I am also thankful to the Telecommunications Research Institute of Ontario (TRIO) and to Carleton University for their generous financial assistance which supported the present work and for giving me an opportunity to continue my education.

Contents

1	Introduction	1
1.1	Thesis objective	3
1.2	Thesis organization	5
2	Indoor Radio Channel Model	6
2.1	Introduction	6
2.2	Mathematical modeling of the channel	6
2.3	Characterization of the impulse response	7
2.3.1	Distribution of the path amplitudes	8
2.3.2	Distribution of arrival times	10
2.3.3	Interdependence within path variables	11
2.3.4	Mean excess delay and rms delay spread	12
2.3.5	Other channel parameters	13
2.4	Channel model	14
2.5	Summary	18
3	Adaptive Equalization and Diversity Combining	19
3.1	Introduction	19
3.2	Diversity combining	19
3.3	Techniques to combat ISI	24
3.4	Equalization	26
3.5	Joint equalization and diversity combining	28
3.6	Other filter structures	31
3.7	Summary	33
4	Filter Structures for Equalization and Diversity Combining	34
4.1	Introduction	34

4.2	Classes of equalizer/combiner filter structures	36
4.2.1	Diversity combiner before equalizer	36
4.2.2	Equalizer before diversity combiner	41
4.2.3	Joint equalization and diversity combining	44
4.3	System description	46
4.4	Summary	47
5	Simulation Results	49
5.1	Introduction	49
5.2	Simulation results	49
5.2.1	Effect of channel model	50
5.2.2	Effect of fast fading	50
5.2.3	Effect of ray arrival rate and channel delay spread	53
5.2.4	Effect of training sequence length for a slow fading channel	56
5.2.5	Effect of equalizer span	58
5.2.6	Effect of co-channel interference	60
5.3	Comparisons between different structures	60
5.4	Summary guideline for appropriate filter structure	64
6	Conclusions	66
6.1	Thesis summary and conclusions	66
6.2	Future work	68
	References	69
	Appendix: Program Listing	73

List of figures

Fig. 2.1	(a) Exponential delay profile (rms delay spread = T) and (b) Two ray model with rays at $t = 0$ and $t = T$	17
Fig. 3.1	Optimum equalizer/combiner	30
Fig. 4.1	Optimum linear combiner/equalizer (FIL1).....	35
Fig. 4.2	Forward and backward filters used in the filter structures.....	37
Fig. 4.3	Estimation of channel impulse response $h(t)$ for FIL2 (shown above). t_s is the sampling rate at the output of the matched filter.	39
Fig. 4.4	FIL3 (above) and training of FIL3(below)	40
Fig. 4.5	FIL4 in decision directed mode of adaptation	42
Fig. 4.6	FIL5.....	43
Fig. 4.7	Joint equalization and diversity combining (FIL8).....	45
Fig. 5.1	Performance of filter structures for a two path channel.....	51
Fig. 5.2	Performance of filter structures with and without feedback for an exponential power profile.....	52
Fig. 5.3	Effect of doppler frequency on the different equalizer/combiner structures ..	54
Fig. 5.4	Effect of delay spread on a decision feedback equalizer.	55
Fig. 5.5	Effect of training length on the filter structures.....	57
Fig. 5.6	Effect of equalizer span on the performance of the equalizers.	59
Fig. 5.7	Effect of co-channel interference on the filter structures using dual diversity with (a) single interferer and with (b) 6 co-channel interferers (SNR= 30 dB)....	61

List of Abbreviations and Acronyms

ABER — average bit error rate
AMPS — Advanced Mobile Phone Service
AWGN — additive white gaussian noise
BER — bit error rate
BPSK — binary phase shift keying
CCI — cochannel interference
CDMA — code division multiple access
CW — continuous wave
dB — decibel
DFE — decision feedback equalizer
FDMA — frequency division multiple access
FIL_k — filter structure k
FSE — fractionally spaced equalizer
GHz — gigahertz = 10^9 Hz
GSM — Global Systems for Mobile communications
GWSSUS — gaussian wide-sense stationary uncorrelated scattering
HCMTS — High Capacity Mobile Telephone System
HF — high frequency
Hz — Hertz
IF — intermediate frequency
ISI — intersymbol interference
Kbits/s — kilo (10^3) bits per second
Ksymbols/s — kilo symbols per second
KHz — kilohertz
km/hr — kilometers per hour
LMS — least mean square
LOS — line of sight
Mbits/s — mega (10^6) bits per second
MHz — megahertz = 10^6 Hz
Msymbols/s — mega symbols per second
MLSE — maximum likelihood sequence estimation

MMSE — minimum mean squared error
ms — milliseconds = 10^{-3} seconds
MSE — mean squared error
NMT — Nordic Mobile Telephone
ns — nanoseconds = 10^{-9} seconds
OBS — obstructed
pdf — probability density function
QPSK — quadrature phase shift keying
RF — radio frequency
RLS — recursive least squares
RMS — root mean square
Rx — receiver
sec — second
SINR — signal to interference plus noise ratio
SIR — signal to interference ratio
SNR — signal to noise ratio
SRC — spectral raised cosine
SRRC — square root raised cosine
TACS — Total Access Communication Systems
TDMA — time division multiple access
Tx — transmitter
 μ s — microsecond = 10^{-6} second
VLSI — very large scale integration

List of Mathematical Symbols and Notation

- B — number of feedback filter taps
- B_c — coherence bandwidth
- B_d — Doppler spread
- d — distance of portable from transmitting base station
- $E[.]$ — expectation operator
- E_b — energy per bit
- f_c — carrier frequency
- $f(t)$ — combined impulse response of physical channel and pulse shaping filters
- $\mathcal{F}[.]$ — Fourier transform
- $g_k^{(l)}(t)$ — complex impulse responses of the k th channel of the l th user
- $h(t; \tau)$ — low-pass equivalent time-variant physical channel response at time t to an impulse applied at $t-\tau$.
- $H(f; \tau)$ — Fourier transform of $h(t; \tau)$
- L — number of forward filter taps
- $\ln(.)$ — natural logarithm
- $\log(.)$ — base 10 logarithm
- m — parameter of Nakagami distribution
- M — number of receiver antennas
- N — number of cochannel interferers; also used for number of multipath components
- N_0 — single sided noise power spectral density
- $P(d)$ — signal power at distance d
- P_b — bit error probability
- S/H — sample and hold
- $S_H(.)$ — Doppler power spectrum
- $s(t)$ — transmitted signal
- t — continuous time variable
- T — symbol interval
- T_c — coherence time
- T_s — receiver sampling interval
- $w(t)$ — array representing a linear combiner
- $X(f)$ — frequency response of raised cosine pulse

$\alpha(t; \tau)$ — continuous time multipath response attenuation function
 α_r — magnitude of r th complex channel tap coefficient
 β — roll-off factor of the SRC pulse
 $\delta(t)$ — Dirac delta or unit impulse function
 λ — wavelength; also used for path arrival rate of cluster channel model
 $\phi_h(t)$ — multipath delay profile for a particular response $h(t)$
 $\phi(t)$ — common multipath delay profile expression
 μ — mean of lognormal distribution
 ν — magnitude of the strong component in Ricean fading
 π — constant 3.141592654.....
 σ^2 — variance of process
 $\theta(t; \tau)$ — continuous time multipath response phase function
 τ — time delay variable
 τ_k — arrival times of rays of channel impulse response
 τ_m — mean delay spread
 τ_{rms} — rms delay spread
 τ_s — sampling rate of digital receive filter; also used for the tap spacing of the channel

Chapter 1

INTRODUCTION

The past decade has seen a phenomenal growth in wireless communications. Cordless telephones and cellular systems are now being widely used and the demand for these services is increasing rapidly. The trend is expected to continue with the number of cellular subscribers reaching 22 million by 1995 [1] in the U.S. while a European market [2] of 50 million users of mobile communication services is predicted for the year 2000.

The bright prospect has given a tremendous impetus to the development of new wireless communications services. The first generation technology of the 80s was characterized by analog voice transmission as in cordless telephones and most of the current cellular systems. In cellular systems such as AMPS (Advanced Mobile Phone Systems) in North America, TACS (Total Access Communication Systems) in the United Kingdom, NMT (Nordic Mobile Telephone) in Scandinavia and HCMTS (High Capacity Mobile Telephone System) in Japan, network control messages, e.g. for handoff or power control, are carried on the same channel that carries voice communication, interrupting speech transmission and thus limiting system capacity.

To meet the requirements of higher capacity, increased quality and increased range of applications, new systems have already begun to appear in the market [3]. The second generation networks will mainly carry digital speech and the wireless terminals will have enhanced capabilities of exchanging signalling and control information with the remainder of the network. Second generation vehicular mobile radio systems, such as the pan-european GSM (Global Systems for Mobile communications) system and the emerging dual-mode North-American IS-54 standard will have some channels that will be dedicated for the exchange of network control information between mobile terminals and the network infrastructure.

The overwhelming demand for wireless communication services has given sufficient reason for researchers and planners to look into a third generation of services to meet the communication demands of the next century. The third generation will herald the beginning of a new era of personal and portable communication services. This will allow an integrated access to cellular, cordless, satellite and wireline services. As a service con-

cept, personal communications [4] represents the ability to initiate or receive calls, i. e. voice or other information such as data, facsimile, video etc., anywhere and to maintain one's identity throughout the network. The service can be looked upon as one passing information to the individual via their own personal number rather than to devices as in the case with today's wireline or mobile radio telephones.

To meet the expected traffic demand, the capacity of the networks have to be increased drastically. This is possible with the use of microcellular system engineering techniques. Cellular systems of the '80s such as the AMPS, TACS, NMT and HCMTS are based on the concepts of frequency reuse and cell splitting. These concepts allows the system capacity to increase considerably, albeit at the cost of a greater complexity. Frequency reuse refers to the use of the same channel frequencies at different locations, sufficiently far apart, to reduce the co-channel interference to required levels. Instead of covering the entire coverage area by a single base station with a relatively high transmitter power i. e., a single cell approach, transmitters of lower power are distributed at various sites somewhat uniformly throughout the coverage area. Each new subarea covered by the low power transmitter is called a cell. Since each cell can now handle a traffic equal to that handled by the base station with the centralized approach, the capacity of the system is increased greatly. Thus, to meet any increase in traffic demand, it is necessary to use cell splitting to increase the number of cells per coverage area. However, there is a trade-off to increasing the level of cell splitting because there is a degradation of transmission quality resulting from more frequent handoff.

Third generation systems envisage the use of a portable unit with a lightweight, low-volume battery using power levels from 1 to 10 milliwatts. This is required both from the viewpoint of convenience and with regards to health and safety from radiation hazard in a multi-user environment. Such low power units can be operated only with very small cell sizes; very small microcells or picocells are being considered [5] for indoor communications which might encompass a room, an office or a house and its immediate environment. Cells with coverage areas smaller than the current ones for outdoor radio will also be required to accommodate the higher average traffic intensity and higher total traffic per day for indoor communications than for vehicular radio which are occupied only a fraction of the day. Base stations for personal portable communications can however be much cheaper than those for vehicular radio as the latter requires higher power base stations to

be mounted on tall buildings as against ceilings or lamp posts for the former.

To achieve the vision of universal and portable communications, a low power system must therefore be designed to complement current and future vehicular radio, mobile satellite, air-to-ground systems as well as wireline networks in such a way that all of these can be integrated into a unified access architecture. It is also expected that the traffic carried over future systems will see an increasing shift from voice to multimedia services incorporating voice, video and data. Thus, the system can be envisioned as one carrying data with a variety of requirements and grades of services over a host of channel conditions.

1.1 Thesis objective

This thesis is aimed at designing a receiver for such radio links. Within a coverage area, one has to determine first the process by which the common channel resource is shared. For indoor wireless communications, two different channel access schemes are envisaged:

- 1) *Time Division Multiple Access (TDMA)*: This technique is actually a combination of FDMA (frequency-division-multiple access) and TDMA; it is thus also referred to as FD-TDMA. The total frequency spectrum is allocated to an arrangement of ports in such a way as to minimize cochannel interference. Each of these ports divides the allocated portion of the spectrum into sets of time-divisioned multiplexed channels i.e. each channel uses the entire spectrum for a specific time-slot. According to the bandwidth of the TDM channels, one may distinguish two sub-classes of this technique:
 - a) *Narrowband TDMA*: When the allocated channel bandwidth for each channel is smaller than the coherence bandwidth of the channel, the fading is essentially flat¹. Typically, indoor narrowband TDMA schemes using ~20 Mbps use a large number of channels (24 or 48) [6] with data rates on each channel of 400 or 800 kbits per second (kbps). This technique has the advantage of simplified transceiver design; however, the relatively large number of channels complicates the channel allocation procedure and the transmission rates for each user are limited and typically fixed.
 - b) *Wideband TDMA*: When the channel bandwidths are larger than the coherence

1. The terms flat fading, frequency selective fading and coherence bandwidth are explained in the next chapter.

bandwidth of the channel, the fading becomes frequency selective, necessitating the use of intersymbol interference (ISI) mitigation techniques such as equalization. Though the transceiver design becomes slightly complicated, the channel allocation scheme now involves fewer channels and users have access to a flexible combination of transmission data rates to accommodate a variety of different application such as voice, video, facsimile etc.

2. Code division multiple access (CDMA): This approach is based on direct sequence spread spectrum modulation. Each user is assigned a unique spread spectrum code sequence. Each user uses the entire spectrum (half the spectrum for transmission and the other half for reception) and is distinguished from other users based on the code assigned. These code sequences must have low (ideally zero) cross correlation properties to ensure low interference between users' signals.

The CDMA system is relatively more complicated and requires solving several additional problems such as code synchronization and power control. This thesis considers the wide-band TDMA system, which can effectively use equalization and diversity combining techniques to achieve high performance. Although the channel condition does not restrict the scope of the applicability of the work, an indoor channel is assumed unless otherwise stated.

As has already been mentioned, future systems will carry a host of services over a variety of channel conditions. In subsequent chapters, it will be explained that the main disturbances to be overcome in the wide-band TDMA indoor channel are shadowing, frequency selective fading and cochannel interference. These impairments are overcome by the use of a suitable combination of antenna diversity combining - both microscopic and macroscopic, equalization and channel coding. Channel coding is particularly effective for fast fading channels. For coding to be effective for slow fading channels, it is necessary to use interleaving or to use a sufficiently high order of explicit diversity. It may not be possible to use the former option for the real-time nature of applications involving video or voice traffic of third generation systems. Channel coding is, however, far more powerful than explicit time or frequency diversity [7] and can be used in conjunction with a high order space diversity system.

This thesis investigates the performance and complexity of some equalization and diversity combining methods for wireless communications. As will be seen in Chapter 3,

the use of both of these techniques, both separately and jointly, has been widely reported in the literature. Both theoretical and simulation performance of systems using equalization and diversity combining by employing optimal receiver structures over a variety of channels have been studied in the literature. Performance of systems using different filter structures, both optimal and sub-optimal, can vary depending upon the channel condition and system parameters, such as delay spread, training sequence length and length of the equalizer. The purpose of the thesis is to study these performance differences with the aim of recommending appropriate filter structures for given channel-system conditions. It is likely that the equalizer/combiner filter structures would use the RLS algorithm for adaptation as these systems require fast convergence.

1.2 Thesis Organization

The thesis is organized as follows:

Chapter 2 is a study of the fading multipath channel. First, the issue of characterizing the portable indoor channel is addressed. These parameters and available literature will be used in selecting the appropriate model for this channel that can be used in the simulation study. The literature study will suggest the need to use equalization and diversity combining for an indoor wireless system. It can also be useful for arriving at a suitable signalling rate for a system employing diversity combining and equalization at the receiver to exploit the inherent diversity of the channel.

Chapter 3 provides a review of the background material and a survey of the literature available for equalization and diversity combining over fading channels. The principle of optimum baseband combining, which conjugates equalization and space diversity, is described.

Chapter 4 presents the structures of the different receiver filters using equalization and antenna diversity combining. Some new filter structures are also introduced. These will be seen to have lower complexity than the prevalent filters without losing on performance for many channel/system conditions.

In chapter 5, computer simulations are used to study the performance of systems employing the different filters. The chief results are presented and some discussions are presented on the use of the appropriate filter structure for a given channel condition.

Finally, chapter 6 summarizes the work of the thesis. The conclusions are presented and some possible directions for future research are suggested.

Chapter 2

INDOOR RADIO CHANNEL MODEL

2.1 Introduction

The performance of communication systems is dependent mainly on the transmission medium. Multipath fading, doppler spread and interference from other users are some of the chief impairments of the wireless channel. Little can be done to eliminate multipath disturbances. However, if the channel is well characterized, the effect of these impairments can be reduced by proper design of transmitter and receiver to “match” the channel. Moreover, if a good model is found, it can greatly aid in predicting system performance through computer simulation and analysis (if possible).

In this chapter, a brief overview of the wireless channel model is presented. Though the present work finds applicability in any fading channel, modeling of the channel is studied with respect to results that are available in the literature for indoor radio propagation. A general model that fits the scenario of both the indoor as well as the outdoor radio channel is later described. This model has been used for the simulation study in the thesis.

2.2 Mathematical modeling of the channel

In most cases, one considers a *statistical* impulse response model for the fading channel. This is because in most cases the number of potential reflectors is too large to incorporate into a purely deterministic model. Also, the locations of potential scatterers vary considerably as transmitters and receivers are moved from one place to another. Thus if one transmits an extremely short pulse, ideally an impulse, over a time-varying multipath channel repeatedly, the received sequence will be a train of pulses where the number of pulses, their sizes and the relative delays between them would vary from experiment to experiment.

Towards this end, we can model the channel as a time-varying filter with a dis-

crete low-pass complex impulse response given by: [7], [8]

$$h(t; \tau) = \sum_{k=0}^{N(t)-1} \alpha_k(t) \delta(\tau - \tau_k(t)) e^{j\theta_k(t)} \quad (2.1)$$

where t and τ are the observation time and application time of the impulse, respectively. $N(t)$ is the number of multipaths received, $\alpha_k(t)$, $\tau_k(t)$ and $\theta_k(t)$ are randomly time varying amplitude, arrival time and phase respectively. When there are a continuum of multipath components such as in the tropospheric scatter channel, the equivalent lowpass channel is more appropriately written as

$$h(t; \tau) = \alpha(t; \tau) e^{-j2\pi f_c \tau} \quad (2.2)$$

where f_c is the frequency of the carrier. The above model is a wide-band model which has the advantage that, because of its generality it can be used to obtain the response of the channel to the transmission of any transmitted (low-pass) signal $s(t)$ by convolving $s(t)$ with $h(t)$ and adding noise.

The impulse response is time-varying. For radio communications, as the user moves from one place to another, he/she is likely to experience a different impulse response at different points in space. These impulse responses are likely to be similar at points close by with respect to the carrier wavelength because the principle reflectors and scatterers are likely to be the same. These are the "small scale variations" in the channel. Another variation in the impulse response could be due to the fact that two receivers at the same antenna separation record very different channel responses. This is due to the variation in the geometry of the reflectors and results in "mid-scale variations". This type of variation is called *shadowing* and can lead to significant deterioration in performance unless compensated for by macroscopic diversity techniques. Yet another type of variation could be due to the various path losses that the signal undergoes as the receiver moves from one antenna separation to another. This is referred to as "large scale variations" and is typically characterized by a d^n degradation, where d is the base-mobile distance and n is usually a constant for the channel under consideration.

2.3 Characterization of the impulse response

Traditionally, under characterization of the impulse response of the channel, pri-

primary interest is given to the distribution of path amplitudes. However, for the indoor wireless channel, distribution of the arrival sequence has also received attention in the literature [8], [9]. Other parameters of interest are the distribution of path phases, interdependencies with path variables, temporal variations of the channel, large scale path losses, delay spread and the frequency dependence of the statistics.

2.3.1 Distribution of the path amplitudes

1) *Rayleigh distribution*: This is a well accepted model for small scale rapid amplitude fluctuations when there is no Line of Sight (LOS) component. In the model used at arriving at the distribution, it is assumed that the transmitted signal arrives at the receiver via N directions, each component having a (complex) signal strength given by $\alpha_n(t) e^{-j\theta_n(t)}$ that can be described by a time-variant phasor with an envelope $\alpha_n(t)$ and a phase $\theta_n(t)$. The envelopes are assumed to be equal as relatively large changes in the medium are required for a small difference. The phases for the different paths can however vary significantly because a single wavelength difference between the paths can cause the phase to change by 2π . The resultant is the vector sum of these components, which can be written as $\alpha \sum e^{-j\theta_n}$. As the path lengths can change in an unpredictable manner, the phase is modeled as a uniformly distributed random variable in $(0, 2\pi]$. The central limit theorem is now invoked in writing the sum as a complex gaussian random variable, the magnitude of which is Rayleigh distributed with probability density function given by [7]

$$Pr(r) = \frac{r}{\sigma^2} \exp\left(-\frac{r^2}{2\sigma^2}\right) \quad r \geq 0 \quad (2.3)$$

where σ is the most probable value of the random variable r .

Measurements have shown a good fit to Rayleigh distributions at 900 MHz, 1.8 GHz, 2.3 GHz, 21.6 GHz and 37.2 GHz [12], [13].

2) *The Rician distribution*: This model is the more appropriate model when there is a path stronger than the others, which most often is the LOS component. Now, the channel is the sum of a scattered Rayleigh vector in addition to the path with a deterministic amplitude and phase. The sum of these two gives the non-zero mean Gaussian distribution, the

magnitude of which is Ricean [7]

$$Pr(r) = \frac{r}{\sigma^2} \exp\left(-\frac{r^2 + v^2}{2\sigma^2}\right) I_0\left(\frac{rv}{\sigma^2}\right) \quad r \geq 0 \quad (2.4)$$

where I_0 is the zeroth-order modified Bessel function of the first kind, v is the magnitude (envelope) of the strong component and σ^2 is proportional to the power of the Rayleigh component.

Measurements have shown a good fit to the Ricean characteristic at 910 MHz [10] (even in the absence of a LOS component) and at 21.6 GHz and 37.2 GHz when directional antennas are used i.e when there is a LOS component [13].

3) The Nakagami distribution: When there is very little motion in a building within which radio waves are transmitted between fixed locations, the fading distribution has been found [10] to differ substantially from the Rayleigh characteristic even without the presence of a dominant direct path. This is the result of the establishment of a physically stationary spatial pattern of field maxima and minima. In such a case, the Nakagami distribution (also called the m -distribution), based on a model which permits the length of the scatter vectors to be random (unlike the model of the Rayleigh distribution where the length of the vectors are equal) characterizes the fading distribution more accurately [6]. The Nakagami distribution has also been found to characterize outdoor channels which have considerable time variation. The pdf of the envelope of the received wave is then given by [11]:

$$Pr(r) = \frac{2m^m r^{2m-1}}{\Gamma(m) \Omega^m} \exp\left\{-\frac{mr^2}{\Omega}\right\} \quad r \geq 0 \quad (2.5)$$

where $\Gamma(m)$ is the Gamma function, $\Omega = E(r^2)$ and $m = E(r^2)/\text{var}(r^2)$, with the constraint $m \geq 1/2$. Nakagami is a general fading distribution that reduces to Rayleigh for $m = 1$ and to the one sided Gaussian distribution for $m = 1/2$. It also approximates, with high accuracy, the Ricean distribution, and approaches the lognormal distribution under certain conditions [8].

Ray tracing techniques showed that the fast fading component of the CW envelope is Nakagami distributed [14].

4) *The lognormal distribution:* This distribution is encountered more often when considering the mid-scale variations in the channel. The pdf of the distribution is given as

$$Pr(r) = \frac{1}{\sqrt{2\pi}\sigma r} \exp\left(-\frac{(\log(r) - \mu)^2}{2\sigma^2}\right) \quad r \geq 0 \quad (2.6)$$

where μ and σ are the mean and variance of the distribution in dB. Thus, $\log(r)$ is a normal random variable. There have been a number of measurements where even the fast fading component of the signal was observed to be lognormally distributed [15], [16]. That the mean of local data are lognormally distributed has been established [16]. Another interesting observation is that the global variation of the impulse response over two office buildings has also been shown to be lognormal [15].

2.3.2 Distribution of arrival times

Here, the distribution of the τ_k in equation (2.1) is studied. Distribution of the arrival times has been modeled as Poisson, modified Poisson and piecewise functions.

1) *Poisson model:* Here the excess delay arrivals (rays arriving after the first ray) follow a Poisson distribution i.e. the ray arrival events are completely random. Following measurements [9] at 1.5 GHz, a variation of this model, the double Poisson model, has been suggested for the indoor mobile radio channel, in which multipath components were modeled to arrive in clusters with a mean cluster arrival rate Λ and a mean ray arrival rate within a cluster λ . However, the application of the double Poisson model to multipath data collected in several factory environments has been unsuccessful and the Poisson model was found to be satisfactory only for insensitive receivers (i.e. high threshold values) [16]

2) *Modified Poisson – The Δ -K model:* This model was suggested first for the outdoor mobile radio channel by Turin *et al* [17] and later for the indoor mobile radio channel also [15][18]. This model has two states S_1 and S_2 with different arrival rates viz. λ_0 and $K\lambda_0$ respectively. Arrival of a path at time t causes the model to be in state S_2 in $(t, t + \Delta)$ and a failure of a single path arrival during this interval puts it back in state S_1 . It was observed [15] that the parameter K was mostly less than 1, indicating that paths are more evenly distributed than the standard Poisson model would indicate. Application of this model to impulse response data collected in several factory environments at 1.3 GHz has not been

successful [16].

3) Piecewise functions: Measurements by Rappaport *et al* [16] have found the modeling of the interarrival times as either of the above two models unsatisfactory. Instead, the probability of a path arrival is computed by counting the total number of detected multipath components at a particular excess delay time and dividing by the total number of possible multipath components for each excess delay interval. Based on this computation, the authors have divided the probability density function into three piecewise continuous functions for the LOS topography and into two functions for the OBS (Obstructed) topography.

The distribution of the phase sequences θ_k in equation (2.1) has not received much attention in literature. It is usually considered uniform, which may be valid as the phase changes by 2π over a distance of only 30 cm at 1 GHz.

2.3.3 Interdependence within path variables

Of interest here are the correlations within a profile and the correlations between spatially separated profiles.

1) Correlations within a profile: Correlations exist within a profile due to the grouping property of local structures which also causes the arrival times to be correlated as in the Δ -K model. Analysis of impulse response data, at a carrier frequency of 1.3 GHz, collected in several factory environments [16] has shown that for the LOS topography correlation on amplitudes of initial paths is small. The amplitude of the LOS path is uncorrelated with the amplitude of the path at 7.8 ns, anticorrelated with the amplitude of the component at 15.6 ns and uncorrelated with subsequent multipath components. For excess delays greater than zero, amplitudes of two components are uncorrelated if the time difference between the arrivals are greater than 25 ns. For the OBS topography, the multipath component amplitudes become uncorrelated for excess delay separations of 25 ns. In either case, the correlations exist only if the paths arrive with excess delays less than 100 ns; otherwise the path amplitudes are uncorrelated. In another set of measurements [15] of office buildings, average correlations of 0.2-0.3 were observed between the first few paths.

2) *Correlations between spatially separated profiles*: The impulse response profiles collected in nearby areas tend to be correlated since the channel's structure does not change over short distances. It was observed [16] that for the LOS topography, multipath component amplitudes become uncorrelated at spatial separations of $\lambda/2$ and become slightly anticorrelated for separations of $2.5-3\lambda$ at excess delays less than 100 ns; for excess delays greater than 100 ns multipath components are uncorrelated for separations greater than $\lambda/2$. For the OBS topography, the multipath component amplitudes are uncorrelated for all receiver separations and excess delays. High correlation coefficients (0.7-0.9) were observed [15] between multipath components for spatial separations of about 2cm which decreases to about 0.1-0.2 for separations of 20 cm. The correlations showed a fair amount of insensitivity to antenna separation but depended on the excess delay. Paths with lower excess delays were more highly correlated than the later ones.

2.3.4 Mean excess delay and rms delay spread

These quantities can be defined as follows [8]:

$$\tau_m = \frac{\sum_k (t_k - t_A) a_k^2}{\sum_k a_k^2} \quad (2.7)$$

$$\tau_{rms}^2 = \frac{\sum_k (t_k - \tau_m - t_A)^2 a_k^2}{\sum_k a_k^2} \quad (2.8)$$

where t_A is the time of arrival of the first path, τ_m and τ_{rms} are the mean excess delay and the rms delay respectively. Thus, these quantities are the first moment and the square root of the second moment of the power delay profile ($|h(t)|^2$) respectively. As will be seen later, τ_{rms} is both a measure of the implicit diversity offered by the channel and also an indication of the potential for intersymbol interference (ISI).

There have been several measurements reported in the literature about the rms delay spreads of the indoor wireless channel. The first multipath delay spread measurements were reported by Devasirvatham [19] who made extensive impulse response measurements at 850 MHz. It was found that the rms delay spread could be as high as 420 ns

(in a large building) when there was no direct line of sight between the transmit and receive antennas; however, when there was a good direct path, the mean rms delay spread was found to vary from 25 ns in a medium size office building to about 125 ns in a large building. Rms delay spreads of upto 50 ns (total delay spread of upto 200 ns) were reported [9] at 1.5 GHz in a two storey medium size building. Median delay spreads of 96 ns for the LOS topography and 105 ns for the OBS topography were reported for factory sites [16]. Extensive measurements [15] made in two office buildings report rms delay spreads typically between 10 and 50 ns with mean values between 20 and 30 ns.

The delay spreads for indoor radio channels with a single base station can thus vary significantly. This can considerably complicate the design of the receiver. Moreover, the large propagation losses would necessitate the radiation of an unreasonable amount of power which is not feasible for the portable. Large signal power transmissions can also cause interference with systems in neighboring buildings. To overcome these problems, it was proposed [20] to either 1) divide the building into very small cells (or picocells) with adjacent cells using different frequency bands or 2) continue using large cells but replace the single antenna of each central station by a distributed antenna system or leaky feeder that winds its way through the hallways of the building. It was found that, with these techniques, the rms delay spread becomes limited to 50 ns, with a typical mean value of 25 ns, even in large buildings. The frequency allotment which could otherwise have been a problem with the picocell approach is somewhat counterbalanced by using wideband TDMA.

2.3.5 Other channel parameters

1) Number of multipath components: The number of multipath components that arrive at the receiver in an open building/factory environment [16] has been found to be Gaussian distributed with a mean that is uniformly distributed between 9 and 35 for LOS and between 11 and 36 for OBS topographies. The standard deviations were linearly related to the respective means. This distribution was also observed by Hashemi [15].

2) Large scale path losses: This is useful for determining the coverage area for an indoor communication system. The path loss can be modeled by an inverse exponent law i.e. $P(d) = P_0 d^{-n}$ where P_0 is the power at $d = 1\text{m}$ [8]. The exponent n depends on the environment and there have been measurements reporting values ranging from 1.2 in an open hangar to $n = 6$ in buildings with steel partitions [21]. Another method of characterizing

path losses is measuring the loss due to various structures that can be encountered along the path of transmission such as different kinds of walls, number of floors between transmitter and receiver and the loss encountered due to penetration of the signal into the building from outside it.

3) Temporal variations in the channel: The indoor channel statistics change with time due to motion of people and equipment and is therefore non stationary. Specifically, it is shown [10] that the channel is wide sense stationary only if data is collected over short intervals of time. A measure of the channel's temporal variation is the width of its spectrum when a single sinusoid is transmitted. This has been found never to exceed 4 Hz [10]. A user walking at the regular pace of 5 km/hr in a medium sized building has been found to encounter a completely new channel every 72 ms [20].

2.4 Channel model

From the above survey of the indoor wireless channel, we can see that none of the above characterizations has yet been unanimously recognized as representative of the channel. Simpler models, on the lines of the gaussian wide sense stationary uncorrelated scattering (GWSSUS) [22] model, can be used to give a meaningful performance analysis of the system. This model can be characterized in the following way [6], [7]: Assuming $h(t;\tau)$ is wide-sense stationary and the attenuation and phase shift associated with path delay τ_1 is uncorrelated with the attenuation and phase shift associated with path delay τ_2 , the autocorrelation function of $h(t;\tau)$ may be written as

$$E [h (\tau_1;t) h^* (\tau_2;t + \Delta t)] = \Phi_h (\tau_1;\Delta t) \delta (\tau_1 - \tau_2) \quad (2.9)$$

If we let $\Delta t = 0$, the resulting autocorrelation function $\Phi_h(\tau;0) = \Phi_h(\tau)$ is simply the mean square value of the impulse response or the average power output of the channel as a function of the time delay τ . For this reason, $\Phi_h(\tau)$ is called the multipath delay profile or the delay power spectrum of the channel.

The time-variant multipath channel can also be characterized in the frequency domain. The autocorrelation function of $H(f;t)$, the Fourier transform of $h(\tau;t)$ with respect to τ , in the frequency variable f , is referred to as the *spaced-time space-frequency correlation function* of the channel and is denoted by $\Phi_H(\Delta f;\Delta t)$. The assumptions of uncorrelated

scattering and wide-sense stationarity make $\Phi_H(\Delta f, \Delta t)$ dependent on frequency and observation time differences only. By setting $\Delta t = 0$, the autocorrelation function in the frequency variable $\Phi_H(\Delta f)$ is found to be the Fourier transform of the multipath delay profile $\Phi_H(\tau)$, whose width B_c (called the *coherence bandwidth* of the channel) is reciprocal of the rms multipath delay spread.

Apart from the above variations of the channel characteristics in the frequency domain, we have already seen that temporal variations of the channel have been measured (differences when Δt is nonzero). Such variations are manifested as Doppler shifts. The Doppler effects of the time variations of the channel are studied by defining a new function, the *Doppler power spectrum* $S_H(\xi)$ as the Fourier transform with respect to Δt in the Doppler frequency variable ξ of $\Phi_H(\Delta f = 0; \Delta t)$. The range of values over which $S_H(\xi)$ is essentially non-zero is called the *fading bandwidth* or *Doppler spread* B_d of the channel. The reciprocal of this quantity is the coherence time T_c of the channel and is a measure of the width of $\Phi_H(\Delta t)$.

Based on measurements, an equivalent delay power spectrum was derived [23] for the indoor channel. It is a WSSUS model with an exponential delay profile and was used for performance analysis of decision feedback equalization for the indoor channel. For the microwave LOS radio channel, a two beam model was fitted [24] to the general N scatter model of eqn. (2.1). Several researchers have used models of two or more beams for performance analysis of systems operating on these channels [6], [25], [26], [27], [29], [31].

In this thesis, the above WSSUS model has been adopted for the performance analysis of the indoor wireless communication systems. This was because, as was noted above, there is no single characterization which accurately describes the indoor channel satisfactorily. It was shown [23] that the exponential delay profile model gives results that were quite close to the performance obtained by using the “cluster model” [9]. The exponential delay profile was used since this profile was obtained by measurements by several researchers [9], [15], [23].

The simplified models that are used can be written as a variation of eqn. (2.1) with the number of paths and the time difference between successive paths being constant; the channel impulse response can now be written as

$$h(t) = \sum_{k=0}^K a_k \delta(t - k\tau_s) \exp(-j\theta_k) \quad (2.10)$$

where τ_s is the constant time difference between the paths, a_k is Rayleigh distributed and θ_k is a uniformly distributed random variable in $\{0, \pi\}$. Each of the a_k is equal to the magnitude of the k th channel tap gaussian and are generated as

$$a_k = \sqrt{x_k^2 + y_k^2} \quad (2.11)$$

where x_k and y_k are gaussian distributed random variables with variances given by

$$\left(\frac{1}{2}\right) E[a_k^2] = E[x_k^2] = E[y_k^2] \quad (2.12)$$

Results of decision feedback equalizer (DFE) analysis [28] using a discrete channel model have shown that for the DFE performance to be insensitive to path arrival rate, the latter normalized by the rms delay spread should be at least 2 i.e. $\lambda \tau_{rms} \geq 2$ paths per delay spread interval, where λ is the average arrival rate of paths per second. Accordingly, two paths per rms delay spread interval was used for most of the simulations of the thesis i.e. in eqn. (2.10), $\tau_s = \tau_{rms}/2$.

For the slow fading channel, a complex Gaussian random variable, $x_k + jy_k$, of appropriate variance was generated anew for each tap k for every channel realization. For each channel realization, a frame is transmitted consisting of a training sequence and information symbols, throughout which the channel was assumed constant. For the fast fading channel, the “sum of sines” model [32] was used to generate varying channel during the transmission of both training and information symbols with maximum Doppler frequencies upto 100 Hz. At the beginning of the frame for the fast fading channel, the starting phases of each of the eighteen oscillators are initialized with a realization of a uniformly distributed random variable in $\{0, \pi\}$. Though such high Doppler frequencies may not be experienced in an indoor system at a frequency of 1 GHz, it may a possibility with future systems offering broadband services at operating frequencies of upto 60 GHz.

Two channel models are considered: 1) the exponential power profile model and 2) Two ray model. These are shown in Fig. 2.1. The rms delay spread of the latter is half the tap spacing of the second ray. The profile of the exponential delay profile can be written as

$$\Phi_h(\tau) = \frac{1}{\tau_{rms}} e^{\frac{-\tau}{\tau_{rms}}} \quad \text{(exponential)} \quad (2.13)$$

(a)

(b)

Fig. 2.1 (a) Exponential delay profile (rms delay spread = T) and (b) Two ray model with rays at $t = 0$ and $t = T$

and that of the double spike model can be written as

$$\Phi_h(\tau) = \frac{1}{2} (\delta(\tau) + \delta(\tau - 2\tau_{rms})) \quad (2 - \text{ray}) \quad (2.14)$$

At five times the rms delay spread, the value of the exponential profile becomes 0.67% of its value at $t = 0$. This was the length of the channel used for various delay spreads in the simulation study. The spacing between adjacent taps, as has already been mentioned, was for most cases set to $\tau_s = \tau_{rms}/2$.

2.5 Summary

In this chapter, we have discussed the efforts at characterization of indoor wireless channels based on measurements. There seems to be little agreement on the characterization of the indoor channel. Contradictory reports might be due to inadequate data or differences in sensitivities of the measuring equipment. It may be a while before the channel is properly characterized (if at all, considering the differences in geometry of different locations). A simpler model, which finds applicability to both the indoor and outdoor radio channels, like a two ray or an exponential model, has been used in this thesis to study the performance of different filter structures.

Chapter 3

ADAPTIVE EQUALIZATION AND DIVERSITY COMBINING

3.1 Introduction

One of the major problems in high speed data communications over multipath channels is the intersymbol interference that results due to the finite coherence bandwidth of the channel. This problem is further compounded in the case of wireless communications by the fading that is characteristic of these channels. The former impediment is overcome by techniques such as adaptive equalization, maximum likelihood sequence estimation, spread spectrum, using higher level modulation or multi-carrier modulation while the latter is overcome by the use of diversity. In this chapter, available literature on diversity combining and adaptive equalization over the frequency selective fading channel and also work done on using both equalization and space diversity combining together is studied.

3.2 Diversity Combining

Several techniques exist to provide the receiver replicas of the same transmitted signal independently to overcome the effect of deep fades that is characteristic of multipath fading channels. These may be classified into macroscopic and microscopic diversity. Macroscopic diversity alleviates the effects of shadowing by selection of the best portable to port path among a number of ports or base stations. Microscopic diversity involves providing independent channels between a single portable and a port and is effective in overcoming the effects of fading (flat or frequency selective) and combating cochannel interference.

Another classification is based on the method of providing independent signal paths to the receiver. These include frequency diversity, time diversity, space diversity, angle-of-arrival diversity and polarization diversity. Of these, the first two use up the scarce frequency spectrum. Moreover, frequency diversity will need excessive band-

width to achieve decorrelation with the small delay spreads typically encountered in the indoor channel. Time diversity may require excessive time separation to achieve decorrelation in the quasi-static indoor channel. Hence antenna diversity, and particularly space diversity is attractive especially considering that at operating frequencies of 1 - 2 GHz, the decorrelation distance is only 5 - 15 cm. For future systems operating at 20 - 60 GHz, this distance is correspondingly reduced further. Another advantage with space diversity is that its order is restricted only by the complexity of the receiver and not by available spectrum. Thus, it is feasible to think of a base station with several orders of diversity reception and employing adaptive retransmission techniques and a dual order diversity at the portable in a high quality wireless communication system to reduce the error rates to acceptable limits.

Another classification involves the method used to combine the diversity signals; the most common combining techniques are selection, switched, equal-gain and maximal ratio. Briefly, selection diversity chooses the best of M diversity branches and hence requires M receivers; switched diversity is simpler as the antenna merely switches to another diversity branch when the current signal level falls below a threshold, thus requiring only one receiver. In equal gain combining, the cophased diversity branches are added together and in maximal ratio combining the cophased branches are added in the ratio of their branch signal powers. Maximal ratio combining is the optimum (yields maximum signal to noise ratio at the output and hence minimizes the bit error rate, BER) linear combining technique for a flat fading channel with only thermal noise impairment [30]. Yet another classification is predetection, which is done without loss of the phase information, and postdetection, which is done following some non-linear operation, combining techniques. The former is preferred as, although cophasing is required, they do not involve multiple carrier recovery loops (for coherent detection) for bit timing. For long delay spreads however, it may be important to use postdetection combining techniques in the form of equalization or explicit space diversity combining along with using equalization.

Diversity combining can significantly improve the bit error rate performance of communication systems over the flat fading channel [7]. When the channel is frequency selective, i. e. when the delay spread becomes significant with respect to the symbol interval, intersymbol interference places a lower limit on the achievable bit error rate. Unacceptable irreducible bit error rates, i. e. larger than 10^{-4} and as high as 10^{-2} , have been observed [26] for different modulation schemes and raised cosine pulse excess band-

widths when the rms delay spread exceed 5% of the symbol interval (corresponding to 2 Mbps for a channel with 25 ns rms delay spread). However, studies [29] of (predetection) maximal ratio combining with coherently detected BPSK and QPSK in a slowly-varying frequency-selective environment have shown that diversity can be used to raise the maximum permissible data rate. For an outage probability of 10^{-4} at a BER of 10^{-4} , it was shown that the data rate can be increased from 75 kbps for no diversity to 11 Mbps with 8 branch diversity for an rms delay spread of 30 ns. It was shown that as the number of antennas increase, the change in the maximum permissible delay spread for a given outage rate at any BER decreases, stabilizing at about 0.1 times the data rate for 8 branch diversity. It was also shown in the same paper that maximal ratio combining can yield twice the data rate as selection diversity, that QPSK affords nearly three times the maximum data rate as BPSK and that increasing the signal bandwidth correspondingly increases the maximum data rate. Similar results were also obtained earlier by Glance and Greenstein [31] who studied the BER performance for a frequency selective fading channel in the presence of cochannel interference with maximal ratio combining. The analysis showed that the average BER was independent of the shape of the delay power spectrum for normalized (to symbol duration T) delay spreads less than 0.2. Moreover, for these delay spreads there was only a marginal degradation in the average BER from the flat fading case.

The above results were obtained by assuming a fixed sampling time. Since the maximal ratio combiner weights and the output SNR both vary with sampling time in a frequency-selective channel, an 'ideal' maximal ratio combiner would determine both the best sampling time and the weights for that sampling time and track both the best sampling time and weights. Hence maximal ratio combining can be quite complex in wide-band systems. Selection diversity is not as complex and Valenzuela [33] studied its performance for coherently detected 4-QAM and an indoor channel with rms delay spread of 25 ns. Increasing the diversity order M from 1 to 2 with selections based on received energy, reduced the outage probability (for a BER of 10^{-4}) from 0.012 to 0.0004 at 2 Mbps, from 0.037 to 0.008 at 4 Mbps and from 0.15 to 0.05 at 8 Mbps. It was also noted that for the more complicated postdetection approach, the outage rate would be the square of the single-antenna outage probability; for 4 Mbps the outage would have been $0.037 \times 0.037 = 0.0014$. This indicates that the received energy, albeit simple to measure, is not always an ideal indicator of system performance.

Another predetection combining technique that is particularly useful for digital

radio with flat fading in the presence of co-channel interference is optimum combining [34]. Combining is achieved by a set of weights (one weight per antenna branch, i.e. “memoryless” filters) which adapt by means of an adaptive algorithm with information of the desired signal. Such a scheme exploits the correlation amongst the interfering signals at the antenna branches to arrive at a set of weights which yield a higher signal-to-noise ratio at the output of the combiner than maximal ratio combining, which is optimum only for the case of uncorrelated noise/interference corrupting the data streams. Optimum combining has shown to be beneficial for both spread spectrum systems [34] and TDMA systems [35].

A variation of optimum combining (MMSE combining) for frequency selective channels has been presented by Clark *et al* [25], where the taps are optimized by minimizing a function dependent on the impulse response of the channel. It was shown that MMSE combining can be effectively used to counter the effects of ISI and frequency selective fading. A simple analysis to determine the number of antenna branches, A_r , required to remove the effects of ISI and CCI has been presented by Falconer *et al* [36]. If we represent the linear combiner weights as an array $\mathbf{w}(t)$ where

$$\mathbf{w}(t) = \begin{bmatrix} w_1 \\ \cdot \\ \cdot \\ w_{A_r} \end{bmatrix} \quad (3.1)$$

and the sampled array output is expressed as [36]

$$\mathbf{w}^H \mathbf{g}^{(l)}(t) \big|_{t=iT} = \begin{cases} 1 & i=0, l=0 \\ 0 & i \neq 0, l=0 \\ 0 & l=1, 2, \dots, N \end{cases} \quad (3.2)$$

where

$$\mathbf{g}^{(l)}(t) = \begin{bmatrix} g_1^{(l)}(t) \\ \cdot \\ \cdot \\ g_{A_r}^{(l)}(t) \end{bmatrix} \quad (3.3)$$

and $g_k^{(l)}(t)$ are the complex impulse responses of the k th channel (seen by antenna element k) of the l th user of a total of $N+1$ users. The channels are frequency selective and the channel impulse responses are assumed to be time-limited to $|t| \leq DT$. Hence, in the time domain there are $2D+1$ nonzero samples in each $g^{(l)}(t)$, and equation (3.3) represents a set of $(2D+1)(N+1)$ equations in the A_r unknowns $w_1, \dots, w_{A_r}$ which can be expressed as

$$\mathbf{w}^* \mathbf{G} = \mathbf{u}^* \quad (3.4)$$

where

$$\mathbf{G} = [\mathbf{G}^{(0)} \quad \mathbf{G}^{(1)} \quad \dots \quad \mathbf{G}^{(N)}] \quad (3.5)$$

$$\mathbf{G}^{(l)} = [\mathbf{g}^{(l)}(-DT) \quad \dots \quad \mathbf{g}^{(l)}(0) \quad \dots \quad \mathbf{g}^{(l)}(DT)] \quad (3.6)$$

and $\mathbf{u}$ is a $(2D+1)(N+1)$ dimensional vector with a single "1" in the D th row and zeros elsewhere.

Thus the ISI and CCI can be eliminated by this "memoryless" array, provided that

$$A_r \geq (2D+1)(N+1) \quad (3.7)$$

and that the A_r rows of the matrix $\mathbf{G}$, defined in (3.4), are linearly independent. If the equation is not an equality, then there exist a set of solutions of which the solution which minimizes the noise variance at the output of the linear combiner (mmse solution) is chosen. It is in thus choosing an optimum solution that the MMSE combiner yields significant performance advantages over a scheme without diversity or even over the conventional diversity schemes such as selection or maximal ratio combining.

Postdetection diversity is also an attractive scheme for mobile radio because the cophasing function is not required and because switching noise due to abrupt phase change is not produced [37] and as has been noted above [33] gives better results than pre-detection combining because of knowledge of a sampling instant for frequency selective channels. In [37], the effect of correlation and unequal average power in the space diversity branches has been quantified. The degradation is 1 dB when the magnitude of the complex correlation coefficient between the two branches is 0.63 with equal branch pow-

ers. When the branch average powers differs by 3 dB, the degradation is 1.5 dB when the errors are dominated by AWGN but only 0.5 dB when the errors are dominated by cochannel interference (CCI) or delay spread.

Thus, diversity combining is a powerful technique to combat fading (both flat and frequency selective) and cochannel interference. As rms delay spreads get larger than 0.1 - 0.2 times the data rate, the combiner (except MMSE combiner, but it may be difficult to implement it) fails to perform efficiently and specific measures to combat ISI becomes crucial.

3.3 Techniques to combat ISI

Several techniques have been proposed in the literature to combat ISI caused by signal propagation over different channels. While diversity combining was already seen in the previous section to implicitly reduce ISI, these other techniques are more effective to combat ISI and exploit the inherent diversity when used over frequency selective fading channels. This section deals with the basic idea of the respective techniques and reviews some of their limitations. The selection of a technique for a particular application depends on many factors such as the complexity, bandwidth efficiency, required performance, channel characteristics etc.

Adaptive Equalization: There are two classes of equalization techniques viz. linear equalization, which employs a T-spaced transversal filter, and decision feedback equalization in which the past decisions are fed back to eliminate the effect of past ISI. Adaptive equalizers, first developed by Lucky [38], can use either the zero-forcing filter or use a minimum mean square error criterion to adapt to the desired filter response [7]. The zero-forcing equalizer suffers from noise enhancement which does not exist if the mse criterion is used for adaptation of the filter coefficients. Decision feedback equalization, first developed by Austin [39], is significantly superior to linear equalization when the effect of past decision errors is neglected.

The equalizer structures with a T-spaced filter for the feedforward section is optimal only if the analog filter preceding it is matched to the channel-corrupted pulse response and its output is sampled at the optimum time instant. In practice, the channel response is not known *a priori*, so it is not possible to design an ideal matched filter. In view of this difficulty, it is customary in practical applications to use a fractionally spaced feed-

forward filter preceded by the filter matched to transmitted pulse. The property of fractionally spaced equalizers (FSE) to reduce the sensitivity to the sampling phase can be explained [40] heuristically by noting that variations in the sampling phase have a much smaller effect on the amplitude and delay characteristics in the spectral overlap region of the sampled equalizer input when the tap spacing is smaller than T . In the same paper, the performance advantage gained by combining the ability of a DFE to equalize deep fades with the FSE in an FSE-DFE structure was analyzed.

Maximum Likelihood Sequence Estimation (MLSE): The use of the Viterbi algorithm for decoding sequences corrupted by ISI was studied by Omura [41] and Forney [42]. The Viterbi algorithm is an efficient way to search for the sequence with the maximum likelihood of transmission through the trellis of states representing all possible data sequences and is the optimum data estimator. The chief limitation of this method is the exponential growth in the complexity of the system with increase in the channel memory length and the need for channel estimation.

Spread Spectrum (SS): SS techniques can be used to combat ISI. The idea is to use long duration wideband signals. But since many users are being served at the same time, the SS system can ideally achieve a throughput comparable or much larger than the TDMA system employing short duration pulses of similar bandwidth. By employing a *rake receiver*, the dual advantages of eliminating ISI and exploiting the diversity present in the channel are obtained. Besides a significant increase in the complexity, the limitations of such a system are the unavailability of a sufficient number of orthogonal codes (which restricts throughput) and the perfect channel estimate required for the rake receiver.

High level modulations: A simple approach to reduce the ISI while maintaining the bit rate is to increase the level of modulation (MQAM or MPSK), thereby decreasing the symbol transmission rate. Such a bandwidth-efficient system requires a high-power linear amplifier for accurate transmission. Moreover in the case of fading channel, the accurate estimation of the channel amplitude and phase required for the system becomes very complex. Yet another basic limitation is the incapacity of the system to exploit the inherent diversity of the channel.

Multi-Carrier Modulation (MCM): This technique partitions the channel into a number of narrowband flat fading subchannels. The subchannels are then coded jointly to achieve

diversity and alleviate fading channels. To make the complexity feasible, a block of symbols are passed through a DFT block, the outputs of which are passed through an amplifier. The chief problem with such a system is the non-linearity associated with passing the DFT block through the amplifier.

In this thesis, we have used equalization as the technique for eliminating ISI. Apart from having only a modest increase in complexity, it can easily be used in conjunction with antenna diversity combining.

3.4 Equalization

In the previous chapter, techniques that can be used to reduce the rms delay spreads to about 25 ns were mentioned. With a 25 ns rms delay spread, a 0.1 normalized delay spread amounts to a bit rate of 4 Msymbols/s or 8 Mbps using QPSK for which outage rates of 10^{-4} at a BER of 10^{-4} was obtainable with diversity combining with four branch diversity. Thus, one would not need equalization for bit rates higher than this value, if the order of diversity were not constrained. Equalization would also be necessary for systems which do not use techniques to reduce the rms delay spread to about 25 ns.

Clearly, a four fold diversity order would not be feasible at the portable. Moreover, bit rates higher than this will soon become necessary, given that multimedia wireless communications is becoming a reality. Future systems supporting data rates upto 20 Mbps [43] necessitate the use of wideband channels. Increasing the data rate also has the advantage of an implicit diversity effect due to the availability of multiple fading beams at the receiver. This “hidden” performance advantage can never be obtained with a narrow-band scheme. Also, wideband systems have lower overall system complexity because of fewer frequency channels and hence less handoffs. However, increasing the bit rate results in a higher normalized delay spread causing a potential performance degradation due to intersymbol interference (ISI). To control the deleterious effects of ISI, it becomes necessary to use equalization at the portable to compensate the effect of reducing the number of diversity branches; the base station could also use equalization to enhance performance for a marginal increase in complexity.

Equalization on a single branch antenna can yield diversity equivalent to the number of paths (with equal powers) present in the multipath channel response. The effect of equalization on a fading multipath channel was first studied by Monsen [44], [45], [46], [47] where he showed that the performance of an adaptive DFE was better in the presence

of multipath than in its absence because the implicit diversity gain is larger than any intersymbol interference degradation. It was also shown that in the absence of decision errors, the DFE was superior to the equalizer without feedback. While the forward filter eliminates precursor ISI, i.e. ISI on which decisions have yet to be made, the backward filter operating on the reconstructed data sequence acts to eliminate postcursor ISI. As this can be done exactly with no noise enhancement, it follows that the DFE will exhibit superior ability to equalize deep fades given the premise of absence of decision errors. Under a minimum mean square error criterion (MMSE), this has been established mathematically [44]. Besides mitigating ISI, the receiver could eliminate timing jitter and Doppler shifts, as long as the variation in these rates are less than the rate of adaptation.

Monsen also compared [46] the performance of a linear equalizer with a decision feedback equalizer over a fading channel, while considering the effects of error propagation in the DFE. Since the error propagation effects was calculated to be negligible in a fading channel application and the decision feedback equalizer does not degrade drastically in the presence of frequency domain nulls, it was concluded that the DFE is superior to the linear equalizer without a decision error premise. In another study, [48], Wong *et al* considered the performance of *linear* equalizers over two different models for the microwave digital radio fading channel and found that their performance was only slightly poorer than the matched filter bound. It was suggested that non-linear receiver structures are not necessary.

However, because the backward filter eliminates all the past digit intersymbol interference, the number of transversal filter taps in the forward filters can be reduced relative to the number of taps required in the linear equalizer. This leads to an implementation advantage and improved adaptation since for the same convergence rate the self-noise due to adaptation increases with the number of taps. A thumb rule for calculating the number of backward taps was given by Monsen in the same paper. If LT is the delay spread of the channel, N is the number of forward taps spaced $T/2$ apart, then B , the number of backward taps, is given by:

$$B = \begin{cases} [L + N/2] & N \geq 3 \\ [L + 3/2] & N < 3 \end{cases} \quad (3.8)$$

where $[x]$ is the integer part of x .

The performance of a one-shot receiver operating with an ideal matched filter with

the assumption of zero ISI was compared [45] with the performance of a receiver operating in the presence of ISI and employing equalization. It was observed that even with ISI spreading over four data symbols, the degradation of the DFE from the optimal assumption of no ISI is only 2/3 dB, and even this was found to decrease with increase in the number of equalizer taps. This was confirmed for a practical receiver structure [46] where the ISI penalty in a DFE was found to be less than 1 dB. Hence, sophisticated ISI suppression techniques such as the Viterbi algorithm was not recommended. Thus, equalization can be seen to be a very effective means of exploiting the implicit diversity offered by a channel, an extra advantage being the significantly higher data rates that can be supported by the system.

The performance of adaptive equalization for indoor radio communications has been studied [23], [49] using the "cluster" channel model [9]. With an rms delay of 25 ns, a matched filter bound suggests that rates up to 64 Mbps with QAM modulation can be achieved with an outage of less than 10^{-3} with a BER of 10^{-4} [49]. An equalizer, realizable as a transversal filter, was shown to achieve performance close to this matched filter bound, supporting rates of 4 Mbps (1 Mbps at 100 ns rms delay spread) with an outage probability of 0.6×10^{-3} . An unequalized channel showed significantly higher outage performances, even at data rates as low as 1 Mbps. In another study [23] of decision feedback equalization for the indoor radio channel, it was shown using measured data, the above channel model and also using analysis of an "equivalent delay spectrum" that (for a decay rate of 15 ns) the performance of a BPSK/DFE modem with three forward (fractionally spaced) and three backward taps at 10 Mbps is equal to that of a BPSK modem (without equalizer) at 1 Mbps. These studies clearly indicate the suitability of equalization to exploit the implicit diversity and to counter ISI for the indoor radio channel.

3.5 Joint equalization and diversity combining

In the previous sections, it was seen that both equalization and diversity combining can be used to exploit the implicit diversity in the channel for frequency selective fading channels. Explicit diversity combining is always beneficial for both flat and frequency selective fading channels, whereas equalization can significantly improve performance only over a frequency selective fading channel. These observations have led several authors [27], [28], [45], [47] to consider the use of both equalization and diversity combining over a frequency selective fading channel.

The use of equalization with diversity combining was first investigated by Brady [50], where a fractionally ($T/2$) spaced transversal filter was placed after the filter matched to the pulse. This filter performed the functions of maximal ratio combining, equalization of delay among diversity signals, equalization of multipath distortion and removal of timing jitter. Mosen described [44] a filter structure with one backward filter for cancelling the post-cursor interference and multiple diversity forward receivers. He showed [45] that the optimal filter structure (achieving the minimum mean square error) is one with a filter matched to the received pulse, the outputs of which are summed and operated on by a transversal filter with tap spacings equal to the baud interval which removes ISI due to future symbols. The reconstructed symbols are then sent back through another baud-interval spaced transversal filter which eliminates past digit intersymbol interference. The filter structure is shown in Fig. 3.1, where $f_k^*(-t)$ is the filter matched to the combined response of the k th channel (physical channel and transmit filter). This same filter structure was derived by Balaban and Salz [27] as the optimal filter for performing equalization and diversity combining.

In another paper [47], a single adaptive transversal filter convolved with the filter matched to the pulse was used for both matched filtering and equalization. A theoretical approach was developed for the calculation of the average BER (ABER), including the effects of ISI due to multipath and the finiteness of the transversal filter. This analysis was applied to frequency-selective, Rayleigh fading, indoor wireless links by Despins [6]. Though this filter structure achieves the same error rate performance as the optimal filter structure derived by Balaban and Salz [27], they differ in their method of implementation. While in the latter, it is required to first estimate the channel and then use this estimate in calculating the equalizer coefficients, the parameters of the equalizer and combiner are estimated jointly in the former filter structure [47] during its adaptation.

The performance of systems using both equalization and diversity combining have been studied in several papers [47], [50], [28], [51]. Balaban and Salz analyzed [27] the performance of the outdoor mobile radio using bounds derived for average BER for systems employing equalization and diversity combining for the two path equal power channel. In that paper, the effect of several channel parameters on the BER of an ideal combiner/infinite equalizer was obtained through analysis. It was seen that as the delay between two paths increases from zero, the average BER decreases till it reaches an asymptotic low value for an rms delay spread of $0.6 \sim 0.7$. For this channel, it was shown

Fig. 3.1 Optimum equalizer/combiner

that each equal power path provides an implicit diversity gain equal to one order of diversity. Outage probability did not change with change in the filter roll-off factor. It was shown that in the worst channel conditions when flat-fading occurs, the average probability of error is about 4% in the absence of diversity, and 0.2% with diversity. These figures improve considerably as the normalized delay becomes greater than 0.6. In this case the optimum decision feedback equalizer-combiner yields 0.8% for the average error rate in the absence of diversity and 0.005% with diversity. Error propagation was not considered as dual diversity reduces the average probability of error, in some cases by two orders of magnitude, which is sufficiently low to avoid this problem.

The use of a QPSK/DFE modem using second order diversity was investigated [28] for the indoor radio channel model [9]. It was shown that for a typical indoor environment with rms multipath delay spread of 50 ns and channel bandwidth of 10 MHz, a QPSK/DFE modem with a minimum of three forward taps can provide data rates as high as 13.3 Mbps ($\text{BER} < 10^{-7}$ at $E_b/N_0 = 30$ dB). In another paper [51], the performance of a system using dual diversity combining and equalization was studied for the indoor radio channel model proposed in [16] using the analysis of [27]. For the LOS channel with a normalized delay spread equal to the symbol duration, it was shown that at 17 dB SNR with 20 Mbps QPSK transmission, the semi-infinite decision feedback equalizer achieved an outage rate of 0.0023 at a BER of 10^{-5} .

Results for the troposcatter channel were derived by Monsen [47]. It was seen that increase in the rms delay spread of the channel increased the implicit diversity gain till an "optimum" value of delay spread is reached after which the delay spread increases the ISI which the equalizer cannot counteract due to the limited number of taps. This "optimum" value increased with increase in the diversity order as the explicit diversity tends to decrease the ISI, as was seen in the previous sections. The troposcatter channel could support data rates upto 12.6 Mbps (to keep the normalized delay spread less than T) with a BER of 10^{-6} at 21 dB.

3.6 Other filter structures

From the preceding discussion, it is clear that joint antenna diversity combining and equalization are powerful techniques to reduce the average BER of communication systems over fading multipath channels. They are especially advantageous because they do not use any excess bandwidth which is considered a scarce and precious commodity.

The structure of optimal filter structures has been well established in the literature. However, it is not clear if there exist sub-optimal structures which can achieve similar performance at reduced computational complexity. Given that the trend in the future is towards very high capacity, high quality broadband systems at carrier frequencies upwards of 20 GHz, it seems likely that these systems would employ diversity combining and equalization. Space diversity of order up to four can be feasible at base stations as the coherence distance at these frequencies is very small. Moreover, equalization to avoid ISI due to excessive multipath delays could involve a significant number of forward and backward taps. Hence, it is imperative to study the complexity of the equalizer/combiner structures to evolve ways of reducing their complexity. Also, any of the above structures will need an initial training phase for the weights to converge to its optimal value, hence the effect of speed of convergence of the filter structures is an important issue in considering the most suitable filter structure for implementation. Another point to be considered is the performance of the filter structure for varying Doppler frequencies, though this may not be very important for indoor communication systems operating at 1-2 GHz. For future indoor systems operating at 20 - 60 GHz, this too may become a significant issue for consideration.

Some sub-optimal receivers were studied by Liu et al [52], for the case of the outdoor two path equal power channel. The filter structures suggested in the paper had BERs considerably poorer than the optimal combiner/equalizer filter structure. Moreover, the schemes were compared only on the basis of the effects on their performance due to variable Doppler frequency and delay spread for a fixed format size of 14 header bits and 148 bits. Some of the schemes studied in that paper were (i) DFE with slot-by-slot diversity selection, where a single branch is chosen at the end of the training slot (ii) DFE with multiple switching/training, where the receiver is allowed to switch to another diversity branch depending on the estimate of the received power (iii) DFE with symbol-by-symbol selection; this is selection diversity with selection after each symbol period (iv) DFE with diversity combining/selection where the diversity signals are added and (v) DFE with joint parameter estimation, which is the joint equalization/combining discussed in the previous section. The joint equalizer/combiner i.e. the optimal filter structure was shown to be far superior to the other filter structures considered for the above cases. This structure, however, was shown to have the highest complexity among all the structures. The paper was however limited in scope as they did not study the performance of the filter structure over

a slow fading channel or the effects of training length.

3.7 Summary

In this chapter, the application of diversity combining and equalization to the multipath fading channel was examined. It was seen that apart from mitigating the effects of fading, diversity combining was effective in combating intersymbol interference. Conversely, it was seen that equalization not only combats ISI, but also helps in exploiting the implicit diversity gain present in the multipaths of a frequency selective fading channel. These techniques can be used together in a system. Optimal filter structures differing in implementation details, though achieving the same performance, were described. The BER performances of systems employing diversity combining or equalization or using both were also described.

Equalization and diversity combining techniques are seen to be promising for achieving reliable communications over the fading multipath channels. These techniques can be used with various types of channel access methods (e. g. TDMA, CDMA, etc.) and have the added advantage of not causing bandwidth expansion. Sub-optimal structures for equalization and diversity combining have not been examined in detail in the available literature. However, these can be very useful if they offer advantages in terms of reduction in computational complexity, tracking properties and length of equalizer span required to achieve low BERs. The remainder of the thesis investigates the applicability of different filter structures for equalization and diversity combining and considers the advantages of using these structures for different channel/system conditions.

Chapter 4

FILTER STRUCTURES FOR EQUALIZATION/COMBINING

4.1 Introduction

In this chapter, we present several filter structures that can be used for equalization and diversity combining. Of a total of eight structures, three have already been proposed in the literature. The others are modifications of the above which do not compromise on performance but reduce complexity. For some conditions of the channel, they offer performance advantages also. In the next chapter, these filters are compared for different values of signal to noise ratio, doppler frequency, training sequence length, channel delay spread and cochannel interference for the cases when RLS is used for adaptation.

The theory of joint equalization and diversity combining is discussed in the two part paper [27]. The filter structure which achieves the minimum mean square error at the input to the decision device is derived in that paper (it was also derived previously in [44], [45]) and is shown in Fig. 4.1 (henceforth referred to as FIL1). The optimal diversity combiner/equalizer structure is composed of a set of matched filters whose outputs are first summed and then sampled. The resulting samples are then passed through an infinite tapped delay line with tap coefficients given by the weights w_k .

A different implementation of the optimal filter structure is the one used by Mosen [47]. The filter structure used in that paper is shown in Fig. 4.7 and described later in section 4.2.3. Though the two structures are equivalent in terms of BER performance, they can differ in their performances due to their convergence characteristics and due to their deviations from the optimal solution due to the non-validity of assumptions of ideality, viz. finite equalizer length, finite training size, etc. In this chapter, the modifications that can be made to these conventional filter structures are studied. The advantages of using one of them over the other will be studied in the next chapter.

Fig. 4.1 Optimum linear combiner/equalizer (FIL1)

4.2 Classes of equalizer/combiner filter structures

The problem of equalization and diversity combining suggests three classes of equalization and diversity combining architectures, namely (a) Combiner before equalizer (b) Equalizer before combiner and (c) Joint equalization and diversity combining. These structures are explained in detail below:

4.2.1 Diversity combiner before equalizer

There are three filter structures that was studied under this case of diversity combining before equalization viz. FIL1, FIL2 and FIL3. The structure of the forward and backward filters used for the equalizer is shown in Fig. 4.2.

a) FIL1: This is only slightly different from the filter structure suggested in [27] which is shown in Fig. 4.1. It was analytically shown to be optimum (in the mean square error sense) in that paper. However, for the purpose of deriving the bounds for probabilities of errors it was assumed that the channel is known perfectly and the tap coefficients of the infinite equalizer are calculated accurately from this perfect estimate. FIL1 differs from this optimal structure slightly: though the matched filters are assumed to be estimated perfectly, the equalizer coefficients (which are finite in number unlike numerical analysis for an infinite length equalizer in [27]) are calculated on a symbol by symbol based adaptation using an algorithm such as the RLS or LMS. Thus FIL1 is an ideal combiner but its equalization is imperfect and is affected by the "noise" inherent in the adaptation algorithm and incomplete equalization caused by the finite number of equalizer taps. These errors were assumed to be negligible and the results derived still served as a benchmark for an optimal combiner/equalizer.

b) FIL2: This is a practicably realizable version of FIL1. It differs from FIL1 in that it uses an estimation filter rather than assuming that the tap weights have been accurately estimated. Data signals at each antenna branch are convolved with a filter matched to the transmitted pulse shape. The outputs of the matched filter are sampled at the symbol rate or any integer multiple thereof depending on the channel. The sampled data during the training stage serve as the desired signal for the estimation filter in order to obtain an estimate of the channel impulse response, $f(t)$. The sampled data are then passed through the complex conjugated time-inverted filter, $f^*(-t)$, obtained from the estimated response. The

i^{th} Forward Filter ($i = 1, \dots, M$)

Fig. 4.2 Forward and backward filters used in the equalizer/combiner filter structures

(T_s is the fractional spacing, $T/2$, and T is the symbol duration)

outputs of this filter is then equalized using a symbol spaced or fractionally spaced equalizer.

For some channels, it may be necessary to use a fractionally spaced equalizer as a perfect knowledge of the channel is not available. To this end, it is necessary to estimate the fractionally spaced channel in order to convolve the $T/2$ sampled received signal with the estimated channel. To estimate the $T/2$ spaced channel, the receiver uses a copy of the training sequence with a zero interleaved between each training symbol. For other fractional spacings, one may insert more than one zero between the symbols of the copy of the training sequence. The desired signal, towards which the filter adapts itself, is the received signal sampled at the rate corresponding to the fractional spacing employed. Adaptation is required to be performed at twice (for $T/2$ fractional spacing) the symbol rate, lest the fractionally spaced taps have only zeroes for inputs during the training phase. The filter is shown in Fig. 4.3.

c) FIL3: This structure has a set of combining filters at each of the antenna inputs. Each of the filters operate directly on the sampled received signal (the sampling corresponds to the tap spacing of the digital receiver filter). Their outputs are fed to the digital filter matched to the transmit pulse. The combining filters combine the incoming signals such that their output is the appropriate modulated sample of the training sequence as shown in Fig. 4.4. The number of taps is chosen to be the same as the number of taps in the channel (in practice, one can set it to have a time span larger than the rms delay of the channel).

This filter structure could not converge to the optimum weights because the samples of the received signal are correlated. The chief reason for trying it out was that in the presence of cochannel interference this could combine the signals more effectively than an ideal matched filter, similar to Winter's optimum combining algorithm [34], [35] performing better than a maximal ratio combiner. Such a system can be more effective in a spread spectrum system, similar to the one described in [34], with tap spacing equal to the chip duration of the spread spectrum signal.

Clearly, the above filters may be inefficient when operating over a fast fading channel in a data directed mode of adaptation due to the inherent delay involved in the pulse shaping filters both before and after decision. Hence, these filters are most useful only in a slow fading channel when feedbacks are either not required or when they feed-backs after delays of about ten symbol durations (about five for each of the digital pulse

Fig. 4.3 Estimation of channel impulse response $h(t)$ for FIL2 (shown above). τ_s is the sampling rate at the output of the matched filter.

Fig. 4.4 FIL3 (above) and training of FIL3(below)

shaping filters) are meaningful.

4.2.2 Equalizer before diversity combiner

Under this case, we have studied four filter structures, two of which have been partially studied in the literature before. The problem here is to use a suitable combining algorithm after the equalization. Two of the four studied in this section viz. selection diversity and equal gain combining of the equalized channel outputs was studied for use in the IS-54 format in [52].

a) FIL4: In this filter structure, the individual branches are equalized separately and the branches are then combined according to some inverse function of the average square error that results at the outputs of the equalizers. The filter thus requires that the decision be made separately during training. After the training stage the decision is made on the combined output. However, the errors which train the equalizer weights in the decision directed mode of adaptation continues to be the difference between the decision of the decision device and the equalized output of each branch. Another feature of this filter is the use of two decision feedback branches as shown in Fig. 4.5. This is required as these feedback filters are trained separately in the training mode.

The function that was used for the simulations is given by:

$$f(mse_k) = \frac{1}{(mse_k)^2} \quad (4.1)$$

where mse_k is the mean square error at the output of the equalizer at the k th branch after the training sequence. The estimate of the mean square error that is used is the average of the square of the error at the end of the training sequence. This error estimates at the antenna branches and the corresponding weighting functions are updated at the end of each symbol duration in a decision directed mode of adaptation to reflect the varying nature of the channel.

b) FIL5: This filter structure (shown in Fig. 4.6) differs slightly from the above filter structure of FIL4. Here, after training the equalizer weights at each antenna branch separately, the equalizer outputs are passed through a combining filter which combines the output so as to minimize the mean square error at the input to decision device. Thus, it acts as an

Fig. 4.5 FIL4 in decision directed mode of adaptation

Fig. 4.6 FIL5

optimal combiner for the equalized inputs. In one sense, it may even be thought of a maximal ratio combiner of the equalized output, assuming that the total noise due to ISI and thermal noise is white. This filter structure may have the disadvantage of receiving incompletely equalized training sequences and hence its combining may not be optimal. We also note that we do not need more than a span of $0.5T$ (or whichever fractional spacing the equalizer employs) for the combiner to get the best results. In fact, increasing the number of taps can reintroduce ISI if the other weights do not converge to its optimal value of zero. A single weight (i.e. a memoryless filter) gives a better performance when the training sequence is small. For the simulations that are reported, only a single tap was used for the combiner.

c) FIL6: This is just a case of selection diversity at the branch outputs. The selection criterion is based on a measure of the output power used in [52], viz.

$$P_k(n) = (1 - \lambda) |r(n)|^2 + \lambda P_k(n - 1) \quad (4.2)$$

where $P_k(n)$ is the power at branch k at time instant n and $r(n)$ is the amplitude at the sampling instant.

d) FIL7: This is equal gain combining of the branch outputs. The outputs of the equalizer outputs at the two branches are added together and decision is made on this sum. The last two filter structures mentioned above were studied in [52] assuming the use of an IS-54 standard and high values of Doppler frequency. These have been studied in a greater generality in this thesis.

4.2.3 Joint equalization and diversity combining

This case is the filter structure proposed by Mosen [47] and shown in Fig. 4.7. The filter (*FIL8*) spans a spacing long enough for the equalizer to overcome the effects of intersymbol interference. Thus, with M branch antenna diversity and L taps in the equalizer, the filter can give up to $M*L$ orders of diversity. The filter combines the samples at $M*L$ slots of the time-space span to yield the minimum mean square error. Given that the channel does not yield more diversity than that is present in these slots, this is also clearly an optimal equalizer/combiner filter structure. This filter thus achieves joint estimation of

Fig. 4.7 Joint equalization and diversity combining (FIL8)

the equalizer and combiner parameters. The structural difference from FIL1 only suggests a different adaptation procedure.

4.3 System description

Simulations were carried out for the two channel models discussed in Chapter 2. The physical channel $h(t)$ was modeled as one with an exponentially decaying power profile with an rms delay spread of either $0.5T$ or $0.25T$. For either case, we assume two paths per rms delay spread. A wideband indoor TDMA system with a bandwidth of approximately 10 - 20 MHz operating at a frequency of 1-2 GHz can fit the delay spread description under study. The symbol rate is assumed to be around 10-20 Mbaud, hence even a nominal delay spread of up to 150 - 200 ns will have ISI extending about 4 data symbols, necessitating the use of equalization.

For this channel, the maximum doppler frequency can be expected not to exceed 20 Hz. Given the high symbol rates, the effect of this small doppler frequency becomes negligible. Thus, this channel can be expected to be static for the time duration of a frame length. (A frame length of 1000 symbols at a bit rate of 10 Mbaud gives a total time slot of 10 ms whereas the coherence time of the channel is ~ 50 ms, which is the reciprocal of the Doppler frequency of ~ 20 Hz). The effect of fast fading on the filter structures is nevertheless considered for outdoor radio; the two tap equal power channel model with spikes at $t = 0$ and at $t = T$ is used and the data rate is 24,000 symbols/s.

The entire system was simulated using the low pass equivalents [7] of the system parameters. For the slow fading case with Doppler frequency of 0 Hz, two independent Gaussian random numbers are generated for each of the real and imaginary components of the channel taps with a variance equal to the power of the respective path. For the fast fading channel, we use the "sum of sines" channel model [32], [53], where the tap coefficient is modelled as a sum of nine sinusoids with distinct Doppler frequencies. The maximum Doppler frequency can take on values from 0 Hz to 100 Hz, to reflect the normal vehicular speeds when the frequency is about 1 GHz. Additive White Gaussian Noise (AWGN) was simulated by generating independent gaussian random numbers for the real and imaginary components.

The modulation used was either coherent BPSK or QPSK. The transmitted signal can then be represented as

$$s(t) = \sum_n x_n g(t - nT) \quad (4.3)$$

where $x_n = a_n + jb_n$ is the n th transmitted symbol; $g(t)$ is the impulse response of the transmit filter and T is the symbol period. We assume that the transmitted data symbols have the following correlation relationship:

$$E\{x_p x_q^*\} = \begin{cases} \sigma_x^2, & p = q \\ 0, & p \neq q \end{cases} \quad (4.4)$$

where $E[\cdot]$ denotes expectation; and $*$ denotes the complex conjugate and σ_x^2 is 1 for BPSK and 2 for QPSK modulation.

The pulse used for transmission was the raised cosine pulse, which was evenly divided between the transmitter and receiver. The frequency responses of the transmit filter $g(t)$ is given by

$$\mathcal{F}[g(t)] = \sqrt{X(f)} \quad (4.5)$$

where $\mathcal{F}[\cdot]$ denotes the Fourier transform and $X(f)$ is a raised cosine frequency response with a roll-off factor β between 0 and 1 [7].

The time domain response of the square root raised cosine pulse is derived for T normalized to unity as

$$g(t) = \frac{(1-\beta) \sin[\pi t(1-\beta)]}{\pi t(1-\beta)} - \left(\frac{4\beta}{\pi}\right) \left\{ \frac{4\beta t \sin[\pi t(1-\beta)] + \cos[\pi t(1+\beta)]}{16\beta^2 t^2 - 1} \right\} \quad \beta \neq 1, t \neq 0, |t| \neq \frac{1}{4\beta} \quad (4.6)$$

where $0 \leq \beta \leq 1$ is the roll-off factor, which was chosen to be 0.35, i.e. a relatively tight roll-off. The first term of the above equation becomes $(1-\beta)$ for $t = 0$ and 0 for $\beta = 1$. The second term for $|t| = 1/4\beta$ is equal to

$$\left(\frac{-\beta}{\sqrt{2}\pi}\right) \left\{ (\pi + 2) \cos\left(\frac{\pi}{4\beta}\right) + (\pi - 2) \sin\left(\frac{\pi}{4\beta}\right) \right\}$$

The system described in this section will be used with the filter structures for performance evaluation in the next chapter.

4.4 Summary

In this chapter, the structure of some filters performing equalization and diversity combining was studied. The filter structures are classified into three classes depending on the order in which equalization and diversity combining are performed. Filter structures FIL1 and FIL8 have been analytically shown to be optimal filter structures in the literature. Filters FIL6 and FIL7, viz. selection diversity and equal gain combining of the equalized channel outputs have been studied for the case when one uses the IS-54 format with a high value of Doppler frequency over the two path model. These and other filter structures are studied in detail under different channel conditions in the next chapter. The channel variables include the channel model, Doppler frequency, the channel delay spread, the training sequence length, length of the equalizer and the presence of co-channel interference. Based on this simulation, a brief outline of the appropriate filter structure that can be used for the fading channel will be drawn out.

Chapter 5

SIMULATION RESULTS

5.1 Introduction

In the previous chapter we have discussed the structures of the filters that were used for performing equalization and diversity combining. In this chapter the performance of the different filters will be studied through computer simulation so as to determine the appropriate filter structure that can be used for a given channel condition. The performances of the different filter structures are compared for various channel-system configurations. A guideline to use the appropriate filter structure for a given condition can be drawn out from the discussion.

5.2 Simulation results

The simulations were carried out for different conditions. Though the study was essentially done for the indoor wireless channel which may be modeled by an exponentially decaying power profile. Some simulations were also carried out for the outdoor mobile IS-54 description of the worst case 'two equal power rays' channel model. The latter channel is a rather harsh channel with respect to the speed of convergence for an equalizer. The equalizer can however extract the diversity order of two that is inherent in this channel. The effect of doppler frequency on the different filter structures was also studied for this two ray channel model. The effect of the channel delay spread and the number of paths per symbol interval was studied for an equalizer operating in the indoor channel environment. The effect of the length of the training sequence and the number of equalizer taps was studied for the filter structures with the exponential channel model. Both linear and decision feedback equalizers are studied. Decision directed adaptation was used only for the case when the effect of fast fading was studied. For the DFE, ideal decision feedback is assumed.

The average bit error rate is evaluated by averaging over the bit error rates for each of 10000 - 15000 channel realizations. Bit error rates are reported for those cases when the number of bit errors exceed 100. As the study was entirely through simulation,

some of the results are affected by the non-idealities of the random number generators, due to the finite length of the equalizer and insufficient adaptation and consequently yield bit error rates higher than those which analysis [27] would have predicted, especially at low bit error rates. The results are still meaningful since the same data, for both channel and noise samples, are used in the comparison of the different filter structures.

Filter structures FIL3 and FIL6 were found to yield BERs higher than the other filter structures and hence the results obtained using these filters were not reported for most cases.

5.2.1 Effect of channel model

The filters were compared for two different channel models. The channel is assumed static during the transmission of the data bits and hence a long training sequence followed by a long sequence of data bits is assumed.

i) *Two ray model*: This is a model that has spectral nulls and hence the use of a decision feedback equalizer can give a significantly better bit error rate performance. The filter structures are trained with a long training sequence (50 symbols with RLS adaptation) using an equalizer with four symbol spaced taps for the forward part and four decision feedback taps. The linear equalizer also has four forward precursor and four postcursor symbol spaced taps. The BER performances of the filter structures are shown in Fig. 5.1.

ii) *Exponentially decaying power profile model*: For the exponential channel model with an rms delay spread of $0.5T$, the profile extends up to $2.5T$ to attain about 1% of the power at $t = 0$. Hence, an equalizer with five forward fractionally spaced taps and either three symbol spaced backward taps with decision feedback or five postcursor fractionally spaced taps for the linear equalizer was used. The bit error rate performance is shown in Fig. 5.2.

It can be seen that apart from FIL3 and FIL6, all the filter structures have almost identical bit error rates, irrespective of the channel model. Hence, the performance of these two filter structures which result in significantly higher bit error rates than with the others are not shown for most cases.

5.2.2 Effect of fast fading

To consider the use of the above filters for fast fading environments, it becomes

(a)

(b)

Fig. 5.1 Performance of filter structures for a two path channel
 a) Decision feedback equalization b) Linear Equalization

(a) DFE

(b) Linear equalizer

Fig. 5.2 Performance of filter structures with and without feedback for an exponential power profile

necessary to consider two issues: the length of the training sequence for a given fraction of information bits (say 10%) and the rate of fading itself. In this section the performance of the filter structures under varying Doppler frequencies is studied using the two tap model and is shown in Fig. 5.3. FIL2 and FIL3 cannot be used over a fast fading channel because adapting the estimation filter in the decision directed mode of adaptation based on decisions on the output of the equalizer would involve a significant increase in delay and complexity.

The other filter structures are however able to track the channel variations. The performance was studied for filters FIL5, FIL7 and FIL8. For short subframe (time allotted to a given user in a single frame) sizes resulting in short training lengths, the performance of FIL5 and FIL8 are nearly identical, with FIL8 performing only marginally better than FIL5. It can be seen that FIL5 results in a much lower BER than the equal gain combining structure, FIL7, especially at low Doppler frequencies. At higher Doppler frequencies all the filters result in large BERs and are almost identical. However, as the training length increases, the relative performance difference between the filters becomes apparent. As training sequence length approaches 50 symbols (for a total subframe size of 500), FIL8 yields significant performance improvements over FIL5 or FIL7 at low Doppler frequencies. This is because the larger subframe size necessitates accurate decision directed mode of adaptation, which is not efficiently performed by FIL7 and FIL5.

5.2.3 Effect of ray arrival rate and channel delay spread

The effect of channel delay spread for the two ray model has already been analyzed for the two ray model in [27]. In this section, this effect is simulated for the exponential decay profile. The simulation was performed for the case of a single antenna. For each rms delay spread, the number of paths was varied from a few paths occurring uniformly in many rms delay spread intervals to many paths occurring uniformly in a single rms delay spread interval. For each case, the time span of the taps in the equalizer was chosen to be larger than the channel delay spread. This was found to be sufficient to eliminate most of the ISI present in the received sequence.

As can be seen from Fig. 5.4, the variation in the bit error rate for a single branch equalizer is quite significant as the rms delay varies. In each case the delay spread of the channel extended up to a value for which the effective channel power reaches 1% of its value at $t = 0$, which occurs at $t = 5\tau_{rms}$. As the rms delay increases from zero to a high

Fig. 5.3 Effect of doppler frequency on the different equalizer/combiner structures (Two tap channel model)

Fig. 5.4 Effect of delay spread on a decision feedback equalizer.

value, two opposing effects influence the performance of the equalizer. One is the positive effect of increased effective diversity and the other is the negative effect of intersymbol interference. At a low value of rms delay spread the effective diversity order is the more important phenomenon and the bit error rate decreases. This continues till the rms delay spread reaches a value of $\tau_{rms} = T$ (with an effective channel length of about 4.5 symbol durations) after which the increase in performance of the equalizer becomes very marginal. This phenomenon where the performance reaches a limit was also observed in [27], where the bit error rate does not decrease beyond the value reached at $\tau_{rms} \sim 0.3T$. This phenomenon is not the same as predicted through analysis [54] that the bit error rate can ideally decrease continuously with increase in the channel delay spread till it reaches the performance of an Additive White Gaussian Noise (AWGN) channel. The phenomenon in [27] is due to the few number of paths (two) which are ideally resolved even at small values of delay spread and hence there is no performance enhancement by increasing the delay spread. However, in the present simulation the performance saturation is due to the poor resolution of the equalizer which is not preceded by an ideal matched filter as in the previous cases.

The figure also shows the effect on the performance of the system with variation in the number of paths that are present in the channel delay profile. It is seen that for most delay spreads, the performance of the system is best when there is about one fading path per symbol duration. Increase in the number of paths does not increase the implicit diversity, rather the $T/2$ spaced FSE is no longer able to synthesize the optimal filter. An exception is when the rms delay spread $\tau_{rms} = 0.25$, when the entire channel spread extends only upto $1.125T$. In this case, a single path per symbol interval gives performance close to flat fading, and more paths tend to give a small increase in performance.

The performance of a system using an equalizer with three T spaced forward taps and two backward taps for a channel with an rms delay spread of $2.0T$ was reported in [28]. For this setup, it was shown that the bit error rate is independent of the arrival rate provided that the average number of paths arriving per rms delay spread interval is greater than two. The channel was modeled as the "cluster" channel model and the path arrivals were Poisson distributed.

5.2.4 Effect of training sequence length for a slow fading channel

For the data rates considered in the thesis, the indoor wireless channel exhibits

Fig. 5.5 Effect of training length on the filter structures

slow fading characteristics. The operational frequency is expected to be about 1-2 GHz and the bandwidth about 10 MHz. Assuming that the channel changes completely at a distance separation of $\lambda/2$, we can expect a completely new channel once every 15 - 30 cm. With the mobile moving at a rate of as high as 2 m/s, we can expect the channel to change every 75 - 150 ms. At a bit rate of 5 - 10 Msymbols/s, we can assume a constant channel for nearly 375,000 - 1,500,000 symbols. To make the maximum use of bandwidth, it is preferable to send a very large number of bits in a single TDMA frame. Thus, one could use a long training sequence and use the LMS algorithm to minimize complexity. The length of the frame, however, would have to take into consideration the nature of the application viz. voice, data or video and a host of other technical issues and hence it may be necessary to restrict the length of the frame for real time applications and thence the size of the header. Moreover, the goal of PCS is to integrate the indoor wireless systems with the outdoor which require short frame sizes due to the presence of the effect of Doppler frequency. Our aim in this section is to see what the minimum header sequence size might be to achieve satisfactory performance when the RLS algorithm is used for adaptation. The exponential delay profile channel with an rms delay spread of twice the symbol duration was used for the simulation.

Varying the training sequence length with an equalizer span on either side of the central tap equal to the channel delay spread of ten symbol durations drastically affected the performance of the conventional filter structure viz. FIL8 as shown in Fig. 5.5. Performance of the new filter structures is considerably better, since the adaptation is on a fewer number of taps and the rate of convergence for the RLS algorithm is proportional to the number of taps. Thus, if it becomes necessary to restrict the training sequence length to a size less than 50 symbols, the new filter structures can be used to lower the bit error rates of the communication system. The effect would be magnified if a higher order of diversity is used as the adaptation algorithm would require to adapt a larger number of taps for FIL8 while a higher order of diversity does not effect the speed of convergence of the other filter structures.

5.2.5 Effect of equalizer span

The length of the equalizer is also an important issue for the portable. While it is feasible for the base station to have equalizers with a long span, the length of the equalizer at the portable is constrained by the complexity and power dissipation due to excess taps.

The effect of this phenomenon is studied in this section, where the performance is compared for different number of equalizer taps. From Fig. 5.6 it can be seen that joint equalizer/combiner has a better BER performance than the other filter structures. This is because insufficient equalization of the individual branches can significantly affect the combining strategy for those filters which perform equalization before diversity combining, whereas for those performing diversity combining before equalization, fewer number of taps for the single equalizer common to all the antenna branches, results in high ISI at the input to the decision device, again resulting in poor performance. The joint equalizer/combiner, being a minimum mean square error equalizer/combiner adapts to the solution which weights each antenna tap optimally and hence gives the best performance.

5.2.6 Effect of co-channel interference

Co-channel interference due to the presence of more than a single user operating at a given frequency at neighboring cells can significantly affect the performance of a TDMA system. This deterioration in the performance of the system using one of the above filter structures for reception is studied in this section. The co-channel interferers are simulated by lower power transmitters but operating over a channel statistically identical with the desired signal. Fig. 5.7 shows the relative performance of the different filter structures.

It can be seen that FIL8 is the only filter structure that can exploit the correlation of the interfering signal at the antenna branches. Thus, only FIL8 is able to suppress the interference and thus performs nearly 3 dB better than the other structures. If however, the number of interfering signals is considerably greater than the number of antenna elements as is the case in (b), the filter structures have nearly identical performance. It may hence be essential to determine the statistics of the co-channel interference in order to determine the filter structure most appropriate for the system.

5.3 Comparisons between different structures

For the slow fading channel, each of the filters yield very similar BERs given that the training sequence length is sufficient to ensure convergence. This is true of most diversity combining schemes and it is generally concluded that the order of diversity is far more significant than the type of combining strategy in terms of mitigating the effects of signal fading [55]. When the channel conditions are such that the delay spread is too long for the equalizer or if the training sequence is too short, filter structures have been found to differ

Fig. 5.6 Effect of equalizer span on the performance of the equalizers.

(a)

(b)

Fig. 5.7 Effect of co-channel interference on the filter structures using dual diversity with (a) single interferer and with (b) 6 co-channel interferers (SNR= 30 dB)

quite significantly. These differences constitute the merits or demerits of the respective filter structures. They can be summarized as follows:

i) FIL3 was found to yield rather unsatisfactory performance, only a few simulations were carried for these filter structures. FIL6 which is selection diversity based on the measure described in the last section was also seen to perform somewhat worse than the other filter structures, and hence simulations for this structure was not performed for all cases. The remainder of the discussion is with respect to the other filter structures.

ii) The filter structures differ significantly with respect to the computational complexity. This is demonstrated in Table 1 where the use of N_1 (T/2 spaced) forward taps, N_2 backward taps, M branch diversity and the use of conventional RLS algorithm for adaptation is assumed. For FIL2, the estimation filter and the equalizer may be T-spaced and N_1' is equal to $N_1/2$ if the taps are symbol spaced or equal to N_1 if they are fractionally spaced. N_1' is used as the number of taps in the equalizer since performance of a fractionally spaced equalizer is almost comparable to the symbol spaced equalizer for this filter structure, especially at low signal to noise ratios. K is the number of taps in the estimation filter. Thus the new filter structures have considerably lower computational complexity, while offering performance gains for many cases.

Table 8: Computational complexity of the filter structures.

Filter	Number of multiplications	$N_1=10, K=5,$ $N_2=3, M=4$
FIL1	N. A. (Since channel is assumed known)	—
FIL2	$2.5(MK^2 + (N_1' + N_2')^2) + 4.5(MK + N_1' + N_2')$	536/1072
FIL4	$2.5M(N_1 + N_2)^2 + 4.5M(N_1 + N_2) + 2M$	1932
FIL5	$2.5M(N_1 + N_2)^2 + 4.5M(N_1 + N_2) + 7M$	1952
FIL6	$2.5M(N_1 + N_2)^2 + 4.5M(N_1 + N_2) + 5M$	1944
FIL7	$2.5M(N_1 + N_2)^2 + 4.5M(N_1 + N_2)$	1924
FIL8	$2.5(MN_1 + N_2)^2 + 4.5(MN_1 + N_2)$	4816

iii) The new filter structures FIL4 and FIL5 can be very easily used for *macroscopic* diver-

sity with the numerical processing distributed between the antennae. Only the baseband signal needs be sent to the central processor, which can make a decision using very little computation. On the contrary, FIL8 would require that the entire computation be done at the central station which for high order diversity, may be infeasible.

iv) Even with microscopic diversity using an RLS algorithm for adaptation, FIL4, FIL5 or FIL7 could be implemented with parallel processing which is not possible with FIL8. Thus, for high order diversity systems the processing speed may become a bottleneck for FIL8, which is not the case with FIL2, FIL4, FIL5, FIL6 or FIL7.

The above two advantages can be of importance in systems offering a high grade of service. A simple way of meeting the performance requirement is to increase the order of explicit diversity. Given the high operating frequency, diversity orders of four seem reasonable, for which the new filter structures can yield significant reduction in computational complexity.

v) Given that the frequency spectrum is always a precious resource, header sequences may be minimized in future systems. Under this case, schemes which perform equalization before diversity combining viz. FIL4, FIL5 and FIL7 perform significantly better than FIL8. As the order of diversity is increased, FIL8 will take greater time to converge whereas the convergence time for the other filter structures is independent of the order of diversity. This phenomenon is because the convergence time for a filter using the RLS algorithm is about twice the number of taps involved in the adaptation.

vi) FIL8 performs better when the equalizer span is less than the channel delay spread. Thus, this may be an appropriate scheme for portables, which have the requirements of low power dissipation which necessitates the use of short equalizers

vii) For low Doppler frequencies of the order of 10 Hz, FIL8 can perform somewhat better than the other filter structures when the training sequences are long due to a longer subframe size. This is because of the greater ability of FIL8 to perform decision directed adaptation compared to the other filter structures. Thus, when using short training sequence it is considerably better to use FIL5 because of its lower computational complexity and use FIL8 only when large subframe sizes are used. For high values of Doppler frequency, all the filters result in identically high average BERs. It may thus be appropri-

ate to rely on channel coding or other schemes which meet the performance requirement.

viii) FIL8 is the only filter structure that can exploit the correlation of the interfering signal in the antenna branches in order to suppress it effectively. But this can be true only if the number of interfering signals is less than the number of antenna branches. But in practice, the number of interferers can be considerably greater than the number of antenna branches. It may thus be of importance to determine the statistics of the cochannel interference with respect to the number of diversity branches that are to be used. FIL8 can thus offer considerable gains over the other filter structures, when in the presence of shadowing, the cochannel interference can attain large values with respect to the desired signal power.

5.4 Summary guideline for appropriate filter structure

In brief, for high order diversity systems, filter structures FIL2, FIL4, FIL5 or FIL7 can offer significant reduction in complexity with almost no loss in performance. Hence, it would be of considerable advantage to use one of the above filter structures for indoor systems, especially for the base stations which may use a high order of diversity, as most systems are to be operated at high frequencies. These structures would be of added importance if the indoor system uses macroscopic diversity in conjunction with microscopic diversity. Moreover, for high order diversity systems involving a large number of taps involved in the adaptation, it may take a considerable time for weights of FIL8 to converge. Thus, unless the interference statistics are such that they are equivalent to only a few interfering signals, filters FIL2, FIL4, FIL5 or FIL7 are highly promising options for performing equalization and diversity combining.

For outdoor base stations the doppler frequency is an additional problem to be considered. The filter structure to be used depends on the frame size that is used for transmission. For long sub-frame sizes, where the role of decision directed adaptation is crucial, a system using FIL8 can give performance advantages over those using the other filters. Most systems, such as the IS-54 and GSM, use short subframe sizes of about 150 - 160 bits and hence FIL5 should be preferred to FIL8 because of its low computational complexity. However, the major impediment for outdoor mobile radio is fast fading which means Doppler frequency of the order of 80 - 100 Hz has to be combated. Equalization and diversity combining are not appropriate techniques for improving performance of sys-

tems when the Doppler frequency is high. For high values of Doppler frequency, a scheme such as coding may be a more attractive option.

For the portables, reduction of computational complexity is of primary concern. It thus depends on the length of the equalizer required to achieve satisfactory error rate performance. For long delay spreads, it becomes advantageous for the equalizer to use the new filter structure FIL5 to reduce the complexity while achieving the best performance. This filter structure is also useful when the system uses a short training sequence. However, for short delay spreads, it may be best to use FIL8 as only a few number of taps are required. This does not increase the complexity significantly and can perform satisfactorily even when delay spreads extend to moderately larger values. Moreover, this filter structure can be robust to the presence of co-channel interference.

In this section, the use of the appropriate filter structure for the outdoor/indoor environments for the base station and the portable has been discussed. While at the base station, the newer filter structures such as FIL5 or FIL4 are recommended, while at the portable FIL8 is ideal. Fading, shadowing and co-channel interference can thus be effectively suppressed by the use of a large diversity (both microscopic and macroscopic) order of low-complexity filters at the base station, while using a filter structure of low diversity order and few equalizer taps that also suppresses cochannel interference efficiently at the portable. The next chapter summarizes the thesis and suggests some directions for future research in this area.

Chapter 6

CONCLUSIONS

6.1 Thesis summary and conclusions

This research finds motivation in the worldwide quest to achieve universal portable and personal communications. In order to realize this vision, a low power tetherless access system must be designed to complement current radio systems into a unified architecture. Two options under investigation are FD-TDMA and CDMA. This thesis examines the TDMA option, which seems attractive for its flexibility in offering a variety of services in an environment where data rates up to 20 Mbps [43] are predicted for future applications.

Towards this end, the thesis began with an investigation into the characterization of the channel so as to use a suitable model for the simulation work. Literature available for the indoor radio channel model was studied and the exponential delay profile model was found as one suitable for the modelling of the indoor radio channel. The major impairments of the indoor radio channel are shadowing, frequency selective fading and co-channel interference.

The use of equalization and diversity combining to overcome these effects was studied in chapter 3. The literature available on the use of equalization and diversity combining for the frequency selective fading channel was studied to gain a clear idea of the practical gains that can be available by using these techniques for the indoor radio channel. It was concluded that equalization and diversity combining can both be used for the third generation systems which require high throughputs and quality. The thesis problem was to study the performance of different equalizer/combiner filter structures for the high performance third generation systems of the near future. It is likely that these systems, especially the indoor systems operating at high carrier frequencies, will employ a high order of diversity and it is also assumed that these filter structures will use the RLS algorithm for fast convergence. This necessitates a search for low complexity filter structures which do not compromise on performance. It was also felt necessary to have a structure whose implementation need not be specific to the order of diversity, but

to allow a modular structure for each antenna branch which can be easily scaled up for higher diversity orders.

The structures of some existing as well as some new low-complexity schemes which are minor modifications of the conventional schemes were introduced in chapter 4. Filter structures were divided into three possible classes depending upon the order in which equalization and diversity combining are performed.

The average BER performance of systems using the above filter structures was studied through computer simulation in chapter 5. Some of the filter structures that were studied were developed previously in the literature, but their performance under various channel parameters have not been reported. The thesis work was used at determining the appropriate filter structure for a given channel condition. The previous chapter also compared the merits of the various filter structures and reported the conditions under which a filter structure performs best or better than the others. The results can be summarized as follows:

- Filter structures FIL2, FIL4, FIL5 and FIL7 result in performance close to the conventional filter structure FIL8 with a significant reduction in complexity when the RLS algorithm is used for adaptation.
- The relative performance of the filters does not depend on the channel model under consideration. The filter structures have nearly identical BERs over any channel, assuming sufficient convergence, slow fading and the absence cochannel interference.
- At high values of Doppler frequency, the systems have similar performance. For low values of Doppler frequency, FIL5 and FIL8 have similar performance for short subframe sizes whereas for longer subframe sizes, FIL8 results in significantly lower BER.
- When the number of taps in the equalizer is constrained, the system using FIL8 has a lower BER than the other filter structures
- Filters FIL2, FIL4, FIL5 and FIL7 have much shorter convergence times than FIL8 and the difference in convergence times increases with increase in diversity order.
- In the presence of a dominant interferer, FIL8 has the best interference suppression capabilities.
- Filters FIL2, FIL4, FIL5 and FIL7 have a modular structure which can make

implementation of different orders of diversity systems in VLSI feasible.

On the basis of the above results, it was felt that it can be advantageous to employ a long equalizer having a high order of diversity (both microscopic and macroscopic) for the base station using filter structure FIL5 whereas a portable having only a few taps with a low diversity order can use filter structure FIL8.

6.2 Future work

It has been established in the thesis that significant gains can be achieved by minor modifications of the filter structures for some channel conditions. It may be of advantage if these structures can be studied analytically to allow for system performance prediction.

FIL4 requires that the equalizer outputs be multiplied by a weighting function before the combiner section. In this thesis, the weighting function chosen was the square of the reciprocal of the mean square error. It may be advantageous to analytically derive the optimal weight function, if it exists, in terms of previously calculated parameters such as the average error.

It was seen that FIL8 performs best in the presence of a few number of interferers, however it is the most computationally complex filter structure. For a practical system, it will be beneficial to use a combination of the filter structures to achieve optimum performance at low complexity. For e.g., a system can use two of the antennae with the FIL8 structure and the rest including this FIL8 structure in an overall FIL4 or FIL5 configuration. This can have the advantage of suppressing any dominant interferer and achieving good performance with an overall low computational complexity.

It can be seen that some of the filters studied are modular in architecture, i.e., if it is needed to increase the order of diversity, it amounts to adding a single antenna to the remaining block whereas a filter such as FIL8 would need a completely new combiner/equalizer filter structure corresponding to the new order of diversity. It is of interest to try an implementation of some of the above filter structures exploiting the modularity of the filter structure, either on a processor or in VLSI to study the practical range of data rates over which these can be operated on. The parallelism with which the new filters can be operated and the resultant increase in throughput could then be studied objectively.

References

- [1] H. Shosteck, "Price and demand - The factors affecting growth of the cellular market", in *Proceedings of the fourth Nordic Seminar on digital mobile radio communications*, (Oslo, Norway), June 1990, pp. 1.5.1 - 1.5.7.
- [2] D. McFarlane, "European personal communications using microcells", in *Proceedings of the IEE Colloquium on Microcellular Mobile Radio*, London, U. K., February 1989.
- [3] D. J. Goodman, "Trends in cellular and cordless communications", *IEEE Comm. Mag.*, June 1991, pp. 31-40.
- [4] D. Cox, "Personal communications: a viewpoint", *IEEE Communications Magazine*, vol. 28, no. 11, Nov. 1990, pp. 10-20.
- [5] D. Akerberg, "Properties of a TDMA picocellular office communication system", in *Proceedings of the 1989 IEEE Vehicular Technology Conference*, (San Francisco, CA), May 1989, pp. 186-191.
- [6] Charles L. B. Despins, Coding and optimum baseband combining for wideband TDMA indoor wireless channel, Ph. D. dissertation, Dept. of Systems and Computer Engineering, Carleton University, Ottawa, ON, Canada, August 1991.
- [7] John G. Proakis, *Digital Communications*, McGraw Hill, Singapore 2nd edition, 1989.
- [8] H. Hashemi, "The indoor radio propagation channel", *Proceedings of the IEEE*, vol 81, No 7 July 1993, pp 943 - 967.
- [9] A. Saleh and R. Valenzuela, "A statistical model for indoor multipath propagation", *IEEE J. Select. Areas Commun.*, Vol. SAC-5, no. 2, Feb 1987, pp. 128-137.
- [10] R. Bultitude, "Measurement, characterization and modeling of indoor 800/900 MHz radio channel for digital communications", *IEEE Commun. Magazine*, vol 25, no 6, June 1987, pp 5 - 12.
- [11] M. Nakagami, "The m-distribution -- A general model of intensity distribution of rapid fading," in *Statistical Methods of Radio Wave Propagation*, Pergamon Press, 1960.
- [12] A. F. Toledo and A. M. D. Turkmani, "Propagation into and within buildings at 900, 1800 and 2300 MHz", in *Proc. IEEE Veh. Technol. Conf.*, VTC '92, Denver, Colo., May 1992, pp 633-636.
- [13] G. A. Kalivas, M. El-Tanany and S. A. Mahmoud, "Millimeter wave channel measurements for indoor wireless communications", in *Proc. IEEE Veh. Technol. Conf.*, VTC '92, Denver, Colo., May 1992, pp 609-612.

- [28] T. A. Sexton and K. Pahlavan, "Channel modeling and adaptive equalization of indoor radio channels", *IEEE J. Select. Areas Commun.*, Vol. 7, No.1, Jan 1989, pp 114-121
- [29] J. Winters and Y. Yeh, "On the performance of wide-band digital radio transmission within buildings using diversity", in *Proceedings of the 1985 IEEE Global Telecommunications Conference*, (New Orleans, Lo.), Nov 1985, pp. 991-996.
- [30] D. G. Provan, "Linear diversity combining techniques", *Proc. IRE.*, vol. 47, June 1959, pp 1075-1102.
- [31] B. Glance and L. J. Greenstein, "Frequency selective fading effects in digital mobile radio with diversity combining", *IEEE Trans. Commun.*, vol. COM 31, No. 9, Sept. 1983, pp. 1085-1094
- [32] W. Jakes, Jr., *Microwave Mobile Communications*, New York: John Wiley and Sons, 1974.
- [33] R. Valenzuela, "Performance of quadrature amplitude modulation for indoor radio communications", *IEEE Trans. Commun.*, vol. COM 35, no. 11, Nov. 1987, pp. 1236-1238.
- [34] Jack H. Winters, "Optimum combining in digital mobile radio with cochannel interference", *IEEE Trans. Veh. Technol.*, Vol. VT 33, No. 3, August 1984, pp 144-155.
- [35] S. A. Hanna, M. S. El Tanany, S.A. Mahmoud and S. R. Todd, "Applications of adaptive antenna combining to digital radio communications within buildings", 1992 IEEE International Conference on Selected Topics in Communications, pp 107-110.
- [36] D. D. Falconer *et al.*, "Advances in equalization and diversity for portable wireless systems", *Digital Signal Processing*, 3, 1993, pp 148-162.
- [37] F. Adachi, K. Ohno and M. Ikura, "Postdetection selection diversity reception with correlated, unequal average power rayleigh fading signals for $\pi/4$ -shift QDPSK mobile radio", *IEEE Trans. Veh. Technol.*, vol. 41, no. 2, May 1992, pp 199-210.
- [38] R. W. Lucky, "Automatic equalization for digital communications", *Bell Sys. Tech. Journal*, vol. 45, pp. 255 -286, April 1965.
- [39] M. E. Austin, "Decision feedback equalization for digital communication over dispersive channels", MIT Lincoln Laboratory, Lexington Mass., Tech. Report no. 437, Aug. 1967.
- [40] R. D. Gitlin and S. B. Weinstein, "Fractionally spaced equalization: an improved transversal filter", *Bell System Technical Journal*, vol. 60, no. 2, Feb. 1981, pp. 275 - 296.
- [41] J. Omura, "Optimal receiver design for convolutional codes and channels with

- [28] T. A. Sexton and K. Pahlavan, "Channel modeling and adaptive equalization of indoor radio channels", *IEEE J. Select. Areas Commun.*, Vol. 7, No.1, Jan 1989, pp 114-121
- [29] J. Winters and Y. Yeh, "On the performance of wide-band digital radio transmission within buildings using diversity", in *Proceedings of the 1985 IEEE Global Telecommunications Conference*, (New Orleans, Lo.), Nov 1985, pp. 991-996.
- [30] D. G. Proiman, "Linear diversity combining techniques", *Proc. IRE.*, vol. 47, June 1959, pp 1075-1102.
- [31] B. Glance and L. J. Greenstein, "Frequency selective fading effects in digital mobile radio with diversity combining", *IEEE Trans. Commun.*, vol. COM 31, No. 9, Sept. 1983, pp. 1085-1094
- [32] W. Jakes, Jr., *Microwave Mobile Communications*, New York: John Wiley and Sons, 1974.
- [33] R. Valenzuela, "Performance of quadrature amplitude modulation for indoor radio communications", *IEEE Trans. Commun.*, vol. COM 35, no. 11, Nov. 1987, pp. 1236-1238.
- [34] Jack H. Winters, "Optimum combining in digital mobile radio with cochannel interference", *IEEE Trans. Veh. Technol.*, Vol. VT 33, No. 3, August 1984, pp 144-155.
- [35] S. A. Hanna, M. S. El Tanany, S.A. Mahmoud and S. R. Todd, "Applications of adaptive antenna combining to digital radio communications within buildings", 1992 IEEE International Conference on Selected Topics in Communications, pp 107-110.
- [36] D. D. Falconer *et al.*, "Advances in equalization and diversity for portable wireless systems", *Digital Signal Processing*, 3, 1993, pp 148-162.
- [37] F. Adachi, K. Ohno and M. Ikura, "Postdetection selection diversity reception with correlated, unequal average power rayleigh fading signals for $\pi/4$ -shift QDPSK mobile radio", *IEEE Trans. Veh. Technol.*, vol. 41, no. 2, May 1992, pp 199-210.
- [38] R. W. Lucky, "Automatic equalization for digital communications", *Bell Sys. Tech. Journal*, vol. 45, pp. 255 -286, April 1965.
- [39] M. E. Austin, "Decision feedback equalization for digital communication over dispersive channels", MIT Lincoln Laboratory, Lexington Mass., Tech. Report no. 437, Aug. 1967.
- [40] R. D. Gitlin and S. B. Weinstein, "Fractionally spaced equalization: an improved transversal filter", *Bell System Technical Journal*, vol. 60, no. 2, Feb. 1981, pp. 275 - 296.
- [41] J. Omura, "Optimal receiver design for convolutional codes and channels with

- memory via control theoretical concepts", *Inform. Sci.*, vol. 3, 1971, pp. 243-266.
- [42] G. D. Forney Jr., "Maximum likelihood sequence estimation of digital sequences in the presence of intersymbol interference", *IEEE Trans. Inform. Theory*, vol. IT-18, May 1972, pp. 363 - 378.
 - [43] R. Steele, "The cellular environment of lightweight handheld portables", *IEEE Communications Magazine*, vol. 27, no. 7, July 1989, pp. 30-40.
 - [44] Peter Monsen, "Feedback equalization for fading dispersive channels", *IEEE Trans. Inform. Theory*, vol. IT-17, Jan. 1971, pp. 56-64.
 - [45] Peter Monsen, "Digital Transmission performance on fading dispersive diversity channels", *IEEE Trans. Commun.*, vol. COM-21, no. 1, Jan. 1973, pp. 33-39.
 - [46] Peter Monsen, "Adaptive equalization of the slow fading channel", *IEEE Trans. Commun.*, vol. COM-22, no.8, Aug. 1974, pp. 1064-1075.
 - [47] Peter Monsen, "Theoretical and measured performance of a DFE modem on a fading multipath channel", *IEEE Trans Commun.*, vol. COM-25, no. 10, October 1977, pp 1144-1153.
 - [48] W.C. Wong and L. J. Greenstein, "Multipath fading models and adaptive equalizers in microwave digital radio", *IEEE Trans. Commun.*, vol. COM-32, no. 8, Aug. 1984, pp. 928-934.
 - [49] R. A. Valenzuela, "Performance of adaptive equalization for indoor radio communications", *IEEE Trans. Commun.*, vol. 37, no. 3, March 1989, pp. 291-293.
 - [50] D. M. Brady, "An adaptive coherent diversity receiver for data transmission through dispersive media", *IEEE Internat. Commun. Conf. Rec.* (San Francisco, Calif), June 1970.
 - [51] T. Shah and K. S. Shanmugan, "Dual diversity combining and equalization for indoor radio channels", *Proc. of the 1992 Global Telecomm. Conf.*, pp 1819-1824
 - [52] Q. Liu et al, "Performance of dual diversity decision feedback equalizers", *Proceedings of the Vehicular Technology Conference, VTC-1993*, pp. 637-640.
 - [53] S. N. Crozier, "Short block data detection techniques employing channel estimation for fading, time-dispersive channels", Ph. D. dissertation, Dept. of Systems and Computer Engineering, Carleton University, Ottawa, ON, Canada, May 1990.
 - [54] M. V. Clark et al, "Matched filter performance bounds for diversity combining receivers in digital mobile radio", *IEEE Trans. Vehic. Technol.*, vol. 41, Nov. 1992, pp. 356-362.
 - [55] D. Cox, "Antenna diversity performance in mitigating the effects of portable radiotelephone orientation and multipath propagation", *IEEE Trans. Commun.*, vol. COM-31, no. 5, May 1983, pp. 620-628.

Appendix

Program Listing

/* bpsk_src.h - this is the header file for monsen.c */

```
#include <stdio.h>
#include <stdlib.h>
#include <math.h>
```

```
#define N 1          /* number of interferers */
#define M 2          /* number of antennae */
#define S 710        /* number of symbols for a given channel */
#define SACT (S-10)  /* BER computation is over these many symbols */
#define C 15000       /* number of channel */
#define Ns 9          /* number of frequencies for generating Doppler */
#define Fd 10         /* maximum Doppler frequency */
#define Tb (1.0/24000) /* Signalling rate */
#define BQ 2          /* 1 for bpsk, 2 for qpsk */
#define CHAN 1         /* 1 for EXP, 2 for 2 path */
#define DOPP 1         /* 1 for slow i.e. 0 Hz, 2 for doppler */

#define rms_delay 2.0 /* delay spread of channel profile */
#define chan_spacing (rms_delay/2.0) /* spacing between adjacent channel taps */
#define chan_taps ( (int) (5.0 * (rms_delay)/(chan_spacing)+0.5) ) /* number of
                                                                    channel taps */
#define margin 10     /* margin used to avoid border effects during simulation */
#define sampling 0.25 /* this is the sampling rate for the simulation (0.25T) */
#define beta 0.35     /* roll-off factor for src pulse */
#define pi 3.14159
#define chan_taps_eff ((int)((chan_spacing/sampling))*(chan_taps-1)+1)
/* number of channel taps including zeros at sampling interval
not containing any channel ray */
#define tx_taps ( 1 + (int) ( 80 * 0.125/sampling + 0.1) )
/* number of taps in rx or tx filter */

#define SNR 20
#define SIR 30

#define eq_spacing 1.0 /* equalizer spacing */

#define AD_LEN 50      /* this is the adaptation length */
#define TR_LEN 20      /* first 20 symbols are fixed */

#define FOR 5 /* ((int)(5*rms_delay/eq_spacing + 0.5)) */
/* number of forward precursor taps */
#define BACK 5 /* ((int)(FOR*eq_spacing + 0.5)) */
/* number of backward taps */
#define POST 0 /* FOR */
/* number of forward postcursor taps */
#define COMB 1
#define SIZE (FOR + POST + BACK) /* SIZE = COMB */
```



```

#define mu1 0.5
#define mu2 0.4
#define mu3 0.5
#define mu4 0.1
#define lambda 0.99

#define sm 0.6
typedef struct cx {
    double re,im;
} cx;

unsigned short n_seed1[2][3] = { { 10, 11, 10 }, { 11, 10, 11 } };
unsigned short n_seed2[2][3] = { { 10, 12, 45 }, { 33, 87, 55 } };

double train[TR_LEN] = { 0.0, 1.0, 0.0, 1.0, 0.0, 0.0, 1.0, 0.0, 1.0, 0.0, 0.0,
1.0, 1.0, 1.0, 1.0, 0.0, 1.0, 0.0, 0.0, 0.0 };

double FREQ[Ns] = { 0.983, 0.932, 0.850, 0.739, 0.603, 0.446, 0.274, 0.092,
1.000 };
/* frequencies used to generate Doppler */

double WEIGHT[2][Ns] = { { 0.448206, 0.342997, 0.185540, 0.0, -0.185540, -
0.342997, -0.448206, -0.485071, 0.242536 }, { 0.185540, 0.342997, 0.448206,
0.485071, 0.448206, 0.342997, 0.185540, 0.0, 0.242536 } };
/* weights of the oscillators */

```

```
/* MONSEN.C - This implements a communication system using the joint equal-
izer combiner at the receiver end */
```

```
/* Include header file which contains constants */
```

```
/* and functions defined in cx_routines.c */
```

```
#include "bpsk_srrc.h"
```

```
extern cx cxadd();
```

```
extern void cxmat_add();
```

```
extern cx cxneg();
```

```
extern void cxmat_neg();
```

```
extern cx cxconj();
```

```
extern void cxmat_conj();
```

```
extern double cxmod();
```

```
extern cx cxmult();
```

```
extern cx cxmat_conv();
```

```
extern double erand48();
```

```
extern double mat_conv();
```

```
extern cx cxmat_mult();
```

```
extern cx cd_div();
```

```
extern cx cd_mul();
```

```
extern void cxmat_col();
```

```
extern void cxmat_div();
```

```
extern void cxmat_col();
```

```
extern void cxrow_mat();
```

```
extern void cxcol_row();
```

```
extern void cxmatrix_div();
```

```
extern void cxmatrix_neg();
```

```
extern void cxrow_scal();
```

```
/* Functions defined in this program following main declared here */
```

```
void bpsk();
```

```
void qpsk();
```

```
cx dd_bpsk();
```

```
cx dd_qpsk();
```

```
void srrc();
```

```
void mf_srrc();
```

```
void time_rect();
```

```
cx exp_channel();
```

```
double gauss();
```

```
void doppler();
```

```
void changephase();
```

```
void rls_update();
```

```
void lms_update();
```

```
/* Variables with larger space requirement is put in global memory */
```

```
cx u_comp[N][S + margin]; /* data bits */
```

```
cx u_shaped[N][ (int) ((S+margin)/sampling) + chan_tap_eff];
```

```
/* sampled tx filter output */
```

```
cx ant_inp[M][ (int) ((S+margin)/sampling)]; /* input to antenna */
```

```
cx mf_out[M][ (int) ((S+margin)/sampling)]; /* input to equalizer */
```

```
cx out[S + margin + BACK]; /* output of decision device */
```

```

cx src_pulse[tx_taps]; /* tx and rx filter */
double PHASE1[M][Ns][chan_taps_eff];
double PHASE2[M][Ns][chan_taps_eff]; /* random phase for generating Doppler */
unsigned short o_seed[M][3], o_seed2[M][3];

FILE *eq_err, *res, *weight;

main()
{
    /* Variables defined on the stack */
    int i, j, k, c, l, i_temp, i2_temp, m;
    int offset[N];
    float s;
    double v, v2, ber;
    double power;
    double temp, count, err_mod[M], mse[M], mse_av[M][AD_LEN];
    cx c_temp, y[M], e[M], factor[M], cx_temp, sum, error;
    cx mu_tr, mu_df;
    cx Y[M*SIZE], W[M*SIZE], P[M*SIZE-(M-1)*BACK][M*SIZE-(M-1)*BACK];
    /* Y is signal, W is equalizer tap weights, P is the inverse of the
       correlation matrix */
    double exp_variance[chan_taps_eff]; /* channel tap powers */
    double stdev[chan_taps_eff]; /* standard deviation of channel taps */
    cx channel[N][M][chan_taps_eff]; /* channel for given iteration */
    cx chan_temp[chan_taps_eff];
    cx chan_out;

    unsigned short inp_seed[2][N][3];
    unsigned short c_seed1[N][M][chan_taps_eff][2][3];
    unsigned short c_seed2[N][M][chan_taps_eff][2][3];
    cx mf_time_rect[tx_taps];
    cx w_for[M][FOR]; /* precursor forward filter taps */
    cx w_post[M][POST + 1]; /* postcursor forward filter taps */
    cx w_back[BACK + 1]; /* decision feedback taps */
    cx post[M][POST + 1]; /* signal data in postcursor taps */

    double ber_final = 0.0;
    double noise_power = 1.0/(pow(10.0, SNR/10.0)*2.0);
    double int_power = 1.0/(pow(10.0, SIR/10.0));
    if ( N > 2 ) int_power = int_power/((double)(N-1));

    /* Check and open the data file, for parameters */
    if ( ( eq_err = fopen ( "eq_error.m", "wb" ) ) == NULL ) ||
        ( ( res = fopen("src_tr20_mu.07.m", "wb") ) == NULL ) ||
        ( ( weight = fopen("wt_ori.m", "wb") ) == NULL ) ) {
        printf("%s", "file opening error\n");
        exit(1);
    }

    /* The tx filter impulse response is calculated and stored in src_pulse[] */

```

```

c_temp.im = 0.0;
power = 0.0;
j = (tx_taps - 1)/2;
for ( i = 0; i < tx_taps; i++) {
    if ((i-j) != 0) {
        c_temp.re = (1-beta) * sin (pi*(i-j)*sampling*(1-beta));
        c_temp.re = c_temp.re / (pi*(i-j)*sampling*(1-beta));
    }
    else c_temp.re = 1 - beta;
    if (fabs((i-j)*sampling) > 1/(4*beta)) * 1e-10)
        srcc_pulse[i].re = (4*beta/pi)*
            (4*beta*(i-j)*sin(pi*(i-j)*(beta+1)*sampling)*sampling -
             cos(pi*(i-j)*(beta+1)*sampling)) /
            (16 * pow((double)beta,2.0) *
             pow((double)(i-j)*(double)sampling,2.0)-1.0);
    else srcc_pulse[i].re =
        (beta*sqrt(2.0)/(2.0*pi))*(cos(pi/(4*beta))*(pi+2) +
        sin(pi/(4*beta))*(pi-2));
    srcc_pulse[i].re += c_temp.re;
    power += pow (srcc_pulse[i].re , 2.0);
    srcc_pulse[i].im = 0.0;
}

power = sqrt(power);
for ( i = 0; i < tx_taps; i++){
    srcc_pulse[i].re = srcc_pulse[i].re/power;
}

mu_tr.re = mu1;
mu_df.re = mu3;
mu_tr.im = mu_df.im = 0.0;
temp = sqrt ( 1.0 / tx_taps);
for ( i = 0; i < tx_taps; i++){
    mf_time_rect[i].re = temp;
    mf_time_rect[i].im = 0.0;
}

/* the following 3 loops initialize seeds for the input data bits, for the *
oscillators used for generating Doppler frequency and for the channel taps */
for ( k = 0; k < 2; k++) {
    for ( i = 0; i < N; i++) {
        for ( j = 0; j < 3; j++)
            inp_seed[k][i][j] = (2*i+1)*(3+j)*(pow(2.0,(double)k));
    }
}

for ( k = 0; k < M; k++) {
    for ( j = 0; j < 3; j++) {
        o_seed[k][j] = (3 + 2*j)*(pow(1.0,(double)k));
        o_seed2[k][j] = (5 + j)*(pow(2.0,(double)k));
    }
}

k = chan_spacing/sampling;

```

```

k = (chan_taps-1)*k+1;
for ( i = 0; i < N; i++ ){
    for ( j = 0; j < M; j++ ){
        for ( l = 0; l < k; l++ ){
            for ( i_temp = 0; i_temp < 2; i_temp++ ){
                for ( i2_temp = 0; i2_temp < 3; i2_temp++ ){
                    c_seed1[i][j][l][i_temp][i2_temp] =
                        (i+1)*(j+1)*(l+1)*
                        (i_temp+1)*(i2_temp+1);
                    c_seed2[i][j][l][i_temp][i2_temp] =
                        (i+2)*(j+1)*(2*l+1)*
                        (i_temp+4)*(3*i2_temp+1);
                }
            }
        }
    }
}

```

/* calculate the power of the channel taps depending on the channel model */

```

temp = power = 0.0;
k = chan_spacing/sampling;
k = (chan_taps-1)*k+1;
for ( i = 0; i < k; i++ ){
    if ( fabs((temp/chan_spacing) - (int)(temp/chan_spacing)) >
        sampling/(chan_spacing*2) )
        exp_variance[i] = 0.0;
    else{
        if ( CHAN == 1 ) exp_variance[i] = exp(-temp/rms_delay);
        else exp_variance[i] = 1.0;
    }
    temp += sampling;
    power += exp_variance[i];
}

for ( i = 0; i < k; i++ ){
    exp_variance[i] = exp_variance[i]*0.5/power;
    stdev[i] = sqrt(exp_variance[i]);
}

```

/* the following loop which is the system simulation for a given channel is executed over C channel iterations to average the bit error rate */

```

for ( c = 0; c < C; c++ )
{
    for ( i = 0; i < M; i++ ){
        for ( j = 0; j < FOR; j++ ){
            w_for[i][j].re = 0.0;
            w_for[i][j].im = 0.0;
        }
        mae[i] = 0.0;
        for ( k = 0; k < POST; k++ )
            w_post[i][k].re = w_post[i][k].im = post[i][k].re = post[i][k].im =
0.0;
    }
}

```

```

    }

    for ( i = 0; i < BACK; i++ ){
        w_back[i].re = w_back[i].im = 0.0;
    }

    for ( i = 0; i < S; i++)
        out[i].re = out[i].im = 0.0;

    for ( i = 0; i < M*SIZE-(M-1)*BACK; i++ ){
        for ( j = 0; j < M*SIZE-(M-1)*BACK; j++){
            if ( i == j ) P[i][j].re = 1000.0;
            else P[i][j].re = 0.0;
            P[i][j].im = 0.0;
        }
    }

    /* Generate channel impulse response for slow fading channel */
    if ( DOPP == 1 ){
        k = chan_spacing/sampling;
        k = (chan_taps-1)*k+1;
        for ( i = 0; i < N; i++){
            for ( j = 0; j < M; j++){
                for ( l = 0; l < k; l++ )
                    channel[i][j][l] = exp_channel(exp_variance[i],
                    &c_seed1[i][j][l][0][0],
                    &c_seed2[i][j][l][0][0]);
            }
        }
    }
    else changephase();
    k = chan_spacing/sampling;
    k = (chan_taps-1)*k+1;
    offset[0] = k-1;
    for ( i = 1; i < N; i++){
        temp = erand48(o_seed);
        j = (int) (temp * ((double)(k - 1)) + 0.5);
        offset[i] = k-1;
    }

    for ( i = 0; i < N; i++){
        for ( j = 0; j < k; j++){
            u_shaped[i][j].re = 0.0;
            u_shaped[i][j].im = 0.0;
        }
    }

    /* Random data bits are generated here */
    for ( i = 0; i < N; i++){
        for ( j = 0; j < S; j++){
            v = erand48(&inp_seed[0][i][0]);
            v2 = erand48(&inp_seed[1][i][0]);

```

```

        if ( i == 0 ) power = 1.0;
        else power = int_power;
        if ( BQ == 1 ) bpsk( v, power, &u_comp[i][j]);
        else qpsk(v,v2,power,&u_comp[i][j]);
    }
}

/* the data bits are modulated here */
power = 1.0;
for ( j = 0; j < TR_LEN; j++){
    v = train[j];
    v2 = erand48(&inp_seed[1][0][0]);
    if ( BQ == 1 ) bpsk( v, power, &u_comp[0][j]);
    else qpsk(v,v2,power,&u_comp[0][j]);
}

/* the modulated signal is passed through a tx filter */
for ( i = 0; i < N; i++)
    srrc(&u_comp[i][0], &u_shaped[i][offset[i]]);

i_temp = (int) (S / sampling);

for ( j = 0; j < M; j++){
    for ( k = 0; k < i_temp; k++){
        ant_inp[j][k].re = 0.0;
        ant_inp[j][k].im = 0.0;
    }
}

i_temp = (int) (S / sampling);
l = chan_spacing/sampling;
l = (chan_taps-1)*l+1;

/* the signal is passed through the channel and noise is added */
if ( DOPP == 1){
    for ( j = 0; j < M; j++){
        for ( k = 0; k < i_temp; k++){
            for ( i = 0; i < N; i++){
                chan_out = cmat_conv( &u_shaped[i][k],
                                     &channel[i][j][0], 1 );
                ant_inp[j][k] = cxadd(chan_out,ant_inp[j][k]);
            }
            c_temp.re = gauss(&n_seed1[0][0],&n_seed2[0][0],
                             noise_power);
            c_temp.im = gauss(&n_seed1[1][0],&n_seed2[1][0],
                             noise_power);
            ant_inp[j][k] = cxadd(c_temp, ant_inp[j][k]);
        }
    }
}

/* for doppler, the channel is updated at every sampling instant */
else {

```

```

for ( j = 0; j < M; j++){
    for ( k = 0; k < i_temp; k++){
        for ( i = 0; i < N; i++){
            doppler (stdev, &channel[i][j][0], 1, k, j);
            chan_out = cxmat_conv(&u_shaped[i][k],
                                &channel[i][j][0], 1);
            ant_inp[j][k] = cxadd(chan_out, ant_inp[j][k]);
        }
        c_temp.re = gauss(&n_seed1[0][0], &n_seed2[0][0],
                           noise_power);
        c_temp.im = gauss(&n_seed1[1][0], &n_seed2[1][0],
                           noise_power);
        ant_inp[j][k] = cxadd(c_temp, ant_inp[j][k]);
    }
}

i_temp = (int) S / sampling;

/* this is the receiver filter, matched to the tx pulse */
for ( i = 0; i < M; i++){
    mf_arrrc(&ant_inp[i][0], &mf_out[i][0]);
}

/* COMBINER + EQUALIZER */
/* Training phase */
/* sum and cx_temp are first initialized with the backward filter output */
/* c_temp stores the sum of the pre and postcursor forward filter outputs */
/* sum adds c_temp of each branch forward filter to the backward filter */
/* e1 or contains the difference between desired and the present sum */

i = 0;
l = 0;
while ( i < AD_LEN ) {
    sum = cx_temp = cxmat_conv ( &out[i], w_back, BACK );
    for ( j = 0; j < M; j++){
        c_temp = cxmat_conv ( &mf_out[j]((int)(i/eq_spacing)),
                              &w_for[j][0], FOR );
        c_temp = cxadd(c_temp, cxmat_conv(&post[j][0], &w_post[j][0],
                                          POST) );
        y[j] = cxadd(cx_temp, c_temp);
        e[j] = cxadd ( u_comp[0][i], cxneg(y[j]));
        sum = cxadd ( sum, c_temp);
        err_mod[j] = pow ( cxmod(e[j]), 2.0);
        mse[j] = (err_mod[j] + i*mse[j])/(double)(i+1);
        mse_av[j][1] = (err_mod[j] + i*mse_av[j][1])/(double)(i+1);
    }

    temp = 0.5;
    for ( j = 0; j < M; j++){
        for ( k = 0; k < FOR; k++)

```



```

        temp += pow(cxmod(mf_out[j][(int)(i/eq_spacing) + FOR -
                                1 - k]), 2.0);
    for ( k = 0; k < POST; k++ )
        temp += pow(cxmod(post[j][POST - 1 - k]), 2.0);
    }
    for ( k = 0; k < BACK; k++ )
        temp += pow(cxmod(out[i + BACK - 1 - k]), 2.0);

    error = cxadd ( u_comp[0][i], cxneg(sum) );

    /* the following loops will form the signal matrix and the tap weight */
    /* matrix and call the update algorithm to update the tap weights */

    m = 0;
    for ( j = 0; j < M; j++ ){
        for ( k = 0; k < FOR; k++ ){
            Y[m] = mf_out[j][(int)(i/eq_spacing) + FOR - 1 - k];
            W[m] = w_for[j][k];
            m++;
        }
        for ( k = 0; k < POST; k++ ){
            Y[m] = post[j][POST - 1 - k];
            W[m] = w_post[j][k];
            m++;
        }
    }
    for ( k = 0; k < BACK; k++ ){
        Y[m] = out[i + BACK - 1 - k];
        W[m] = w_back[k];
        m++;
    }
    rls_update(W,P,Y,error,m);
    /* lms_update(W,Y,cd_div(error,temp),mul, m); */

    /* update over: so copy updated matrix data into filter variables */

    m = 0;
    for ( j = 0; j < M; j++ ){
        for ( k = 0; k < FOR; k++ ){
            w_for[j][k] = W[m];
            m++;
        }
        for ( k = 0; k < POST; k++ ){
            w_post[j][k] = W[m];
            m++;
        }
    }
    for ( k = 0; k < BACK; k++ ){
        w_back[k] = W[m];
        m++;
    }

    /* postcursor filter inputs have to be shifted by one position */

```

```

for ( j = 0; j < M; j++){
    for ( k = 0; k < POST - (int)(1/eq_spacing); k++){
        post[j][k] = post[j][k + (int) (1/eq_spacing)];
    }
    if ( POST > 1)
        post[j][POST - (int)(1/eq_spacing)] =
            mf_out[j][(int)(1/eq_spacing)];
    if ( POST > 0)
        post[j][POST - 1] = mf_out[j][(int)(1/eq_spacing) +
            (int)(1/eq_spacing)-1];
}

/* Backward filter has input data */

out[i + BACK] = u_comp[0][i];
i++;
}

/* Training phase over, decision directed mode of adaptation */

l = 0;
while ( l < SACT ) {
    sum.re = sum.im = 0.0;
    for ( j = 0; j < M; j++){
        c_temp = cxmat_conv ( &mf_out[j][(int)(1/eq_spacing)],
                               &w_for[j][0], FOR );
        c_temp = cxadd(c_temp, cxmat_conv(&post[j][0], &w_post[j][0],
                                           POST) );
        sum = cxadd(sum, c_temp);
    }
    c_temp = cxmat_conv ( &out[i], w_back, BACK );
    sum = cxadd ( c_temp, sum);

    if (BQ == 1 ) c_temp = dd_bpsk(sum);
    else c_temp = dd_qpsk(sum);
    error = cxadd ( c_temp, cxneg(sum) );
    out[i+BACK] = u_comp[0][i];
    temp = 0.5;
    for( j= 0; j < M; j++){
        for ( k = 0; k < FOR; k++ )
            temp += pow(cxmod(mf_out[j][(int)(1/eq_spacing) + FOR -
                1 - k]), 2.0);
        for ( k = 0; k < POST; k++ )
            temp += pow(cxmod(post[j][POST - 1 - k]), 2.0);
    }
    for ( k = 0; k < BACK; k++ )
        temp += pow(cxmod(out[i + BACK - 1 - k]), 2.0);

    m = 0;
    for ( j = 0; j < M; j++ ){
        for ( k = 0; k < FOR; k++ ){
            Y[m] = mf_out[j][(int)(1/eq_spacing) + FOR - 1 - k];
            W[m] = w_for[j][k];
            m++;
        }
    }
}

```

2 of/de 2

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

```

    }
    for ( k = 0; k < POST; k++ ){
        Y[m] = post[j][POST - 1 - k];
        W[m] = w_post[j][k];
        m++;
    }
}
for ( k = 0; k < BACK; k++ ){
    Y[m] = out[i + BACK - 1 - k];
    W[m] = w_back[k];
    m++;
}

/* rls_update(W,P,Y,error,m); */
/* lms_update(W,Y,cd_div(error,temp),mu3, m); */

m = 0;
for ( j = 0; j < M; j++ ){
    for ( k = 0; k < FOR; k++ ){
        w_for[j][k] = W[m];
        m++;
    }
    for ( k = 0; k < POST; k++ ){
        w_post[j][k] = W[m];
        m++;
    }
}
for ( k = 0; k < BACK; k++ ){
    w_back[k] = W[m];
    m++;
}

for ( j = 0; j < M; j++){
    for ( k = 0; k < POST - (int)(1/eq_spacing); k++)
        post[j][k] = post[j][ k + (int)(1/eq_spacing)];
    if ( POST > 1)
        post[j][POST - (int)(1/eq_spacing)] =
            mf_out[j][(int)(1/eq_spacing)];
    if ( POST > 0)
        post[j][POST - 1] = mf_out[j][(int)(1/eq_spacing) +
            (int)(1/eq_spacing) - 1];
}

if ( u_comp[0][i].re != c_temp.re ) l++;
if ( u_comp[0][i].im != c_temp.im ) l++;
i++;
}

ber = 1 / (double)(BQ*(i - AD_LEN));
ber_final = (c * ber_final + ber)/(c + 1.0);
printf("%d %e %e", c, ber, ber_final);
printf("\n");
}
}

```

```
/* main program over, functions will now be described */
```

```
/* BPSK modulator */
```

```
void bpsk(ran_num, power, u_comp)
```

```
double ran_num;
```

```
double power;
```

```
cx u_comp[];
```

```
{
```

```
if (ran_num > 0.5)
```

```
    (*u_comp).re = power;
```

```
else
```

```
    (*u_comp).re = -power;
```

```
(*u_comp).im = 0.0;
```

```
}
```

```
/* QPSK modulator */
```

```
void qpsk(ran_num1, ran_num2, power, u_comp)
```

```
double ran_num1, ran_num2;
```

```
double power;
```

```
cx u_comp[];
```

```
{
```

```
if (ran_num1 > 0.5)
```

```
    (*u_comp).re = power;
```

```
else
```

```
    (*u_comp).re = -power;
```

```
if (ran_num2 > 0.5)
```

```
    (*u_comp).im = power;
```

```
else
```

```
    (*u_comp).im = -power;
```

```
}
```

```
/* BPSK demodulator */
```

```
cx dd_bpsk(y)
```

```
cx y;
```

```
{
```

```
cx o;
```

```
if ( y.re > 0.0 ) o.re = 1.0;
```

```
else o.re = -1.0;
```

```
o.im = 0.0;
```

```
return o;
```

```
}
```

```
/* QPSK demodulator */
```

```
cx dd_qpsk(y)
```

```
cx y,
```

```
{
```

```
cx a,b,c,d,u,r;
```

```
double x;
```

```
.
```

```
a.re = b.re = 1.0;
```

```

a.im = d.im = 1.0;
b.im = c.im = -1.0;
c.re = d.re = -1.0;

```

```

u = cxmult(cxconj(a),y);
x = u.re;
r = a;
u = cxmult(cxconj(b),y);
if (u.re > x){
    x = u.re;
    r = b;
}
u = cxmult(cxconj(c),y);
if (u.re > x){
    x = u.re;
    r = c;
}
u = cxmult(cxconj(d),y);
if (u.re > x){
    x = u.re;
    r = d;
}
return r;
}

```

/* tx filter with time rectangular impulse response: never used in program */

```

void time_rect(u,u_sh)
cx u[];
cx u_sh[];
{
    int i;
    cx sign;
    cx r;

    sign.re = ( (*u).re > 0 ) ? 1 : -1;
    sign.im = ( (*u).im > 0 ) ? 1 : -1;
    r.re = fabs((*u).re/tx_taps);
    r.im = fabs((*u).im/tx_taps);

    for ( i = 0; i < tx_taps; i++ ){
        u_sh[i].re = sign.re * pow(r.re, 0.5);
        u_sh[i].im = sign.im * pow(r.im, 0.5);
    }
}

```

/* generate channel from variance matrix and 2 arrays for erand random num */

```

cx exp_channel(exp_variance, cs1,cs2)
double exp_variance;
unsigned short cs1[2][3], cs2[2][3];
{
    cx coeff;
    coeff.re = gauss(&cs1[0][0],&cs2[0][0],exp_variance);
    coeff.im = gauss(&cs1[1][0],&cs2[1][0],exp_variance);
}

```

```

return coeff;
}

/* this is the square root raised cosine shaping filter */
void srrc(u, u_shaped)
cx u[];
cx u_shaped[];
{
    cx temp[tx_taps];
    int i,j,k,l,m,n,next,sampling_rec;

    sampling_rec = (int)(1.0/sampling);

    for (i = 0; i < tx_taps; i++){
        temp[i].re = 0.0;
        temp[i].im = 0.0;
    }

    i = next = 0;
    m = tx_taps - 1;
    l = m/2;

    while (i < l){
        temp[m] = u[next];
        next++;
        i++;
        for (j = 0; j < sampling_rec - 1; j++){
            for (k = 0; k < m; k++)
                temp[k] = temp[k + 1];
            i++;
            temp[k].re = temp[k].im = 0.0;
        }
        for (k = 0; k < m; k++)
            temp[k] = temp[k + 1];
    }

    i = 0;
    l = sampling_rec * S - 1;

    while (i < l){
        temp[m] = u[next];
        next++;
        u_shaped[i] = cxmat_conv(temp, srrc_pulse, tx_taps);
        i++;
        for (j = 0; j < sampling_rec - 1; j++){
            for (k = 0; k < m; k++)
                temp[k] = temp[k + 1];
            temp[k].re = temp[k].im = 0.0;
            u_shaped[j+i] = cxmat_conv(temp, srrc_pulse, tx_taps);
            i++;
        }
        for (k = 0; k < m; k++)
            temp[k] = temp[k + 1];
    }
}

```

```

    }

    l = sampling * S;

    while (i < l){
        temp[m].re = temp[m].im = 0.0;
        u_shaped[i] = cxmat_conv(temp, srrc_pulse, tx_taps);
        i++;
        for ( j = 0; j < sampling_rec - 1; j++){
            for ( k = 0; k < m; k++ )
                temp[k] = temp[k + 1];
            temp[k].re = temp[k].im = 0.0;
            u_shaped[i] = cxmat_conv(temp, srrc_pulse, tx_taps);
            i++;
        }
        for ( k = 0; k < m; k++ )
            temp[k] = temp[k + 1];
    }
}

/* this is matched filter to square root raised cosine pulse */
void mf_srrc(u, u_sh)
cx u[];
cx u_sh[];
{
    cx temp[tx_taps];
    int i,j,k,l,m,n,next,sampling_rec;

    sampling_rec = (int)(1.0/sampling);

    for (i = 0; i < tx_taps; i++){
        temp[i].re = 0.0;
        temp[i].im = 0.0;
    }

    i = next = j = 0;
    m = tx_taps - 1;
    l = m/2;

    while (i < l){
        temp[m] = u[next];
        next++;
        for ( k = 0; k < m; k++ )
            temp[k] = temp[k + 1];
        i++;
    }
    i = 0;
    l = sampling_rec * S - 1;

    while (i < l){
        temp[m] = u[next];
        next++;
    }
}

```



```

        if (fabs((i*sampling)/eq_spacing - (int)((i*sampling)/eq_spacing)) <
            (sampling/(2*eq_spacing)) ){
            u_sh[j] = cxmat_mult(temp, srrc_pulse, tx_taps);
            j++;
        }
        for ( k = 0; k < m; k++ )
            temp[k] = temp[k + 1];
        i++;
    }

    l = sampling_rec * S;

    while (i < l){
        temp[m].re = temp[m].im = 0.0;
        next++;
        if (fabs((i*sampling)/eq_spacing - (int)(i*sampling/eq_spacing)) <
            (sampling/(2*eq_spacing)) ){
            u_sh[j] = cxmat_mult(temp, srrc_pulse, tx_taps);
            j++;
        }
        for ( k = 0; k < m; k++ )
            temp[k] = temp[k + 1];
        i++;
    }
}

/* From two random seeds generate a gaussian rv with var variance using erand */
double gauss(seed1, seed2, var)
unsigned short seed1[];
unsigned short seed2[];
double var;
{
    double r,s,t;

    r = erand48(seed1);
    s = erand48(seed2);
    if ( r <= 1.000000e-10) r = 1.000000e-10;
    if ( s <= 1.000000e-10) s = 1.000000e-10;
    t = sin(2 * 3.14159 * s) * sqrt (2 * var * log(1/r));
    return t;
}

/* this is the standard rls update algorithm, y is signal matrix, c is filter */
/* tap weights, k is the Kalman gain vector, P is inverse of autocorrelation */
/* matrix */
void rls_update(c,p,y,e,size)
cx c[];
cx *p;
cx y[];
cx e;
int size;
{

```

```

cx *temp, *k;
cx *tempo;
cx a;
double b;

temp = (cx *) malloc(sizeof(cx)*size);
k = (cx *) malloc(sizeof(cx) * size);
tempo = (cx *) malloc(sizeof(cx) * size * size * sizeof(cx));

cxmat_conj(y,k,size);

cxmat_col(p, k, temp, size);

a = cxmat_mult(y, temp, size);
b = a.re + lambda;
cdmat_div(temp, b, k, size);

cxrow_mat(y, p, temp,size);
cxcol_row(k, temp, tempo, size);
cxmatrix_neg(p, tempo, p, size);
cdmatrix_div(p, lambda, p, size);

cxrow_scal(k,e, temp, size);
cxmat_add(c,temp, c, size);
free((char *) temp);
free((char *) k);
free((char *) tempo);
}

/* the standard lms update algorithm */
void lms_update(c,y,e,mu,size)
cx c[];
cx y[];
cx e;
int size;
double mu;
{
cx *temp;
temp = (cx *) malloc(sizeof(cx) * size);
cxmat_conj(y,temp,size);
cxrow_scal(temp,e, temp, size);
cdmat_mult(temp,mu,temp, size);
cxmat_add(c, temp, c, size);
free((char *) temp);
}

/* this function is called every symbol interval to update the channel tap */
/* weight when doppler is present. It uses the sum of nine sinusoids to */
/* calculate the channel tap weight */
void doppler(a,b,c,d,e)
double a[];

```

```

cx b[];
int c,d,e;
{
int i,j,k;

for ( i = 0; i < c; i++)
    b[i].re = b[i].im = 0.0;

for ( j = 0; j < c; j++){
    for ( i = 0; i < Ns; i++){
        b[j].re += a[j] * 1.4142 * WEIGHT[0][i] *
            cos ( (2 * pi * FREQ[i] * Fd * d * sampling * Tb) +
                PHASE1[e][i][j]);
        b[j].im += a[j] * 1.4142 * WEIGHT[1][i] *
            cos ( (2 * pi * FREQ[i] * Fd * d * sampling * Tb) +
                PHASE2[e][i][j]);
    }
}

/* this initializes the phase of the oscillators at the beginning of each */
/* iteration */
void changephase()
{
int i,j,k,l;
k = (int)((chan_spacing/sampling))*(chan_taps-1)+1;

for ( l = 0; l < M; l++){
    for ( i = 0; i < Ns; i++){
        for ( j = 0; j < k; j++){
            PHASE1[l][i][j] = pi * erand48(&o_seed1[l][0]);
            PHASE2[l][i][j] = pi * erand48(&o_seed2[l][0]);
        }
    }
}
}

```

```
/* cx_routines.c - this program contains all the complex routines required */
/* in monsen.c */
```

```
#include <math.h>
typedef struct cx {
    double re,im;
} cx;
```

```
/* Add two complex numbers */
```

```
cx cxadd(a, b)
cx a;
cx b;
{
    cx c;
    c.re = a.re + b.re;
    c.im = a.im + b.im;
    return c;
}
```

```
/* Add two complex row matrices */
```

```
void cxmat_add(a, b, c, size)
cx a[];
cx b[];
cx c[];
int size;
{
    int i;
    for ( i = 0; i < size; i++ ){
        c[i].re = a[i].re + b[i].re;
        c[i].im = a[i].im + b[i].im;
    }
}
```

```
/* negate a complex number */
```

```
cx cxneg(a)
cx a;
{
    cx b;
    b.re = -a.re;
    b.im = -a.im;
    return b;
}
```

```
/* negate a complex row vector */
```

```
void cxmat_neg(a, b, size)
cx a[];
cx b[];
int size;
{
    int i;
    for ( i = 0; i < size; i++ ){
        b[i].re = - a[i].re;
        b[i].im = - a[i].im;
    }
}
```

```

}

/* find conjugate of a complex number */
cx cxconj(a)
cx a;
{
  cx b;
  b.re = a.re;
  b.im = -a.im;
  return b;
}

/* conjugate a matrix */
void cxmat_conj(a, b, size)
cx a[];
cx b[];
int size;
{
  int i;
  for ( i = 0; i < size; i++ ){
    b[i].re = a[i].re;
    b[i].im = -a[i].im;
  }
}

/* find modulus of complex number */
double cxmod(a)
cx a;
{
  double b;
  b = a.re * a.re + a.im * a.im;
  b = sqrt(b);
  return b;
}

/* multiply two complex numbers */
cx cxmult(a, b)
cx a;
cx b;
{
  cx c;
  c.re = a.re * b.re - a.im * b.im;
  c.im = a.re * b.im + a.im * b.re;
  return c;
}

/* find middle element of convolution of two real vectors */
double mat_conv(a, b, size)
double a[];
double b[];
int size;
{
  int i;
  double c;

```

```

c = 0.0;
for ( i = 0; i < size; i++)
    c += a[i] * b[size - i - 1];
return c;
)

/* find middle element of convolution of two complex vectors */
cx cxmat_conv(., b, size)
cx a[];
cx b[];
int size;
{
    int i;
    cx d;
    cx c;

    c.re = 0.0;
    c.im = 0.0;
    for ( i = 0; i < size; i++){
        c.re += a[i].re * b[size - i - 1].re - a[i].im * b[size - i - 1].im;
        c.im += a[i].re * b[size - i - 1].im + a[i].im * b[size - i - 1].re;
    }
    return c;
}

/* multiply two vectors */
cx cxmat_mult(a, b, size)
cx a[];
cx b[];
int size;
{
    int i;
    cx d;
    cx c;

    c.re = 0.0;
    c.im = 0.0;
    for ( i = 0; i < size; i++){
        c.re += a[i].re * b[i].re - a[i].im * b[i].im;
        c.im += a[i].re * b[i].im + a[i].im * b[i].re;
    }
    return c;
}

/* multiply vector by a real number */
cx cd_mult(a,b)
cx a;
float b;
{
    cx c;
    c.re = a.re * b;
    c.im = a.im * b;
    return c;
}

```

```

    }

    /* divide a vector by a real number */
    cx cd_div(a,b)
    cx a;
    float b;
    {
    cx c;
    c.re = a.re / b;
    c.im = a.im / b;
    return c;
    }

    /* multiply a square matrix by a real number */
    void cdmatrix_mult(a,b,c,size)
    cx a[],c[];
    double b;
    int size;
    {
    int i;
    for ( i = 0; i < size; i++ ) {
        c[i].re = a[i].re * b;
        c[i].im = a[i].im * b;
    }
    }

    /* divide a square matrix by a real number */
    void cdmatrix_div(a,b,c,size)
    cx a[],c[];
    double b;
    int size;
    {
    int i;
    for ( i = 0; i < size; i++ ) {
        c[i].re = a[i].re / b;
        c[i].im = a[i].im / b;
    }
    }

    /* multiply a column vector by a square matrix */
    void cxmatrix_col(a, b, c, size)
    cx *a;
    cx b[];
    cx c[];
    int size;
    {
    int i,j;
    for ( i = 0; i < size; i++ ){
        c[i].re = 0.0;
        c[i].im = 0.0;
    }
    }

```

```

for ( i = 0; i < size; i++ ){
    for ( j = 0; j < size; j++ ){
        c[i] = cxadd(c[i], cxmult(*(a+size*i+j),b[j]));
    }
}

/* multiply a row vector by a square matrix */
void cxrow_mat(a, b, c, size)
cx a[];
cx *b;
cx c[];
int size;
{
    int i,j;
    for ( i = 0; i < size; i++ ){
        c[i].re = 0.0;
        c[i].im = 0.0;
    }

    for ( i = 0; i < size; i++ ){
        for ( j = 0; j < size; j++ )
            c[i] = cxadd(c[i], cxmult(a[j],*(b+j*size+i)));
    }
}

/* multiply a row vector by a column vector */
void cxcol_row(a, b, c, size)
cx a[];
cx b[];
cx *c;
int size;
{
    int i,j;

    for ( i = 0; i < size; i++ ){
        for ( j = 0; j < size; j++ )
            *(c+size*i+j) = cxmult(a[i],b[j]);
    }
}

/* multiply a row vector by a scalar */
void cxrow_scal(a, b,c , size)
cx a[],c[];
cx b;
int size;
{
    int i;
    for ( i = 0; i < size; i++ ){
        c[i] = cxmult(a[i], b);
    }
}

```



```

)
)

/* negate square matrix */
void cxmatrix_neg(a,b,c,size)
cx *a;
cx *b;
cx *c;
int size;
{
  int i,j,k;

  for ( i = 0; i < size; i ++ ){
    for ( j = 0; j < size; j ++ ){
      k = i*size+j;
      *(c+k) = cxadd(*(a+k), cxneg(*(b+k)) );
    }
  }
}

/* divide square matrix by a real number */
void cdmatrix_div(a,b,c,size)
cx *a;
double b;
cx *c;
int size;
{
  int i,j,k;

  for ( i = 0; i < size; i ++ ){
    for ( j = 0; j < size; j ++ ){
      k = i*size+j;
      *(c+k) = cd_div(*(a+k), b);
    }
  }
}

```

END

0	2	4	0	6	4	9	5
----------	----------	----------	----------	----------	----------	----------	----------

FIN