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Equalization Requirements and Solutions for Fixed Broadband Wireless Access Systems

by

Ranjiv Singh Saini, B. Eng.

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

Master of Engineering

Ottawa-Carleton Institute for Electrical and Computer Engineering

Faculty of Engineering

Department of Systems and Computer Engineering

Carleton University

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the Faculty of Graduate Studies and Research
acceptance of the thesis

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the degree of Master of Engineering

Dr. David Falconer, thesis supervisor

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Abstract

Data communication at millimeter wave frequencies is under close evaluation for the delivery of a host of multimedia type of high bit rate applications. The work reported in this thesis sought to investigate the characteristics, in terms of the parameters such as multipath delay spread, of the wireless broadband millimeter wave outdoor radio channel in the presence of directional antennas at both transmitter and receiver, and their influence on the design of the receiver for LMCS/LMDS systems. The result of this study is significant in identifying the suitable equalization schemes and optimum equalizer structures for the broadband mm-wave wireless communication systems deployed outdoors.

A measurement campaign aimed at acquiring knowledge concerning radio propagation at 29.5 GHz frequencies was carried out in Ottawa area and the channel impulse response estimates were obtained by crosscorrelating the received sampled complex envelope with the replica of the original transmitted sequence. This impulse response data was used in this thesis to estimate the amount of multipath spread in the measured responses and to evaluate the complexity of the decision feedback equalizer in terms of the number of feedforward taps and the feedback taps required at different symbol rates for achieving the optimum solution. The results of this study are important from the viewpoint - whether an equalizer is required in an LMCS system receiver or not. At many locations strong signals were received from non-line-of-sight directions; Relatively simple DFE's appear effective at the symbol rates investigated. For example, an equalizer can be avoided at 10 Msps; (3,1) and (5,1) DFEs seem sufficient at 25 Msps and 50 Msps respectively when using a symbol spaced forward filter; and (4,1) and (8,1) at 25 Msps and 50 Msps respectively when using a FS-DFE. A comparative study of various adaptation schemes has been carried out from the point of view of the computational load and the speed of convergence of equalizer parameters. It is found that maximum-likelihood-sequence-estimation (MLSE) with acceptable complexity (with a trellis of 4 or maximum of 16 states) instead of the DFE is also be a viable alternative.

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Table of Contents

Abstract.....	iii
Acknowledgments.....	iv
Table of Contents.....	v
List of Figures.....	viii
List of Tables.....	xi
List of Acronyms and Abbreviations.....	xii
List of Mathematical Symbols and Notations.....	xv
Chapter 1.....	1
INTRODUCTION.....	1
1.1 Background.....	1
1.2 Thesis Objective.....	1
1.3 Thesis Organization.....	3
1.4 Thesis Contributions.....	3
Chapter 2.....	5
LMCS - AN OVERVIEW.....	5
2.1 Background.....	5
2.2 Rationale and Introduction to LMCS/LMDS Systems.....	7
2.3 Broadband Wireless Applications.....	12
2.4 Assumed Service Requirements.....	13
2.5 Propagation Environments / Impairments.....	14
Chapter 3.....	16
SYSTEM DESCRIPTION AND THEORETICAL BACKGROUND.....	16
3.1 Introduction.....	16
3.2 Considerations influencing performance of an Equalizer.....	17
3.3 A Brief Description of Decision Feedback Equalization.....	21

3.3.1	Symbol Spaced / Fractionally Spaced DFE	23
3.4	Adaptive Algorithms.....	26
3.5	Time variations of the channel.....	27
3.6	Maximum-Likelihood Sequence Estimation (MLSE)	29
3.6.1	Performance of Viterbi Algorithm	30
Chapter 4	.....	32
EHF MEASUREMENTS AND DATA PROCESSING.....		32
4.1	Introduction.....	32
4.2	EHF Measurement Details.....	33
4.3	Wideband Data Processing	37
Chapter 5	.....	43
ANALYSIS OF MEASURED BROADBAND CHANNEL IMPULSE RESPONSES AND DISCUSSION		43
5.1	Introduction.....	43
5.2	Parkwood Hills Impulse Responses Analysis Results	44
5.2.1	10 Mega symbols per second.....	50
5.2.2	25 Mega symbols per second.....	60
5.2.3	50 Mega Symbols Per Second	66
5.3	Kanata Impulse Responses Analysis Results.....	71
5.3.1	10 Mega Symbols per second	76
5.3.2	25 Mega symbols per second.....	80
5.3.3	50 Mega symbols per second.....	85
5.4	Adaptation Algorithms Complexities and Comparison.....	98
5.4.1	LMS Algorithm:	98
5.4.2	Block adaptation with DMI (direct matrix inversion) using Covariance method:99	
5.4.3	RLS Algorithm:	100

5.4.4 Fast RLS Algorithm:	100
5.5 Time Variation of the Channel	101
5.6 Maximum-Likelihood Sequence Estimation	102
Chapter 6	108
CONCLUSIONS	108
6.1 Summary of the results	108
Appendix A	112
EQUALIZER MSE EVALUATION	112
Appendix B	119
IMPULSE RESPONSE ANALYSIS RESULTS	119
LIST OF REFERENCES	129

List of Figures

FIGURE 2.1	Local Multipoint Communication Service (LMCS) Concept	8
FIGURE 3.1	Block diagram of a baseband data transmission system	18
FIGURE 3.2	A Typical DFE structure.....	24
FIGURE 4.1	A typical delay profile before thresholding	37
FIGURE 4.2	Delay profile of Figure 4.1 on page 37 after thresholding	41
FIGURE 5.1	Histogram of angle of occurrence of strongest signal at all sites	44
FIGURE 5.2	CDF of multipath spread for receive directions of and	46
FIGURE 5.3	CDF of the multipath spread for best receive directions at all sites	47
FIGURE 5.4	Sample level Vs. delays for channel at 6 degree at site 11b23002.....	50
FIGURE 5.5	Contour plots of SNR _{out} variation with feedforward and feedback filter taps, at best sampling phase at 10 M symbol per second.....	51
FIGURE 5.6	SNR _{out} variation with sampling phase for channel impulse response of Figure 5.4 at 10 M symbol per second	52
FIGURE 5.7	SNR _{out} variation with sampling phase for channel impulse response of Figure 5.4 with FS-DFE at 10 M symbols per second	56
FIGURE 5.8	Contour plots of SNR _{out} variation with forward filter and feedback taps, at 10 Msps with forward filter T/2 spaced.....	57
FIGURE 5.9	Contour plots of SNR _{out} variation with feedforward and feedback filter taps, at best sampling phase for 25 M symbols per second.....	59
FIGURE 5.10	SNR _{out} variation with sampling phase for channel impulse response of Figure 5.4 for 25 M symbol per second.....	60
FIGURE 5.11	SNR _{out} variation with sampling phase for channel impulse response of Figure 5.4 for FS-DFE at 25 M symbols per second.....	62
FIGURE 5.12	Contour plots of SNR _{out} variation with forward filter and feedback taps, at 25 Msps with forward filter T/2 spaced.....	63

FIGURE 5.13	Contour plots of SNR _{out} variation with feedforward and feedback filter taps, at best sampling phase for 50 M symbols per second.....	65
FIGURE 5.14	SNR _{out} variation with sampling phase for channel impulse response of Figure 5.4 at 50 M symbol per second	66
FIGURE 5.15	SNR _{out} variation with sampling phase for channel impulse response of Figure 5.4 for FS-DFE at 50 M symbols per second.....	67
FIGURE 5.16	Contour plots of SNR _{out} variation with forward filter and feedback taps, at 50 Msps with forward filter T/2 spaced.....	68
FIGURE 5.17	CDF of the multipath spread at zero degree receive directions for Kanata data.....	70
FIGURE 5.18	Sample level Vs. delays for channel at 0 degree at site kab29044	72
FIGURE 5.19	Contour plots of SNR _{out} variation with feedforward and feedback filter taps, at best sampling phase for 10 M symbol per second	73
FIGURE 5.20	SNR _{out} variation with sampling phase for channel impulse response of Figure 5.18 at 10 M symbol per second (Kanata data).....	75
FIGURE 5.21	SNR _{out} variation with sampling phase for channel impulse response of Figure 5.18 for FS-DFE at 10 M symbols per second (Kanata data).....	76
FIGURE 5.22	Contour plots of SNR _{out} variation with forward filter and feedback taps, at 10 Msps with forward filter T/2 spaced (Kanata data)	77
FIGURE 5.23	Contour plots of SNR _{out} variation with feedforward and feedback filter taps, at best sampling phase for 25 M symbols per second (Kanata data).....	79
FIGURE 5.24	SNR _{out} variation with sampling phase for channel impulse response of Figure 5.18 at 25 M symbol per second (Kanata data).....	80
FIGURE 5.25	SNR _{out} variation with sampling phase for channel impulse response of Figure 5.18 for FS-DFE at 25 M symbols per second (Kanata data).....	81

FIGURE 5.26	Contour plots of SNR _{out} variation with forward filter and feedback taps, at 25 Msps with forward filter T/4 spaced (Kanata data)	82
FIGURE 5.27	Contour plots of SNR _{out} variation with feedforward and feedback filter taps, at best sampling phase for 50 M symbols per second (Kanata data)	84
FIGURE 5.28	SNR _{out} variation with sampling phase for channel impulse response of Figure 5.18 at 50 M symbol per second (Kanata data).....	85
FIGURE 5.29	Contour plots of SNR _{out} variation with forward taps and feedback taps, at 50 Msps with forward filter T/2 spaced (Kanata data)	86
FIGURE 5.30	SNR _{out} variation with sampling phase for channel impulse response of Figure 5.18 for FS-DFE, at 50 M symbols per second.....	87
FIGURE 5.31	Histogram of output SNR (dB) for the channels of Parkwood Hills and Kanata with a DFE of (1,0) at 10 Msps, symbol spaced case.....	89
FIGURE 5.32	Histogram of output SNR (dB) for the channels of Parkwood Hills and Kanata with a FS-DFE of (2,0) at 10 Msps.	90
FIGURE 5.33	Histogram of output SNR (dB) for the channels of Parkwood Hills and Kanata with a symbol spaced DFE of (3,1) at 25 Msps.	91
FIGURE 5.34	Histogram of output SNR (dB) for the channels of Parkwood Hills and Kanata with a FS-DFE of (4,1) at 25 Msps.	92
FIGURE 5.35	Histogram of output SNR (dB) for the channels of Parkwood Hills and Kanata with a symbol spaced DFE of (5,1) at 50 Msps.	93
FIGURE 5.36	Histogram of output SNR (dB) for the channels of Parkwood Hills and Kanata with a FS-DFE of (8,1) at 50 Msps.	94
FIGURE 5.37	Probability of symbol error bounds for MLSE and DFE QPSK signalling.....	103
FIGURE 5.38	Histogram of the SNR gain of MLSE over DFE for P_e of $1.0e-3$	104

List of Tables

TABLE 5.1	Summary of DFE structures at 25 Msps and 50 Msps.....	95
TABLE 5.2	Computational Complexity with LMS algorithm	96
TABLE 5.3	Computational Complexity with Block Adaptation	97
TABLE 5.4	Computational Complexity of RLS Algorithm.....	97
TABLE 5.5	Computational Complexity of fast RLS Algorithm.....	98
TABLE 5.6	Computations of time variation of channel.....	99
Table B.1	Output SNR (dB) for various tap combinations for Symbol spaced and Fractionally Spaced DFE.....	116

List of Acronyms and Abbreviations

ABR	Available bit rate
ADSL	Asynchronous digital subscriber loop
ATM	Asynchronous transfer mode
AWGN	Additive white gaussian noise
BPSK	Binary phase shift keying
C/N	Carrier to noise ratio
CBR	Constant bit rate
CDF	Cumulative distribution function
CDMA	Code division multiple access
CIR	Complex impulse response
CITR	Canadian Institute for Telecommunications Research
CRC	Communications Research Centre
CW	Continues wave
dB	decibel
dBi	dB of an antenna gain relative to isotropic gain
dBm	dB of power relative to one milliwatt
DAVIC	Digital audio video council
DFE	Decision-feedback equalization
DMI	Direct matrix inversion
DQPSK	Differential quadrature phase shift keying
EHF	Extremely high frequency
FCC	Federal Communications Commission
FDM	Frequency division multiplex

FDMA.....Frequency division multiple access
 FM.....Frequency modulation
 FS-DFE .....Fractionally spaced decision-feedback equalizer
 FTTH.....Fiber-to-the-home
 HEMT .....High electron mobility field effect transistors
 HFC.....Hybrid-fiber-coax
 Hz.....Hertz
 ISI.....Inter-symbol-interference
 LMCS.....Local Multipoint Communication Service
 LMDS .....Local Multipoint Distribution Service
 LMS .....Least mean square
 LOS .....Line-of-sight
 GHz.....Giga Hertz
 GaAS.....Gallium Arsenide
 MAP .....Maximum a posteriori probability
 Mbps .....Mega bits per second
 MF.....Matched filter
 MHz .....Mega Hertz
 ML.....Maximum likelihood
 MLSE.....Maximum-likelihood sequence estimation
 mm .....Millimeter
 MMDS .....Multichannel Multipoint Distribution Service
 MMSE.....Minimum-mean-square error
 MSE .....Mean-square error
 Msps.....Mega symbols per second

ns	Nano second
OFDM	Orthogonal frequency division multiplex
Pe	Probability of symbol error
PN	Pseudo noise
PRBS	Pseudo random bit sequence
QAM	Quadrature amplitude modulation
QPSK	Quadrature phase shift keying
RF	Radio frequency
RLS	Recursive least square
SHF	Super high frequency
SNR	Signal-to-noise power ratio
TDM	Time division multiplex
TDMA	Time division multiple access
TRIO	Telecom Research Institute of Ontario
VA	Viterbi Algorithm
VBR	Variable bit rate
XPI	Cross polarization isolation

List of Mathematical Symbols and Notations

$\{b_j\}$	.....Tap gains of the feedback filter of DFE
B_D	.....Doppler spread
d_n	.....Data sequence
d_{min}	.....Minimum distance between two sequences
$e(n)$	.....estimation error
$\{f_j\}$	.....Tap gains of the feedforward filter of DFE
f_m	.....Maximum doppler shift
$g(t)$	.....Transmitter filter pulse shape
$G(f)$	.....Transfer function of the receive filter
$h(t))$	.....Channel impulse response
J	.....Mean-square error
J_{min}	.....Minimum-mean-square error
$\log_2(x)$	.....log to the base 2 of x
$N_0/2$	.....Two sided noise power spectral density
OP_{FF}	.....Output of the feedforward filter of DFE
OP_{FB}	.....Output of the feedback filter of DFE
P	.....Cross correlation vector of tap inputs of DFE and desired signal
P_{fa}	.....Probability of false alarm
$Pr(e)$	.....Probability of symbol error

Q	Samples per symbol period
$Q(x)$	Error function
$r(t)$	Receive signal
$rc(t)$	Raised cosine impulse response
R	Correlation matrix of tap inputs of the forward and feedback taps
S	Number of samples per symbol interval
$s(t)$	Transmitted signal
T	Symbol duration
T_c	Coherence time
$T-sp$	Symbol spaced DFE
v	Velocity of an object
W_{opt}	Optimum tap weights of DFE
$x(t)$	Pulse presented to the receiver
$y(n)$	Signal at the input of the detector at time n
$\Psi(t)$	Resultant pulse shape including the effects of channel as well as transmit and receive filters
β	Raised cosine roll-off factor
λ	Wavelength
ζ	Some threshold level
σ	Standard deviation of noise
σ_m	Median level
σ_d^2	Variance of the desired response

σ_N^2	Variance of background thermal noise
τ_0	Optimum sampling phase
$\eta(t)$	Additive white Gaussian noise
$( )^H$	Hermitian transposition
$\sum_a^b$	Summation from a to b
$Re(a)$	Real part of a complex variable a
$E[a ^2]$	Mean square of a
$ a $	Absolute value of a
π	3.141592654
$\delta(.)$	Dirac delta function
a^*	Complex conjugate of a
$\otimes$	Continuous or discrete time convolution operator

Chapter 1

INTRODUCTION

1.1 Background

Due to phenomenal increase in demands for the multimedia type of high bit rate services, there has been a significant activity to find cost effective alternatives to other relatively expensive broadband communication technologies like fiber-to-the-home (FTTH), hybrid-fiber-coax (HFC) and Asynchronous Digital Subscriber Loop (ADSL) on copper wire, for the delivery of broadband applications such as broadcasting multi-channel TV, video-on-demand, high speed internet access, video conferencing and data transfer etc. Explorations into the broadband communications have been furthered through many independent research projects to make use of ATM (asynchronous transfer mode) to transport high speed data over a wireless network. Broadband communication between a nodal point and a number of fixed subscribers is being contemplated without having to build the expensive wired networks. The proposed systems termed as LMCS (Local Multipoint Communication Service) could be used for providing bidirectional broadband wireless access to fixed subscribers through millimeter wave frequencies, promise lower infrastructure cost and time to market. Independent investigations and standardization efforts have been going on in many parts of the world for years.

1.2 Thesis Objective

One of the main technical challenges in realizing such systems is how to deal

with the radio aspect which includes the problem of how to combat the deleterious effects of time varying multipath dispersion and fading prominent in any outdoor environments due to a number of static and moving obstacles, scatters, and reflectors. The information regarding these factors is absolutely necessary before many design parameters such as modulation schemes, equalization, coding and diversity etc. can be finalized.

The main objective of this thesis is to investigate, based upon the channel characteristics, if an equalizer is required in the LMCS system receiver over a range of bit rates and if any, then of what type and what complexity? The impulse responses needed for this analysis were made available from outdoor measurements done in Ottawa and Kanata areas which involved the measurement of radio channel characteristics between a typical cell site transmitter and many receive locations employing directional antennas.

After postulating the equalizer structure at different symbol rates (for both baud rate decision feedback equalizer (DFE) and Fractionally Spaced DFE), it is aimed to do a comparative study of various equalizer adaptation algorithms and propose the most suitable one for the present set of conditions. A receiver based on the Maximum-Likelihood-Sequence-Estimation (MLSE) is the optimum solution but complexity involved in its implementation for channels with time dispersion is unreasonably large unless some sub-optimal measures like reducing the channel memory prior to maximum likelihood estimation are adopted. Here in this thesis we have tried to find the viability of applying the MLSE by approximating the overall impulse response of the channel with its shorter equivalent and computing the computational complexity if we apply the viterbi algorithm (VA) to this impulse response and compare the error bounds obtained by applying DFE and MLSE.

1.3 Thesis Organization

Following this introductory chapter, in Chapter 2, an overview of the LMCS system concept has been presented along with a brief description of the proposed system architecture, assumed system requirements, broadband wireless applications and the problems posed by the propagation environments in the evolution of this technology. Chapter 3 provides a theoretical background necessary for understanding and solving the thesis problem. It includes a description of the problem of intersymbol interference (ISI), system model used for the analysis of the impulse responses, mathematical and theoretical treatment of DFE equalizers, and a brief description of the equalizer adaptation algorithms and their complexities.

Chapter 4 is about the EHF measurements and the subsequent processing of the recorded data that has provided the complex impulse responses for this thesis to work upon. In Chapter 5, we present results on the equalization requirements of LMCS systems, obtained from the analysis of the measured impulse responses. In this chapter we treat the impulse responses of the channel with both baud-rate DFE and fractionally spaced DFE (FS-DFE) at different symbol rates. A comparative study of various equalizer adaptation algorithms from the point of view of their complexities and convergence speeds, based upon the equalizer results obtained, has also been presented. Finally, Chapter 6 summarizes the conclusions from this thesis work.

1.4 Thesis Contributions

The research that we have done here has sought to answer the equalization requirements of the LMCS system receiver, on the basis of the multipath spread present

in the measured outdoor impulse responses. The research has also validated that FS-DFE gives better performance than a baud-rate sampling DFE. It has been concluded from the analysis of the measured impulse responses that when there is a directional antenna at the transmitter and a very narrow beamwidth antenna at the receiver, multipath is reduced, especially in the signals received with maximum power at each location, which simplifies the equalization requirements. Symbol rates of the input signal have been varied to show their effect on the change of DFE structure.

It has also been shown that for some of the data rates it is possible to avoid an equalizer and for some higher data rates, an equalizer would be required for improved system performance. Based on the comparison of various equalizer adaptation schemes in respect of their adaptation speeds and the computational burden associated with them, most suitable of them is identified for LMCS system modems. It has been shown that owing to the fixed subscriber locations, the channel stays relatively constant during the transmission of a time slot carrying one ATM cell, which obviates the need to track the channel during the data transmission once the equalizer is trained. It has also been shown that a receiver based on MLSE using VA is also computationally feasible when the overall system impulse response is approximated by a shorter truncated version of it and the error rate performance of this receiver is shown to be superior than that of DFE.

Chapter 2

LMCS - AN OVERVIEW

Before proceeding to the actual problem addressed by the thesis, we first present an overview of the LMCS system concept, the assumed service requirements, a brief description of the proposed system architecture, and a quick survey of the problems faced for the conception and evolution of this technology which promises a phenomenal growth once realized.

2.1 Background

There is currently a significant activity by the Canadian and US telecommunication companies and service providers focussed on development and deployment of cellular type broadband (generally speaking, systems with user data rates greater than 2 Mega bits per second upto 100s of Mega bits per second) local area radio communication systems to serve both residential and commercial subscribers. A similar type of service known as “Multichannel Multipoint Distribution Service (MMDS)”, also known as wireless cable [1] [2], had been in use in different parts of US to deliver one-way Video broadcast services at SHF range (2.5 GHz frequency) to fixed subscribers, as an alternative to cable TV, but with little or no upstream bandwidth for broadband interactive services. Actually the lower frequencies at which MMDS is operating do not have the requisite bandwidth for the type of services LMCS has envisioned to provide.

While MMDS was drawing much of attention in the United States, the concept of

using millimeter wave radio frequencies for the purpose of delivering cable services was proposed by CellularVision which it called “fiber in the sky” [3]. Later on, the millimeter wave frequency band evinced a lot of interest from telecommunication service providers and equipment manufacturers to provide a host of two-way interactive multimedia and video broadband services which were earlier provided by local exchange carriers and cable operators.

In 1995, the Federal Communications Commission (FCC) in the United States proposed a non contiguous band of 1.3 GHz of wireless spectrum in the 28 GHz Ka-band, with the band splitting as - 850 MHz at 27.5 GHz to 28.35 GHz, 150 MHz at 29.1 GHz to 29.25 GHz and 300 MHz from 31.0 GHz to 31.3 GHz, to distribute bidirectional broadband information and entertainment services called Local Multipoint Distribution Service (LMDS) [4]. This 1300 MHz of the spectrum will be split into two licenses, one for 1150 MHz and one for 150 MHz. Following that Industry Canada also started the licensing process in 28 GHz band with one GHz of spectrum to each service provider (1 per city) from 27.35 GHz to 28.35 GHz [5]. In Canada, These services will be termed as Local Multipoint Communication Services (LMCS). An additional band at about 23 GHz has also been proposed by Industry Canada for private enterprise networks and fixed radio backbone services [6]. There is also a possibility of offering these services at 40 GHz (40.5 GHz to 42.5 GHz) for additional capacity subject to satisfactory propagation. The most attractive fact is the availability of a huge bandwidth (of the order of 1 GHz) to provide a mix of asymmetrical services over radio waves. Similar bandwidths have also been allocated at millimeter frequencies in other parts of the world like Europe (bands centered on 60 GHz are being considered) and Japan for broadband services. So

for data rates of the order of 10's and 100's Mbps, the use of millimeter band seems an appropriate choice because of relative abundance of available bandwidth. Also, with recent advances in solid-state technology such as GaAs and HEMT devices, efficient design of both RF and base band sub-systems is possible at millimeter wave frequencies and high data rates [31] [32].

2.2 Rationale and Introduction to LMCS/LMDS Systems

Local Multipoint Communication Service is an outdoor, terrestrial point to multipoint wireless access system to deliver bidirectional broadband information and entertainment services to subscribers in fixed locations such as homes and businesses or broadband wireless access to fixed networks [7]. The service can be put to use to provide one-way video distribution, interactive video as well as high speed data transmission and telephony services. This proposed technology at 28 GHz is intended to be a cost effective alternative to other broadband communication services such as fiber to the home (FTTH) or curb, hybrid fiber coax (HFC), ADSL on copper twisted pair. The advantage of wireless local access over these competing technologies would be significantly lower infrastructure cost and time to market [8].

LMCS/LMDS will be a cellular structure with cell radii of the order of less than 5 km with eventual evolution to microcellular system (cell radii 0.5 km to 2 km) to cater to increased subscriber densities. Using available technology, a system would be capable of delivering 50 to 100 channels in 1 GHz of bandwidth, whereas the number of channels can be further expanded using digital modulation techniques and video compression. To subscribe to these services, each customer is equipped with small high gain antenna and

FIGURE 2.1 Local Multipoint Communication Service (LMCS) Concept

a transceiver set top box. The proposed systems attempt to completely reuse the allocated bandwidth in a cell and possibly in each sector of the cell. Highly directional subscriber antennas and orthogonal linear polarization transmissions (polarization reuse) in adjacent cells (or sectors) reduce the interference levels in the system. To further reduce the effect of inter cell, inter system interference, frequency interleaving can be used. Figure 2.1 [7][9][10] shows a typical LMCS/LMDS scenario with three overlapping cells which enables the subscribers to have the cell site diversity, fixed subscribers with highly directional antennas and a broadband central office. The link budgets proposed for LMCS systems are based on the assumption of an unblocked path between the base station transmitter and the subscriber antenna, which may not be true in many situations due to the presence of obstacles from built up structures, trees or other path impairments. The line-of-sight or near line-of-sight requirement for reliable communication at these frequencies is presenting a huge challenge to the system designers who are striving hard to maximize coverage at minimum cost.

Area coverage in a LMCS system can be increased by the use of high hub antennas, cell site diversity using overlapping cells, micro cellular structure, use of passive reflectors in some locations, high gain directive hub site antennas in addition to highly directional high gain subscriber antennas. On-channel low power active repeaters have also been recommended to fill the gaps where there is insufficient signal strength due to excessive blockage in the direct path to the nearest hub transmitter. The repeater will amplify and redirect the signal in the direction of the coverage hole [11] [7].

While some of the early system implementations (e.g. CellularVision) have used

analog FM modulation for the limited purpose of broadcasting one-way video, the future implementations of LMCS/LMDS are focussing on digital modulation formats to increase channel capacity to provide data and many more high bit rate services. To maximize the system channel capacity, highly spectrally efficient modulation schemes such as 16-QAM or 64-QAM etc. seem attractive, however, these schemes are less tolerant to interference and channel distortions, and require higher power (C/N) and more linear componentry. On the other hand, low level modulations such as QPSK, 4QAM or DQPSK etc. are more tolerant to interference and channel distortions, have lower power requirements (C/N or SNR), reasonable spectral efficiency [12] and they could be utilized during the initial phases of system deployment with low take rates. Also, these serial modulations along with adaptive equalization should be able to combat inter-symbol-interference (ISI) resulting from the detrimental effects of delay spreads at high bit rates. Another anti-multipath scheme such as OFDM (orthogonal frequency division multiplex) in combination with 16-QAM is also a possible alternative.

The air interface format for LMCS/LMDS may be FDMA (frequency division multiple access), TDMA (time division multiple access), CDMA (code division multiple access) or combinations thereof. The multiple access method for upstream bandwidth may be different than the downstream due to lower bit rate requirements and physical separation of the subscribers. For downstream, the traffic meant for different subscribers can be multiplexed using TDM or FDM. CDMA for downstream may not be a viable alternative because of the huge bandwidth requirements at high bit rates although it can be employed for the upstream due to its spectral efficiency, robustness and multiple access simplicity. The upstream traffic will allow for statistical multiplexing of bursty

subscriber traffic. TDMA is preferable to support services with different bit rates whereas FDMA is suited to be used with fixed bandwidth requirements such as TV broadcasting. A multiple access protocol may be based upon polling or reservation/contention in a time divisioned frame structure [12] [13] [14].

CellularVision, Video/Phone and Texas Instruments are some of the companies to propose preliminary system designs [7] and their systems are briefly discussed here.

CELLULARVISION SYSTEM: The CellularVision LMDS system [7], which it calls “fiber in the sky”, operates with a cell size of approximately 4.8 km to deliver analog FM video signal. The system proposes to reuse entire frequency band in each cell and employs polarization reuse, frequency interleaving and directional subscriber antennas to reduce intera-system interference. It employs omnidirectional transmitter antenna to transmit from the center of the cell and narrow beamwidth subscriber antennas to eliminate multipath reception and to gain sufficient link margin for operation. The return links operate in the guard bands of the downstream broadcast video channels owing to smaller signal bandwidths of the applications such as telephony, pay-per-view requests and home shopping and require less transmitter power than down links.

TEXAS INSTRUMENT SYSTEM: Texas Instrument (TI) system [7] is a two-way digital (employs QPSK modulation) high bit rate system capable of equal data rates in both directions to provide services like video, data, telephony and fully interactive switched broadband multimedia applications. It employs four sectorized hub antennas with cell radii of about 5 kms. Power control is implemented to compensate for excess propagation loss due to rain. Isolation between adjacent cells is achieved by cross polarization transmis-

sion and highly directional subscriber antennas. For applications involving asymmetrical traffic, TDMA is used to multiplex various users.

VIDEO/PHONE SYSTEM: The Video/Phone system [4][7] is a digital system (employs QAM modulation) with 36 sectored hub antenna to provide overall omnidirectional coverage in comparatively smaller cells of 1.8 km radii. Like the TI system, it also employs power control to provide extra gain to overcome attenuation due to rain and foliage. Extra gain is also available due to highly sectored antenna at the hub. Dedicated spectrum is required for uplink transmissions. The system may employ hub diversity in some configurations, to allow users to orient antennas towards best hub locations, as well as dynamic channel assignment.

2.3 Broadband Wireless Applications

Local Multipoint Communication Service at 28 GHz promises to provide wireless access to a host of services ranging from broadcast video to many packet switched and circuit switched applications at competitive cost compared to other alternative technologies. Some of the targeted applications of LMCS include [10] [11] [15]:

NARROWCAST:

- Video-on-demand and Music-on-demand
- High speed Internet access
- Desktop Video conferencing
- Distance Education
- Interactive Multimedia communication

- Telephony services
- High resolution image and large file transfer; CAD
- LAN interconnect, Wireless LAN
- Games and home shopping

BROADCAST:

- Broadcast Video (Multichannel TV)

2.4 Assumed Service Requirements

In the LMCS scenario, the bit rates to and from each subscriber are expected to be variable in the range of 5 to 50 M bits per second, with downstream traffic typically requiring higher bit rates than the up stream traffic. The downstream spectrum of 850 MHz (or 500 MHz) will consist of both public and private portions. The public portion is to broadcast to all the subscribers and will include analog and/or digital TV channels, HDTV, sports etc. and will be mostly constant bit rate (CBR) traffic, it may or may not use ATM transport whereas the private portion of the spectrum will include variable bit rate traffic (VBR), such as video-on-demand and file transfers etc. meant for individual subscribers. ATM protocols and signalling can be used for this portion. In a typical LMCS system, there will be around 200 - 500 residential subscribers per cell, in contrast to fewer business subscribers. The shared downlink bit rates would be around 50 to 100 M bits per second which would be much higher (5 to 10 times) than the uplink traffic rates. For business subscribers, maximum bit rates per customer can be as high as 155 Mbps (OC-3) or 622 Mbps (OC-12).

Antennas with adaptive arrays will be required to achieve high coverage, capacity and higher data rates (owing to less multipaths and increased power margins) [16].

2.5 Propagation Environments / Impairments

One of the major challenges in delivering LMCS to the subscribers is to overcome the fragile propagation environment at millimeter wave frequencies. At these frequencies, the signal is so much attenuated by the built up structures that it becomes almost necessary to provide line-of-sight (LOS) path from hub to each subscriber for maintaining sufficient signal strength over individual radio links. In situations where the radio path between a subscriber and all neighboring transmitters have blockage due to buildings, trees or terrain obstructions, the service may be degraded or even lost. The signal wavelength at these frequencies is so small that any clutter or foliage in the radio paths become potential scatters of the signal energy resulting in significant excess propagation loss and multipaths. The wind induced motion of the foliage can result in significant fluctuation in the received signal levels. Even when LOS paths are available, excess loss due to rain attenuation must be accounted for in the link budgets.

It may be required to provide overlap of cells to maintain LOS links with subscribers for attaining more than 90% coverage of a single service area. Advances in solid-state technology (GaAs and HEMT devices) are under way for efficient design of both RF and baseband subsystems at millimeter wave frequencies and at high data rates. Efforts to reduce the cost of EHF components are absolutely necessary to make these services common place. Many of the currently planned systems may be operating with various transmission margins that are arbitrarily chosen in the absence of detailed infor-

mation on propagation characteristics. In order to properly model the channel at the millimeter frequencies through experimental and analytical research, many propagation measurement campaigns have been conducted in the past or are presently going on, to help an overall efficient and cost effective system design [7] [17] - [27]. In this thesis we also talk about a similar measurement campaign conducted in the Ottawa area.

In summary, we can say that if the system designers are able to overcome all the limitations imposed by the propagation environments, LMCS/LMDS can play a major role in deployment of future telecommunication networks.

Chapter 3

SYSTEM DESCRIPTION AND THEORETICAL BACKGROUND

3.1 Introduction

The radio channel between the transmitter and the receiver places fundamental limitation on the performance of wireless communication systems. The presence of reflecting objects and scatterers in the channel result in multiple versions of the transmitted signal that arrive at the receiving antenna displaced with respect to one another in time and in spatial orientation. In the case of multipath propagation, the bandwidth of the channel can be quantified by the ‘coherence bandwidth’ which is related to the specific multipath structure of the channel [29].

Depending upon the channel coherence bandwidth and the information transmission rate, the communication systems can be classified as 1) narrowband systems, where the coherence bandwidth of the channel is larger than the signal bandwidth or 2) wideband systems, where the transmission bandwidth is larger than the channel coherence bandwidth. In the case of narrowband systems the amplitude of the received signal may fluctuate rapidly, but the signal will not be distorted in time. Remedies like antenna diversity, coding and interleaving together with adaptive equalization (optional) can improve the system performance [33] [29]. Whereas, in the case of wideband systems, the received signal will be distorted due to the multipath channel but the received signal strength will fade less over a local area since it may be the sum of several independently

fading paths. The wideband channels are characterized by the frequency selective fading which results from the fact that the channel possesses a constant gain and linear phase response only over bandwidths that are smaller than the bandwidth of the transmitted signal. Multipath propagation often lengthens the time required for the signal to reach the receiver and owing to the small symbol durations at high bit rate applications (as perceived for wideband LMCS/LMDS systems), the multiple received copies of the transmitted signal overflow to the successive bit/symbol periods causing inter-symbol-interference (ISI) which has been recognized as a major obstacle in high speed data transmission.

Equalization is an effective remedy for ISI caused by bandlimited channels and multipath propagation. Figure 3.1 shows an equalizer as a part of a typical data transmission system. An equalizer within a receiver compensates for the average range of expected channel amplitude and delay characteristics. If the channel is frequency selective, the equalizer enhances the frequency components with small amplitudes and attenuates the strong frequencies in the received frequency spectrum in order to provide a flat frequency response and linear phase response. In this thesis, we have employed Decision Feedback Equalization (DFE) and later, maximum likelihood sequence estimation (MLSE) to equalize the measured impulse responses and hence to determine the optimum equalizer structure at different symbol rates.

3.2 Considerations influencing performance of an Equalizer

In this section various factors that influence the performance of an equalizer are outlined. The performance of an equalizer in a particular operating environment is

FIGURE 3.1 Block diagram of a baseband data transmission system

mainly dependent upon following:

i) Performance index or cost function:

The equalizers try to optimize their parameters by operating upon a performance index or a cost function in a defined manner. The cost function can be the mean square error or the probability of symbol error i.e. the optimization process may involve a) the minimization of the mean square error between the information symbol transmitted (or expected information symbol) and the estimate of that symbol at the output of the equalizer or b) maximization of the a posteriori probability (MAP) or maximum likelihood function (ML). Since the most meaningful measure of the performance of a digital communication system is the average probability of symbol error, it is desirable to choose the equalizer coefficients to minimize this performance index. However, the probability of error is a highly non linear function of the equalizer tap coefficients. Consequently, the choice of this cost function for tap weight optimization is impractical [30], however the minimization of the mean square error (MSE) is a linear estimation method, involves only the second order statistics in the estimation process and finds widespread use in optimizing the equalizer tap coefficients. It forms the basis for the most useful adaptation techniques found in the practical systems.

ii) Equalization method / Detection Schemes:

The equalizer performance also depends upon the type of the equalization scheme used. One is based on the maximum-likelihood sequence detection criteria, which is optimum from the probability of error viewpoint and this scheme is known as Maximum Likelihood Sequence Estimation (MLSE). In this case, a block of information

symbols is detected at one time and the cost function is calculated over a period of one block of information symbols.

The second type of commonly used equalization method is based on the use of linear filters with adjustable coefficients and exploits the use of previously detected symbols to suppress the inter-symbol-interference in the present symbol being detected. It is called decision-feedback equalization (DFE) and is a suboptimum channel equalization approach. It is a symbol by symbol detection scheme and each symbol is detected independently.

MLSE detectors incur large delays as they employ probabilistic algorithms in the detection process which affect the channel estimation process. The advantage of using sequence estimation is its optimality, but its complexity increases exponentially with the increase in the time dispersion of the channel. Hence, the selection between the two schemes mainly depends on the complexity and whether the system performance is affected by large decision delays.

iii) Adaptation Techniques:

In most of the communication systems that employ equalizers, the channel characteristics are unknown a priori and in many cases, the channel impulse response is time varying. In such a case, the equalizers are designed to be adjustable to the channel response and for time variant channels, to be adaptive to the time variations in the channel impulse response through the use of adaptive algorithms. The type of the adaptive algorithm used has direct effect on the system performance. Least mean square (LMS), Block adaptation using covariance method and Recursive least square (RLS) and its vari-

ants are some of the adaptation algorithms commonly used, and the choice amongst them depends upon the trade off between the required rate of convergence and the computational complexity that can be afforded.

3.3 A Brief Description of Decision Feedback Equalization

Decision Feedback Equalization (DFE) is one of the techniques to effectively combat the detrimental effects of ISI. The DFE belongs to non linear class of equalizers, and it makes use of data decisions that have already been made to provide improved performance (provided the decisions are correct). Here in this section, both symbol spaced and fractionally spaced DFE have been detailed.

Consider discrete information bearing sequence of symbols $\{d_n\}$ corresponding to input data being applied to the transmitter at symbol intervals of T seconds. The symbols, transmitted using QPSK format, are assumed to be independent with zero mean and unit variance. The data sequence after passing through the transmitter filter is:

$$s(t) = \sum_n d_n g(t - nT) \quad (3.1)$$

where $g(t)$ is the transmitter filter pulse shape that has band-limited frequency response characteristics. In a baseband data transmission system, the output from the transmitter filter is filtered by the channel $h(t)$, (represented by the measured outdoor impulse responses in this thesis) and is corrupted by the additive white gaussian noise (AWGN) $\eta(t)$ with two-sided power spectral density of $N_0/2$ watts/Hz. The received signal has the form:

$$r(t) = \sum_n d_n x(t - nT) + \eta(t) \quad (3.2)$$

where $x(t) = g(t) \otimes h(t)$ the pulse presented to the receiver (here $\otimes$ is the convolution operator). After passing through the receiver front end filter, the signal can be represented as:

$$z(t) = \sum_n d_n \Psi(t - nT) + \eta'(t) \quad (3.3)$$

where $\Psi(t)$ is the resultant pulse shape including the effects of channel as well as transmit and receive filters and $\eta'(t)$ is the bandpass noise. The first term in equation (3.3) represents the desired information symbol along with the effect of intersymbol interference. At time nT , the information symbol d_n is needed to be detected. The AWGN is filtered by the receiver filter and then it has the power spectral density that can be represented by $\frac{N_0}{2}|G(f)|^2$ where $G(f)$ is the transfer function of the receive filter.

TRANSMITTER AND RECEIVER FILTERS: In the system model used for the analysis of the measured impulse responses, transmitter and receive filters have been implemented in the form of square root raised cosine filters, so that their combined response has raised cosine spectral characteristics. So as to determine the equalizer structure at different symbol rates, these filters were designed for symbol rates of 10, 25 and 50 Mega symbols per second (Msps). The corresponding bit rates, if QPSK modulation is used, are 20, 50 and 100 Mega bits per second (Mbps). The impulse response of a raised cosine filter is:

$$rc(t) = \frac{\sin(\pi t/T)}{(\pi t/T)} \cdot \frac{\cos(\pi \beta t/T)}{1 - 4\beta^2 t^2/T^2} \quad (3.4)$$

where, β = roll-off factor and is equal to 0.3 in our analysis, and T = Symbol period

3.3.1 Symbol Spaced / Fractionally Spaced DFE

A typical block diagram of a DFE, shown in Figure 3.2 , consists of two filters, a feedforward filter and a feedback filter. The signal presented at the input of the feedforward filter is given by equation (3.3).

FEEDFORWARD FILTER: In this thesis, both symbol spaced and fractionally spaced (FS) feedforward filters have been considered and their performance compared. An equalizer with a FS feedforward filter effectively compensates for arbitrary channel delay distortion without any noise enhancement and it avoids extreme sensitivity to sampling phase as is associated with conventional symbol-spaced forward filter case. In the case of a FS filter, taps are T' spaced, where $T' = T/S$ and S is the number of samples per symbol interval T . In the case of FS-DFE, the input signal $z(t)$ is sampled every T' seconds whereas for a symbol spaced DFE, it is every T seconds. The output of the FS feedforward filter is:

$$OP_{FF}(n) = \sum_{j=0}^{K_1} f_j^* z(nT - jT'); \quad n = 0, 1, 2, \dots \quad (3.5)$$

where $\{f_j\}$ are the $K_1 + 1$ complex valued tap weight coefficients of the forward filter and asterisk denotes complex conjugation. For a symbol spaced feedforward filter output, T' gets replaced by T in equation (3.5).

FEEDBACK FILTER: The feedback filter in a DFE is used to remove that part of ISI from the present estimate which is caused by previously detected symbols. The feedback

FIGURE 3.2 A Typical DFE structure

filter has symbol spaced tap coefficients and its output is represented by:

$$OP_{FB}(n) = \sum_{j=1}^{K_2} b_j^* d(n-j-D); \quad n = 0, 1, 2, \dots \quad (3.6)$$

where D is the delay of the system and $\{b_j\}$ are the K_2 number of complex valued tap weight coefficients for feedback filter.

SYMBOL BY SYMBOL DETECTOR: The input to the detector at a discrete time n is given by the combination of two linear convolution sums defined by the equations (3.5) and (3.6) i.e.

$$y(n) = OP_{FF}(n) + OP_{FB}(n) \quad (3.7)$$

The detector decides its output based on which one of the four possible symbols in the transmitted signal alphabet is closest to $y(nT)$, henceforth denoted $y(n)$.

PERFORMANCE INDEX: Using mean-square-error as the performance index at the output of the equalizer, the tap weights of both filters are adjusted to minimize the mean square of the estimation error $e(n)$ i.e. minimize

$$MSE = J = E|e(n)|^2 = E|d(n-D) - y(n)|^2 \quad (3.8)$$

where $d(n-D)$ is the desired signal.

Minimization of mean square error tends to reduce the bit error rate. The quantity $E[|e(n)|^2]$ is the power or variance of the error signal. If this quantity is minimized, there are less chances of decision errors at the output of the detector [34].

For an optimum set of taps in the feedforward and feedback filters, the MSE J is minimized to the minimum mean square error (MMSE), J_{min} , given by [30] [35]:

$$MMSE = J_{min} = \sigma_d^2 - \mathbf{P}^H \mathbf{R}^{-1} \mathbf{P} = \sigma_d^2 - \mathbf{P}^H \mathbf{W}_{opt} \quad (3.9)$$

where σ_d^2 is the variance of the desired response (assumed to be 1 in our analysis) and $\mathbf{W}_{opt} = \mathbf{R}^{-1} \mathbf{P}$ is the set of optimum tap weights for feedforward and feedback filters. Here, vector $\mathbf{P}$ is the cross correlation of the tap inputs of forward and feedback filters and the desired signal, and $\mathbf{R}$ is a correlation matrix of tap inputs of forward and feedback filters. This evolution to MMSE involves the evaluation of a set of matrices and equations and have been detailed in the Appendix A.

3.4 Adaptive Algorithms

In most communication systems that employ equalizers, the channel characteristics are unknown a priori and, in many cases the channel response is time-variant. In such a case, the equalizers are designed to be adjustable to the channel response and, for time-variant channels, to be adaptive to the time variations in the channel response. Typically, an equalizer has to be adapted to the unknown multipath channel by sending a known sequence of symbols, termed as training sequence, prior to the actual data transmission so as to let it learn about the unknown channel. After the training mode, the equalizer is shifted to decision directed mode. In a packet based modem, such as for access to ATM networks as envisioned for LMCS/LMDS applications, in order to increase the efficiency of transmission, the overhead due to the training part should be minimized [36] which requires the use of fast convergent algorithms. The number of iterations required to achieve a particular value of output SNR or to converge to a particular value of mean-square-error (MSE) can be used as a yard stick to measure the conver-

gence speed of an algorithm. The performance measure generally considered is the speed of convergence of the MSE (or output SNR) at the slicer output, the final value of the MSE achieved and the stability and the computational complexity of the algorithms. A number of adaptive algorithms are available in the literature e.g. Least-Mean-Square (LMS), Recursive Least-Squares (RLS) [30] [34] [35], fast Kalman algorithm [40] and lattice RLS algorithm [41] [42], Square root RLS and modified square root RLS [43], Block adaptation (with direct matrix inversion (DMI)) using covariance method [35] [45] and Cyclic equalization [44], and the complexities and the adaptation speeds of some of them are compared in literature e.g. [36] and [37].

The LMS algorithm is relatively simple and very stable, provided the step size is chosen properly, but it takes significantly longer time for convergence than the RLS, Fast Kalman and Lattice RLS algorithms. Its convergence speed is dependent to a large extent on the channel characteristics whereas the performance of RLS and Lattice RLS algorithms is almost independent of the channel. The RLS algorithms eventually tend to be unstable when implemented in a finite precision environment, the fast RLS being most prone to instability of all. The convergence speeds and the computational complexities of some of these algorithms are compared in Chapter 5 in the LMCS context. Based upon the results obtained in that chapter, we decide the best suitable for LMCS applications.

3.5 Time variations of the channel

Delay spread and coherence bandwidth define the time dispersive nature of the channel but they do not offer any insight into its time varying nature. In the LMCS/LMDS scenario, although the transmitting and receiving antennas do not involve any

mobility, which is in contrast to the mobile radio scenario, the channel between base station and customer antenna may change owing to the movement of the objects in the channel such as wind induced fluttering of tree leaves or motion of vehicles etc. These time variations can be characterized by the doppler spread and coherence time, and these two parameters are used in Chapter 5 to analyze the time variability of the radio channel in the LMCS context, and hence to decide whether it is necessary to employ some algorithm in the equalizer to constantly track the channel once the training of the equalizer tap coefficients is done.

The doppler spread B_D is a measure of the spectral broadening caused by the time rate of change of the channel and has maximum value defined by $f_m = v/\lambda$ where v is the velocity of the moving object in the channel and λ is the wavelength of the electromagnetic wave at the frequency of transmission. The coherence time, T_c , is a statistical measure of the time duration over which the channel impulse response is essentially invariant and may be related to the maximum doppler shift through [29]

$$T_c = \sqrt{9/(16\pi f_m^2)} \quad (3.10)$$

If the baseband signal bandwidth is much larger than the doppler spread, then its effects are negligible at the receiver or in other words, if the reciprocal bandwidth of the baseband signal is greater than the coherence time, then the channel will change during the transmission of the signal and the received signal will be distorted. So these criteria were used in deciding if we need to track the channel.

3.6 Maximum-Likelihood Sequence Estimation (MLSE)

Non-linear equalization techniques are preferred for channels with severe amplitude distortion because of the potential of linear equalizers to enhance the noise in such cases, in the vicinity of the spectral nulls. One of the nonlinear equalization technique discussed in the previous sections, decision feedback equalization, decodes the output on a symbol-by-symbol basis. Another equalization scheme to fall under the nonlinear category for combating the effects of ISI is MLSE which produces the data decisions at its output that are most likely to be correct, utilizing the entire received signal sequence and is implemented with a recursive processor called Viterbi Algorithm (VA) [46] [30] [34].

The limitations to the applicability of MLSE to practical receivers is the computational complexity associated with the VA especially when the delay spread of the channel is large [46]. For a channel with dispersion of M symbol intervals, its computational and storage requirements are proportional to L^M (4^M for QPSK) operations per symbol, where L is the symbol alphabet size. MLSE can be viewed as a problem in estimating the state of a discrete time finite state machine. The channel state at any instant of time is estimated by the receiver based on M most recent input samples. In such a case, the number of possible state vectors is L^M . With retention of a M -element state path trajectory for each possible current state, in which each element requires $\log_2 L^M$ bit storage capacity, the total storage capacity not including the computational burden, is substantial, being $L^M M \log_2 L^M$. Since the channel has L^M states, a L^M trellis (which is an ensemble of all the possible states at all the possible times) is used by the receiver to model the

channel over time. Also, MLSE requires knowledge of the channel characteristics in order to compute the matrices for making the decision. In the VA, the number of transition matrices which must be computed at each step is L^{M+1} . Hence we find that it is not an attractive choice for channels with long impulse response owing to its huge computational load unless some means are employed to truncate the channel impulse response without incurring too much performance loss i.e the number of states may be so large that a reduced state system is the only implementation option.

The structure of the receiver for the maximum likelihood sequence estimation of the entire transmitted sequence perturbed by finite ISI and AWGN consists of a whitening matched filter, a symbol rate sampler, and the recursive VA processor [46] [47]. The combination is simple to implement and feasible if the channel impulse response is not too long. So, in practical situations, a near optimum solution is to use a linear equalizer to shape the channel IR to some desired channel whose overall response is short and is similar to the channel spectrum and then apply the VA to the outputs of this channel [48] [49].

3.6.1 Performance of Viterbi Algorithm

The performance of the maximum-likelihood viterbi sequence detector depends principally on the minimum distance d_{min} , between the allowed transmitted sequences. Forney in [46] has given that the symbol error probability for MLSE using VA is upper-bounded and estimated by

$$Pr(e) \leq KQ[d_{min}/2\sigma] \quad (3.11)$$

where d_{min}^2 - minimum energy of any non zero signal, σ^2 - the spectral density of noise, and K - a small constant are defined in [46]. In Chapter 5, we have estimated the error bounds for VA using a shorter IR and compared the performance with that of DFE.

Chapter 4

EHF MEASUREMENTS AND DATA PROCESSING

4.1 Introduction

During the past 4-5 years there has been a significant interest in the provision of broadband communication services through wireless systems in the EHF portion of the radio spectrum. More recently this interest has shifted from indoor broadband system design to outdoor broadband wireless communications called Local Multipoint Communication services as described before.

In this chapter, we briefly describe a propagation measurement campaign aimed at measuring the impulse responses in the outdoor to characterize the radio channel in the band of frequencies allocated for LMCS/LMDS system operation. For communications in the troposphere, radio channels are linear and can be modeled as linear filters. Regardless of the type of system, if wide bandwidths are required, it is necessary to first carry out these kinds of experiments, as briefly described here, to determine and characterize the equivalent impulse response of such linear channel models before the design of proposed systems can be carried out. These impulse response characteristics are ought to influence many system design parameters such as modulation schemes, equalization, coding and diversity etc.

While these measured impulse responses are currently being investigated by various researchers from different angles, these were also used in this thesis to analyze the

amount of multipath propagation under the given set of system operating conditions (e.g. use of highly directional antennas) and to investigate, based upon these impulse responses, appropriate equalizer structures suitable for LMCS system modems.

4.2 EHF Measurement Details

A set of EHF measurements were conducted under the project, sponsored by CITR, led by Dr. R Bultitude of CRC along with a team comprising of researchers from CRC, Carleton University and Industrial partners like Nortel, Newbridge and Stentor etc. with the objective of acquiring knowledge concerning radio propagation at EHF in typical urban commercial and urban residential areas and to help model the outdoor radio channel at these frequencies.

The experiments were carried out to derive data for statistical analysis of a host of propagation related phenomenon such as determination of Receiver power versus range; determine multipath characteristics; distribution of received power at several ranges; cross polarization isolation (XPI) for adjacent cell or adjacent sectors; presence of specular reflections as a function of antenna height; area coverage and subsequent determination of coverage probabilities in specific operating scenarios. This involved measurements of radio channel characteristics between a typical cell site transmitter and many receive locations in areas of Parkwood Hills, Ottawa and city of Kanata, a suburb of Ottawa.

While the Parkwood Hills is primarily a residential area, the sites surveyed in Kanata were a mix of residential and open industrial park type. In the absence of access to ideal locations for measurements which would ideally have been the roof tops where

the subscriber antennas are likely to be located, the measurements were carried out at three to four sub locations at each site which included along roadways, in the drive ways, near houses, etc. The problem here is that the clutter surrounding these locations and that surrounding the buildings owned by the subscribers, as well as a significant portion of the propagation path to the cell station from these sites may sometimes be different e.g. different path losses, multipath spread, XPI etc. At each location about three different antenna heights were tried so as to derive data about the optimum receive antenna heights. The primary purpose of the measurements was to:

- Derive statistics on spatial variations of the received power and propagation loss, foliage attenuation and loss due to diffraction around the building edges, wind and traffic induced temporal variations of channels.
- Derive statistics on multipath dispersion due to reflection and scattering. With the use of narrow beamwidth fixed transmit and receive antennas, it is anticipated that the multipath propagation should not be very severe but to be sure, one of the aims of the measurements was to verify this fact. These statistics will help to determine the equalization requirements of the system and have been used in this thesis.
- Estimate XPI (Cross Polarization Isolation) which is of interest as many of the currently proposed systems plan to operate adjacent cells or adjacent sectors in the same cell on the same frequency but with orthogonal transmit polarization to prevent co-channel interference. And it is known that reflection from structures, scattering from foliage and rain drops etc. can reduce the isolation.
- Derive data on variation of the received power as a function of angle of arrival.

- Determine the percentage of area coverage.

Both narrowband as well as broadband measurement were carried out.

In Parkwood Hills, three different types of transmitters from roof top of Chesterton towers were tried. Transmitter 1 and 2 were for CW output while transmitter 3 was for wideband measurements. Transmitter 1 (Transmitter 2) had a vertically (horizontally) polarized horn antenna with antenna gain of 18 dBi, elevation beamwidth ± 11.5 degrees, azimuth beamwidth of ± 8 degrees. The output power was +23.8 dBm. For wideband measurements, a 63 bit pseudo random noise code at 40 M bps was BPSK modulating an RF carrier at 950 MHz which was then upconverted to an IF of 5.2 GHz and then to final frequency of 29.5 GHz before transmission from transmitter antenna. Antenna 3 was a vertically polarized horn antenna with antenna gain of 18 dBi, and beamwidths same as transmitter 1. The transmitter 3 output power was 24 dBm.

The transmitter was always located at the same spot in the roof of Chesterton Towers apartment. The transmitter direction was occasionally changed so as to keep the measurement site within the beam. The receiver antenna is a vertically polarized horn antenna with beamwidth of ± 6 degrees in both elevation and azimuth, with gain of 23 dBi or vertically polarized cassegrainan reflector antenna with beamwidth of ± 1 in both elevation and azimuth and antenna gain of 36 dBi.

The receiver antenna was mounted on a turn table which was strapped to the bucket of a bucket truck. The receive antenna could be adjusted at various heights and was capable of being given one full azimuth rotation to scan for any specular reflections in addition to the direct signal. During the wideband probe, the measurement method

involved transmitting a BPSK signal at 29.5 GHz modulated by a 63 bit long PN sequence generated at 40 Mbps and then receiving the signal by employing a quadrature (I and Q) receiver. The demodulated quadrature signals were sampled at a rate of 100 M samples per second which corresponded to 2.5 samples per bit or 157.5 samples per 63 bit long PRBS stream. The sampling process (either time or angle) was controlled by software. The PRBS rate of 40 Mbps resulted in a resolution of 25 ns in the measured impulse response echoes. Hence, following the I-Q demodulation process at the receiver, the resultant complex envelope of the received signal was sampled and stored. Subsequently, the sampled complex envelope was crosscorrelated with a replica of the transmitted sequence, resulting in an estimate of the low pass equivalent of the radio channel complex impulse response.

During the measurements, a set of preassigned locations were surveyed. At each site, measurements were done at three or four closely located sub-sites, labelled as a, b, c and so on, to assess the variation of various phenomena such as the received power, multipath spreads, cross polarization isolation (XPI) etc. within a close distance. The receive antenna heights were varied at each receive location and at each height, one complete azimuth rotation in steps of six degrees was given to measure the receive power as a function of the angle of arrival. For example, at Parkwood hills mainly the following type of sub sites, along with different antenna heights, were surveyed:

- Above Roadways, antenna heights: 15 ft., 3 ft. + roof height, 5-10 ft. + roof height
- Above driveways of the residence, 15 ft., 3 ft. + roof height
- Above or near roof tops, 3 ft. + roof height, 10 ft. + roof height

In Kanata, an open business and residential area where the measurements were done employing cassegrain receive antennas with beamwidth of ± 1 degree in azimuth, these sites and antenna heights were:

- Above roadways, antenna heights: 23 feet, 27 feet, 29 feet and 30 feet.

Not all the sites had line of site (LOS) path from the measurement site to the transmitter. In Kanata, many measurements were done just looking towards the transmitter antenna (may not be LOS) i.e. with no azimuth rotation of the receive antenna and at some locations the receive antenna also had an azimuth rotation in steps of 2 degrees. This thesis makes use of the impulse responses from both Parkwood Hills and Kanata, after applying a threshold to reduce the impact of noise as briefly described in the next section, to analyze the impact of multipath propagation on system design from an equalizer point of view.

4.3 Wideband Data Processing

Three delay profiles were recorded at each location for analysis. The quadrature signals were sampled 2.5 times per resolution chip resulting in 2048 complex samples (over 6 sequence lengths of 63 bit PRBS stream) per impulse record. From the samples of the received complex envelope, the channel impulse response estimates were obtained after cross correlating the samples with a replica of the original transmitted sequence. A typical delay profile i.e. magnitude of the received complex impulse response (normalized to the maximum amplitude sample) obtained in such a manner is shown in Figure 4.1. The title for the Figure 4.1, 05d28 at 6 deg, represents a particular location of the measurement and the antenna height used. '05' is the location number, 'd' is some

FIGURE 4.1 A typical delay profile before thresholding

sublocation at this site and ‘28’ represents the antenna height in feet. “6 deg” represents 6 degrees from LOS. A fairly sophisticated thresholding technique, described in [28] was applied to each of the recorded impulse responses, in order to reduce the impact of noise on the measured data and to extract the valid impulse response echoes (the process is called thresholding). In this algorithm, the background thermal noise is assumed to be gaussian distributed with variance σ_N^2 and it is required to estimate its level. The amplitude of this noise is assumed to be Rayleigh distributed.

The value of σ_N^2 could be estimated as one half of the mean square value of the sample amplitudes if all of them were due to noise. However, valid echoes along with any impulsive interference can significantly distort the mean value owing to their comparatively large amplitudes. So, in order to mitigate the effect of a few large amplitude data points, σ_N^2 is evaluated based on median of the impulse response samples. The algorithm [28] calculates the noise threshold as follows:

The probability that the noise amplitude N , assumed to be Rayleigh distributed in the absence of any signal, exceeds some level ζ is

$$P(N > \zeta) = \exp(-\zeta^2/2\sigma_N^2) \quad (4.1)$$

If σ_m is the median level of the samples, setting $\zeta = \sigma_m$ and equating (4.1) with 0.5, we get $\sigma_N = 0.85\sigma_m$ i.e. the thermal noise standard deviation is 1.41 dB below the median level in the absence of interference. The purpose here is to set a flexible threshold to achieve a constant false alarm rate which is independent of the signal to noise ratio.

Setting the threshold $\zeta = \eta\sigma_N$, we get from equation (4.1),

$$P(N > \zeta) = P_f = \exp\left(\frac{-\eta^2}{2}\right) \quad (4.2)$$

where P_f is the probability of false alarm per sample i.e. the probability that the noise exceeds the threshold in any given sample without being the valid echo. From equation (4.2), $\eta = \sqrt{-2(\ln(P_f))}$.

Probability of false alarm is further improved upon by noticing that if the echo is present, consecutive measured delay profiles will be highly correlated but there will be little correlation if just noise is present. It is most probable that large noise samples will be present just in one bin, while there is high probability that samples with valid echoes will be present in at least two bins. Out of the three delay profiles recorded at each location, it is most unlikely that any noise sample with large amplitude will be present in all the three delay profiles at the same delay or in two out of three delay profiles. However, in the case of valid echoes, with high probability this condition is true. So a large sample is declared to be valid if and only if 1) the threshold is exceeded at that point in all the three recorded profiles and 2) at least one of the neighboring points satisfy the condition 1). And, it is very unlikely that a noise sample will satisfy these conditions in all the three recorded delay profiles. Using this criterion, the false alarm probability is given by,

$$P_{fa} \approx 2(P_f^3)^2 = 2\exp(-3\eta^2) \quad (4.3)$$

So using equation (4.3), for a particular chosen probability of false alarm (10^{-2} in our processing), the value of η can be known which gives the threshold ζ relative to the largest echo component in the delay profile. So the resultant delay profile, as shown in

following Figure 4.2 , is a truncated version of the recorded profile, in which the probability of a noise spike being interpreted as valid echo has been greatly diminished. In the shown delay profile, the algorithm sets a relative threshold level of about 28 dB down with respect to the maximum amplitude sample and signal below that level is truncated.

FIGURE 4.2 Delay profile of Figure 4.1 on page 38 after thresholding

Chapter 5

ANALYSIS OF MEASURED BROADBAND CHANNEL IMPULSE RESPONSES AND DISCUSSION

5.1 Introduction

In this chapter, the results from the analysis of the measured broadband wireless channel impulse responses have been presented. These impulse responses were measured in Parkwood Hills and Kanata as described in Chapter 4. To evaluate the optimum equalizer structure at different transmission rates, we consider transmitting data at different symbol rates of 10 Mega symbols per second (Msps), 25 Msps and 50 Msps using QPSK modulation format. The propagation environment is the outdoor wireless channel. The results from the evaluation of Parkwood Hills data and Kanata data has been presented in separate sections of this chapter.

The transmit and receive filters, as shown in the system model in Figure 3.1 in Chapter 3, are square root raised cosine filters with roll-off factor of 0.3. It is known that a DFE with symbol spaced feedforward filter is extremely sensitive to the timing error in sampling at the output of the equalizer [30] [37]. A fractionally spaced equalizer (one which is based on sampling the incoming signal at least at the nyquist rate) can alleviate this problem [37] [38]. In our analysis, we have used both symbol spaced and FS-DFE. As stated before, Mean-square error or the output SNR at output of the slicer has been used as the performance index. Having known the equalizer complexity, a study of vari-

ous adaptation algorithms has been carried out in terms of their speeds of convergence and the computational burden associated with each of them, so as to propose the most suitable for the LMCS systems. Also we have looked into whether it is required to track the channel to take care of temporal variations of the channel impulse response due to movement of the tree leaves or vehicles etc. Using the results obtained for minimum mean square error (MMSE) through decision feedback equalization, we have tried to estimate the complexity involved if optimal equalization technique of Maximum Likelihood Sequence Estimation using Viterbi algorithm is employed instead. The upper bounds for the error rate performance for both the schemes have been plotted and compared.

5.2 Parkwood Hills Impulse Responses Analysis Results

For the Parkwood Hills, impulse response data from nine main locations has been analyzed. At each main location, the measurements were done at approximately four to five sub locations (this number variable from location to location); so here, we have data from a total of 52 sites and at each measurement site, as detailed in Chapter 4, the receive antenna was given a full rotation of 360 degrees in azimuth, in increments of 6 degrees, capturing the received signal for each setting of the receiver antenna, thus resulting in a total of 3120 complex impulse responses for the broadband channel.

Not all the locations had line-of-site links with the transmitter; useful reflections were found in other directions in many cases. Usually, the strongest signal was detected in directions of 0° or $\pm 6^\circ$, where 0° means looking directly towards the transmitter antenna. Figure 5.1 shows the histogram of the placement (angle) of receive antenna

FIGURE 5.1 Histogram of angle of occurrence of strongest signal at all sites

with respect to the transmitter antenna for the strongest signal for all the sites (at one of the sites no signal peak was left after thresholding algorithm was applied). It is known that at high data rates such as the ones perceived for LMCS applications, multipath delay spread is a major transmission problem leading to ISI. Also the delay spread varies with the environment and the characteristics of the transmitter and receiver antennas. Since for the present set of measurements, very narrow beamwidth transmitter and receiver antennas were used, the multipath spread of the received signal is expected to be relatively small.

Figure 5.2 shows the CDF of the multipath spread for receive antenna placements of 0° and $\pm 6^\circ$, and Figure 5.3 shows the CDF of the multipath spread for those directions in which the strongest signal (also called “best directions” here for which the equalizer results are reported) was received at the respective sites. The 90th percentile equals 90 ns in both the plots. In Figure 5.3, which is for strongest signal receive directions, there are only two sites (05d28 at 0 deg. and 08d15002 at 174 deg.) out of the 52 measured sites which have relatively larger multipath spread due to the presence of much delayed signal components coming in (A chimney at the top of one building near the receiver location was thought to be the cause for a possible signal reflection at 05d28, although there was LOS available, and in the second case of 08d15002, there was no LOS available and also, at heights of just 15 feet above the ground level, the receive antenna can not clear the roof tops and is surrounded by a number of houses which in addition to blocking the LOS, also become potential source of signal reflections). Barring these two cases, the rest of the impulse responses in Figure 5.3 are well behaved.

FIGURE 5.2 CDF of multipath spread for receive directions of 0° and $\pm 6^\circ$

FIGURE 5.3 CDF of the multipath spread for best receive directions at all sites

Figure 5.3 represents data for 51 sites (instead of 52) as no useful signal was left in one case after thresholding. Small multipath spread implies that at some of the lower transmission rates, the system may not require any equalizer at all. Here in this chapter, we present the results obtained for the equalizer structure (i.e. the total number of taps in the feedforward and feedback filters for optimum performance) for one particular measured impulse response, at three different symbol rates, and the results for the other impulse responses are summarized in Appendix B. For the analysis, the impulse responses have been supplied to the results of Appendix A. The square root raised cosine filters spanning 15 symbols have been designed with a rolloff factor of 0.3. The noise variance level for the evaluation of the all the measured 60 responses corresponding to a particular site (due to antenna rotation in azimuth) has been decided based on the energy in the strongest signal in some direction at that site i.e. the energy in all the impulse responses for the site under consideration was calculated and the maximum of them was identified to fix the noise level 20 dB below that. Then this noise level is used to analyze all the 60 responses of that particular location.

Hence, the noise level used varies with the site and is not a fixed quantity used throughout but the input SNR corresponding to the maximum energy impulse response is always 20 dB (this is just an assumed SNR). The aim here is to get an idea about the size of the equalizer treating each of the sites independently (i.e. assuming that at least the maximum energy signal at each location is of workable quality or level). The noise variance (20 dB down) varies with the symbol rate.

The forward filter of the DFE has been designed to be either a symbol spaced or $T/2$ spaced linear transversal filter (for more than 2 samples per T , no improvement

would expected, given the assumed rolloff factor of 0.3) and the feedback filter is a symbol spaced linear transversal filter. We determine the effect on the minimum-mean-square error (MMSE) due to the change in the feedforward and feedback filter taps for QPSK modulation. Performance can also be measured by the output signal-to-noise ratio, defined as:

$$SNR_{out} = 10\log_{10}\left(\frac{1}{MMSE}\right) \quad (5.1)$$

where received baseband average signal power is assumed to be unity. Note that the maximum possible output SNR is 20 dB, given the maximum 20 dB input SNR assumed.

5.2.1 10 Mega symbols per second

To explain our results, here we present one typical impulse response from one of the measurement sites at one particular receive angle and show the results of our analysis on the equalization requirements for that impulse response at three different symbol rates. Figure 5.4 displays one of the impulse responses, obtained after noise thresholding was applied, and was measured at site 11b23002 when the receive antenna was at 6 degree. For this particular case, the total energy in this impulse response captured at six degree exceeds that in any other impulse response measured at the same site (The impulse response corresponding to 0 deg has slightly lower level). Here, in the designation '11b23002', '11' represents a particular location, 'b' stands for a particular measurement site (or a sub location) at this location, '23' represents the height of the receive antenna from ground level in feet and '002' is some index.

This impulse response represents a multipath spread of 90 nanoseconds and was chosen because it was part of the worst case multipath spreads (except for two of the

FIGURE 5.4 Sample level Vs. delay for channel at 6 degree at site 11b23002

FIGURE 5.5 Contour plots of SNR_{out} variation with feedforward and feedback filter taps, at best sampling phase at 10 M symbol per second

FIGURE 5.6 SNR_{out} variation with sampling phase for channel impulse response of Figure 5.4 at 10 M symbol per second

cases mentioned before which had some delayed multipath component too but we don't consider them as normal behavior). These cases with delayed multipath components are found to be addressable but only by using a DFE with more of taps which can take into the account the delay of the multipath components (as has been observed in the analysis of these impulse responses with delayed components); or instead of using many taps we can implement a delay in the feedback section to take care of the delayed multipath. Since at each main location there were a number of sub locations where the receive antenna was positioned and different antenna heights were tried, it is possible that there is one sublocation at the same subscriber location (or there is a particular antenna height) where there is no delayed multipath component. So instead of using a big DFE we can make adjustments in the antenna placement or its height to take care of this problem. We will see in Appendix B that for each of these two sites, there are other closely placed sites which do not have a too long multipath spread and have simple equalization requirements.

For the 10 mega symbols per second case, the symbol duration is 100 ns. At this symbol rate, as we see in Figure 5.5 , the system may not require an equalizer but there is definitely an improvement in the output SNR, as evident in the contour plots, by using an equalizer over no equalizer ((1,0) DFE). Here, notation (a,b) represents 'a' taps in the forward filter and 'b' taps in the feedback filter of DFE. This figure represents the contour plots of the variation of output SNR (which is signal to mean-square error ratio) with change of taps in the feedforward filter and feedback filter for a DFE which tries to equalize the impulse response shown in Figure 5.4 . DFE used here employs symbol spaced forward filter. We see from this plot that with no equalizer, if we are sampling at

right sampling phase, we can get an output SNR of 18.14 dB, this corresponds to the lower right corner of Figure 5.5 . Using one feedback tap, the system performance improves to 19.17 dB, which is close to the upper bound or the matched filter bound (which is equivalent to the transmission of an isolated pulse in the absence of any interference and ISI) of 20 dB on SNR_{out} for this channel. Hence we conclude that for this symbol rate the system may not need an equalizer for satisfactory output SNR, but the contour plots also suggest that use of an equalizer does improve the system performance, for example with three forward and one feedback tap we get 19.71 dB as output SNR. Hence, there is a trade-off between the performance we require and the complexity we can afford. In the contour plots, the contours are level lines in the units of output SNR (which is dB). In Matlab, CONTOUR(Z) plots a contour plot of a matrix Z treating the values in Z as heights above a plane, so these level lines can appear as a staircase type of structure. The contour plot algorithm examines the range of the SNR_{out} values to be plotted, then divides the total region along height into equispaced levels and draws lines at those levels.

The SNR_{out} values corresponding to different combinations of feedforward filter and feedback filter taps lie on these lines or the planes between these lines. In Figure 5.5 , the contour lines are each 0.2 dB apart and the topmost of them lies at level of 19.8 dB as shown by a label. The level of other lines can be calculated in decrement steps of 0.2 dB. When the number of feedforward taps become greater than or equal to 4, SNR_{out} exceeds 19.8 dB, and achieves a maximum of 19.98 dB at (8,1). It has been observed that the addition of more than one feed back tap does not have a noticeable effect on the output SNR. The feedforward filter for contour plots of Figure 5.5 is symbol

spaced where we take one sample per baud period at its input. Figure 5.6 highlights the importance of sampling the output at correct phase in such a case. The SNR_{out} is maximized when we sample at optimum sampling phase of $\tau_0 = \frac{3T}{10}$ while any sampling phase error deteriorates the performance. In this figure (1,0) represents an equalizer with one forward tap and no feedback tap; (1,1) represents a DFE with one forward and one feedback tap. The same representation is used throughout in the explanation of the results. It can be seen in the results that addition of one feedback tap does improves the sensitivity of the output SNR w.r.t. some sampling phases but it has been observed that the addition of more feedback taps do not improve this noticeably. In the analysis of the 05d28 case (having delayed component by about 300 ns) at 10 Msps, symbol spaced case, it required (3,3) DFE to achieve SNR_{out} performance of 19.24 dB at the best sampling phase, for (3,2) it gives 17.68 dB and for (3,1) - 16.79 dB. For a fractionally-spaced DFE (FS-DFE), the input to the feedforward filter is sampled at intervals of $T' = T/2$ whereas the feedforward filter output is still sampled at intervals of T since data decisions are made at symbol intervals. The significant property of the fractionally spaced equalizer is that it can compensate for any delay distortion without noise enhancement.

Next we present the results obtained for a FS-DFE for the same symbol rate of 10 M symbols per second. Figure 5.7 shows the variation of output SNR with sampling phase for this case. Here the input to forward filter of DFE is sampled twice every baud period though the output is still sampled once per baud interval. It is evident from this plot that the performance does not change too much with the change in sampling phase, as it should be when the forward filter is fractionally spaced. This figure corresponds to

the case when the forward filter is just one symbol duration long but it has two taps in it, that is why the notations (2,0) and (2,1) appear in the plot. This figure underlines the advantage of FS-DFE over symbol spaced DFE due to relative independence of performance from sampling phase. Figure 5.8 is for contour plots of the output SNR achieved in this case. The horizontal axis in this case is for length of feedforward filter in symbol durations and since there are two taps per symbol period so the horizontal axis gets multiplied by 2 when interpreted in terms of number of taps. In this case, output SNR of 19.24 dB and 19.86 is achieved with forward filter of one symbol duration (2 taps) and no feedback tap and then one feedback tap, respectively. The SNR_{out} does improve with addition of more feedforward and feedback taps. Also we found out that for a particular number of feedforward and feedback taps, the FS-DFE performs better than the baud rate DFE. Generally, the price for the better performance of FS-DFE compared to the baud rate DFE is its higher complexity.

From Figure 5.5 - Figure 5.8 , we conclude that for symbol rates of the order 10 M symbols per second, the system may not require any equalizer though the use of one will enhance the performance. And also that the timing recovery system needs to provide an accurate estimate of the proper sampling epoch to the receiver sampling system, a condition which gets relaxed when we are using FS-DFE.

FIGURE 5.7 SNR_out variation with sampling phase for channel impulse response of Figure 5.4 with FS-DFE at 10 M symbols per second

FIGURE 5.8 Contour plots of SNR_{out} variation with forward filter and feedback taps, at 10 Msps with forward filter T/2 spaced

5.2.2 25 Mega symbols per second

Here we increase the symbol rate from 10 Msps to 25 Msps to observe the change in the equalization requirements of the system. Figure 5.9 shows the contour plots for the variation of output SNR as a function of feedforward filter taps and feedback filter taps when the symbol rate is 25 M symbols per second and the feedforward filter is symbol spaced. The channel here again corresponds to the impulse response of Figure 5.4 . The SNR_{out} in this particular figure changes from 13.23 dB to 19.83 dB, with the former value corresponding to the lower left corner and the latter value corresponding to the upper right corner of the plot, when the feedforward filter and feedback filter taps are varied from (1, 0) to (10, 10). The contour levels vary in steps of 1 dB.

From this plot it is observed that for symbol rate of 25 M symbols per second, an equalizer with 3 forward and 1 feedback tap seems an optimum choice which gives an output SNR of 18.0 dB (as shown in Figure 5.9 with an arrow pointing to (3,1)) for an input SNR of 20 dB although the performance does improve by increasing the forward taps, for example, (4,1) gives 18.01, (5,0) gives 18.88, (5,1) gives 19.13. Any further increase in the feedback taps beyond 1 does not result in any appreciable improvement in the performance. So at 25 M symbols per second, if we do not use an equalizer there is a considerable degradation in the performance (with output SNR measuring 13.23 dB).

Figure 5.10 underlines the critical requirement of sampling at correct phase when using symbol spaced forward filter. This plot is for (3,0) and the optimum (3,1) case, where (3,0) has been plotted to highlight the advantage of using one feedback tap.

Next we change the symbol spaced forward filter to one with T/2 spaced taps. Figure 5.11 shows the system performance with change in the output sampling phase.

FIGURE 5.9 Contour plots of SNR_{out} variation with feedforward and feedback filter taps, at best sampling phase for 25 M symbols per second

FIGURE 5.10 SNR_{out} variation with sampling phase for channel impulse response of Figure 5.4 for 25 M symbol per second

It is observed that when the forward filter is two symbol durations or longer (the number of taps in the filter would be two times the symbol spans), the output SNR is approximately independent of the sampling phase. In this figure we also showed the case when the span of feedforward filter is insufficient (curves corresponding to (2,0) and (2,1), in which case the forward filter is just one symbol duration long) and the output SNR does change with the sampling phase.

Figure 5.12 shows the contour plots for FS-DFE at 25 M symbols per second. These plots corresponds to $\tau_0 = 0T$. Here, though we see that for (2,0), i.e. the bottom right corner of plot, the SNR_out value is 19.38 dB but we don't consider that value as the optimum case, as with the change of sampling phase that value changes appreciably, whereas for taps equal to or more than (4,0) or (4,1) (i.e. for a forward filter of two symbol durations or long), SNR_out becomes independent of the sampling instants. The SNR_out values corresponding to (4,0) and (4,1) have been shown with arrows in Figure 5.12. Here also, comparing the results for symbol spaced DFE and FS-DFE, we find the latter superior in performance over the other for same number of taps.

FIGURE 5.11 SNR_{out} variation with sampling phase for channel impulse response of Figure 5.4 for FS-DFE at 25 M symbols per second

FIGURE 5.12 Contour plots of SNR_{out} variation with forward filter and feedback taps, at 25 Msps with forward filter T/2 spaced

5.2.3 50 Mega Symbols Per Second

In the next step, we increased the symbol rates to 50 M symbols per second and examined the equalization requirements for this transmission rate. From the contour plots in Figure 5.13 , we find that with symbol duration of 20 ns at 50 M symbols per second and multipath spread of 90 nanoseconds, the symbol spaced DFE needs to have at least 4 feed forward taps and one feedback tap to achieve an output SNR close to 18 dB when sampling at the best sampling phase of $\tau_0 = \frac{5T}{10}$. Here also any addition of feedback taps beyond 1 does not result in an appreciable performance enhancement, though for a forward filter of 3 taps any increase of feedback taps beyond 2 results in very small change in output SNR. To give an idea we can summarize some of the results as: (3,1) = 16.19 dB, (3,2) = 17.84 dB, (4,1) = 17.96 dB, (4,2) = 17.97 dB, (5,1) = 17.98 dB, (5,2) = 17.99 dB, (6,1) = 18.03 dB, (6,2) = 18.04 dB. Hence (4,1) appears to provide a reasonable performance / complexity trade-off for a symbol spaced DFE. Figure 5.14 shows the performance change with sampling phase.

Figure 5.15 and Figure 5.16 portray the performance variation with sampling phase and the contour plots respectively at 50 M symbols per second for FS-DFE. From Figure 5.15 , we find that for DFE with forward filter of 3 symbol durations i.e. six forward taps and one feedback taps, the output SNR changes by a small amount with sampling phase whereas when the forward filter spans four symbol durations (i.e. 8 taps) or more the performance is independent of the sampling phase. So we recommend a DFE with 4 symbol durations for forward filter and one feedback tap i.e (8,1), when we sample every $T/2$, to achieve improved performance and independence from sampling phase

FIGURE 5.13 Contour plots of SNR_{out} variation with feedforward and feedback filter taps, at best sampling phase for 50 M symbols per second

FIGURE 5.14 SNR_out variation with sampling phase for channel impulse response of Figure 5.4 at 50 M symbol per second

FIGURE 5.15 SNR_out variation with sampling phase for channel impulse response of Figure 5.4 for FS-DFE at 50 M symbols per second

FIGURE 5.16 Contour plots of SNR_{out} variation with forward filter and feedback taps, at 50 Msps with forward filter T/2 spaced

in contrast to symbol spaced DFE. Figure 5.16 gives the output SNR for various combinations of feedforward filter taps and feedback filter taps. Here also the horizontal axis represents the forward filter length in symbol durations, so the number of taps in the filter are twice as many. This figure corresponds to sampling phase of $\tau_0 = 0$ in which case, although the SNR_out value of 18.14 dB for (6,1) is very close to 18.24 dB for (8,1) but we have recommended (8,1) because for this tap combination SNR_out does not vary with sampling phase.

5.3 Kanata Impulse Responses Analysis Results

The measurement setup for Kanata measurements has been described in Chapter 4, where the transmitter antenna was 90 degree sectored horn with 2.5 degree in elevation and receiver antenna was a vertically polarized cassegrain reflector antenna with beamwidths of +/- 1 degree in elevation and +/- 1 in azimuth. For the Kanata measurements, the measured impulse responses evaluated here belong to those resulting from receive antenna direction of 0 degree i.e. looking towards the transmitter. Again, not all the sites had line-of-sight (LOS) link. A total of 111 impulse responses were processed by applying thresholding algorithm to remove spurious peaks due to noise. Since this algorithm sets a flexible threshold, based on the received signal statistics, to achieve a constant false alarm rate, so after processing only in 79 impulse responses there was data above the respective noise floors which implied that no useful signal was received in other 31 sites.

Figure 5.17 shows the CDF of the multipath spread in the received signal for these 79 cases. As in Parkwood Hills case, here also we have small multipath spreads

FIGURE 5.17 CDF of the multipath spread at zero degree receive directions for Kanata data

attributed to the fact that very narrow beamwidth antennas were used. The 90th percentile of multipath spread lies at 80 nanoseconds; the 99th percentile is 100 nanoseconds, which implied that the results from the analysis of Kanata data should be similar to ones for Parkwood Hills.

Only two of the impulse responses of all these 79 had a delayed signal component (though of very small amplitude compared to the strongest signal sample, about 25 dB down) coming in at the receive antenna. Here also we chose an impulse response having a multipath spread of 90 nanoseconds, which is one of the maximum spreads of the remaining 77 impulse responses, for analysis purposes. We present the results of the analysis at the same three symbol rates as we did in the case of Parkwood Hills data to derive the conclusion about the equalization requirements for LMCS/LMDS systems. Though there are impulse responses with multipath spreads of smaller durations also, we decided to chose one from the group which represents the maximum of multipath spreads but without any delayed signal components.

Figure 5.18 displays one of the impulse responses measured at zero degree at site 'kab29044' at Kanata, where 'k' in site name stands for Kanata; 'a' represents a series of measurements; '29' signify the height of the receive antenna in feet and '044' stands for the location number of the measurement campaign. The results for the equalizer structure derived in the following sections are for this typical impulse response. Here, same assumptions and analysis procedure have been used as for Parkwood Hills impulse responses.

FIGURE 5.18 Sample level Vs. delay for channel at 0 degree at site kab29044

FIGURE 5.19 Contour plots of SNR_{out} variation with feedforward and feedback filter taps, at best sampling phase for 10 M symbol per second

5.3.1 10 Mega Symbols per second

Figure 5.19 displays the contour plots of output SNR for various combinations of feedforward and feedback filter taps at 10 M symbols per second corresponding to best sampling phase, the feedforward filter being symbol spaced. Figure 5.20 shows the variation in the SNR_{out} with the sampling phase for this symbol spaced forward filter case. These two plots in Figure 5.19 and Figure 5.20 are very similar to ones for Parkwood Hills in Figure 5.5 and Figure 5.6 at the same data rate. Hence the same conclusions, as drawn for the impulse response shown in Figure 5.4 from Parkwood Hills, hold here and system would not require any equalizer at this symbol rate owing to relatively small multipath spread.

Next we display the results when the forward filter is a T/2 spaced filter. Figure 5.21 shows the relative invariance of the output SNR with sampling phase in this case, where the plots correspond to a forward filter of one symbol duration. Figure 5.22 is for the contour plots of output SNR for a FS-DFE where the horizontal axis represents the forward filter symbol durations. Again, there is a similarity in these results obtained for a FS-DFE with those obtained with one of the Parkwood Hills impulse response shown in Figure 5.7 and Figure 5.8 , at same transmission rate.

Though Parkwood Hills is primarily residential and Kanata a mix of industrial and residential type, the amount of multipath present in the recorded impulse responses from both of these locations is similar and we have obtained similar equalizer structures for both of these cases. Thus we conclude that for symbol rates in the vicinity of 10 mega symbols per second, there will not be an appreciable deterioration in the system performance if we do not use an equalizer though we can still improve upon the output SNR

FIGURE 5.20 SNR_out variation with sampling phase for channel impulse response of Figure 5.18 at 10 M symbol per second (Kanata data)

FIGURE 5.21 SNR_{out} variation with sampling phase for channel impulse response of Figure 5.18 for FS-DFE at 10 M symbols per second (Kanata data)

FIGURE 5.22 Contour plots of SNR_{out} variation with forward filter and feedback taps, at 10 Msps with forward filter T/2 spaced (Kanata data)

if we use one. Also, FS-DFE has better performance than its symbol spaced counterpart. The results from the analysis of other impulse responses are summed up in Appendix B.

5.3.2 25 Mega symbols per second

Next we tried the impulse response of Figure 5.18 at 25 M symbols per second. Figure 5.23 , for a symbol spaced forward filter, shows the SNR_{out} values that can be achieved for various combinations of forward and feedback filter taps when sampling at the right sampling phase. Here, with taps (2,1) we can achieve a reasonable value of 18.29 dB for the output SNR, though the performance still improves with increase in the feedforward taps but the addition of more than 1 feedback taps does not appreciably change the performance. Going back to the results shown in Figure 5.9 of Parkwood Hills, we need to have a combination of (3,1) to achieve an output SNR of greater than 18 dB. This difference in the performance of these cases (although both have same multipath spread) is attributed to the difference of the shape of the impulse response curves. As will be shown in the Appendix B, the number of the taps shuffle between (2,1) and (3,1) for achieving SNR value of 18 dB or better. So, for 25 M symbols per second case with T-spaced forward filter, we recommend a (3,1) DFE equalizer.

Figure 5.24 shows the variation of SNR_{out} with sampling phase for the symbol spaced forward filter case. We have shown plots corresponding to tap combinations of (2,0), (2,1), (3,0), (3,1) and timing phase of '0' is optimum time to sample. Then we present the results when the forward filter of DFE is a fractionally spaced (T/2) filter.

FIGURE 5.23 Contour plots of SNR_{out} variation with feedforward and feedback filter taps, at best sampling phase for 25 M symbols per second (Kanata data)

FIGURE 5.24 SNR_out variation with sampling phase for channel impulse response of Figure 5.18 at 25 M symbol per second (Kanata data)

FIGURE 5.25 SNR_{out} variation with sampling phase for channel impulse response of Figure 5.18 for FS-DFE at 25 M symbols per second (Kanata data)

FIGURE 5.26 Contour plots of SNR_{out} variation with forward filter and feedback taps, at 25 Msps with forward filter T/4 spaced (Kanata data)

Figure 5.25 shows that for a DFE with forward filter of minimum 2 symbol durations (i.e. 4 taps), the output SNR is independent of the sampling phase. Also shown in the same figure is the fact that the output SNR varies appreciably with the sampling phase when the forward filter is of inadequate length of just one symbol duration. Figure 5.26 shows contour plots of SNR_{out} corresponding to the sampling phase of ' $T/4$ '. From the present case and previously presented Parkwood Hills example and also from the results presented in Appendix B from the analysis of all the impulse responses analyzed in this thesis, we conclude that (4,1) is a reasonable tap combination for this channel that should be used for a FS-DFE at 25 M symbols per second. Also, if we compare the output SNR values for symbol spaced and FS-spaced forward filter for taps (4,0) and (4,1), we find FS equalizer performing better than the symbol spaced equalizer, and also we get additional independence from the sampling phase.

5.3.3 50 Mega symbols per second

At 50 M symbols per second with symbol spaced forward filter, the plots in Figure 5.27, drawn for the best sampling phase, show that the equalizer requires a tap combination of at least (5,1) to achieve output SNR of more than 18 dB. This tap combination we finally recommend based on the results obtained from analysis of all the impulse responses for a symbol spaced DFE, the performance subject to the condition of correct sampling phase estimation. For an equalizer of (5,0) or (5,1), Figure 5.28 shows the performance variation with sampling phase and we observe that '0' is the best sampling phase for (5,1).

For FS-DFE case, as evident from Figure 5.29 and Figure 5.30, tap combinations

FIGURE 5.27 Contour plots of SNR_{out} variation with feedforward and feedback filter taps, at best sampling phase for 50 M symbols per second (Kanata data)

FIGURE 5.28 SNR_{out} variation with sampling phase for channel impulse response of Figure 5.18 at 50 M symbol per second (Kanata data)

FIGURE 5.29 Contour plots of SNR_{out} variation with forward taps and feedback taps, at 50 Msps with forward filter T/2 spaced (Kanata data)

FIGURE 5.30 SNR_{out} variation with sampling phase for channel impulse response of Figure 5.18 for FS-DFE, at 50 M symbols per second

of (6,1) or more give SNR_{out} value of more than 18 dB and also we find in this particular case that with (6,1) the value of the SNR_{out} remains mostly constant with sampling phase in contrast to the corresponding example of FS-DFE for Parkwood Hills. Looking at results from analysis of all the impulse responses, we conclude that (8,1), i.e. forward filter of four symbol durations and one feedback tap, would be a judicious choice at this data rate when employing FS-DFE. The recommended number of taps in the previous sections, for all the symbol rates and for symbol spaced and FS equalizers, are based upon the statistical results for a large number of channels analyzed. Since it is not possible to present graphically (as we did in the previous sections) the analysis results for all the channels here, the readers are recommended to refer to Appendix B for further details, where all the results in terms of output SNR as a function of different tap combinations have been compiled in the tabular form at all the three symbol rates.

The overall DFE statistics, for symbol spaced as well as FS-DFE, corresponding to the recommended number of taps at all the three transmission rates are also presented here through histograms from Figure 5.31 to Figure 5.36 (for the best sampling phases). Each of these histograms show the output SNR achieved for all the analyzed channels of Parkwood Hills and Kanata corresponding to a particular (recommended) tap combination of the DFE at a particular symbol rate. In these plots, to deal with the cases which are not “well behaved”, as we have already mentioned, we need to have more taps (for example for 05d28 case, at 50 Msps, it requires about 16 taps in the feedback filter to achieve output SNR of greater than 18 dB) in the feedback filter (or some delay to be implemented). Also, corresponding to these sites there may be some close location to position the antenna where there is no delayed component or a different antenna height

can be tried. As we can see that the multipath spread at location 05c28002, which is at the same main location of '05' as is 05d28, is only 70 ns in contrast to 370 ns for 05d28. The equalization requirements for a spread of 70 ns are (1,0), (2,1), (4,1) at 10, 25, 50 Msps respectively, for an output SNR of better than 18 dB at the best sampling phases for T-spaced FF case. Whereas, in the case of 05d28, the corresponding tap combinations are (3,3), (7,8) and (14,16) at 10, 25, 50 Msps, respectively for better than 18 dB output SNR requirement. Hence, we find that the equalization requirements of '05c28002' are much simpler than '05d28'. So it is possible that by making some adjustments in the receive antenna height or placement we can have a better view of the transmitter antenna and avoid a big equalizer. A similar explanation holds true for cases like '08d15002', where the antenna height is just 15 feet high (probably not clearing the roof tops) and we can see that by raising it to 26 feet as in '08d26002' at the same location, the multipath spread is reduced to 80 ns from 170 ns.

FIGURE 5.31 Histogram of output SNR (dB) for the channels of Parkwood Hills and Kanata with a DFE of (1,0) at 10 Msps, symbol spaced case.

FIGURE 5.32 Histogram of output SNR (dB) for the channels of Parkwood Hills and Kanata with a FS-DFE of (2,0) at 10 Msps.

FIGURE 5.33 Histogram of output SNR (dB) for the channels of Parkwood Hills and Kanata with a symbol spaced DFE of (3,1) at 25 Msps.

FIGURE 5.34 Histogram of output SNR (dB) for the channels of Parkwood Hills and Kanata with a FS-DFE of (4,1) at 25 Msps.

FIGURE 5.35 Histogram of output SNR (dB) for the channels of Parkwood Hills and Kanata with a symbol spaced DFE of (5,1) at 50 Msps.

FIGURE 5.36 Histogram of output SNR (dB) for the channels of Parkwood Hills and Kanata with a FS-DFE of (8,1) at 50 Msps.

5.4 Adaptation Algorithms Complexities and Comparison

In this section we carry out the complexity (the number of complex multiplications and complex additions) comparisons of a few widely used adaptation algorithms, using the results obtained for the DFE structures from the analysis of all the available impulse responses at different symbol rates (presented in previous sections and Appendix B). Then we recommend a suitable algorithm that is optimum in terms of the complexity and the speed of convergence in the present scenario. For this comparison, assume 'M' is the total number of feedforward plus feedback taps in the DFE (symbol-spaced or FS-DFE). Here we sum up the results that will be used in the later computations:

TABLE 5.1 Summary of DFE structures at 25 Msps and 50 Msps

Symbol Rate (Msps)	DFE Type	Number of Taps (forward, feedback)	M
25	T-spaced	(3,1)	4
25	FS-DFE	(4,1)	5
50	T-spaced	(5,1)	6
50	FS-DFE	(8,1)	9

5.4.1 LMS Algorithm:

It requires total of ' $2M+1$ ' complex multiplications and ' $2M$ ' complex additions per iteration and converges to the optimum solution in $\sim 10M$ iterations [35],[30],[34]. Table 5.2 details the computational complexity of this algorithm for different values of 'M' that have been listed in Table 5.1.

TABLE 5.2 Computational Complexity with LMS algorithm

M	DFE Type	Number of complex multiplications		Number of complex additions	
		per iteration (2M+1)	for convergence ~10M(2M+1)	per iteration 2M	for convergence ~10M(2M)
4	T-sp	9	360	8	320
5	FS	11	550	10	500
6	T-sp	13	780	12	720
9	FS	19	1710	18	1620

5.4.2 Block adaptation with DMI (direct matrix inversion) using Covariance

method:

In this adaptation scheme, the available data are divided into individual blocks [35]. The equalizer coefficient vector is determined once at the end of training sequence and is held fixed during the transmission of the data block. So the adaptation process proceeds on a block by block rather than on a sample by sample basis. Assume the equalizer span of M symbols, training period of N symbols, and Q samples per symbol period. For block adaptation with DMI, total number of complex multiplications and additions per training symbol is (including ones for Cholesky factorization for equation solving) [39]:

$$\text{Total\_Multiplications / iteration} = 2M + \frac{M}{2N} \left(\frac{M}{Q} - 1 \right) + \frac{1}{6N} (M^3 + 9M^2 - 4M) \quad (5.2)$$

$$\text{Total\_Additions / Iteration} = \frac{2M}{N} (N - 1) + \frac{M}{N} \left(\frac{M}{Q} - 1 \right) + \frac{1}{6N} (M^3 + 6M^2 - 7M) \quad (5.3)$$

Then with this algorithm, the equalizer converges to optimum solution in about 2M iterations, i.e. a good choice of training sequence length is $N = 2M$ for convergence. Table 5.3 gives the computational complexity (rounded to nearest integer value) using equations (5.2) and (5.3), assuming $N=2M$, $Q=1$ for symbol spaced DFE and $Q=2$ for

FS-DFE

TABLE 5.3 Computational Complexity with Block Adaptation

M	DFE Type	Number of complex multiplications		Number of complex additions	
		per iteration (see equation 5.2)	for convergence ($2M * \text{column 3}$)	per iteration (see equation 5.3)	for convergence ($2M * \text{column 5}$)
4	T-sp	~13	102	~12	90
5	FS	~16	~159	~14	~138
6	T-sp	~21	~245	~19	~227
9	FS	~32	~577	~30	~530

5.4.3 RLS Algorithm:

RLS algorithm requires about $2M$ iterations to converge i.e. has the convergence properties almost the same as the Block adaptation. Its computational complexity [35], [37],[30],[34] is computed in Table 5.4:

TABLE 5.4 Computational Complexity of RLS Algorithm

M	DFE Type	Number of complex multiplications		Number of complex additions	
		per iteration ($2M^2+6M$)	for convergence $2M(2M^2+6M)$	per iteration (M^2+4M+2)	for convergence $2M(M^2+4M+2)$
4	T-sp	56	448	34	272
5	FS	80	800	47	470
6	T-sp	108	1296	62	744
9	FS	216	3888	119	2142

5.4.4 Fast RLS Algorithm:

Fast RLS algorithm has the same convergence properties as the RLS algorithm but its complexity is of the order of M , though the algorithm tends to be unstable under certain environments. It also converges in $\sim 2M$ iterations and its computational complex-

ity [35],[37],[40] is shown in Table 5.5:

TABLE 5.5 Computational Complexity of fast RLS Algorithm

M	DFE Type	Number of complex multiplications		Number of complex additions	
		per iteration ($10M+4$)	for convergence $2M(10M+4)$	per iteration ($12M+5$)	for convergence $2M(12M+5)$
4	T-sp	44	352	53	424
5	FS	54	540	65	650
6	T-sp	64	768	77	924
9	FS	94	1692	113	2034

As we see, of all the adaptation algorithms, Block Adaptation with DMI, using covariance method, has the least computational load for convergence. Although the LMS algorithm is simple yet it is computationally uneconomical due to its slow convergence. Hence we recommend Block DMI with Cholesky solution for adaptation when the channel is nearly time-invariant. We next investigate the time variation of the channel during the transmission of one ATM cell.

5.5 Time Variation of the Channel

As per LMCS/LMDS specific DAVIC 1.2 recommendations, the upstream channel is divided into fixed length TDMA slot frames and each TDMA frame is sub-divided into time slots (of the type - poll, contention or reserve time slots) of 68 bytes long (4 bytes of header, 48 bytes of pay load (ATM cell), 10 bytes of R-S coding, and one byte of guard interval plus some other overheads). We assume that the equalizer is trained after every time slot of 68 bytes and investigate here if the channel stays stationary (time invariant) during one time slot interval, or if we need to track the channel during the

transmission of the time slot. We consider two main causes that can result in channel variation: 1) Fluttering of tree leaves and 2) Moving vehicles between transmitter and receiver. We assume an average wind speed of 30 Km/hr, vehicle speed of 50 Km/hr, transmission frequency of 28 GHz, slot length of 68 bytes and QPSK modulation. Using the equations given in Chapter 3 in section 3.5, we make the following computations:

TABLE 5.6 Computations of time variation of channel

Symbol Rate (Msps)	Slot Tx time, T (μs)	Wind/Vehicle speed (Km/hr)	Max. Doppler shift, f_m (Hz)	Coherence Time, T_c (μs)
10	27.2	wind/30	777.78	544.03
10	27.2	vehicle/50	1296.29	326.31
25	10.88	wind/30	777.78	544.03
25	10.88	vehicle/50	1296.29	326.31
50	5.44	wind/30	777.78	544.03
50	5.44	vehicle/50	1296.29	326.31

From Table 5.6, we note that the slot transmission time $T < T_c$, the coherence time of the channel which is the measure of the time over which channel impulse response is essentially invariant. Hence, it may not be necessary to track the channel after we train the equalizer (if we have to, then the LMS algorithm is a good option for tracking).

5.6 Maximum-Likelihood Sequence Estimation

As described in Chapter 3, the computational complexity of the VA would be substantial at some of the higher symbol rates, such as 50 Msps or higher when the channel dispersion can be as high as 4-5 symbol durations or more. So here we try to find out the complexity along with the performance of the VA if the suboptimal techniques of using a shorter impulse response are employed.

As we have discussed in section 3.6 , the less complex approach is to partially equalize the channel so that the overall sampled response seen by the VA is significantly different from zero over only a small number of samples and any remaining ISI is considered to be noise. The linear portion of the DFE (the forward filter) can be assumed to work as a prefilter and in that case the initial sample of the overall desired impulse response (DIR) is rendered as large as possible compared to the additive noise. The overall sampled impulse response seen by the VA is of the type $\{F_k\}$, $k=0,1,...,F_B$, where F_B are the total number of the feedback taps in the DFE and $F_0 \approx 1$ [48] [49]. The VA makes decisions on the assumption that the DIR is the actual overall channel response. The length of the DIR is much less than the actual channel impulse response. A receiver structure employing DIR has been shown in [48].

If the DFE has only one feedback tap, then the overall impulse response presented to VA has only two samples, thus has much reduced complexity and results in much simplified processing. For QPSK signalling, it will result in a trellis with 4 states, or 16 states if we assume 2 feedback taps in DFE.

Hence, we can think of the cascade of the transmit filter, the channel, the receive filter and forward filter of the DFE optimally conditioning the impulse response of the channel to approximate an impulse response of limited duration for which maximum-likelihood sequence estimation of the data is implementable in practice [48]. From this structure we can get the bounds for the probability of error when MLSE is employed with shortened impulse response [49]. The estimation of the overall system impulse response with the shortened impulse response does lead to some performance degrada-

tion. The upper bound on performance is computed by using the probability of error equation suggested by Forney [46], by computing MMSE and the minimum coding distance of DIR. The MMSE includes the additive noise plus any residual ISI and is the effective noise seen by the VA. The P_e in this case is approximated by:

$$P_e \leq K Q[d_{min}/2\sigma] = K Q\left(\sqrt{\frac{d_{min}^2}{4MMSE}}\right) \quad (5.4)$$

where K is a constant depending upon the error structure of the channel and d_{min} is the minimum euclidean distance between all possible pairs of sequences with differing first information symbols emerging from the DIR.

Also from the optimum structure of the DFE, which minimizes the MSE at the equalizer output, an estimate of the error rate performance of DFE can be made [50]. In the case of DFE, the MMSE may be roughly related to the performance in terms of the probability of symbol error for QPSK signalling via:

$$P_e \approx 2Q\left(\sqrt{\frac{|F_0|^2/2}{MMSE}}\right) \left(1 - \frac{1}{2}Q\left(\sqrt{\frac{|F_0|^2/2}{MMSE}}\right)\right) \quad (5.5)$$

The probability of error bounds for MLSE and DFE using the above equations has been plotted in Figure 5.37, for the impulse response of the Parkwood Hills channel 11b23002 discussed earlier, and the MMSE calculated at 50 Msps with a FS-DFE of (8,1). For comparison in Figure 5.37, we have also plotted the matched filter bound of the probability of symbol error for QPSK modulation which is equivalent to transmission of an isolated pulse. We see from this plot that the MLSE has a performance edge over DFE. This performance edge is at the cost of added complexity of MLSE over DFE. So the complexity of the MLSE is equal to the complexity of DFE (as it includes a DFE

(front end filter)) plus number of states times computations per state. Similarly, we calculated the performance bounds for VA and DFE using the impulse responses corresponding to the best directions at 52 location from Parkwood Hills under the same set of conditions, i.e a FS-DFE of (8,1). Figure 5.38 shows the histogram of the gain in dB of the MLSE detection over DFE for a P_e of 10^{-3} . As we see for VA, in most of the cases we need 1.7 dB to 1.8 dB less SNR per bit to achieve the same performance as DFE. Hence we conclude that maximum likelihood sequence estimation using VA can be employed in the receiver to achieve better performance if we can pay for added complexity of MLSE (over DFE).

FIGURE 5.37 Probability of symbol error bounds for MLSE and DFE for QPSK signalling, drawn for impulse response of 11b23003 at 50 Msps with MMSE calculated with a (8,1) FS-DFE

FIGURE 5.38 Histogram of the SNR gain of MLSE over DFE for P_e of 1.0×10^{-3} at 50 Msps

Chapter 6

CONCLUSIONS

6.1 Summary of the results

The work reported in this thesis had the objective of investigating the amount of multipath spread in a large number of millimeter wave outdoor broadband wireless channel impulse responses and to determine if an equalizer is required in the LMCS system receiver and if any, then what complexity?

For acquiring the complex impulse response (CIR) of the channel a spread spectrum based channel measurement was adopted by the measurement team from CRC, using a 63 bit pseudonoise sequence generated at 40 Mbps and transmitted over 29.5 GHz using a biphas modulation process. The channel CIR estimates were obtained by cross correlating the measured complex envelope and replica of original sequence. These CIRs were further processed by applying a thresholding algorithm to reduce the impact of noise on measured data and to extract the valid echoes.

The multipath spread in CIRs was found to be relatively small, in at least one direction with strong signal (relative to other directions) at most of the locations owing to use of very narrow beamwidth receive antenna. Not all the sites had line-of-sight communication with the transmitter antenna. At some sites stronger signals than the one at zero degree were received in other directions. Measured data from 52 locations in the Parkwood Hills resulted in 3120 CIRs to be evaluated, due to the azimuth rotation of the receive antenna in steps of 6° at each location. From Kanata data, 111 CIRs were evalu-

ated which came from the measurements done while looking in the transmitter direction. In this case after the noise thresholding was applied, only 79 responses had the data samples above their respective noise floors which implied that no useful signal was received at other 32 locations. The 90th percentile of the multipath spread for the Parkwood Hills (corresponding to impulse response with highest energy at each location) and Kanata data has been found to be 90 ns and 80 ns, respectively.

A DFE had been chosen for the analysis. From the measured responses, we determined the optimum number of feedforward and feedback filter taps for achieving the minimum mean-square-error or for maximizing the output SNR, for QPSK modulation at different symbol rates of 10 Msps, 25 Msps and 50 Msps assuming an input SNR of 20 dB. As implied by the small multipath spreads, it has been found in our analysis that the LMCS systems may not require an equalizer at all at low transmission rates such as 10 Msps, though the system performance does improve if we use one, e.g., in the given example from Parkwood Hills data, the output SNR changes to 19.17 dB with (1,1) DFE from 18.14 dB for no equalizer at 10 Msps. The equalization requirements of the system do become evident at higher symbol rates. For a symbol spaced DFE, at 25 Msps, it has been found that for achieving an output SNR of more than 18 dB, the system requires a (3,1) equalizer and at 50 Msps we need a (5,1) equalizer. For FS-DFE case, at 25 Msps and 50 Msps we recommend an equalizer with (4,1) and (6,1) or (8,1) taps, respectively. As we can see in Figure 5.9 at 25 Msps, using a (3,1) equalizer results in about 5 dB improvement in the performance than using no equalizer. Similar results can be quoted to show the need of equalization at 50 Msps. This stems from the fact that we can resolve more and more components (rays) of the multipath structure at higher data rates and

hence need to have an equalizer. For the symbol spaced feedforward DFE cases, it can be generalized from the analysis that the number of the taps in the FF filter is about (or one more than) the length of the impulse response divided by the symbol period and for the FS cases it is about twice that. It has been observed in the analysis that the addition of more than one feedback tap does not significantly change the output SNR. The effect of the quantization error in representation of the equalizer coefficients was not taken into account while calculating output SNR. We can think of 10-12 bit quantizer, as in the telephone modems, for actual filtering and about 16 bit quantizer for each coefficient in the adaptation phase [34]. Hence, overall we find that relatively simple DFE appears effective at the symbol rates investigated. The results have also confirmed the superiority of the performance of FS-DFE over the baud rate sampling DFE. Through the plots of the output SNR Vs. timing phase, we have shown that the performance of the symbol spaced DFE is very sensitive to the sampling phase error whereas by using a FS forward filter we gain in terms of freedom from the sampling phase error and also achieve better output SNR. In the analysis at 25 Msps and 50 Msps with FS-DFE, it has also been seen that the forward filter of the DFE must be of adequate duration to achieve the independence from the sampling phase and to avoid the loss of performance e.g see Figure 5.15 and Figure 5.25 .

The analysis of the Kanata data yielded similar equalizer structure as suggested by the analysis of Parkwood Hills data. Hence, over all we can say that a relatively simple DFE appears effective for the symbol rates investigated. From the comparison of complexity of various adaptation algorithms, we have concluded that Block adaptation with DMI using covariance method is preferable over the RLS algorithm for training the

equalizer. Assuming a particular speed for fluttering of tree leaves and vehicular movement, we have shown that the channel remains invariant during a time slot of 68 bytes and that there is no need to track the channel after training the equalizer coefficients. We have also found that a receiver based on MLSE using VA is also possible with acceptable complexity when the suboptimal measures like approximating the overall system impulse response with its shorter version are employed and its performance has been shown to be better than that of DFE.

Appendix A

EQUALIZER MSE EVALUATION

In this thesis, for analyzing all the measured broadband wireless impulse responses, in order to estimate the optimum size of the decision feedback equalizer at different symbol rates for LMCS system receivers, the mean-square-error at the output of the equalizer is used as the performance index. It is desired to find the optimum tap weights for feedforward and feedback filters, using Weiner-Hopf theory for linear transversal filters, that can result in minimum-mean-square-error (MMSE) [30],[34],[35]. This appendix describes the sequence of steps that were implemented for this analysis to calculate the MMSE. We assume the absence of any decision errors in the analysis.

Recalling from Chapter 3, signal presented to the input of the feedforward filter of the decision feedback equalizer is represented as:

$$z(t) = \sum_n d_n \Psi(t - nT) + \eta'(t) \quad (\text{A.1})$$

where $\{d_n\}$ is the sequence of transmitted symbols and $\eta'(t)$ is the bandpass noise. The $\{d_n\}$ are QPSK symbols represented by $\frac{1}{\sqrt{2}}(\pm 1 \pm j)$. It is assumed that information symbols are independent with zero mean and unit variance i.e.

- $E(d_n) = 0$
- $E(|d_n|^2) = \sigma_d^2 = 1$

It is desired to minimize the mean square error J given by:

$$J = E|d(n-D) - y(n)|^2 \quad (\text{A.2})$$

where $d(n-D)$ is the desired signal and

$$y(n) = \sum_{j=0}^{K_1} f_j^* z(nT - jT' - \tau_0) + \sum_{j=1}^{K_2} b_j^* d(n-j-D) \quad (\text{A.3})$$

where τ_0 is the sampling delay (at the input of the DFE we could sample once or more in a symbol interval with time shifts τ_0 so as to maximize the output SNR), and

$\{f_j\}$ are complex valued forward filter tap settings,

$\{b_j\}$ are complex valued feedback filter tap settings, and

$K_1 + 1$ and K_2 are the total number of tap weights in the feedforward and feedback filters respectively.

(A.3) can also be written as:

$$y(n) = \mathbf{w}^H \mathbf{u}(n) \quad (\text{A.4})$$

where $(\ )^H$ is the hermitian transposition (transpose of the complex conjugate) and the quantities in bold letters represent vectors. Also in the above equation,

$$\mathbf{w}^H = [f_0 \ f_1 \ f_2 \ \dots \ f_{K_1} \ b_1 \ b_2 \ b_3 \ \dots \ b_{K_2}]^* \quad (\text{A.5})$$

is the vector of forward plus feedback filter tap coefficients and

$$\mathbf{u}(n) = \begin{bmatrix} \mathbf{z}(n) \\ \mathbf{d}(n-1-D) \end{bmatrix} \quad (\text{A.6})$$

is the vector comprising of received signal samples and previously detected symbols. In (A.6):

$$\mathbf{z}(n) = [z(nT) \quad z(nT-T') \quad z(nT-2T') \quad \dots \quad z(nT-K_1T')]^T \quad (\text{A.7})$$

and

$$\mathbf{d}(n-1-D) = [d(n-1-D) \quad d(n-2-D) \quad \dots \quad d(n-K_2-D)]^T \quad (\text{A.8})$$

For finding the optimum tap weight coefficients for DFE, we can set the gradient of the mean square error J equal to zero. Equation (A.2) can also be written as:

$$\begin{aligned} J &= 1 - 2\text{Re}(\mathbf{w}^H \mathbf{P}) + \mathbf{w}^H \mathbf{R} \mathbf{w} \\ &= 1 + (\mathbf{w} - \mathbf{R}^{-1} \mathbf{P})^H \mathbf{R} (\mathbf{w} - \mathbf{R}^{-1} \mathbf{P}) - \mathbf{P}^H \mathbf{R}^{-1} \mathbf{P} \end{aligned} \quad (\text{A.9})$$

where

$$\mathbf{P} = E[\mathbf{u}(n) d^*(n-D)] \quad (\text{A.10})$$

and

$$\mathbf{R} = E[\mathbf{u}(n) \mathbf{u}^H(n)] \quad (\text{A.11})$$

In these equations, $\mathbf{P}$ is a $(K_1 + K_2 + 1) \times 1$ vector from the cross correlation of the tap inputs of forward and feedback filters and the desired signal, and $\mathbf{R}$ is $(K_1 + K_2 + 1) \times (K_1 + K_2 + 1)$ correlation matrix of tap inputs of forward and feedback filters.

The mean square error in (A.9) is minimized when the taps weights in the DFE have optimum settings given by:

$$\mathbf{w}_{opt} = \mathbf{R}^{-1} \mathbf{P} \quad (\text{A.12})$$

And the minimum mean square error (MMSE) for these set of taps is:

$$J_{min} = MMSE = 1 - \mathbf{P}^H \mathbf{R}^{-1} \mathbf{P} = 1 - \mathbf{P}^H \mathbf{w}_{opt} \quad (\text{A.13})$$

The computation of J_{min} requires the knowledge of the correlation matrix $\mathbf{R}$ and the cross correlation vector $\mathbf{P}$. These can be computed as follows:

$$\begin{aligned} \mathbf{R} &= E[\mathbf{u}(n) \mathbf{u}^H(n)] \\ &= E \left[\begin{pmatrix} \mathbf{z}(n) \\ \mathbf{d}(n-1-D) \end{pmatrix} \begin{pmatrix} \mathbf{z}^H(n) & \mathbf{d}^H(n-1-D) \end{pmatrix} \right] \end{aligned} \quad (\text{A.14})$$

on solving which can be written as:

$$\mathbf{R} = \begin{bmatrix} \mathbf{R}_1 & \mathbf{R}_2 \\ \mathbf{R}_2^H & \mathbf{I} \end{bmatrix} \quad (\text{A.15})$$

where $\mathbf{I}$ is an identity matrix (of size $K_2 \times K_2$), and matrices $\mathbf{R}_1$ (of size $(K_1 + 1) \times (K_1 + 1)$) and $\mathbf{R}_2$ (of size $(K_1 + 1) \times K_2$) are given as:

$$\begin{aligned}
R_1 &= E[z(n) z^H(n)] \\
&= E \left[\begin{pmatrix} z(nT) \\ z(nT - T') \\ \vdots \\ z(nT - K_1 T') \end{pmatrix} \begin{pmatrix} z(nT) & z(nT - T') & \dots & z(nT - K_1 T') \end{pmatrix}^* \right] \quad (A.16)
\end{aligned}$$

On evaluating (A.16), we may represent the autocorrelation matrix R_1 as:

$$R_1 = \begin{bmatrix} r(0, 0) & r(0, 1) & \dots & r(0, K_1) \\ r^*(0, 1) & r(1, 1) & \dots & r(1, K_1) \\ \vdots & \vdots & \ddots & \vdots \\ r^*(0, K_1) & r^*(1, K_1) & \dots & r(K_1, K_1) \end{bmatrix} \quad (A.17)$$

where the quantity $r(j, k)$ is given by:

$$\begin{aligned}
r(j, k) &= E[z(nT - jT') z^*(nT - kT')] \\
&= E \left[\left(\sum_{i_1} d_{i_1} \Psi(nT - i_1 T - jT') + \eta'(nT - jT') \right) \times \right. \\
&\quad \left. \left(\sum_{i_2} d_{i_2} \Psi(nT - i_2 T - kT') + \eta'(nT - kT') \right)^* \right] \quad (A.18)
\end{aligned}$$

equivalently,

$$\begin{aligned}
r(j, k) &= \sum_{i_1} \sum_{i_2} E[d_{i_1} d_{i_2}^*] \Psi(nT - i_1 T - jT') \Psi^*(nT - i_2 T - kT') \\
&\quad + E[\eta'(nT - jT') \eta'^*(nT - kT')] \quad (A.19)
\end{aligned}$$

which can be further simplified to:

$$r(j, k) = \sum_i \Psi(iT - jT') \Psi^*(iT - kT') + \phi((j - k)T') \quad (\text{A.20})$$

where $\phi(t)$ is the autocorrelation function of the power spectral density (PSD) of band pass noise at the receiver filter output.

In (A.17), $r(j, k) = r^*(k, j)$.

Next we find matrix R_2 as:

$$R_2 = E \left[\mathbf{z}(n) \mathbf{d}^H(n-1-D) \right] \\ = E \left[\begin{pmatrix} z(nT) \\ z(nT - T') \\ \vdots \\ z(nT - K_1 T') \end{pmatrix} \begin{pmatrix} d(n-1-D) & d(n-2-D) & \cdots & d(n-K_2-D) \end{pmatrix}^* \right] \quad (\text{A.21})$$

On solving, we have:

$$R_2 = \begin{bmatrix} \Psi((D+1)T) & \Psi((D+2)T) & \cdots & \Psi((D+K_2)T) \\ \Psi((D+1)T - T') & \Psi((D+2)T - T') & \cdots & \Psi((D+K_2)T - T') \\ \Psi((D+1)T - 2T') & \Psi((D+2)T - 2T') & \cdots & \Psi((D+K_2)T - 2T') \\ \vdots & \vdots & & \vdots \\ \Psi((D+1)T - K_1 T') & \Psi((D+2)T - K_1 T') & \cdots & \Psi((D+K_2)T - K_1 T') \end{bmatrix} \quad (\text{A.22})$$

The vector $\mathbf{P}$ is given by:

$$\begin{aligned}
\mathbf{P} &= E[\mathbf{u}(n) d^*(n-D)] \\
&= E \begin{bmatrix} z(nT)d^*(n-D) \\ z(nT-T')d^*(n-D) \\ \vdots \\ z(nT-K_1T')d^*(n-D) \\ d(n-1-D)d^*(n-D) \\ d(n-2-D)d^*(n-D) \\ \vdots \\ d(n-K_2-D)d^*(n-D) \end{bmatrix} \quad (\text{A.23})
\end{aligned}$$

On solving, we can write $\mathbf{P}$ as:

$$\mathbf{P} = \begin{bmatrix} \Psi(DT) \\ \Psi(DT-T') \\ \Psi(DT-2T') \\ \vdots \\ \Psi(DT-K_1T') \\ 0 \\ 0 \\ \vdots \\ 0 \end{bmatrix} \quad (\text{A.24})$$

Appendix B

IMPULSE RESPONSE ANALYSIS RESULTS

In Chapter 5, we have presented the results for the equalization requirements of the LMCS systems at three different symbol rates using two different impulse responses, one from each of Parkwood Hills and Kanata set of measured data. Although impulse responses were recorded at many sites in both of these localities, it is not possible, in this thesis, to pictorially present the results of our analysis of all these responses. Hence, in this appendix we present the analysis results in the tabular form, for the best directions only (which have been defined as the direction with the strongest signal) at each of the measurement site, at all the three symbol rates we tried. The table includes results for the tap combinations that we recommended in Chapter 5 in addition to some other tap combinations for comparison. The results are in terms of the output SNR in dB (which is interpreted from the output MMSE) for both symbol spaced and FS DFE. The symbol spaced DFE results correspond to the best sampling phase. To show the impact of using an equalizer, the table also lists the output SNR value that would be achievable if we use no equalizer. The span of the impulse responses, in nanoseconds, is also shown in the table.

Table B.1 Output SNR (dB) for various tap combinations for Symbol spaced and Fractionally Spaced DFE (Contd.)

File Name	IR Span (ns)	SNR in dB at 10 Msps			SNR in dB at 25 Msps						SNR in dB at 50 Msps									
		T-spaced	FR-SP	T-spaced	FR-SP			T-spaced			FR-SP									
					(1,0)	(2,1)	(1,0)	(2,0)	(4,1)	(6,1)	(1,0)	(4,1)	(5,1)	(5,2)	(6,1)	(2,0)	(6,1)	(8,1)		
PARKWOOD HILLS RESULTS																				
01a15001	40	19.36	19.44	19.98	19.99	10.45	17.21	18.89	19.05	16.25	19.82	19.96	9.58	18.69	18.76	18.78	18.87	14.00	18.88	18.92
01a15003	50	19.76	19.83	19.93	19.99	16.12	19.20	19.42	19.43	16.45	19.70	19.88	8.79	17.90	17.97	18.10	18.13	9.97	18.04	18.17
01a32001	80	19.48	19.57	19.96	19.99	8.89	16.74	18.63	18.75	13.37	19.82	19.91	8.26	17.99	18.14	18.16	18.32	9.70	17.98	18.35
01b15001	30	19.74	19.87	19.93	19.99	17.65	18.76	19.68	19.74	14.45	19.86	19.95	11.03	19.02	19.03	19.15	19.10	11.97	19.19	19.31
01b25001	70	19.08	19.64	19.59	19.96	8.31	18.99	19.05	19.14	13.27	19.50	19.92	6.31	18.08	18.23	18.23	18.28	9.94	17.97	18.34
01c15001	50	19.16	19.76	19.53	19.95	14.64	18.98	19.55	19.67	15.13	19.68	19.87	7.85	17.85	18.05	18.08	18.06	8.22	18.11	18.19
01c25002	40	19.10	19.24	19.97	19.99	10.91	17.62	18.99	19.14	15.76	19.79	19.95	9.14	18.54	18.63	18.66	18.74	11.09	18.59	18.77
01d10006	40	19.29	19.77	19.64	19.97	15.67	19.26	19.62	19.75	16.23	19.89	19.91	9.31	18.65	18.73	18.79	18.80	12.32	18.82	18.90
01d5pk00	30	19.97	19.99	19.93	19.99	18.56	19.43	19.88	19.93	18.73	19.99	19.99	11.56	19.17	19.18	19.31	19.29	12.52	19.35	19.38
01e1010	30	19.61	19.68	19.62	19.69	18.48	19.72	19.89	19.97	18.49	19.99	19.99	11.57	19.25	19.26	19.33	19.34	13.44	19.40	19.43
01e5001	40	18.67	18.89	19.90	19.99	11.41	18.25	19.12	19.27	14.90	19.72	19.90	8.14	18.03	18.18	18.19	18.28	13.44	19.40	19.43
01f10001	40	18.58	19.38	19.33	19.89	10.81	18.25	19.26	19.37	14.78	19.64	19.72	7.57	17.98	18.06	18.14	18.13	12.43	18.20	18.32
03a15001	80	19.36	19.52	19.90	19.98	8.93	16.83	18.67	18.79	8.99	19.83	19.87	8.08	17.95	18.13	18.17	18.24	9.61	18.03	18.27
03a20001	90	19.86	19.95	19.94	19.99	17.37	18.67	19.68	19.74	17.74	19.90	19.94	8.21	17.91	18.02	18.14	18.19	10.32	17.97	18.23
03b20001	50	19.64	19.94	19.76	19.99	16.03	19.37	19.69	19.77	16.18	19.75	19.90	8.31	18.21	18.25	18.33	18.38	8.76	18.32	18.42

Table B.1 Output SNR (dB) for various tap combinations for Symbol spaced and Fractionally Spaced DFE (Contd.)

File Name	IR Span (ns)	SNR in dB at 10 Msps				SNR in dB at 25 Msps (dB)								SNR in dB at 50 Msps (dB)							
		T-spaced		FR-SP		T-spaced				FR-SP				T-spaced				FR-SP			
		(1,0)	(1,1)	(2,0)	(2,1)	(1,0)	(2,1)	(3,1)	(3,2)	(2,0)	(4,1)	(6,1)	(1,0)	(4,1)	(5,1)	(5,2)	(6,1)	(2,0)	(6,1)	(8,1)	
03c20001	60	19.82	19.94	19.83	19.98	7.52	16.18	18.45	18.52	12.87	19.81	19.89	7.24	17.95	18.03	18.24	18.27	10.72	18.07	18.28	
05a15002	70	19.79	19.85	19.96	19.99	7.01	17.82	17.83	17.86	8.47	19.88	19.92	7.09	17.04	18.20	18.21	18.33	10.47	18.08	18.35	
05a25002	90	18.35	18.66	19.24	19.82	11.59	17.51	19.13	19.30	14.21	19.85	19.89	8.15	17.77	18.13	18.15	18.13	10.11	18.22	18.23	
05b15002	40	19.36	19.84	19.67	19.98	15.81	19.69	19.69	19.84	17.94	19.60	19.96	9.71	18.80	18.88	18.90	18.95	12.30	18.94	18.98	
05b26002	50	19.83	19.99	19.83	19.99	16.05	19.37	19.66	19.71	17.25	19.67	19.89	7.45	17.70	17.78	17.97	17.97	11.22	17.88	17.99	
05b36002	80	18.35	18.88	19.81	19.97	11.81	18.00	19.15	19.32	14.22	19.78	19.91	8.36	18.14	18.27	18.30	18.35	11.26	18.23	18.41	
05c28002	70	19.00	19.67	19.55	19.95	1.92	19.42	19.51	19.65	6.02	19.92	19.92	5.79	18.00	18.14	18.15	18.21	6.82	17.87	18.34	
05d28	370	2.87	2.94	6.42	7.70	1.82	1.89	1.93	1.94	2.01	3.40	3.78	1.22	1.81	1.87	1.90	1.95	1.47	2.70	3.35	
05e20002	70	18.32	19.15	19.81	19.97	1.57	15.55	18.66	18.80	11.57	19.63	19.88	8.15	17.99	18.14	18.17	18.28	11.32	18.20	18.40	
06a15002	60	19.39	19.45	19.97	19.99	10.01	17.42	18.98	19.09	13.18	19.79	19.89	7.18	18.08	18.16	18.31	18.35	11.13	18.19	18.39	
06a18002	50	19.54	19.91	19.69	19.98	15.54	19.26	19.67	19.77	17.82	19.69	19.89	8.08	18.03	18.17	18.19	18.20	9.91	18.23	18.30	
06a25002	70	19.66	19.96	19.75	19.99	4.69	19.36	19.42	19.48	7.08	19.90	19.91	6.89	17.90	18.06	18.07	18.22	9.96	17.78	18.25	
08a15002	50	19.51	19.55	19.98	19.99	15.64	18.78	19.02	19.03	15.81	19.81	19.92	7.84	18.21	18.27	18.41	18.43	11.03	18.34	18.46	
08a28002	90	18.48	19.34	19.40	19.93	13.58	18.80	19.39	19.59	15.28	19.93	19.95	7.38	17.95	18.17	18.18	18.17	9.91	18.26	18.29	
08a31003	60	19.40	19.48	19.99	19.99	9.97	17.12	18.83	18.97	13.67	19.83	19.92	9.23	18.63	18.72	18.75	18.79	10.14	18.38	18.45	

Table B.1 Output SNR (dB) for various tap combinations for Symbol spaced and Fractionally Spaced DFE (Contd.)

File Name	IR Span (ns)	SNR in dB at 10 Msps				SNR in dB at 25 Msps								SNR in dB at 50 Msps							
		T-spaced		FR-SP		T-spaced				FR-SP				T-spaced				FR-SP			
		(1,0)	(1,1)	(2,0)	(2,1)	(1,0)	(2,1)	(3,1)	(3,2)	(2,0)	(4,1)	(6,1)	(1,0)	(4,1)	(5,1)	(5,2)	(6,1)	(2,0)	(6,1)	(8,1)	
08b15002	40	18.24	19.23	19.81	19.97	12.79	18.69	19.30	19.47	14.90	19.82	19.95	9.23	18.63	18.72	18.75	18.79	12.76	18.72	18.83	
08b29002	70	19.54	19.59	19.97	19.99	16.32	19.05	19.23	19.25	16.47	19.81	19.93	8.06	17.00	18.28	18.29	18.30	9.91	18.24	18.38	
08b34002	90	19.68	19.87	19.86	19.99	17.03	18.95	19.75	19.85	17.29	19.93	19.94	8.32	18.06	18.20	18.30	18.33	9.91	18.15	18.33	
08c27002	50	19.44	19.53	19.80	19.95	14.47	19.41	19.74	19.80	15.72	19.65	19.85	8.60	18.21	18.24	18.32	18.37	11.35	18.32	18.45	
08c34002	80	19.69	19.77	19.95	19.99	8.53	16.56	18.56	18.68	13.64	19.86	19.93	8.31	18.13	18.28	18.29	18.39	10.44	18.11	18.41	
08d15002	170	6.72	7.56	10.22	10.87	1.55	4.89	8.02	8.66	8.65	9.49	10.76	1.01	3.45	8.66	8.74	8.80	10.17	15.09	15.18	
08d26002	80	18.16	19.26	19.72	19.94	12.78	18.41	19.35	19.51	14.39	19.77	19.90	8.11	17.98	18.14	18.17	18.28	10.99	18.15	18.35	
11a15002	70	19.19	19.77	19.57	19.97	1.85	13.41	16.45	16.45	6.31	19.90	19.91	6.13	17.93	18.08	18.09	18.15	10.48	17.72	18.25	
11a23002	70	19.63	19.66	19.96	19.99	16.41	19.10	19.29	19.31	16.52	19.78	19.92	8.05	17.04	18.24	18.25	18.28	9.46	18.22	18.36	
11b23002	90	18.14	19.17	19.24	19.86	13.23	15.65	18.00	18.01	19.38	19.67	19.92	1.32	17.96	17.98	17.99	18.00	10.23	18.14	18.24	
11crt001	90	19.96	19.96	19.55	19.82	17.71	18.57	19.67	19.72	18.28	19.91	19.96	8.37	18.10	18.22	18.25	18.25	8.73	18.21	18.30	
12a19001	50	19.76	19.87	19.84	19.92	10.65	16.47	17.27	17.42	13.61	17.71	17.85	5.75	15.95	16.96	17.00	17.06	10.74	17.05	17.21	
12a29001	70	19.60	19.64	19.93	19.98	16.52	19.15	19.32	19.33	16.64	19.82	19.94	8.22	17.17	18.37	18.38	18.41	9.47	18.39	18.46	
12b19002	70	19.60	19.91	19.74	19.99	5.06	19.33	19.39	19.45	6.97	19.92	19.93	6.94	18.02	18.17	18.17	18.31	10.15	17.89	18.32	
12b29002	70	19.74	19.82	19.93	19.99	2.90	17.86	17.86	17.89	8.23	19.89	19.92	7.93	18.07	18.22	18.23	18.32	10.15	18.10	18.34	

Table B.1 Output SNR (dB) for various tap combinations for Symbol spaced and Fractionally Spaced DFE (Contd.)

File Name	IR Span (ns)	SNR in dB at 10 Msps				SNR in dB at 25 Msps						SNR in dB at 50 Msps								
		T-spaced		FR-SP	T-spaced			FR-SP			T-spaced			FR-SP						
		(1,0)	(1,1)	(2,0)	(2,1)	(1,0)	(2,1)	(3,1)	(3,2)	(2,0)	(4,1)	(6,1)	(1,0)	(4,1)	(5,1)	(5,2)	(6,1)	(2,0)	(6,1)	(8,1)
12c19004	70	19.81	19.82	19.98	19.99	16.94	19.35	19.49	19.52	16.96	19.81	19.94	8.32	17.45	18.30	18.34	18.36	10.99	18.29	18.44
2c29002	70	19.70	19.89	19.85	19.98	1.28	18.80	18.82	18.85	7.50	19.92	19.93	7.54	18.06	18.21	18.22	18.34	10.11	17.97	18.35
12d15002	40	18.74	18.96	19.91	19.99	11.33	17.34	18.29	18.47	14.91	18.93	19.09	8.79	17.91	18.01	18.03	18.12	12.74	17.99	18.18
12e15002	90	17.17	18.75	19.03	19.76	3.45	11.05	14.55	14.94	7.18	13.45	17.60	2.72	8.21	11.18	11.48	11.24	10.78	15.61	15.63
22hva000	90	19.84	19.98	19.91	19.99	17.38	19.42	19.72	19.86	17.40	19.93	19.96	8.31	17.98	18.13	18.18	18.21	9.95	18.07	18.23
2a40003	90	19.57	19.68	19.93	19.99	17.19	17.91	19.62	19.67	18.64	19.92	19.94	8.46	18.18	18.28	18.34	18.33	9.39	18.25	18.33
KANATA RESULTS																				
kab23001	70	19.66	19.73	19.96	19.99	8.52	17.34	18.60	18.70	14.27	19.81	19.90	8.04	18.02	18.18	18.20	18.31	11.22	18.18	18.37
kab23002	90	19.89	19.90	19.98	19.99	9.73	18.50	18.56	19.69	16.64	19.38	19.90	4.59	17.66	17.71	18.03	18.18	10.61	17.49	18.29
kab23004	80	18.64	18.92	19.91	19.99	11.71	17.94	19.16	19.32	14.21	19.84	19.93	8.36	18.14	18.25	18.26	18.34	10.97	18.19	18.40
kab23005	70	19.35	19.76	19.69	19.98	6.51	19.0	19.12	19.19	6.71	19.93	19.94	6.73	18.10	18.24	18.25	18.34	8.144	18.04	18.37
kab23007	70	18.48	19.31	19.62	19.92	5.46	18.41	18.46	18.57	5.65	19.90	19.92	5.46	18.02	18.16	18.19	18.30	6.14	18.10	18.40
kab23008	70	18.94	19.08	19.96	19.99	10.73	17.66	19.01	19.15	13.69	19.78	19.90	8.23	18.09	18.23	18.26	18.34	11.14	18.20	18.38
kab23010	70	19.82	19.99	19.84	19.98	7.33	19.02	19.04	19.08	7.55	19.90	19.91	7.34	17.95	18.10	18.11	18.25	8.02	18.27	18.37
kab23011	70	18.24	19.25	19.85	19.98	12.58	18.40	19.36	19.51	14.31	19.81	19.92	8.29	18.13	18.28	18.28	18.41	11.49	18.30	18.48

Table B.1 Output SNR (dB) for various tap combinations for Symbol spaced and Fractionally Spaced DFE (*Contd.*)

File Name	IR Span (ns)	SNR in dB at 10 Msps				SNR in dB at 25 Msps						SNR in dB at 50 Msps								
		T-spaced		FR-SP		T-spaced				FR-SP		T-spaced				FR-SP				
		(1,0)	(1,1)	(2,0)	(2,1)	(1,0)	(2,1)	(3,1)	(3,2)	(2,0)	(4,1)	(6,1)	(1,0)	(4,1)	(5,1)	(5,2)	(6,1)	(2,0)	(6,1)	(8,1)
kab23013	50	19.65	19.95	19.73	19.99	15.66	19.29	19.67	19.74	17.71	19.63	19.88	7.60	17.75	17.81	17.96	17.97	8.23	17.89	18.02
kab23014	70	19.62	19.68	19.97	19.99	8.73	17.15	18.64	18.75	8.77	19.86	19.90	8.09	18.03	18.17	18.19	18.30	9.37	18.09	18.32
kab23015	50	19.45	19.91	19.64	19.98	15.38	19.24	19.69	19.81	18.84	19.70	19.92	8.29	18.13	18.34	18.34	18.34	8.48	18.30	18.41
kab23016	70	19.66	19.83	19.86	19.98	7.54	18.36	18.39	18.48	7.69	19.91	19.92	7.61	18.07	18.21	18.22	18.33	9.74	18.10	18.35
kab23017	70	19.75	19.91	19.87	19.99	7.48	18.66	18.67	18.70	7.65	19.92	19.93	7.58	18.06	18.21	18.21	18.33	9.65	18.07	18.36
kab23019	70	19.40	19.83	19.67	19.98	6.46	19.34	19.41	19.49	6.69	19.92	19.93	6.52	17.97	18.12	18.12	18.21	7.90	17.90	18.26
kab23020	70	19.51	19.63	19.91	19.98	8.37	17.15	18.56	18.66	8.51	19.84	19.87	7.72	17.86	18.01	18.04	18.14	8.82	17.95	18.16
kab23022	70	19.11	19.73	19.55	19.96	5.96	19.54	19.64	19.77	6.19	19.91	19.92	6.04	17.97	18.11	18.12	18.18	7.06	17.81	18.30
kab23023	80	18.23	19.11	19.81	19.97	12.44	18.20	19.24	19.42	14.57	19.83	19.93	8.37	18.19	18.28	18.29	18.37	11.13	18.24	18.43
kab23025	70	19.70	19.93	19.8	19.99	7.14	19.00	19.02	19.06	7.35	19.92	19.93	7.22	17.99	18.13	18.14	18.29	9.05	17.98	18.32
kab23026	80	18.86	19.04	19.94	19.99	11.46	17.85	19.13	19.29	14.11	19.83	19.93	8.25	18.05	18.16	18.16	18.25	10.77	18.10	18.35
kab23028	70	19.11	19.68	19.57	19.96	6.03	19.23	19.32	19.43	6.25	19.93	19.94	6.22	18.11	18.25	18.25	18.30	7.30	17.99	18.40
kab23029	80	18.33	19.16	19.79	19.97	12.43	18.19	19.26	19.44	14.54	19.84	19.94	8.41	18.18	18.29	18.30	18.38	10.81	18.33	18.48
kab23031	90	17.15	18.63	19.40	19.84	11.85	17.81	19.25	19.38	13.43	19.68	19.72	7.18	17.23	17.55	17.76	17.63	7.87	17.33	17.79
kab23032	70	19.29	19.76	19.65	19.97	6.38	19.09	19.15	19.22	6.60	19.92	19.93	6.43	18.03	18.18	18.19	18.24	7.67	18.00	18.30

Table B.1 Output SNR (dB) for various tap combinations for Symbol spaced and Fractionally Spaced DFE (Contd.)

File Name	IR Span (ns)	SNR in dB at 10 Msps			SNR in dB at 25 Msps							SNR in dB at 50 Msps								
		T-spaced		FR-SP	T-spaced				FR-SP			T-spaced				FR-SP				
		(1,0)	(1,1)	(2,0)	(2,1)	(1,0)	(2,1)	(3,1)	(3,2)	(2,0)	(4,1)	(6,1)	(1,0)	(4,1)	(5,1)	(5,2)	(6,1)	(2,0)	(6,1)	(8,1)
kab23035	50	19.40	19.47	19.98	19.99	15.29	18.60	18.91	18.98	15.57	19.75	19.89	7.56	18.12	18.17	18.30	18.31	10.01	18.26	18.38
kab29043	50	19.68	19.70	19.97	19.99	15.97	19.01	19.22	19.22	16.04	19.76	19.90	7.82	18.07	18.14	18.27	18.30	9.42	18.20	18.33
kab29044	90	18.32	19.18	19.34	19.86	13.59	18.29	19.38	19.58	18.79	19.57	19.91	6.57	17.74	18.14	18.17	18.15	7.39	18.19	18.23
kab29046	90	17.87	18.98	19.01	19.61	12.67	18.11	19.28	19.47	15.21	19.89	19.91	8.09	17.78	18.10	18.13	18.10	8.25	18.10	18.23
kab29047	50	19.44	19.50	19.92	19.99	15.32	18.63	18.93	18.99	15.58	19.76	19.89	7.49	18.13	18.17	18.31	18.32	9.91	18.24	18.37
kab29049	50	19.80	19.95	19.86	19.98	16.39	19.41	19.65	19.69	17.38	19.71	19.90	8.14	18.10	18.19	18.31	18.34	8.38	18.23	18.35
kab29050	70	19.36	19.68	19.74	19.98	7.03	18.22	18.23	18.31	7.23	19.89	19.90	7.10	18.01	18.17	18.19	18.29	8.81	18.08	18.30
kab29052	40	18.30	19.33	19.78	19.96	13.07	18.95	19.37	19.54	14.61	19.81	19.95	8.86	18.46	18.58	18.58	18.64	12.18	18.57	18.68
kab29056	50	19.88	19.93	19.94	19.97	16.53	19.39	19.58	19.59	16.90	19.75	19.91	8.11	18.08	18.12	18.28	18.29	8.66	18.18	18.30
kab29057	50	19.77	19.94	19.83	19.99	16.01	19.34	19.62	19.67	17.17	19.67	19.88	7.63	17.81	17.92	18.09	18.09	7.80	18.04	18.13
kab29059	50	19.72	19.73	19.90	19.98	16.08	19.07	19.26	19.26	16.13	19.77	19.91	7.89	18.10	18.17	18.30	18.33	9.43	18.22	18.35
kab29060	50	19.74	19.77	19.92	19.99	16.12	19.12	19.32	19.32	16.22	19.75	19.90	7.87	18.03	18.11	18.24	18.27	9.16	18.17	18.30
kab29065	50	19.89	19.93	19.95	19.96	16.60	19.41	19.58	19.60	16.93	19.76	19.91	8.18	18.15	18.19	18.33	18.34	10.03	18.26	18.38
kab29066	60	19.38	19.86	19.63	19.98	15.38	19.25	19.69	19.82	15.69	19.80	19.93	8.32	18.13	18.27	18.28	18.27	10.31	18.32	18.37
kab29068	70	19.84	19.94	19.91	19.97	7.82	18.46	18.48	18.57	7.96	19.91	19.93	7.76	18.02	18.17	18.17	18.27	8.17	17.98	18.31

Table B.1 Output SNR (dB) for various tap combinations for Symbol spaced and Fractionally Spaced DFE (*Contd.*)

File Name	IR Span (ns)	SNR in dB at 10 Msps				SNR in dB at 25 Msps								SNR in dB at 50 Msps							
		T-spaced		FR-SP		T-spaced				FR-SP				T-spaced				FR-SP			
		(1,0)	(1,1)	(2,0)	(2,1)	(1,0)	(2,1)	(3,1)	(3,2)	(2,0)	(4,1)	(6,1)	(1,0)	(4,1)	(5,1)	(5,2)	(6,1)	(2,0)	(6,1)	(8,1)	
kab29069	90	18.25	19.00	19.29	19.88	14.91	16.57	19.21	19.30	17.04	19.78	19.81	7.63	16.84	16.93	17.46	17.97	8.60	16.98	18.06	
kab30037	70	19.00	19.23	19.89	19.98	15.31	18.63	18.90	19.06	16.12	19.88	19.96	7.73	16.62	18.24	18.34	18.29	10.11	18.21	18.30	
kab30038	60	19.48	19.91	19.65	19.98	15.43	19.22	19.66	19.77	15.60	19.73	19.89	8.10	17.85	18.11	18.12	18.11	9.77	18.20	18.25	
kab30040	50	19.71	19.98	19.77	19.99	16.13	19.42	19.72	19.81	17.34	19.65	19.90	8.29	18.20	18.23	18.30	18.36	9.06	18.29	18.41	
kab30041	60	18.26	19.26	19.75	19.95	13.20	18.58	19.38	19.55	14.95	19.84	19.94	8.58	18.16	18.34	18.35	18.34	11.44	18.34	18.40	
kad23013	70	18.98	19.65	19.54	19.96	5.83	19.40	19.49	19.63	6.05	19.92	19.93	5.91	17.99	18.14	18.14	18.18	6.88	17.85	18.32	
kad29062	50	19.44	19.51	19.98	19.99	15.33	18.65	18.95	18.99	15.60	19.75	19.89	7.49	18.11	18.15	18.30	18.30	9.80	18.21	18.35	
kgb27012	30	19.67	19.73	19.77	19.98	17.28	17.72	19.44	19.45	19.83	19.96	19.99	11.95	19.34	19.34	19.43	19.39	15.60	19.44	19.45	
kgb27014	40	19.16	19.27	19.67	19.96	14.93	17.80	18.98	19.12	15.45	19.73	19.92	8.23	18.00	18.15	18.16	18.28	8.89	18.16	18.40	
kgb27018	50	19.80	19.95	19.85	19.97	16.18	19.36	19.62	19.66	17.17	19.68	19.88	7.78	17.94	17.98	18.15	18.16	8.01	18.06	18.18	
kgb27021	40	18.28	19.32	19.78	19.96	13.03	18.93	19.36	19.53	14.62	19.80	19.94	8.85	18.46	18.58	18.59	18.64	12.19	18.57	18.68	
kgb27024	40	19.76	19.78	19.80	19.98	17.19	19.43	19.52	19.52	17.26	19.90	19.97	9.80	18.81	18.85	18.91	18.98	11.69	18.87	19.02	
kgb29002	40	18.57	19.48	19.71	19.95	13.63	19.25	19.46	19.62	14.35	19.81	19.94	8.77	18.40	18.52	18.53	18.61	11.86	18.54	18.64	
kgb29004	40	18.94	19.11	19.95	19.98	14.57	17.90	19.06	19.22	15.53	19.80	19.95	9.13	18.55	18.64	18.66	18.70	11.55	18.59	18.75	
kgb29005	40	18.72	19.54	19.64	19.93	14.01	19.27	19.49	19.65	14.30	19.30	19.94	9.02	18.50	18.61	18.63	18.68	12.10	18.64	18.73	

Table B.1 Output SNR (dB) for various tap combinations for Symbol spaced and Fractionally Spaced DFE (Contd.)

File Name	IR Span (ns)	SNR in dB at 10 Msps		SNR in dB at 25 Msps						SNR in dB at 50 Msps										
		T-spaced		FR-SP	T-spaced			FR-SP			T-spaced			FR-SP						
		(1,0)	(1,1)		(2,0)	(2,1)	(1,0)	(2,1)	(3,1)	(3,2)	(2,0)	(4,1)	(6,1)	(1,0)	(4,1)	(5,1)	(5,2)	(6,1)	(2,0)	(6,1)
kgb30028	50	19.72	19.95	19.79	19.96	15.95	19.33	19.64	19.70	17.43	19.65	19.88	7.63	17.85	17.91	18.07	18.08	7.73	17.99	18.10
kgb30043	40	19.39	19.48	19.9	19.97	15.89	16.86	19.10	19.10	16.32	19.76	19.93	9.38	18.58	18.65	18.69	18.74	11.78	18.61	18.84
kgb30044	80	18.39	19.30	19.66	19.94	13.32	18.47	19.33	19.53	14.69	19.89	19.95	8.49	18.21	18.28	18.31	18.42	11.17	18.29	18.48
kgd27023	40	19.46	19.52	19.56	19.92	16.01	17.10	19.12	19.12	16.33	19.80	19.96	9.42	18.63	18.71	18.72	18.78	10.13	18.64	18.82
kgl30044	50	19.13	19.25	19.96	19.99	14.64	18.17	18.98	19.11	15.19	19.76	19.89	7.19	18.11	18.21	18.32	18.35	10.42	18.29	18.40
krb27003	50	19.59	19.95	19.70	19.99	15.74	19.33	19.72	19.82	18.48	19.71	19.91	8.26	18.14	18.27	18.28	18.29	8.67	18.27	18.38
krb27004	30	19.92	19.99	19.95	19.99	18.33	19.74	19.88	19.96	18.35	19.98	19.99	11.47	19.22	19.23	19.30	19.31	13.42	19.37	19.41
krb27005	40	19.61	19.93	19.76	19.99	16.58	19.72	19.77	19.89	17.36	19.57	19.96	9.75	18.82	18.90	18.94	18.97	11.72	18.98	19.00
krb27009	40	18.81	19.59	19.57	19.94	14.26	19.33	19.52	19.68	15.33	19.92	19.93	8.94	18.54	18.64	18.67	18.67	12.03	18.67	18.75
krb27011	40	18.26	19.30	19.75	19.95	13.01	18.83	19.34	19.50	14.52	19.78	19.93	8.78	18.39	18.50	18.52	18.59	11.99	18.52	18.66
krb27020	40	18.19	19.27	19.78	19.96	12.84	18.81	19.32	19.49	14.59	19.79	19.94	8.68	18.37	18.49	18.50	18.56	11.89	18.49	18.63
krb27021	40	18.65	18.92	19.86	19.98	14.02	17.80	19.06	19.23	15.43	19.77	19.92	9.56	18.68	18.75	18.79	18.82	12.51	18.76	18.91
krb27023	40	18.56	19.43	19.60	19.92	13.77	18.91	19.40	19.57	14.42	19.78	19.92	9.35	18.56	18.64	18.71	18.77	12.64	18.74	18.87
krb27024	50	19.28	19.82	19.58	19.96	14.97	19.10	19.61	19.74	18.66	19.63	19.89	8.12	18.03	18.24	18.25	18.24	8.28	18.23	18.33
krb27031	40	18.35	19.34	19.74	19.96	13.29	18.85	19.36	19.54	14.84	19.83	19.96	9.58	18.76	18.83	18.86	18.87	13.30	18.85	18.94

Table B.1 Output SNR (dB) for various tap combinations for Symbol spaced and Fractionally Spaced DFE

File Name	IR Span (ns)	SNR in dB at 10 Msps				SNR in dB at 25 Msps								SNR in dB at 50 Msps							
		T-spaced		FR-SP		T-spaced				FR-SP				T-spaced				FR-SP			
		(1,0)	(1,1)	(2,0)	(2,1)	(1,0)	(2,1)	(3,1)	(3,2)	(2,0)	(4,1)	(6,1)	(1,0)	(4,1)	(5,1)	(5,2)	(6,1)	(2,0)	(6,1)	(8,1)	
krb27035	40	19.50	19.90	19.70	19.99	16.05	19.68	19.74	19.86	16.28	19.95	19.95	9.04	18.63	18.73	18.74	18.74	10.88	18.76	18.79	
krb27037	30	19.95	19.99	19.96	19.99	18.46	19.72	19.89	19.97	18.47	19.96	19.99	11.51	19.24	19.24	19.32	19.33	13.36	19.39	19.42	
krb27038	20	18.69	19.33	19.61	19.96	14.73	17.88	19.28	19.53	17.51	19.98	19.99	15.62	19.43	19.69	19.74	19.73	16.50	19.81	19.85	
krb27047	30	19.70	19.75	19.77	19.98	17.43	17.83	19.48	19.49	19.83	19.97	19.99	12.23	19.38	19.38	19.48	19.41	13.05	19.48	19.49	
krb27049	30	19.93	19.98	19.95	19.99	18.71	19.19	19.88	19.89	19.10	19.98	19.99	11.74	19.25	19.26	19.36	19.30	12.10	19.39	19.41	
krb30026	30	19.94	19.99	19.96	19.99	18.67	19.38	19.89	19.92	14.65	19.94	19.99	11.80	19.23	19.24	19.36	19.32	12.54	19.39	19.42	
krb30029	40	18.29	19.32	19.77	19.96	13.09	18.87	19.35	19.52	14.67	19.81	19.95	9.03	18.53	18.63	18.65	18.70	11.53	18.68	18.77	
krb30060	30	19.96	19.97	19.98	19.99	18.44	18.85	19.81	19.82	19.17	19.98	19.99	11.15	19.13	19.15	19.24	19.24	11.29	19.29	19.32	
krb30061	20	18.84	19.19	19.78	19.98	15.15	16.82	18.90	19.18	16.70	19.97	19.99	14.53	19.70	19.84	19.87	19.88	17.20	19.92	19.92	
krb39030	40	18.70	19.54	19.65	19.93	13.95	19.30	19.49	19.66	14.27	19.81	19.94	8.90	18.47	18.58	18.60	18.64	11.97	18.60	18.69	

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IMAGE EVALUATION TEST TARGET (QA-3)

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