MEASUREMENT AND MODEL BASED CHARACTERIZATION
OF INDOOR WIRELESS CHANNELS

BY

HARK-SANG KIM
B.Eng., KYUNGPOOK NATIONAL UNIVERSITY, TAEGU, KOREA (1987)
M.Eng., KYUNGPOOK NATIONAL UNIVERSITY, TAEGU, KOREA (1989)

SUBMITTED IN PARTIAL FULFILLMENT OF THE REQUIREMENTS
FOR THE DEGREE OF DOCTOR OF ENGINEERING
IN ELECTRICAL ENGINEERING
UNIVERSITY OF MASSACHUSETTS LOWELL

Signature of Author: _______________________________ Date: ______________

Signature of Dissertation Director: ________________________________
Kavitha Chandra, Eng.D., Associate Professor

Signatures of Other Dissertation Committee Members:

________________________________________
Charles Thompson, Ph.D., Professor

________________________________________
Vineet Mehta, Eng.D., Adjunct Faculty, MIT Lincoln Labs.
MEASUREMENT AND MODEL BASED CHARACTERIZATION OF INDOOR WIRELESS CHANNELS

BY

HARK-SANG KIM

ABSTRACT OF A DISSERTATION SUBMITTED TO THE FACULTY OF THE DEPARTMENT OF ELECTRICAL AND COMPUTER ENGINEERING IN PARTIAL FULFILLMENT OF THE REQUIREMENTS FOR THE DEGREE OF DOCTOR OF ENGINEERING IN ELECTRICAL ENGINEERING UNIVERSITY OF MASSACHUSETTS LOWELL 2003

Dissertation Supervisor: Kavitha Chandra, Eng.D.
Associate Professor, Department of Electrical and Computer Engineering
ABSTRACT

In wireless communication systems, the transmitted signal is distorted by various phenomena that are intrinsic to the structure and contents of the wireless channel. Among these, multipath fading is a dominant source of distortion in indoor wireless communications. The effects of multipath can be determined from the channel impulse response (CIR). The structure of the CIR is a function of the channel geometry and its contents as well as the transmission frequency and bandwidth. The multipath effect can be pronounced with increase in transmission bandwidth. The transition of existing narrow band cellular systems to wideband communication will require a deeper fundamental understanding of the wireless channel and its impact on the performance of wireless systems.

In this work, the measurement and modeling of narrow and wideband channel impulse responses is undertaken. The channels considered are of the indoor type, with a focus on classroom and laboratory environments in the University. The measurements are made using a frequency sweep method. From the channel measurement and frequency selective fading features caused by multipath components are identified. The wideband characteristics of indoor wireless channel are investigated through delay parameters and the pole-zero characteristics of the channel transfer function. The narrow band characteristics are examined using the band limited CIR and its intersymbol interference (ISI) power and performance in orthogonal frequency division multiplexing (OFDM) systems. The dynamics of poles and zeros in the case of ideal 1-D lossless medium are discussed and time varying characteristics are studied using a
3-D empty room CIR which calculated by method of image.

The reverberation part of the CIR is classified into two regions, the coherent and the diffuse part. The coherent region is composed of line of sight (LOS), if it exists, and a number of first and second reflections and is highly affected by geometrical configuration of transmitter and receiver. The diffuse region is formed by multiple reflections and depends more on overall structure of the channel such as size of the room and material properties of boundary surfaces. The communication performance is affected by the high energy coherent region.
ACKNOWLEDGEMENTS

First of all I am grateful to my advisor Professor Kavitha Chandra who encouraged and supported me to obtain doctoral degree. From the beginning of my doctorate she guided and pointed me in the right direction. She put in a lot of effort for me and waits long time with the patience. I really appreciate her support and guidance and will bear a debt of gratitude rest of my life.

I would like to thank Professor Charles Thompson who directed, helped and rescued me when I was fall into the dilemma. He also taught me background knowledge of dissertation work. He was real mentor to me in both research and life.

I would like to express my gratitude to Dr. Vineet Mehta for motivating me to finish and for all the comments and remarks. I would like to use this opportunity to thank Professor Krishnan for educating me in probability and statistical communication theory. Special thanks to all members of Center for Advanced Computation and Telecommunications.

I want to express many thanks to my mother, brothers, sisters and my in-laws for all the support and encouragement and to dedicate this degree to my father who in the heaven. Thanks to my daughter Hee-jeo Kim and my son Nam-key Kim for really doing well when I had spent most of the time working.

Finally I am extremely thankful to my wife Kyung-jin Lee who has given me support all along. I am very grateful for her love, encouragement and devotion.
# TABLE OF CONTENTS

<table>
<thead>
<tr>
<th>Section</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>ABSTRACT</td>
<td>ii</td>
</tr>
<tr>
<td>ACKNOWLEDGEMENTS</td>
<td>iv</td>
</tr>
<tr>
<td>TABLE OF CONTENTS</td>
<td>v</td>
</tr>
<tr>
<td>LIST OF TABLES</td>
<td>ix</td>
</tr>
<tr>
<td>LIST OF ILLUSTRATIONS</td>
<td>x</td>
</tr>
<tr>
<td>LIST OF SYMBOLS</td>
<td>xiii</td>
</tr>
<tr>
<td>LIST OF ACRONYMS</td>
<td>xvi</td>
</tr>
</tbody>
</table>

**Chapter 1.** Introduction

1.1 Background on Wireless Systems  
1.2 Channel Measurements  
1.3 Calculation of Channel Impulse Response  
1.4 Wireless Propagation Models  
1.5 Thesis Objectives  

**Chapter 2.** Indoor Wireless Channel Measurements

2.1 Introduction  
2.2 Channel Measurement System  
2.2.1 Agilent 8753ES Vector Network Analyzer  
2.2.2 Transmitting and Receiving Antenna  
2.2.3 Measurement System
Chapter 4. Numerical Analysis of Wideband Channel Impulse Response

4.1 Introduction 79
4.2 Channel Delay Parameters 80
  4.2.1 Measurement Environments 82
  4.2.2 Calculation and Analysis of Delay Parameters 84
4.3 Pole Zero Calculation using Singular Value Decomposition 89
  4.3.1 Poles and Zeros 92
  4.3.2 Reconstruction of CIR based on Pole-Zero Model 95
4.4 Summary 97

Chapter 5. Analysis of Narrow Band Channel Impulse Response

5.1 Introduction 99
5.2 Narrow Band CIR and Its Intersymbol Interference (ISI) Power 101
  5.2.1 The Effect of Transmission Bandwidth 102
  5.2.2 Impact of Carrier Frequencies and Bandwidth 105
    5.2.2.1 Carriers in 1.8 - 2.2 GHz Band 107
    5.2.2.2 Carriers in 2.3 GHz Region 108
    5.2.2.3 Carrier in 2.4 GHz Region 109
5.3 The Influence of Channel Features on an OFDM System 110
  5.3.1 Principle of OFDM 111
  5.3.2 The Channel Effect on OFDM System 120
    5.3.2.1 Carriers in 1.8 - 2.2 GHz Band 123
    5.3.2.2 Carriers in 2.3 GHz Region 124
5.3.2.3 Carrier in 2.4 GHz Region  

5.4 Summary  

Chapter 6. Conclusions and Future Work  

6.1 Conclusions  

6.1.1 Channel Measurements  

6.1.2 Computational Models and Characterization of CIR: Wideband analysis  

6.1.3 Narrowband Analysis  

6.2 Future Work  

REFERENCES  

BIOGRAPHY
LIST OF TABLES

2-1. Technical specifications of Agilent 8753ES VNA 18

4-1. Measured mean excess delay (MED) [nsec] 85
4-2. The difference between MED and location of maximum peak (DMM) [nsec] 86
4-3. Measured RMS delay spread [nsec] 86
4-4. Measured 10 dB maximum excess delay [nsec] 87
4-5. Measured coherence bandwidth [MHz], $\alpha = 2\pi$ 88

5-1. ISI powers for $f_c = 1.9$ GHz 108
5-2. ISI powers for $f_c = 2.35$ GHz 109
5-3. ISI powers for $f_c = 2.45$ GHz 110
LIST OF ILLUSTRATIONS

2-1. Three different methods to get channel impulse response 17
2-2. Functional block diagram of vector network analyzer 19
2-3. Antenna radiation pattern in anechoic chamber 21
2-4. Measurement system 22
2-5. Super Gaussian window with different powers 26
2-6. Inverse Fourier transform of super Gaussian window 27
2-7. Anechoic chamber 29
2-8. The complex transfer function in anechoic chamber 30
2-9. The complex transfer function in multipath environment 32
2-10. Channel impulse response inside the anechoic chamber 33
2-11. Classroom (Ball Hall 313) 34
2-12. The magnitude diagram in a completely empty classroom 36
2-13. The group delay in the empty classroom 37
2-14. Normalized CIR in the empty classroom 38
2-15. Normalized CIR in the classroom with furniture 42
2-16. Normalized CIR during the session 43
2-17. Normalized CIR during the class intermission and time-varying condition 44
2-18. CACT laboratory (Falmouth 203) 47
2-19. The magnitude diagram in CACT 48
2-20. The group delay in CACT 49
2-21. Normalized CIR in a CACT (Falmouth 203)  

3-1. Structure and wave propagation in 1-D lossless channel  
3-2. Channel impulse response of 1-D space  
3-3. Channel impulse response of the bounded 1-D space  
3-4. The effects of change in reflection coefficients, $a$, on pole-zero location  
3-5. The effects of changing channel dimension, $n$, of 1-D sample space, $a=0.7$  
3-6. The effects on zeros by receiver movement in 1-D sample space  
3-7. The concept of 2-D image method  
3-8. The channel impulse response in an empty room  
3-9. The fading envelope of the empty room  
3-10. The real and the imaginary component of received signal  

4-1. The schematics of channel  
4-2. The pole and the zero of a completely empty classroom  
4-3. The detailed local plot of zero of a completely empty classroom  
4-4. The measured and reconstructed channel impulse response  

5-1. The super Gaussian band pass filter, $S(f; w, 2)$  
5-2. Narrow band complex channel impulse response  
5-3. The normalized channel impulse response with different bandwidth  
5-4. The normalized ISI power of narrow band CIRs  
5-5. The narrow band CIR at 1.9 GHz  
5-6. The narrow band CIR at $f_c = 2.35$ GHz  

xi
5-7. The narrow band CIR at 2.45 GHz with 200 MHz bandwidth  109
5-8. The schematic diagram of OFDM system  113
5-9. The constellation of QPSK signal  115
5-10. The band limited complex channel impulse response  116
5-11. The constellation of detected signal of known channel  118
5-12. The constellation of detected signal  119
5-13. The BER by the change of number of multicarrier  120
5-14. The BER by the change of SNR  121
5-15. The BER at 1.9 GHz carrier frequency  123
5-16. The BER at 2.35 GHz carrier frequency  124
5-17. The BER at 2.45 GHz center frequency with 200 MHz bandwidth  125
LIST OF SYMBOLS

\( \hat{H}(f) \)  Measured transfer function with magnitude, \( \hat{H}_m(f) \) and phase, \( \hat{\phi}(f) \)  
\( \hat{h}(n) \)  Measured channel impulse response (CIR)  
\( h(t, \tau) \)  Multipath channel impulse response  
\( A_k(t) \)  Amplitude of \( k \)th multipath component  
\( \theta_k(t) \)  Phase of \( k \)th multipath component  
\( \tau_k(t) \)  Time delay of \( k \)th multipath component  
\( h_s(t) \)  Impulse response of system with input, \( \delta(t) \)  
\( c_n(t_i) \)  System output with pseudo noise input \( n(t_i) \) at time \( t_i \)  
\( H_s(f_i) \)  Transfer function with unit magnitude input \( S(f_i) \) in frequency domain  
\( h_s(t) \)  Channel impulse response of frequency sweep method  
\( S_{ij} \)  S-parameter  
\( a_i \)  Input signal at port \( i \) of vector network analyzer (VNA)  
\( b_i \)  Output signal at port \( i \) of VNA  
\( H(\omega) \)  Complex transfer function with magnitude, \(|H(\omega)|\) and phase, \( \phi(\omega) \)  
\( \lambda \)  Wavelength  
\( t_g \)  Group delay of phase difference, \( \Delta \phi \) and frequency difference, \( \Delta f \)  
\( S(x; w, n) \)  Super Gaussian window with half width \( w \) and power \( n \)  
\( \bar{d} \)  Mean free path of space with volume, \( V \) and enclosed surface, \( S \)  
\( H_{1D}(z) \)  1-D channel transfer function with reflection coefficient of boundary, \( a \)  
\( h_{1D}(i) \)  1-D channel impulse response
\( H_{1D}(z) \) 1-D reciprocal transfer function

\( N(z) \) Numerator polynomial of transfer function, \( H(z) \)

\( D(z) \) Denominator polynomial of transfer function, \( H(z) \)

\( H'(z) \) Stabilized transfer function using reciprocal pole

\( r^{-1} e^{j\theta} \) Reciprocal pole of pole, \( r e^{j\theta} \)

\( T_{i,j} \) Transmitter in image grid \((i, j)\)

\( \Pi(x, x^s, t) \) Hertz potential of source at \( x^s \) and receiver at \( x \)

\( f_j \) \( j \)th image component amplitude

\( X_K \) Distance between receiver and \( K \)th image

\( E_j \) \( j \)th component of discrete time electric field

\( \beta_j \) \( j \)th component distance projected on the \( \bar{X}_K \)

\( M_K \) \( X_K \) divided by sampling distance

\( h_c \) Critical height of rough surface

\( \theta_i \) Incidence angle in surface

\( \rho_s \) Scattering loss factor of rough surface

\( \Gamma_r \) Rough surface reflection coefficient

\( |E_S(x, x^s)| \) Fading envelope with real, \( E_s \), and imaginary component, \( E_{si} \)

\( \bar{E}_S \) Mean process of fading envelope

\( E_S - \bar{E}_S \) Residual process of fading envelope

\( P(\tau) \) Power delay profile of channel

\( m_{\tau} \) Mean excess delay of channel

\( \sigma_{\tau} \) RMS delay spread of channel

\( B_c \) Coherence bandwidth of channel with decay rate \( \alpha \)
\( D_m \)  
Maximum reliable data rate of channel

\( Z_i \)  
ith nonzero zero of transfer function, \( H(z) \)

\( P_i \)  
ith pole of transfer function, \( H(z) \)

\( h \)  
Vector notation of channel impulse response

\( A^t \)  
Pseudoinverse of channel transfer matrix \( A \)

\( C \)  
Coefficient matrix of channel

\( W \)  
Singular value matrix with diagonal components \( s_k \)

\( \varepsilon \)  
Error between original CIR, \( h(n) \) and reconstructed CIR, \( \hat{h}(n) \)

\( P_{ISI} \)  
Normalized ISI power of narrow band CIR, \( h_{NB} \) with data rate, \( d_t \)

\( l_{max} \)  
Location of maximum peak, \( h_{NB_{max}} \), in \( h_{NB} \) with sampling rate, \( t_s \)

\( H_{NB} \)  
vector notation of narrow band transfer function

\( \hat{x} \)  
Received signal at OFDM system with input vector, \( x \)

\( \varepsilon_{BER} \)  
Bit error rate with data rate \( d_t \)

\( \sigma_n^2 \)  
Channel noise power
### LIST OF ACRONYMS

<table>
<thead>
<tr>
<th>Acronym</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>1-D</td>
<td>1 Dimensional</td>
</tr>
<tr>
<td>1G</td>
<td>1st Generation</td>
</tr>
<tr>
<td>2-D</td>
<td>2 Dimensional</td>
</tr>
<tr>
<td>2G</td>
<td>2nd Generation</td>
</tr>
<tr>
<td>3-D</td>
<td>3 Dimensional</td>
</tr>
<tr>
<td>3G</td>
<td>3rd Generation</td>
</tr>
<tr>
<td>ADSL</td>
<td>Asymmetric Digital Subscriber Line</td>
</tr>
<tr>
<td>AGC</td>
<td>Automatic Gain Controller</td>
</tr>
<tr>
<td>AMPS</td>
<td>Advanced Mobile Phone System</td>
</tr>
<tr>
<td>AWGN</td>
<td>Additive White Gaussian Noise</td>
</tr>
<tr>
<td>BER</td>
<td>Bit Error Rate</td>
</tr>
<tr>
<td>BPSK</td>
<td>Binary Phase Shift Keying</td>
</tr>
<tr>
<td>BW</td>
<td>BandWidth</td>
</tr>
<tr>
<td>CACT</td>
<td>Center for Advanced Computation &amp; Telecommunications</td>
</tr>
<tr>
<td>CCK</td>
<td>Complementary Code Keying</td>
</tr>
<tr>
<td>CDMA</td>
<td>Code Division Multiple Access</td>
</tr>
<tr>
<td>CIR</td>
<td>Channel Impulse Response</td>
</tr>
<tr>
<td>CP</td>
<td>Cyclic Prefix</td>
</tr>
<tr>
<td>DAB</td>
<td>Digital Audio Broadcasting</td>
</tr>
<tr>
<td>dB</td>
<td>decibel</td>
</tr>
<tr>
<td>Abbreviation</td>
<td>Description</td>
</tr>
<tr>
<td>--------------</td>
<td>-------------</td>
</tr>
<tr>
<td>dBm</td>
<td>decibel milliwatt, $0 \text{ dBm} = 1 \text{ mW}$</td>
</tr>
<tr>
<td>DCP</td>
<td>Detaching CP</td>
</tr>
<tr>
<td>DFT</td>
<td>Discrete Fourier Transform</td>
</tr>
<tr>
<td>DMM</td>
<td>Difference between MED and the location of Maximum peak</td>
</tr>
<tr>
<td>DQPSK</td>
<td>Differential Quadrature Phase Shift Keying</td>
</tr>
<tr>
<td>DS</td>
<td>Direct Sequence</td>
</tr>
<tr>
<td>DSSS</td>
<td>Direct Sequence Spread Spectrum</td>
</tr>
<tr>
<td>DUT</td>
<td>Device Under Test</td>
</tr>
<tr>
<td>DVB</td>
<td>Digital Video Broadcasting</td>
</tr>
<tr>
<td>FCC</td>
<td>Federal Communications Commission</td>
</tr>
<tr>
<td>FDMA</td>
<td>Frequency Division Multiple Access</td>
</tr>
<tr>
<td>FDTD</td>
<td>Finite Difference Time Domain</td>
</tr>
<tr>
<td>FFT</td>
<td>Fast Fourier Transform</td>
</tr>
<tr>
<td>FHSS</td>
<td>Frequency Hopping Spread Spectrum</td>
</tr>
<tr>
<td>GFSK</td>
<td>Gaussian Frequency Shift Keying</td>
</tr>
<tr>
<td>GHz</td>
<td>Giga Hertz</td>
</tr>
<tr>
<td>GMSK</td>
<td>Gaussian Minimum Shift Keying</td>
</tr>
<tr>
<td>GSM</td>
<td>Global System for Mobile communication</td>
</tr>
<tr>
<td>HPA</td>
<td>High Power Amplifier</td>
</tr>
<tr>
<td>ICI</td>
<td>Inter Channel Interference</td>
</tr>
<tr>
<td>IEEE</td>
<td>Institute of Electrical and Electronics Engineers</td>
</tr>
<tr>
<td>IF</td>
<td>Intermediate Frequency</td>
</tr>
<tr>
<td>IFFT</td>
<td>Inverse FFT</td>
</tr>
<tr>
<td>Abbreviation</td>
<td>Full Form</td>
</tr>
<tr>
<td>--------------</td>
<td>-----------</td>
</tr>
<tr>
<td>IMT</td>
<td>International Mobile Telephone</td>
</tr>
<tr>
<td>IR</td>
<td>InfraRed</td>
</tr>
<tr>
<td>IS-54</td>
<td>Interim Standard 54</td>
</tr>
<tr>
<td>IS-136</td>
<td>Interim Standard 136</td>
</tr>
<tr>
<td>ISI</td>
<td>Inter Symbol Interference</td>
</tr>
<tr>
<td>ISM</td>
<td>Industrial, Scientific and Medical</td>
</tr>
<tr>
<td>ITU</td>
<td>International Telecommunication Union</td>
</tr>
<tr>
<td>KHz</td>
<td>Kilo Hertz</td>
</tr>
<tr>
<td>LAN</td>
<td>Local Area Network</td>
</tr>
<tr>
<td>LOS</td>
<td>Line Of Sight</td>
</tr>
<tr>
<td>MAC</td>
<td>Medium ACcess</td>
</tr>
<tr>
<td>MAP</td>
<td>Maximum A Porteriori</td>
</tr>
<tr>
<td>Mbps</td>
<td>Mega bit per second</td>
</tr>
<tr>
<td>MDP</td>
<td>Maximum excess Delay of certain Power level</td>
</tr>
<tr>
<td>MED</td>
<td>Mean Excess Delay</td>
</tr>
<tr>
<td>MHz</td>
<td>Mega Hertz</td>
</tr>
<tr>
<td>ML</td>
<td>Maximum Likelihood</td>
</tr>
<tr>
<td>NLS</td>
<td>Non-Line of Sight</td>
</tr>
<tr>
<td>NMT</td>
<td>Nordic Mobile Telephony</td>
</tr>
<tr>
<td>NTT</td>
<td>Nippon Telephone and Telegram</td>
</tr>
<tr>
<td>OFDM</td>
<td>Orthogonal Frequency Division Multiplexing</td>
</tr>
<tr>
<td>PAM</td>
<td>Phase Amplitude Modulation</td>
</tr>
<tr>
<td>PD</td>
<td>Parallel Detector</td>
</tr>
<tr>
<td>Abbreviation</td>
<td>Description</td>
</tr>
<tr>
<td>--------------</td>
<td>-------------------------------------------</td>
</tr>
<tr>
<td>PDA</td>
<td>Personal Data Assistant</td>
</tr>
<tr>
<td>PDP</td>
<td>Power Delay Profile</td>
</tr>
<tr>
<td>ppm</td>
<td>part per million</td>
</tr>
<tr>
<td>P/S</td>
<td>Parallel to Serial converter</td>
</tr>
<tr>
<td>QAM</td>
<td>Quadrature Amplitude Modulation</td>
</tr>
<tr>
<td>QOS</td>
<td>Quality Of Service</td>
</tr>
<tr>
<td>QPSK</td>
<td>Quadrature Phase Shift Keying</td>
</tr>
<tr>
<td>R1</td>
<td>Receiver location 1</td>
</tr>
<tr>
<td>R2</td>
<td>Receiver location 2</td>
</tr>
<tr>
<td>R3</td>
<td>Receiver location 3</td>
</tr>
<tr>
<td>R4</td>
<td>Receiver location 4</td>
</tr>
<tr>
<td>RC1</td>
<td>Receiver location 1 in CACT</td>
</tr>
<tr>
<td>RC2</td>
<td>Receiver location 2 in CACT</td>
</tr>
<tr>
<td>RC3</td>
<td>Receiver location 3 in CACT</td>
</tr>
<tr>
<td>RC4</td>
<td>Receiver location 4 in CACT</td>
</tr>
<tr>
<td>RDS</td>
<td>RMS Delay Spread</td>
</tr>
<tr>
<td>RF</td>
<td>Radio Frequency</td>
</tr>
<tr>
<td>RMS</td>
<td>Root Mean Square</td>
</tr>
<tr>
<td>SMA</td>
<td>SubMiniature A</td>
</tr>
<tr>
<td>SNR</td>
<td>Signal to Noise Ratio</td>
</tr>
<tr>
<td>SOHO</td>
<td>Small Office Home Office</td>
</tr>
<tr>
<td>S/P</td>
<td>Serial to Parallel converter</td>
</tr>
<tr>
<td>SS</td>
<td>Spread Spectrum</td>
</tr>
</tbody>
</table>
SVD  Singular Value Decomposition
T    Transmitter
TDMA Time Division Multiple Access
Tr   Transmitter in CACT
UMass University of Massachusetts
UMTS Universal Mobile Telecommunication Service
UWB  Ultra Wide Band
VNA  Vector Network Analyzer
WBS  Wireless Broadband System
WCDMA Wideband CDMA
WLAN Wireless LAN
WLL  Wireless Local Loop
WWW  World Wide Web
1. Introduction

1.1 Background on Wireless Systems

Research on aspects of wireless communications related to consumer applications has been active for at least the last thirty years. In 1970, the FCC made available a 75 MHz band in the 806-881 MHz range for mobile telephony. This relatively large capacity system paved the way for significant innovations in cellular networks, personal communications services and more recently for data services on wireless networks. The evolution is often captured in terms of first (1G), second (2G) and third (3G) generation systems. The first generation systems encompass the analog cellular systems that are based on frequency-division multiple access (FDMA). These systems were standardized in the United States as Advanced Mobile Phone System (AMPS), as Nordic Mobile Telephony (NMT) in the European Nordic countries and the Nippon Telephone and Telegraph (NTT) system in Japan. The AMPS system supports 832 channels at 30 KHz carrier spacing. The 2G systems that evolved from the mid to late eighties focussed on digital modulation techniques and time-division multiple access (TDMA) schemes. The Global System for Mobile Communications (GSM) European standard was the first 2G implementation. It was designed to support slow rate date services in the 2.4-9.6 Kbps range. It uses 200 KHz carrier spacings and Gaussian
minimum shift keying (GMSK) modulation that can support bit rates of 270.8 Kbps. The Interim Standard 54 (IS-54 / IS-136) in North America is the counterpart to GSM. It is however reverse compatible to AMPS while using TDMA with 30 KHz carrier spacing and $\pi/4$ phase-shifted quadrature differential phase shift keying $\pi/4$ DQPSK and bit rates of 48.6 Kbps. These systems provided a much-needed increase in the capacity. In 1993 a higher capacity 2G system based on direct sequence code division multiple access (DS-CDMA) technology proposed by Qualcomm Inc. was implemented in the United States IS-95 standard. These systems have been shown to provide around an order of magnitude increase in capacity over AMPS technology\textsuperscript{[1]} \textsuperscript{[2]}.

The advent of the world-wide-web (WWW) and commercialization of the Internet during the mid nineties set the tone for expected functionality of the third generation systems. Wireless communication no longer meant mobile plain voice telephony service\textsuperscript{[3]}. Wireless services are expected to include voice mail, e-mail, instant messaging, multimedia messaging, web and intranet access. These services are to be supported on digital devices such as cellular phone, PDA, the personal computer and its peripherals. Spread spectrum technology on which CDMA is based was considered as the most viable means of achieving the higher data rates required for multimedia wireless transmission. However, 3G standards based on the GSM TDMA model are also being considered. The 3G standards have been addressed in the International Mobile Telephone 2000 (IMT-2000) specifications put forth by the International Telecommunications Union (ITU). These systems were targeted to
operate globally in the 2 GHz frequency range. The evolution of IS-54 and GSM to 3G is specified as wideband CDMA (WCDMA) or Universal Mobile Telecommunications Service (UMTS) systems. The IS-95 extension to CDMA is referred to as CDMA2000. These two systems are expected to be co-existing platforms in the current decade. The 3G systems use the 1.8-2.2 GHz frequency band and maximum bandwidth is 140 MHz[^4].

Most, if not all of the aforementioned standards activities have focussed on the support of enhanced mobile telephony services. In the last five years, the demand for fixed wireless systems has arisen, particularly in the context of setting up ad-hoc networks for various consumer and business related applications. The concept of the wireless local area network (WLAN) is a case in point. The application of WLANs may be found in campus offices, small office home office (SOHO), hospitals, residences, warehouses, manufacturing facilities, parking lots, buliding-to- building complexes and limited outdoor regions. In the last two years, major computer and telecommunication systems providers have developed cost effective wireless terminal cards and access points for deploying WLANs. It is instigated in large part by the availability of the Industrial, Scientific and Medical (ISM) frequency bands in the 902-928, 2400-2483.5 MHz and 5.725-5.825 GHz ranges. The FCC allows operation of low-power spread-spectrum (SS) devices in these frequency bands. The IEEE 802.11 working group focusses on the standardization of SS devices for WLANs. The first wireless LAN standard was issued by IEEE in 1997, based on international consensus, except Europe.
This was the IEEE 802.11 wireless LAN standard. The 802.11 specifies the physical and medium access (MAC) layers for operation of WLANs and addresses the direct sequence spread spectrum (DSSS) and frequency hopping spread spectrum (FHSS) access methods for the radio medium. The 802.11 supports both asynchronous and synchronous data transfer mode. The asynchronous data transfer is applied to non-time sensitive applications such as e-mail and file transfers. The synchronous mode supports time bounded applications like video and packetized voice\cite{1}. The IEEE 802.11 supports 3 different MAC layers: 2.4 GHz radio FHSS, 2.4 GHz DSSS and infrared (IR) with 1 and 2 Mbps data rate. The IEEE 802.11b can support data rate of 5.5-11 Mbps in the same MAC layer with 20 MHz bandwidth and 3 cell frequency reuse pattern. The IEEE 802.11a in 5 GHz band, supports 6-54 Mbps data rate with maximum 200 MHz band, 48 subcarriers, orthogonal frequency division multiplexing (OFDM) and 4 cell reuse pattern. The IEEE 802.11g standard for WLANs, which will extend the data rate of the IEEE 802.11b to 54 Mbps from current level of 11 Mbps using same 2.4 GHz physical layer, has been approved by the IEEE working group. Two more approval steps remain before IEEE 802.11g is completed. Final approval is expected in June 2003. More detailed information about IEEE 802.11 is available at the IEEE standards association homepage\cite{5}.

The orthogonal frequency division multiplexing (OFDM) modulation scheme is used for European digital audio broadcasting (DAB) and digital video broadcasting (DVB) standard and it is also a strong candidate modulation scheme for digital
television broadcasting in North America\textsuperscript{[6]} and next generation wireless LAN, IEEE 802.11g, and wireless broadband systems (WBS). This scheme has attracted much interest in the last few years because of its robustness to multipath fading and its intrinsic characteristics of multicarrier modulation for supporting maximum data rate.

The complementary open standard referred to as Bluetooth has also evolved for supporting wireless communication by various kinds of digital devices that are small in size and of relatively low cost. The Bluetooth wireless specification includes both link and application layer definitions for product developers and is aimed at supporting data, voice and content centric applications. These systems also operate in the unlicensed, 2.4 GHz ISM band to support worldwide compatibility. Bluetooth uses a spread spectrum, frequency hopping, full duplex signal at up to 1600 hops/sec. There are 78 frequency hopping channels with 1 MHz bandwidth to give a high degree of interference immunity. A maximum of 7 connections of devices can be established and maintained. There are transmitter power limitation regulations to support small frequency reuse pattern. The maximum output power is specified at 100 mW for class 1, 2.5 mW for class 2 and 1 mW for class 3 devices. Bluetooth supports up to 1 Ms/s symbol rate with Gaussian frequency shift keying (GFSK) modulation scheme\textsuperscript{[7]}. The specification of Bluetooth is available at the official Bluetooth website, http://www.bluetooth.com.

1.2 Channel Measurements

In 1966, Meadows et al. measured outdoor multipath propagation over a line of
sight radio link using the frequency sweep technique[8] and they found the height of the reflective ionosphere layer. Janssen et al.[9] [10] measured wide band indoor channel impulse responses using a coherent wideband measurement system in an office environment at frequencies of 2.4, 4.75, 11.5, and 17 GHz and analyzed the effect on bit error rate (BER) for a maximum of 50 Mbps data rate using the binary phase shift keying (BPSK) modulation scheme. The RMS delay spread (RDS) was determined to be on the order of 20 nanoseconds for indoor sites and Ricean K-factor was 1.5 dB and -1.1 dB for line of sight (LOS) and nonline of sight (NLS) indoor sites, respectively. The effect of people moving in the room was measured by Walker et al.[11] in a laboratory environment at the 2.4 GHz industrial, scientific and medical (ISM) band. They calculated the fading distribution and showed that it approached the Ricean distribution[12] and calculated level crossing rates and average fade durations. The power delay profile was measured in the same frequency band and mean excess delay, RMS delay, maximum excess delay and coherent bandwidth were calculated. McDonnell et al.[13] measured RMS delay spread (RDS) in indoor LOS environments at 5.2 GHz and concluded that maximum RDS in a room depends on the dimensions of the room and the reflection coefficients of the walls. Similar work was done by Smulders and Wagemans in the 58 GHz region[14]. Maeda et al.[15] measured and investigated the effect of different kinds of enclosure material in a rectangular room and compared it with the results of theoretical ray tracing method. As expected, the BER was high when the materials of the enclosures of the room had high reflection coefficients.
Heddergott and Truffer\textsuperscript{[16]} proposed a stochastic model of indoor wideband channel based on directional measurements of the power delay profile, the power azimuth profile and the power elevation profile in the 24 GHz band. The data were collected by measuring different propagation scenarios with directional measurements through the indoor channel. They measured the non-LOS transmission and fit the statistical characteristics of the delay spread profile using an exponential distribution and azimuth and elevation spread profiles as Laplacian distributions\textsuperscript{[17]}. In 2001, Foerster\textsuperscript{[18]} proposed indoor wireless propagation models for ultra wide band (UWB) based on the arrival of multipath impulses using a modified Poisson and 2 state Markov model, so called $\Delta-K$ model. Using this model the BER performance of various M-array phase amplitude modulation (PAM) systems with and without rake receivers were analyzed.

1.3 Calculation of Channel Impulse Response (CIR)

A number of computational models have been developed to calculate the CIR. The finite difference time domain (FDTD) method was used by Lauer et al.\textsuperscript{[19]} to calculate the channel characteristic with consideration of antenna properties. The FDTD method is based on Maxwell’s partial differential equations\textsuperscript{[20]} of electromagnetism that describe the evolution of electric and magnetic fields in time and space. There were a number of reasons to use FDTD. Time domain pulse is used as the source pulse, and a wide frequency range is solved with only one simulation. It is
generally useful when the resonant frequencies were not known exactly and broadband results are desired. And it also provides the electric and magnetic fields directly. There were some limitations in using FDTD in wireless wave propagation simulation. To calculate the electric and magnetic field, FDTD requires that the three dimension computational grid dimensions resolve the smallest wavelength and smallest structure inside the channel. The computational complexity of FDTD increases by the cube of the frequency increase and the increase of the channel dimensions. Another weakness of the FDTD was the stability. The stability of the computational algorithm was a function of the grid space and the time step used to calculate the time integration.

Derschm et al. and Yang et al. calculated the CIR for indoor wireless propagation using the ray tracing method. Derschm et al.\[^{21}\] calculated multiple reflection response in a corridor and Yang et al.\[^{22}\] calculated the CIR in an office environment. The ray tracing technique was developed to generate more realistic computer graphic images. The technique started at the viewing point and traced rays until they reached a light source. In a radio wave, the light source was replaced by the transmitter. Unlike light, radio waves can experience significant diffraction. To model diffraction, the ray tracer must take every ray that approached the edge of a surface and break it into several new diffracted rays. To consider all of the different ray paths, a variety of different ray tracing algorithm are developed. The method of images finds the paths by iteratively placing ghost transmitters behind each specular boundary surface. The image approach is able to find the exact ray path\[^{23}\] between points by first imaging the source in the plane.
of each wall, one at a time, and checking that the path between the image and receiver intersects the wall in the physically existing segment. In 1979, Allen and Berkley\cite{24} calculated the acoustic room impulse response using the image method. The exact image theory for the Sommerfeld half-space problem was constructed by Lindell and Alanen\cite{25} \cite{26} with both vertical magnetic and electric dipole case in 1984.

### 1.4 Wireless Propagation Models

There are a number of approaches to model the wireless channel. Bello and Nelin\cite{27} proposed a frequency selective fading channel model in 1962. And its effect on the communication system was investigated by Glance and Greenstein \cite{28} and the relationship between performance and shape of delay spread presented. The well known Clarke’s model\cite{29} based on the statistical characteristics of the received electromagnetic fields was proposed in 1968. Based on the experimental data from several factories, statistical properties of path gain coefficients, path interarrival times and number of paths are modeled by Yegani et al.\cite{30}. Okumura et al.\cite{31} proposed path loss models based on the empirical received power decay rate. Howard and Pahlavan find the path loss model and delay spread of the indoor case based on channel measurements using the frequency sweep method\cite{32}.

Turin\cite{33} analyzed the effects of multipath and fading in direct sequence code division multiple access (DS-CDMA) systems with the assumption that all users link with equal energies. Sengupta et al.\cite{34} estimated the channel impulse response in
CDMA systems with an array antenna using the maximum likelihood (ML) approach. And they used the CIR information to modify the spreading code in transmission and detection processes.

1.5 Thesis Objectives

The objective of this thesis work is to improve the understanding of indoor channels through measurements and models. Of particular interest is the impact of different types of interference in the channel on the structure of the channel response. It is known that temporal changes in the environment caused by motion of humans causes significant variations in the channel response. The degree of this response relative to stationary interferers will be investigated. Measurements of CIRs will be obtained in various controlled environments and the results will be analyzed to examine the impact of interfering structures in the channel. A model based characterization of the measurements will be undertaken to capture the influence of channel parameters. In particular, a pole-zero model of the transfer function is considered. The relationship of poles and zeros to channel parameters will be examined considered simple one-dimensional analytical models. The insights obtained from this analysis will be applied to the measurements and also to the results of computational models that provide estimates of channel impulse responses in empty rooms. The image source method for predicting the CIR is of particular interest. Both wide and narrowband analysis of the CIR is undertaken. The appropriateness of classical parameters for describing channel
performance is examined using the measured data. Finally, the performance of orthogonal frequency multiplexing systems and their sensitiveness to channel frequency features is examined.

This understanding will support higher data rates and higher quality of service (QOS) on wireless networks. In particular the influence of carrier frequencies in the gigahertz range and bandwidths in excess of a megahertz will be investigated for indoor environments. The effects of reflection, diffraction and scattering of the transmitted electromagnetic wave are exacerbated at these higher frequencies. In indoor environments, the reflections from the bounding surfaces can create significant multipath interference and a time and space varying CIR. These features place a tremendous burden on error control and equalization functions to maintain the QOS. The present state-of-the-art in wireless systems uses rather coarse descriptions of the CIR such as the RMS delay spread and exponential decay rate to select the operating regime of the aforementioned functions.

Most of the current and future wireless standards and specifications uses a wide spectrum to implement higher data rate and QOS. By the increase of data rate, intersymbol interference (ISI) and interchip interference originate from multipath fading is critical factor which actually limiting data rate. In the case of indoor wireless channel, it depends on the structure and size of the room, the internal and external materials of the building, the dynamics and characteristics of objects. To overcome these problems, most of state-of-the art wireless system adopt orthogonal frequency
division multiplexing (OFDM) modulation scheme which has a high spectral efficiency and low multipath distortion. The OFDM uses orthogonality to lower from high data bit rate to the low data stream by orthogonal transform, discrete Fourier transform.

There are a number of approaches to characterize and model wireless channel. Some of them is based on the statistical characteristics of the received electromagnetic energies such as Clarke’ model. the other typical approach is based on the magnitude of the electromagnetic field strength as function of distance between transmitter and receiver such as path loss model. These models are useful to estimate channel variation and bit error rate for time division multiple access (TDMA) and code division multiple access (CDMA). In OFDM system more tight channel information is needed to detect signal in receiver. Channel impulse response itself is highly related with multipath fading. And we can easily extract the information which will be used in wireless communication system as part of detector.

To implement high data rate communication system, the understanding of the channel is critical factor to overcome multipath fading which physically limits data rate. In this work, The indoor wireless channels are measured in a 3 different places with various scenrio and its frequency selective fading will be characterized and dominant multipath components are identified. The dynamics of pole and zero are investigated in the case of 1-D lossless space and channel variation will be studied using CIRs which generated by image method. The characteristics of indoor wireless channel will be investigated both wideband and narrow band case. The OFDM scheme is used to
evaluate channel capacity using 3 different communication systems. Through these work, indoor wireless channels are characterized and gave basis to overcome multipath fading effectively which is major obstacle of the progress of wireless communication technology.
2. Indoor Wireless Channel Measurements

2.1 Introduction

In this chapter, the methodology and results for measuring the multipath features of indoor channels are discussed. A vector network analyzer (VNA) is made use of to capture the frequency response of the channel. In particular, the frequency sweep method is used, wherein the VNA executes a frequency sweep for generating the source signal and measures the frequency response of the channel across the sweep frequency range. The complex transfer function of the communication channel is represented by

\[ \hat{H}(f) = \hat{H}_m(f) e^{j\hat{\phi}(f)} \]

in polar coordinates. Where \( \hat{H}_m \) is measured magnitude and \( \hat{\phi}(f) \) is the measured phase signal.

The impulse response of the channel \( \hat{h}(n) \) is calculated by the inverse Fourier transform of the transfer function.

\[ \hat{h}(n) = \text{IFFT}(\hat{H}(f)) \]  

In Equation (2-1) both \( n \) and \( f \) represent discrete samples in time and frequency. The values of \( \hat{h}(n_i) \) at delay index \( n_i \) is the sum of all multipath components of the channel that arrive in the interval \([n_{i-1}, n_i)\). A general representation of the time-varying multipath CIR is of the form,
\[ h(t, \tau) = \sum_k A_k(t) e^{j\theta_k(t)} \delta(\tau - \tau_k(t)) \]  

(2-2)

Where \(A_k\) and \(\theta_k\) are the magnitude and phase of the \(k^{th}\) multipath component from the transmitter to the receiver. The time delay for the \(k^{th}\) arrival is \(\tau_k\). The values of \(A_k\), \(\theta_k\) and \(\tau_k\) are all dependent on time \(t\). The discrete counterpart of Eq. (2-2) is obtained by replacing \(\tau\) by samples that are uniformly spaced on the time delay axis.

In this work, the amplitude and the phase measurements of the channel transfer function are measured in three different indoor environments. These correspond to an anechoic chamber, a classroom and the Center for Advanced Computation & Telecommunications (CACT) research laboratory. The measurements from the anechoic chamber allow the evaluation of accuracy of measurements obtained in multipath environments and calibration of the radiation patterns of antenna used in the experiments. To further understand the impact of different types of interference, the measurements made in a classroom are conducted under four different scenarios that range from an empty classroom, a classroom with furniture, when the class is in session with participants seated and during a period when occupants are in motion in the classroom.

The configuration of the measurement system will be described in the next section. The measurement results and calculated wideband channel impulse responses are shown in Section 2.3.
2.2 Channel Measurement System

There are at least three different methods that can be applied to determine the CIR for a given transmitter and receiver configuration and environment. One approach is to transmit an impulse like signal at the transmitter and measure the received power at the receiver. This method has been applied in relatively low frequency regions for acoustic applications. It is a direct time domain method for measuring the CIR. The second method is a spread spectrum technique and is referred to as the correlation method \[^{35}\]. It uses the fact that the autocorrelation of Gaussian random noise is an impulse function. Random noise is transmitted and the received signal is cross correlated with the input noise to yield the impulse response. The third method is the frequency sweep method. Here, signals are transmitted by sweeping through a finite range of frequencies and the complex channel response captured for the entire frequency range. The broader the range of frequencies considered, the closer is the approximation of the measurements to the channel transfer function. All of these methods are depicted in Fig. 2-1.

The frequency sweep method is used to measure indoor wireless channel characteristics in this thesis work. The measurement system consists of an Agilent 8753ES 30KHz- 3GHz vector network analyzer (VNA). The VNA allows measurement of the complex transmission coefficient, $S_{21}$ which will be explained in next section. For the calibration of the measurement system, the Agilent 85056D mechanical calibration kit was used. The Astron AXQ24SM-A quarter wave omni-directional monopole antenna was used for transmitting and receiving. The transmitting antenna is
attached to port 1 of the VNA through a 3 ft subminiature A (SMA) cable and the receiving antenna is connected to port 2 through a 30 ft cable. A short cable is used in the transmitting side to maximize the radiating power into the transmitting antenna.

\[ \delta(t) \quad \text{Channel} \quad h(t) \quad h_\delta(t) \]

(a) Time domain method

\[ n(t_i) \quad \text{Channel} \quad h_n(t) \quad c_n(t_i) \]

(b) Correlation method: \( h_n(t) = n(t)^* c_n(t) \)

\[ S(f_i) = 1 \quad \text{Channel} \quad H(f) \quad H_s(f_i) \]

(c) Frequency domain method: \( H(f) = H_s(f_i) \)

Figure 2-1. Three different methods to get the channel impulse response.

2.2.1 Agilent 8753ES Vector Network Analyzer

The Agilent 8753ES VNA allows complete characterization of RF components and channels by measuring their effect on the amplitude and phase of swept frequency and power, test signals\(^{[36]}\) \(^{[37]}\). The 8753ES includes an integrated synthesized swept signal source, test set and tuned receiver. The built-in S parameter test set provides a full range of magnitude and phase measurement in both the forward and the reverse
directions. And full two port calibration is possible that includes any adapters attached to the ports. The key performance and characteristics of Agilent 8753ES VNA which was used in the indoor wireless channel measurements are summarized in Table 2-1. The dynamic frequency range of the VNA is specified from 30 KHz to 3 GHz. The maximum output power of the generated signal is 10 dBm. The impedance of the input and output terminals are 50 Ω.

**Table 2-1. Technical specifications of Agilent 8753ES VNA**

<table>
<thead>
<tr>
<th>Categories</th>
<th>Sub-categories</th>
<th>Specifications</th>
</tr>
</thead>
<tbody>
<tr>
<td>Test-port Output</td>
<td>Frequency range</td>
<td>30KHz - 3GHz</td>
</tr>
<tr>
<td></td>
<td>Frequency resolution</td>
<td>1Hz</td>
</tr>
<tr>
<td></td>
<td>Frequency accuracy</td>
<td>± 10ppm at 25 ° ± 5 ° C</td>
</tr>
<tr>
<td></td>
<td>Power range</td>
<td>-85 to + 10 dBm</td>
</tr>
<tr>
<td></td>
<td>Sweep power resolution</td>
<td>0.01 dB</td>
</tr>
<tr>
<td></td>
<td>Impedance</td>
<td>50 Ω</td>
</tr>
<tr>
<td>Test-port Input</td>
<td>Frequency range</td>
<td>30KHz - 3GHz</td>
</tr>
<tr>
<td></td>
<td>Average noise level</td>
<td>&lt; -110dBm (10kHz BW)</td>
</tr>
<tr>
<td></td>
<td>Maximum input level</td>
<td>10 dBm</td>
</tr>
<tr>
<td></td>
<td>Impedance</td>
<td>50 Ω</td>
</tr>
<tr>
<td></td>
<td>Frequency response</td>
<td>± 1.0dB</td>
</tr>
<tr>
<td></td>
<td>Harmonics measurement accuracy</td>
<td>± 1.5dB</td>
</tr>
<tr>
<td>Magnitude Characteristics</td>
<td>Dynamic accuracy</td>
<td>10Hz IF BW</td>
</tr>
<tr>
<td></td>
<td>Reference range</td>
<td>± 500dB</td>
</tr>
<tr>
<td></td>
<td>Reference resolution</td>
<td>0.001 dB</td>
</tr>
<tr>
<td></td>
<td>Stability</td>
<td>0.02dB/ ° C</td>
</tr>
<tr>
<td>Phase Characteristics</td>
<td>Dynamic accuracy</td>
<td>10Hz IF BW</td>
</tr>
<tr>
<td></td>
<td>Range</td>
<td>± 180 °</td>
</tr>
</tbody>
</table>

The functional block diagram of the VNA is shown in Fig 2-2. Network analyzers differ in form and function from another commonly used RF measurement equipment.
Other equipment normally measure the unknown signal but the network analyzer measures a known signal which was generated by the sweep generator or external signal source\cite{38}. The VNA is composed of a sweep generator, a processor, S-parameter test set and display and control unit. The sweep generator produces a frequency swept signal in time to probe the device under test (DUT). In the channel measurements, the DUT is the combination of the transmitting system, channel and receiving system. The 8753ES VNA can generate signals ranging in frequency from 3 KHz - 3 GHz.

![Functional block diagram of vector network analyzer.](image)

The channel is characterized using complex scattering parameters referred to as S-parameters, which are developed to characterize linear networks at high frequencies.
S-parameters define the ratios of reflected and transmitted travelling waves measured at the DUT’s ports. The relation between the input and output signals may be described as

\[
\begin{bmatrix}
  b_1 \\
  b_2 
\end{bmatrix} = \begin{bmatrix}
  S_{11} & S_{12} \\
  S_{21} & S_{22}
\end{bmatrix} \begin{bmatrix}
  a_1 \\
  a_2
\end{bmatrix}
\]

where \(a_1\) is input signal of port 1, \(a_2\) is input at port 2 and \(b_1\) and \(b_2\) are output signals at each of the ports. \(S_{11} = \frac{b_1}{a_1}\big|_{b_2=0}\) and \(S_{21} = \frac{b_2}{a_1}\big|_{b_2=0}\) are the reflection and transmission coefficients determined by placing a matched load at port 2. Similarly \(S_{22} = \frac{b_2}{a_2}\big|_{a_1=0}\) and \(S_{12} = \frac{b_1}{a_2}\big|_{a_1=0}\) are same coefficients when port 1 is matched.

The S-parameter test set is composed of an RF attenuator, switch and directional coupler to measure the reflected and transmitted signals. The swept frequency signal is sent to port 1 of the S-parameter test set. A part of the swept signal is reflected back to port 1 and the rest is transmitted to the air through the transmitting antenna which is connected to port 1. The receiving antenna detects the swept signal and it is fed to port 2 of the S-parameter test set. The complex transmission coefficient is defined as \(S_{21}(\omega) = |H(\omega)| e^{j\phi(\omega)}\), where \(|H(\omega)|\) is the magnitude and \(\phi(\omega)\) is the phase at frequency \(\omega\) radians/sec represents the transfer function of the DUT. The VNA computes both magnitude and phase responses of the DUT and displays these functions on the display unit.
2.2.2 Transmitting and Receiving Antenna

Astron wireless technologies, Flextron portable, Model AXQ24SM-A, quarter wave omni-directional monopole antenna are used for transmitting and receiving. The vertical and horizontal radiation pattern of the antenna are measured in the anechoic chamber at 2.45 GHz. Fig. 2-3 shows the measured results of the radiation pattern. In the figure, the dotted outer circle represents 0 dB, the inner circle is at −20 dB and the origin represents −40 dB. The angular dotted lines are located at latitudes of 15°. The radiation intensity in the horizontal plane is nearly uniform in angle. The maximum deviation is 1.8 dB. The shape of the vertical radiation pattern shows that dominant radiation takes place at right angles to the antenna orientation.

![Horizontal radiation pattern](image1.png)  ![Vertical radiation pattern](image2.png)

(a) Horizontal radiation pattern  (b) Vertical radiation pattern

Figure 2-3. Antenna radiation pattern in anechoic chamber.
2.2.3 Measurement System

For measurement convenience, the VNA is mounted on a plastic moving cart. As shown in Fig. 2-4, the transmitting and receiving antenna are connected to the VNA with asymmetric structure to maximize transmitting power. The length of SMA cable which is connected to port 1 and port 2 of VNA is 3 and 30 ft respectively. The power loss of the 3 ft cable, model RG58C/U, and 7 mm adaptor was −1.54 dB. The supplied signal power to the transmitting antenna for radiation is 8.46 dBm with the setting at maximum VNA output power. The power loss at the receiver side is −12.54 dB. But the loss in the receiver side cable and adaptor does not significantly affect the signal to noise ratio (SNR) of channel measurements. The SNR is mainly affected by the transmitting power and the power in the channel interference noise.

Figure 2-4. Measurement system.
2.2.4 System Calibration

In indoor wireless channel measurements using VNA, the phase variations are sensitive to the transmitter and receiver separation distance. The relationship between the phase and the distance travelled by the electromagnetic wave is given in Equation (2-4). Here the travelling length is fixed at \( l \) meters and \( \lambda \) is the transmission signal wavelength.

\[
\phi = \frac{2 \pi \times l}{\lambda} \tag{2-4}
\]

A calibration process is needed to eliminate the affect on the phase by the connection cable. The VNA built-in calibration process contains offset opens, offset shorts and broadband loads from the Agilent 85056D mechanical calibration kit. The calibration process eliminates the phase variation by the connection cable and also eliminates power loss in the measurements of the relative magnitude in decibel unit. After the calibration, the measured magnitude and phase are only affected by transmitting and receiving antenna and wireless communication channel.

2.3 Indoor Wireless Channel Measurements

In this section, the measurement of channel transfer functions using the frequency sweep method in the VNA are described. The group delays of various channel environments are calculated based on the measured phases. Measured complex transfer functions are filtered to suppress sidelobes and inverse Fourier transformed to obtain the
CIRs.

Complex indoor wireless channel transfer functions are measured in three different places: (A) the anechoic chamber; (B) a classroom and in (C) CACT research laboratory. The anechoic chamber is in principle expected to absorb all the multipath components. Based on the measurement results of the anechoic chamber, the accuracy of measurements and effect of multipath and noise will be verified. For the other two cases, measurements are conducted at varying transmitter-receiver distances and different interference scenarios. For the classroom these scenarios are: (B1:) an empty classroom; (B2) a classroom with furniture; (B3) classroom in session, where reception includes line of sight (LOS) components; (B4) classroom during movement of occupants, where reception is typically non-LOS (NLOS). For case B2, the room includes metal student chairs with desks. In case B3, four students are seated and another person representing the instructor is in front of the room. This room contains three windows and one front door which are open in all cases. For case B4, representative of class intermission time, the transmission is blocked by movement of people in the room. For measurements in CACT, the environment consists of typical office equipment, such as partitions, desks, tables, chairs and computers and other equipment such as lab workbenches, bureaus etc..

2.3.1 Group delay

The phase linearity of the communication channel can be specified in terms of
group delay\textsuperscript{[37]}. The group delay is the rate of change of phase shift with respect to angular frequency, as the wave progresses through the wireless channel. Traditionally, group delay has been used to describe the propagation delay and examine deviation from conventional linear phase characteristics of systems. This parameter also contains valuable information about wireless wave propagation delay and distortion through nonlinear communication channels.

Mathematically, group delay, $t_g$, is the derivative of the phase response and is represented by

$$
t_g = -\frac{\Delta\phi}{360 \cdot \Delta f}
$$

(2-5)

where $\Delta\phi$ is the phase difference in degrees at two frequencies separated by $\Delta f$ in hertz. The $\Delta f$ is commonly called the aperture of the measurement. The 8753ES VNA can calculate group delay from its phase response measurements, but the group delay which will be shown is calculated from the channel measurement results, where the measurement aperture is 1875 KHz measurement aperture. The group delay may be considered as the time delay of the amplitude envelope of a sine wave at frequency $\omega$.

2.3.2 Super Gaussian Window

The Fourier transform of a signal with sharp discontinuities causes Gibbs effect\textsuperscript{[40]}. In the measurements, the data memory size and digitizing speed limitations of the measuring equipment often make it impossible to properly sample a function over
its entire domain. Consequently, window effects or sharp discontinuities become unavoidable. The common solution for such problems is to place an artificial window over the sampled data signal that gradually reduces the signal to zero at the endpoints. There are many windows such as Kaiser, Lanczos, Hamming, Hanning, triangular and Blackman to name just a few. There is no clear guide on how one chooses a window. In this work, a window is used to reduce the Gibbs’ phenomena and effectively suppress sidelobes of the channel transfer function. For this the super Gaussian window is selected which was first realized as apodized apertures at Lawrence Livermore National Laboratory and used to minimize diffraction effects in high power laser beams used in fusion research\cite{[41]}.

![Super Gaussian window with different powers.](image)

Figure 2-5. Super Gaussian window with different powers.

The super Gaussian window, $S$ is expressed as
\[ S(x; w, n) = e^{-|x/w|^n} \]  

(2-6)

where \( n \) is called power and \( w \) the half width. \( S(w) = \frac{1}{e} \approx 0.3679 \) and these points are located at -4.34 dB and sample index of \( w \) is 376. Fig. 2-5 shows \( |S(x; w, n)| \) function for several different values of \( n \). In the figure, \( x \) axis is data index number and \( y \) axis is the magnitude of the window. As can be seen in the figure, by increasing the power \( n \) the flatness of the window is increased and as \( n \) goes to \( \infty \) the window leads to rectangular shape.

![Figure 2-6. Inverse Fourier transform of super Gaussian window.](image)

The normalized inverse Fourier transforms of the super Gaussian window are shown in Fig. 2-6. For the case of \( n = 2 \) the sidelobes are suppressed. There is no loss and distortion in the mainlobe and the function is a simple Gaussian function,
\[ S(\chi; w, 2) = e^{-|\chi/w|^2} \]. To calculate the channel impulse responses based on the measured complex channel transfer function, this Gaussian window is used as the window and bandpass filter.

### 2.3.3 Calculation of Channel Impulse Response

The measured complex channel transfer functions are saved in the vector network analyzer’s built-in disk drive and transferred to the computer. Indoor wireless channel impulse responses are calculated numerically from measured complex channel transfer functions. As mentioned earlier, the calculation is based on the inverse Fourier transform. The log scaled magnitude of measured channel complex transfer function is converted to a linear scale. After that the polar coordinate, magnitude and phase values, are transformed to rectangular coordinate in the form of real and imaginary values. This data set is inverse Fourier transformed to get a complex channel impulse response.

The measurements for both magnitude and phase signals are saved with maximum frequency resolution available with the 8753ES VNA. This yields 1600 data points, for the allowable frequency sweep range. The frequency interval between sampled data is 1875 KHz. The data is zero padded with 448 points for the convenience of utilizing a fast fourier transform. The hypothetical frequency sweep range due to zero padding is 30 KHz - 3.84 GHz and time domain resolution is \( 1 / 3.84 \text{ GHz} = 0.2604167 \text{ nsec} \). The range resolution, of 7.8125 cm, is acquired by multiplying the speed of the electromagnetic wave with time domain resolution.
2.3.4 Measurements in Anechoic Chamber

Channel measurements are obtained in an anechoic chamber to investigate the accuracy of the measurement system and to check the effects of multipath. The measurements are conducted at the UMass Lowell anechoic chamber, located in the
basement floor of Pasteur Hall. The sideview and structure of the anechoic chamber is shown in Fig. 2-7. The size of the cubical region is 3.05 m in height, width and depth. The distance between transmitter and receiver bed is 4.57 m. All the inner surfaces of the chamber are covered by pyramidal absorber-ferrite tile, Echosorb, except on the test bed where normally the receiving antenna is located. Both transmitting and receiving antenna are vertically installed.

Figure 2-8. The complex transfer function in anechoic chamber.
The measured complex transfer function magnitude, phase and the group delay inside the anechoic chamber are presented in Fig. 2-8 (a-c). The $x$ axis is frequency in hertz and $y$ axis of (a) is relative received power in dBm unit, that of (b) is phase in degrees and $y$ axis of (c) is group delay in degrees/Hz. The magnitude diagram shows the effect of frequency selective channel attenuation. Ideally, one should observe a relatively flat response profile for the anechoic chamber. However, antenna characteristics will influence this pattern. There are two pass bands, one is between 0.5-0.75 GHz and the other is in the 1.6-2.65 GHz range. In these pass bands, the phase is observed to be linearly dependent on frequency. This feature is more clearly seen in the group delay response, where the group delay calculated from Eq. (2-5) has a zero value. The highest transfer ratio is observed around 2.45 GHz region. To compare the anechoic chamber characteristics with that of a general multipath environment, a complex transfer function and group delay measurements made in the region outside the anechoic chamber are shown in Fig. 2-9. Although the magnitude profile for this case is similar to that of the anechoic chamber, one sees an increase in the number of spectral nulls and the lower frequency pass band is shifted to 0.27-0.63 GHz. Increased number of fluctuations observed in the group delay characteristic indicate the nonlinear effects in the channel.

The calculated CIR for the anechoic chamber is shown in Fig. 2-10. The $x$ axis is the time in seconds and $y$ axis is relative magnitude in decibels. The distance between transmitter and receiver antenna is 4.57 $m$. The calculated wave travelling time is 15.24
nsec and line of sight (LOS) peak is detected at 15.1 nsec. The error is less than 1 sample point interval, 0.2604 nsec. The second peak arrives at 21.9 nsec and its magnitude is \(-23.8\) dB with respect to the LOS peak. The background noise level obtained by averaging is \(-51.9\) dB. A number of weak multipath components arrive in the interval between the second peak and around 100 nsec. The multipath delay profile decays with a slope of \(-0.15256\) dB/nsec. These reference parameters are useful for relative evaluation of CIRs obtained in other environments.

Figure 2-9. The complex transfer function in multipath environment.
If the anechoic chamber perfectly absorbs all multipath components then the CIR of the chamber would be a single LOS peak. The calculated CIR inside the anechoic chamber does not give this ideal result. But it is reasonably flat. The second arrival which experiences a 23 dB drop may be attributed to the test bed in anechoic chamber.

![Channel Impulse response inside the anechoic chamber.](image)

**Figure 2-10.** Channel Impulse response inside the anechoic chamber.

### 2.3.5 Channel Impulse Response in a Classroom

The response of a classroom in different scenarios is examined next. The classroom represents a typical rectangular indoor environment. These measurements will give some insight for implementation of future wireless systems for educational applications. Ball Hall 313 is selected to measure wave propagation and its dimensions are shown in Fig. 2-11.
The classroom is rectangular in shape of dimension \((7.26, 5.74, 2.85)\) m. Three of the
walls are made of concrete. The fourth wall is partially covered with window as indicated in the Fig. 2-11. A flat metallic heater cover extends on the bottom one third of the windowed wall. The concrete floor is covered by a carpet and the ceiling is made of celotex acoustic tiles and fluorescent lighting. A $1.22 \times 3.66$ m size blackboard is in the middle of the front wall. There are two wooden doors, one is in front wall and the other is on the right side wall.

The magnitude and the phase of channel transfer function with respect to frequency domain are measured in the classroom for the four different environments stated before at four different receiver locations. The transmitting antenna, $T$, is located at the front right corner $(0.91, 0.91, 1.04)$ m, if the observer faces the black board. The four receiver positions selected are shown as R1, R2, R3 and R4. They were randomly selected. R1 is in front of the section where students will be seated, at $(5.84, 2.34, 0.81)$ m, R2 is in the middle of the student side at $(1.22, 3.68, 0.81)$ m, R3 is beside the window at $(5.69, 1.83, 0.79)$ m and near the back wall and R4 is on the instructor’s table at $(2.97, 1.6, 0.81)$ m.

The measured magnitude and the group delay of a completely empty classroom at the four different locations are shown in Fig. 2-12(a-d) and Fig. 2-13 (a-d). In the magnitude diagram, x axis is frequency in hertz and y axis is relative magnitude of received energy in dBm. The y axis of the group delay is magnitude of group delay in degree/Hz.
Figure 2-12. The magnitude diagram in a completely empty classroom.

The magnitude diagrams show the effects of frequency selective channel attenuation. Focussing on the band in the 1-1.5 GHz range, one can see that at R1 the channel is highly attenuated in this band. In location R2 the same frequency band is less attenuated. The width of the attenuated band is increased for location R3, but a distinct pass band appears between 850MHz and 1 GHz for this case. Location R4 has the narrowest attenuation in this region and this may be attributed to its closer promixity
to the transmitter relative to the other locations. With respect to the group delay plots in Fig. 2-13(a-d), the location R4 again because of its promixity to the transmitter shows extended regions of the frequency axis where a linear phase assumption may be made.

![Group delay plots](image)

Figure 2-13. The group delay in the empty classroom.

The measured complex transfer functions are filtered by a super Gaussian bandpass filter of order 2, centered at 2.45 GHz center frequency with bandwidth of 1.33 GHz to reduce the effects of sidelobes. The CIRs corresponding to the four
different locations for the case of the completely empty classroom are shown in Figs. 2-14 (a-d). Considering the bandwidth of the filter, the range resolution of the acquired CIR is 0.43 m.

![channel Impulse Response](image1.png)

(a) At R1

![channel Impulse Response](image2.png)

(c) At R3

![channel Impulse Response](image3.png)

(b) At R2

![channel Impulse Response](image4.png)

(d) At R4

Figure 2-14. Normalized CIR in the empty classroom.

In the figure, x axis is the time in seconds and y axis is magnitude in dB. The CIRs may be separated into three different regions. The first part extends from the first peak arrival to about 50-60 nsec and constitutes the coherent region. This region is
composed of line of sight, single and strong double reflections. This region is seen to have significant variation at the different receiver locations. The region of the CIR, where multiply reflected arrivals occur is referred to as the diffuse component and extends from the end of the coherent region upto the noise floor, around −50 dB. A regression analysis of this region, shows that all four locations have nearly same decay rate of −2.27 dB/nsec. This is to be expected, since this region typically depends on the channel dimensions and material parameters of the wall that influence the reflection coefficients. A detailed analysis, relating the dominant features of the CIR with channel features is undertaken next.

2.3.5.1 Analysis of CIR in an Empty Classroom

The origin of the dominant component of the coherent part of the CIR of the completely empty classroom are identified in this section. In the case of location R1, the first peak at 16.9 nsec is identified as the LOS component, based on the transmitter-receiver (T-R) separation of 5.1 m. The path length of the second peak is 5.7 m. The excess delay from LOS component may be due to a single reflection by the floor. Of course, multipath arrivals with delays less than the range resolution of 0.43 m are not resolved in these results. The CIRs show some dominant reflected peaks between 23 and 36 nsecs and they correspond to a travelling length from 6.9 m to 10.8 m. Considering the size of classroom and the location of the receiver, these may be attributed to, as arising from less than 4 reflections.
The mean free path of reverberation in the room is defined by

\[ d = 4 \cdot \frac{V}{S} \]  \hspace{1cm} (2-7)

where \( V \) is the volume of channel space enclosed by surface \( S \). The mean free path of the completely empty classroom is 2.63 \( m \).

Considering the CIR of the empty classroom at location \( R2 \), shown in Fig. 2-14 (b), the T-R distance is 2.8 \( m \). The location of LOS, the first dominant peak, is 8.8 nsec as expected. The second isolated peak arrives at 15.4 nsec and its path length is 4.6 \( m \). This second peak is reflected by the front wall and the incident angle is nearly normal. The third peak arriving at 24.5 nsec is reflected by the metal heater cover in the window side wall with nearly normal incidence angle. It is important to note that the magnitude of this peak is greater than the second isolated peak.

The CIR for location \( R3 \), Fig. 2-14 (c), is affected by the plastic curtain, window and window frame. Line of sight components arrive at 15.4 nsec. The second peak, which is bigger than the LOS, occurs at 17.7 sec and it maybe the reinforcement of signals from different paths which arrives within the range resolution. The distance between the curtain and window glass is 0.27 \( m \). The second peak is the sum of reflections by the window and window frames. There are isolated peaks at 21.4 nsec, 23.5 nsec and 25.8 nsec. The peak at 21.4 is identified as reflected by the ceiling. The peak at 45.8 nsec is reflected by the left side wall.
In location R4, the LOS arrives at 6.8 nsec. Reflections by the floor, front wall and right side wall are detected at 9.9, 10.4 and 12.0 nsec respectively. The signal reflected by the window side and the double reflection by the left and right side wall arrive at 31.8 nsec and 43.9 nsec. After 50 nsec, multiple reflection components arrive at the receiver antenna. Next, a comparative analysis is carried out for the scenario where furniture is added to the empty classroom.

2.3.5.2 Analysis of CIR in the Classroom with Furniture

The CIRs computed for the classroom with furniture are shown in Fig. 2-15(a-d) for the same four receiver locations. In comparison to the empty room scenario, there is no change in the position of the two dominant peaks at location R1 but other singly reflected multipath components are relatively smaller in amplitude in this situation. Some of singly reflected paths in the empty room, are now blocked by the furniture. The location R1 being just in front of student section, amplifies this effect. In the case of location R2, there is no change in the major peak and decay rate. It has nearly same multipath characteristics as in the previous case. However, the depth of the nulls of the CIR at certain delays is decreased. Again this may be attributed to a strengthening of the signal due to the reflections from the furniture. Strong differences in the coherent pattern are observed for location R3 due to the effect of furniture and the random movement of the plastic curtain. In location R4, the multipath pattern is nearly the same except the multipath components which are reflected by the window side and the left
and right wall are blocked or scattered by the classroom furniture. Overall, one can say that the presence of furniture, which serve as scattering objects, reinforce the CIR at time delays where no multipaths occurred in their absence. This effect can serve to degrade the performance since, the ISI power is increased and spectral nulls can arise in the channel frequency response.

Figure 2-15. Normalized CIR in the classroom with furniture.
2.3.5.3 Analysis of CIR during a lecture in the classroom

![Channel Impulse Response (a) At R1](image1)

![Channel Impulse Response (b) At R2](image2)

![Channel Impulse Response (c) At R3](image3)

![Channel Impulse Response (d) At R4](image4)

Figure 2-16. Normalized CIR during the session.

Fig. 2-16(a-c) shows the CIRs during a lecture session. Here four people are seated on the student side and one person serves as an instructor in the front. In this case, at the receiver locations, LOS components are still available and are not blocked by the people. However, one multipath component in the second group of arrivals is blocked by the human body. And new stronger multipath components are detected.
between 36 and 54 nsec in the location \( R1 \). Some of the single and double reflections at location \( R2 \) which arrive between 39 nsec and 64 nsec, are blocked by the body.

Figure 2-17. Normalized CIR during the class intermission and time-varying conditions.

Stronger multipath components are also identified between 25 nsec and 55 nsec which maybe the reflections by the human body. These are not evident in the previous two scenarios. These features again result in increase in ISI power. Some of single reflection components of location R4 also blocked by the body. Depending on the
relative location of humans and transmitter and receiver antenna, a number of dominant multipath components are blocked and a few paths are newly identified.

2.3.5.4 Analysis of CIR during the Class Intermission (Time-varying channel)

The CIRs for this scenario are shown in Fig. 2-17. In this environment, three students are still sitting on his or her chair, two students are walking around and one student is talking to the instructor. The LOS path between transmitter and receiver is in most all cases blocked by the human body. And some of dominant single reflections are blocked also. The coherent region is markedly different for all four locations. The coherent region is typically shorter than in the other three scenarios, however the decay rate in the diffuse region is nearly the same as with other scenarios.

2.3.6 Measurements in University CACT Research Laboratory

The channel measurements described in this section are made in the Center for Advanced Computation and Telecommunications (CACT) which is located in Falmouth 203. The environment is a typical university research laboratory and indoor wireless environment. The room dimensions are (16.56, 10.26, 3.15) m and the lab contains tables, desks, chairs, computers, shelves, partitions, and research equipment. The view and schematic of the laboratory and the placement of transmitter and receiver are shown in Fig. 2-18. The picture is taken from lower left corner to the direction of upper right corner. There are two doors and two windows at the far wall in the picture and another
two windows and unused metal plated door in opposite wall, which is not shown in the picture. The wood floor is covered by plastic tiles and the ceiling is made of celotex acoustic tiles and fluorescent lighting. There are three round metal columns with 8.64 cm radius, aligned near the far wall.

The channel transfer function of CACT is measured at four receiver locations, denoted as RC1, RC2, RC3 and RC4. The transmitter, $Tr$, is located towards the far call and approximately the center of the laboratory and its coordinates are (7.93, 1.13, 1.07) m. The coordinate of RC1 is (6.58, 5.06, 0.69) m, RC2 at (7.44, 3.3, 0.69) m, RC3 at (9.63, 6.38, 0.76) m and RC4 at (10.49, 3.45, 0.76) m. The location of receiver is selected based on the different impediments and obstructions they are characterized by. The location RC1 is blocked from the transmitter by one metal plated partition, one metal frame and fabric plated partition and two metal cabinets. RC2 is blocked by one fabric plated metal frame partition. RC3 is the LOS case and RC4 is blocked by two metal plated partitions.

The measured magnitude diagrams of the channel transfer functions of the four different locations in CACT are shown in Fig. 2-19(a-d) and the group delays are shown in Fig. 2-20(a–d). The locations they correspond to are indicated on the figures. Relative to the classroom measurements, the measurements of the transfer function magnitude inside CACT are highly attenuated at all locations at the frequency region greater than 500 MHz. The magnitude of around 300 MHz is increased in comparison to the results of empty classroom except for location RC4.
There are a number of deeply attenuated regions for this location, one at 1 GHz and two at around 2 GHz and one at 2.8 GHz. A stronger band exists around 1.7 GHz and 2.45
GHz with the magnitude of -60 dBm. The group delay plots show that the high and flat magnitude region, between 250 and 350 MHz, exhibits approximate linear characteristics. In RC2, there are highly attenuated regions from 0.7 GHz to 1.5 GHz and less attenuated narrow bands from 2.1 GHz to 2.5 GHz. The measurements for receiver location RC3, show a flatter region than for any other location in the region from 2.0 GHz to 2.7 GHz and its magnitude is roughly -53 dBm.

Figure 2-19. The magnitude diagram in CACT.
The group delay between 2.5 GHz and 2.8 GHz shifts from zero, indicating a quadratic dependence on frequency. In location RC4, the magnitude changes fast with respect to frequency in all of the regions. There is a wider pass band between 1.1 GHz and 1.8 GHz and the group delay of this region also shifts from zero. The complex transfer functions for CACT are changed dramatically as the receiver position is changed.

Figure 2-20. The group delay in CACT.
The CIRs obtained from the transfer functions are shown in Fig. 2-21. The 2nd order super Gaussian window is used to reduce the sidelobe effect. The CIRs at different locations can again be characterized into coherent, diffuse and background noise regions. At all four locations, the background noise start at around 150 nsec with the nearly same magnitude of $-50\text{dB}$. In this channel, the location RC1, RC2 and RC4 are non-line of sight (NLS) cases and RC3 is line of sight (LOS) case.

Figure 2-21. Normalized CIR in a CACT (Falmouth 203).
A number of dominant multipath components and LOS arrivals are identified in the results for CACT. In location RC1, the T-R distance is 4.2 m and its estimated arrival time is 14.0 nsec. But there is no peak in that time because the LOS component is blocked by several obstacles. Most dominant peak arrives at 19.5 nsec and its travelling length is 5.9 m. It is reflected by the cabinet which is located in the upper right corner of the receiver. A peak arrives at 78.4 nsec and its distance is 23.5 m. Estimated propagation path of this peak is a first reflection by the metal plated partition which is located between RC3 and RC4, after that it is reflected by the unused metal door on the wall and finally reflected by the cabinet which is to the right of the receiver. Another isolated peak is detected at 108.8 nsec and its magnitude is -15.3 dB.

The receiver location RC2 is blocked by one partition which is made of a metal frame and both sides are fabric plated. From Fig. 2-21 (b), the first arrival is detected at 7.0 nsec, 2.1 m, with -1.7 dB magnitude and calculated LOS distance is 2.26 m and travelling time is 7.5 nsec. Although there are some small differences, the estimates are reasonable. The second peak is detected at 10.7 nsec, 3.2 m, at 0 dB. In this case, LOS component is attenuated when it penetrates the partition but the second peak is reflected by the metal partition which is in upper left location. The estimated travelling path of the peak at 55.2 nsec, 16.56 m, is reflected by the computer body which is placed at the lower part of RC3 desk section.

The location RC3 is the LOS case and the T-R distance is 5.5 m. The LOS peak is
detected at 18.7 nsec and calculated distance is 5.61 m. This confirms that the dominant peak is the LOS component. Another peak is detected at 25.5 nsec, 7.7 m, which is reflected by the computer which is at the lower end of RC3 desk.

The partitions which enclose RC4 are all metal plated ones. The distance between transmitter RC4 is 3.5 m but the first arrival of peak is 26 nsec and its travelling distance is 7.8 m. The line of sight (LOS) component is blocked by the partitions. It is estimated that first peak is a multiply reflected incoming wave because the magnitude is just 28.0 dB bigger than the background noise.

In the case of location RC1, all the detected peaks are multiply reflected waves except the first arrival which is singly reflected by the metal plate. In this lab environment, the received magnitude of multiple reflected wave starts around 25 dB greater than the background noise. The same is applicable to location RC2 and RC3. In RC2 case, the coherent region is identified from the first arrival to about 65 nsec. After the peak at 71.6 nsec or 21.5 m, all arrivals are multiply reflected waves and the magnitude of 71.6 nsec is -14.5 dB. Location RC3 also has similar patterns after the peak at 59.5 nsec. The magnitude of the 59.5 nsec peak is -16.6 dB. All the arrivals of RC4 are multiple reflections or a few reflections by higher relative permittivity materials. The slopes between 50 nsec and 150 nsec were -0.1168 dB/nsec for RC1, -0.1624 dB/nsec for RC2, -0.175 dB/nsec for RC3 and -0.00914 dB/nsec for RC4. From these results, the pattern of CIRs of the partitioned large room are different at different places. The received power is no more a function of transmitter and receiver separation distance.
and it highly depends on the structure and materials of the room and geometrical configuration of transmitter and receiver.

2.4 Summary

Channel measurements were conducted based on the frequency sweep method using a vector network analyzer. An Agilent 8753ES vector network analyzer was used to measure complex indoor wireless channel transfer function with Astron AXQ24SM-A quarter wave omni-directional monopole antenna as a transmitter and a receiver. The channel impulse responses are calculated from measured channel transfer functions using inverse Fourier transforms. To reduce the Gibbs’ effect, 2nd order super Gaussian function is adapted as a window.

Complex transfer functions are measured in 3 different places, anechoic chamber, a classroom and CACT research laboratory. The measurements in a classroom are conducted under the four different scenarios, each identifying one or more new components in the channel contributing to multipath.

One of UMass Lowell research laboratory (CACT) is selected to measure channel characteristics. The laboratory, typical indoor environments, contains partition walls, tables, desks, chairs, computers, shelves, and research equipment. Some of them such as partition walls, desks and shelves are composed by high conductive metal plate. From the results in CACT, the received power is no more function of transmitter and receiver separation distance. It is highly depend on the structure and materials of the room and
the geometrical configuration of transmitter and receiver.
3. Channel Characterization using Rational Function Models

3.1 Introduction

In this chapter, the CIRs for an empty room are modeled using rational function models, where the channel transfer function is characterized in terms of its poles and zeros. The transfer function $H(s)$ relates the input and output of a linear system in the complex frequency domain, where $s = \sigma + j\omega$ is the complex frequency variable. The complex transfer function evaluated at $s = j\omega$ represents the frequency response of the system $H(j\omega)$. It is specified as $H(j\omega) = \frac{X(j\omega)}{Y(j\omega)}$, when $Y(j\omega)$ and $X(j\omega)$ are Fourier transforms of the input and output of the system. Note that if $x(t) = \delta(t)$, $X(j\omega) = 1$ and $H(j\omega)$ represents the Fourier transform of the impulse response of the system. The transfer function of a system is a measure of how the system responds to any, and all, frequencies, that it is excited with. A rational function model is often used to characterize the transfer function. This consists of a ratio of two polynomials in $\omega$. The roots of the numerator polynomial are the zeros the frequency response and characterize the distribution of spectral nulls in frequency space. The roots of the denominator are referred to as the poles of the transfer function and represent the characteristic or resonant frequencies of the system. The poles and zeros are the unique roots that remain
after accounting for any pole-zero cancellations. The poles and the zeros are useful for analyzing the dynamics and stability of the system. In the context of the indoor wireless channel impulse response, the zeros are important for determining the frequency selective fading characteristics of the channel and the poles determine the energy in the tail of the channel impulse, its decay rate and therefore its influence on performance issues such as intersymbol interference and background noise.

First, a one dimensional free space model is analyzed to understand the relationship between geometric configuration of transmitter and receiver and poles and zeros of the channel transfer function. The channel transfer function is explicitly derived in terms of the channel parameters. The channel impulse response and poles and zeros are calculated from this model.

The experimental results are also compared with a computational model proposed by Thompson \cite{42} which applies the method of images to calculate the CIR for an empty rectangular channel. The image based solution to the CIR is an approximation to the exact solution of wave equation. It approaches the exact solution as the walls of the room approach rigid conditions. This method can provide better time granularity in distinguishing peaks in the CIR than ray tracing methods. This is important for analyzing performance of high frequency, wideband systems\cite{43}.

Using the calculated and modeled CIRs, the effects of the movement of the receiver are simulated and statistically characterized. The fading statistics are
characterized with respect to the received energy as a function of receiver movement.
Using these results the indoor wireless channel is characterized.

The 1-D lossless channel space and its transfer function are derived in Section 3.2. The measured CIR in a 3-D empty room is compared with the image source model in Section 3.3.

3.2 1-D Lossless Wireless Channel

A simple and ideal 1-D lossless space is considered as an indoor wireless channel to understand the dynamics of poles and zeros of the channel transfer function. The wave propagates along a single axis without loss and is only attenuated when it hits the boundary surfaces. The poles and zeros are a function of the geometric configuration of transmitter and receiver, dimension of the channel space and reflection coefficients of the boundary.

The structure and mechanics of wave propagation in the 1-D lossless wireless channel is shown in Fig. 3-1. The horizontal axis is represented in units of $\Delta x$. The transmitter, $Tr$, is located at the $k$th interval and the receiver, $Re$, is at the $m$th sample interval. The transmitted wave is attenuated at the boundaries and the reflection coefficients of both walls are set to $a$. The assumption is made that both the transmitter and receiver are volumeless. Another assumption is that the wave radiates in two directions, to the boundary $A$, the right traveling wave (RTW) and to the boundary $B$, the left traveling wave (LTW), from the transmitter. The channel extends for $n$ sample
A unit impulse is assumed to be generated from the transmitter. The signal at the receiver consists of the directly received pulse and additional delayed and attenuated echoes generated by reflection from the walls A and B. The impulse response generated by the initial RTW towards A and that generated by the LTW towards B are shown in Figs. 3-2 (a-b). The T-R separation $d = m - k$. Let $r = m + k$. The $x$ axis is time in sample intervals and $y$ axis is the magnitude of received wave amplitude relative to the direct component.

![Figure 3-1. Structure and wave propagation in 1-D lossless channel.](image)

For the right radiation case shown in Fig. 3-2 (a), the first arrival is the LOS component with magnitude of one and the traveling distance in sample time units is $d$. After the first arrival, the rest of the arrivals depend on the location of the receiver, $m$, and $d$. The arrival times of successive reflected components are given in the figure. The magnitude of the received amplitudes is proportional to $a^{nr}$, where $n_r$ is the number of
reflections. In the case of left radiation, the only difference is the magnitude and arrival time of the first arrival. The magnitude of the first arrival is $a$ and the arrival time is $r$.

Figure 3-2. Channel impulse response of 1-D space.

### 3.2.1 Derivation of Transfer Function

The channel transfer function, $H_{1D}(z)$ of the 1-D lossless bounded channel can be extracted from CIR shown in Fig. 3-2. Each peak in the figure is expressed by its
magnitude and its delay.

\[ H_{1D}(z) = z^{-d} + a z^{-(2(n-m)+d)} + a^2 z^{-(2(n-m)+2m+d)} + a^3 z^{-(4(n-m)+2m+d)} + a^4 z^{-(4(n-m)+4m+d)} + a^5 z^{-(6(n-m)+4m+d)} + a^6 z^{-(6(n-m)+6m+d)} + \ldots \]

\[ + a z^{-r} + a^2 z^{-(2(n-m)+r)} + a^3 z^{-(2(n-m)+2m+r)} + a^4 z^{-(4(n-m)+2m+r)} + a^5 z^{-(4(n-m)+4m+r)} + a^6 z^{-(6(n-m)+4m+r)} + a^7 z^{-(6(n-m)+6m+r)} + \ldots \]  

(3-1)

The arrival peaks of right and left radiations are expressed in Equation (3-1). The common term of right radiation \( z^{-d} \) and for the left radiation it is \( a z^{-r} \). Extracting these terms and rearranging the equation,

\[ H_{1D}(z) = (z^{-d} + a z^{-r}) \left( 1 + a z^{-(2(n-m))} + a^2 z^{-(2n)} + a^3 z^{-4n+2m} + a^4 z^{-4n} + a^5 z^{-6n+2m} + a^6 z^{-6n} + \ldots \right) \]  

(3-2)

The term in the second parentheses is simplified further as,

\[ H_{1D}(z) = (z^{-d} + a z^{-r}) \left( 1 + a z^{-(2(n-m))} \right) \left( 1 + a^2 z^{-2n} + a^4 z^{-4n} + a^6 z^{-6n} + \ldots \right) \]  

(3-3)

Representing the third term by the geometric series sum \( \frac{1}{1 - a^2 z^{-2n}} \), the channel transfer function of the 1-D lossless channel is expressed as,

\[ H_{1D}(z) = \frac{z^{-(m-k)} + a z^{-(m+k)} + a z^{-(2(n-m)+(m-k))} + a^2 z^{-(2(n-m)+(m+k))}}{1 - a^2 z^{-2n}} \]  

(3-4)

The order of the denominator is \( 2n \) and numerator order is max \( (m-k), (m+k), (2(n-m)+(m-k)), (2(n-m)+(m+k)) \). The number of poles and zeros are the order of denominator and numerator minus any pole-zero cancellations. From Equation (3-4)
or Fig. 3-2, the denominator order 2n is larger than that of the numerator. The channel impulse response is calculated as

\[
h_{1D}(i) = a^2 h(i - 2n) + \delta (i - (m - k)) + a \delta (i - (m + k)) + a \delta (i - 2(n - m) - (m - k)) + a^2 \delta (i - 2(n - m) - (m + k))
\] (3-5)

If all of the zeros of Eqn. (3-4) are located inside the unit circle then the transfer function is said to be minimum phase\[44\]. The inverse filter of the channel is the reciprocal of Eqn. (3-4) and expressed by

\[
H_{1D}^{-1}(z) = \frac{1 - a^2 z^{-2n}}{z^{-(m-k)} + a z^{-(m+k)} + a z^{-(2(n-m)+(m-k))} + a^2 z^{-(2(n-m)+(m+k))}}
\] (3-6)

The order of the numerator is now greater then the denominator. A time shift of the filter is therefore required to implement a causal time domain inverse filter. The amount of the shift is the difference between 2n and max \(d, r, 2n - r, 2n - d\).

If the magnitude of the number of zeroes of the Equation (3-4) are greater than one, then the inverse transfer function will be unstable. A stabilization process is needed when the poles of Equation (3-6) are located outside the unit circle\[40\]. The unstable filter can be modified to a stable filter that has the same amplitude characteristic. Consider the digital filter with a pole that is outside the unit circle

\[
H(z) = \frac{N(z)}{D(z) (z - r e^{j\theta})}
\] (3-7)
Where \( r \) is greater than 1. The pole \( r e^{j\theta} \) is outside the unit circle. The filter is reformed by a reciprocal pole.

\[
H'(z) = \frac{N(z)}{D(z)} \frac{1}{r (z - r^{-1} e^{j\theta})} = \frac{N(z)}{D(z) (r z - e^{j\theta})}
\]  

(3-8)

It is obtained by replacing the pole \( r e^{j\theta} \) by \( r^{-1} e^{j\theta} \) and introducing a constant \( r \). The pole \( r^{-1} e^{j\theta} \) is called the reciprocal pole of \( r e^{j\theta} \). After stabilization using the reciprocal pole, the amplitude characteristics of \( H'(z) \) is the same as that of the original transfer function \( H(z) \) but the phase characteristics of stabilized transfer function is different from that of \( H(z) \). If the modulation schemes of the communication system depend only on the magnitude such as in binary phase shift keying (BPSK), then the stabilization process is applicable but for modulation schemes that depend on both magnitude and phase such as in quadrature phase shift keying (QPSK) then it cannot be used.

### 3.2.2 Dynamics of Poles and Zeros

The size of the 1-D space is set to \( n = 18 \) sample units for the simulation and the location of transmitter and receiver are \( k = 3 \) and \( m = 10 \) sample units respectively. The reflection coefficients of both boundaries are assumed to be \( a = 0.7 \). The CIR of the 1-D lossless space obtained by substituting these values in Eqn. (3-5) is shown in Fig. 3-3. In the figure, \( x \) axis is time in sample units and \( y \) axis is the magnitude.
The dynamics of poles and zeros are investigated using the transfer function of the 1-D lossless channel. From Equation (3-4), it is seen that the poles are affected by the sample space dimension \( n \) and reflection coefficients \( a \). The locations of the zeros are influenced by \( n, a \), and both the location of transmitter and receiver, \( k \) and \( m \) respectively. The effects on poles and zeros due to a change in reflection coefficient amplitude are shown in Figs. 3-4 (a-b). Fig. 3-4 (a) depicts the \( 2n = 36 \) poles in the Z plane. The magnitude of all of the poles is at \( z = n\sqrt{a} \). The phase angle is uniformly distributed from \( 0:2\pi \). By increasing the reflection coefficient of the boundaries both the pole and zero show outward motion. The maximum order of the numerator is \( 2n - d = 29 \) and minimum order is \( d = 7 \). There are 22 nonzero zeros. The distribution of zeros in the Z plane are depicted in Fig. 3-4 (b). In the case of zeros, the zeros which
are close to the origin move faster and approach the unit circle when the reflection coefficient, $a$ approaches 1.0. Both poles and zeros show no change in angle as $a$ is changed. The solution to the roots of the both numerator and denominator is obtained by the using the function $cpoly$ in the Bell labs port library package.

Figure 3-4. The effect of change in reflection coefficient, $a$, on pole-zero locations.
Figure 3-5. The effects of changing channel dimension, $n$, of the 1-D sample space, $a = 0.7$.

The effect of changing the dimension of the channel is shown in Figs. 3-5(a-b). The poles and the zeros are calculated for $n = 14, 16, 18$, keeping the transmitter and the
receiver locations at 3 and 10 respectively. Since the number of poles is $2n$, there is an increase in number of poles as the channel dimension increases. The magnitude remains same at $z = n\sqrt{\alpha}$. As $n$ increases new poles are added. The locations of the pole are moved along the circle from the real axis to the imaginary axis by the decrease of $n$. The number of nonzero zeros, $2(n-d)$ also increase in number linearly when the location of the transmitter and the receiver is fixed. It is not as readily predictable how the location of zeros change with increase in channel dimension, since the zeros are also strongly related to the placement of the transmitter and receiver.

Next, the effects of mobile movement on poles and zeros are investigated. Movement close to the transmitter considering, $m = 5, 6, 7$, and far from the Transmitter for $m = 12, 13, 14$ are considered. The locations of zeros are given in Figs. 3-6 (a-b) for these two cases. The channel dimension and reflection coefficient are fixed for all cases. The locations of the poles are not affected by the change in receiver locations. When the mobile approaches the right boundary, $B$, the order of the numerator of the transfer function is decreased. This means that the number of zeros are decreased. The zeros move from the real axis towards the imaginary axis when the mobile approaches the right wall. However, some of the zeros are invariant with change in receiver location. There are six such zeros. Two are complex zeros located at $(0.8160, 0.4711)$, $(-0.8160, -0.4711)$ and their complex conjugate values. The other two zeros found invariant lie on the imaginary axis. These zeros are changed by only $10^{-5}$ in distance and therefore considered invariant. The locations of fixed zeros depend on the fixed
components such as the transmitter or receiver location. If both transmitter and receiver are moved then no fixed zeros are detected.

Figure 3-6. The effects on zeros by receiver movement in 1-D sample space.
Based on Eqn. (3-4) and the simulation results one can conclude that the poles or denominator of the transfer function are affected by the sample space dimension \( n \), and the reflection coefficients of the boundaries, \( a \). The zeros or numerator are affected by \( n \), \( a \) and geometrical configuration of transmitter and receiver. In the ideal case if either the transmitter or receiver is fixed, then some invariant zeros may be found that characterize the channel.

3.3 Computation of CIR of an Empty Room

A computational approach for estimating the channel impulse response of an empty rectangular room has been proposed in \([45]\). It is an extension of the results by Allen and Berkley\([24]\) by including the effect of microphone directivity. This result is compared with the measurements obtained for an empty classroom. A series of channel impulse responses are calculated based on the movement of the mobile. The variation of received energy is acquired based on the calculated channel impulse responses. The stochastic characteristics of received energy are also investigated in this section.

3.3.1 Image Method

In the image method, the walls of the enclosure are replaced by point sources of varying strength and location. The component amplitude of each image is chosen such that the transverse electric and magnetic fields are continuous across the boundary. The image source in these cases is the summation of a point source having an amplitude
equal to the plane wave reflection coefficient.

![Figure 3-7. The concept of 2-D image method.](image)

The physical concept of the image method in 2-D rectangular space is given in Fig. 3-7. The rectangular region in the middle is the enclosure which contains the transmitter $T_{0,0}$ and receiver $R$. The 2-D image space and its transmitter and receiver configuration are also shown in the picture [24]. Each grid has an image transmitter point, $T_{i,j}$. When the transmitter is excited, each image point is simultaneously excited, creating spherical waves which propagate away from each image point. The solid line between $T_{0,0}$ and $R$ is the line of sight component and the dashed line which is reflected 2 times has the same wave propagation distance and incident angles as that generated by the image location $T_{1,-1}$. The exact solution is acquired when grid space and number of images are expanded to infinity.
In the general three-dimensional case, the \( j \)th coordinate directional component of the Hertz potential of transmitter location, \( \vec{x}^s \) and receiver location \( \vec{x} \) is expressed by

\[
\Pi_j(\vec{x}, \vec{x}^s, t) = \sum f_j \left( \frac{ct}{X_K} - 1, K \right) \prod_j X_K
\]

where \( c \) is speed of light, \( f_j \) is component amplitude of each image and \( X_K \) is the distance between receiver and \( K \)th image. The sampled \( j \)th component of electric field, \( E_j \) for discrete time and space is given by

\[
E_j(\vec{x}, \vec{x}^s, n) = \sum_{p=1}^{3} \sum_K \frac{\cos(\beta_j) \cos(\beta_p) - \delta_{jp}}{M_K} \times \left\{ \frac{\dot{f}_p(n/M_K - 1, K)}{M_K} + \frac{\ddot{f}_p(n/M_K - 1, K)}{M^2_K} + (3 - \delta_{jp}) \frac{f_p(n/M_K - 1, K)}{M^2_K} \right\}
\]

where \( \beta_j \) is the distance of \( j \)th component projected on the \( \vec{X}_K \), \( M_K \) is nondimensional distance between transmitter and receiver and \( \dot{\ } \) and \( \ddot{\ } \) correspond to the first and second derivatives. More detailed derivation is given in the reference\[43\].

Based on Equation (3-10), the channel impulse response of an empty room is calculated and shown in Fig. 3-8 (a) and for the comparison, the measured channel impulse response of the empty classroom is shown again in Fig. 3-8 (b) and the \( y \) axis of both results is normalized to a maximum of 0 dB. The size of the room, \((7.26, 5.74, 2.85) \) m for the image model is set to be the same as that of the empty
classroom. The transmitter is located at the same place, (0.91, 0.91, 1.04) m and the receiver is at the R1 location, (5.84, 2.34, 0.81) m. The relative permittivities of the three concrete side walls are set to 2.5, the window side wall is set to 3.0, the floor permittivity is set to 3.0 and for the ceiling it is 2.2. All of the permittivity values are approximately chosen to represent the materials of the wall, floor and ceiling in measurement environment. In the case of the window side wall, since there are two different materials, 2/3 of window and 1/3 of metal plate. The relative permittivity of the glass can range from 4.5 to 10 and for the metal plate a value of one is appropriate.

As can be seen in the picture, both computational and measured results have a similar decay rate in the diffuse region. Some similarities exist in the location of LOS peak and the second dominant peak. But the shape and decay pattern of the coherent regions exhibit some differences. The difference may come from the non-ideal
conditions in the classroom and in the transmitter and receiver selectivity functions. The surfaces of the classroom are also not perfectly flat, particularly the carpet covered floor and the window side wall which is composed of different materials. The radiation and receiving pattern of antenna is omni-directional only in the horizontal plane and both antenna are not the volumeless source and detector, assumed in the computations.

The reflections from rough surfaces induces different propagation effects. The roughness of surface is tested by the Rayleigh criterion which defines a critical height, $h_c$

$$h_c = \frac{\lambda}{8 \sin \theta_i}$$  \hspace{1cm} (3-11)

where $\lambda$ is wavelength with incidence angle, $\theta_i$. The scattering effects are introduced by the roughness of the surface. The scattering loss factor, $\rho_s$, of a rough surface is derived by Ament\cite{46} and modified by Boithias\cite{47}. The basic assumption is that the variation of the height is Gaussian distributed with standard deviation, $\sigma_h$.

$$\rho_s = e^{\left[-8 \left(\frac{\pi \sigma_h \sin \theta_i}{\lambda}\right)^2\right]} I_0 \left\{8 \left(\frac{\pi \sigma_h \sin \theta_i}{\lambda}\right)^2\right\}$$ \hspace{1cm} (3-12)

where $I_0$ is the Bessel function of the first kind and zero order. The reflection coefficients of the rough surface, $\Gamma_r$ are measured by Landron et al.\cite{48} and is
Using this scattering phenomenon, it may be explained that the diffuse region of the measured channel impulse response is more smoothly decaying than that obtained by the method of images.

3.3.2 Spatial Variation of Received Energy

The effect of movement of the mobile is investigated using the CIR derived from the image method. The received energy is calculated along the trajectory of the mobile in an empty room. The size of the room is kept the same as the classroom dimensions. The reflection coefficients of the room are assumed to be the same values as in the previous section. The transmitter is also located at the same place (R1). The bandwidth is set to 30 GHz in the computational model, providing a time resolution of 0.03 nsec and a spatial resolution of 1 cm. The mobile is moved 100 sample units, 1 m, from left of location R1 to 1 m to the right of it. The movement is parallel to the x axis.

The received energy is calculated at every sample unit along the mobile path. The dominant LOS component is removed, to consider LNOS effects in the calculation of the received energy. This allows the examination of the worst case situation. The calculation procedure for the received energy is as follows. First, the time domain channel impulse response, $h(n, l)$, is calculated at mobile location index $l$ for the given locations of transmitter and receiver using the method of images given in Eq. (3-10),

$$\Gamma_r = \rho_s \Gamma$$  \hspace{1cm} (3-13)
there \( n \) is time index, representing time samples at 0.03 nsec resolution. Second, the CIR is Fourier transformed to yield \( H(m, l) = FFT[h(n, l)] \), where \( m \) is the frequency index. The real and imaginary parts of the frequency response are multiplied by a bandpass filter, \( H_{NB}(m, l) = H(m, l) F(f_c, w) \). In this simulation, the bandwidth, \( w \) is 14.7 MHz with 2.45 GHz center frequency, \( f_c \). All the components of bandpass filtered transfer function are summed over entire passband range to yield \( E(l) = \sum_m H_{NB}(m, l) \).

This summed value, \( E(l) \), is the time domain component, \( h_{NB}(0, l) = \sum_m H_{NB}(m, l) \), of the band limited channel impulse response and it is used to represent the signal received at the specific receiver location, \( l \). Finally, this function is upsampled using a \( sinc \) function as,

\[
E_S(r) = \sum_l h_{NB}(0, l) \cdot \text{sinc}\left(\pi \left(\frac{r}{N_u} - l\right)\right)
\]

where real variable, \( r \) is upsampling rate \( N_u \) times location index \( l \). The variable \( r \) is related with the locations of both the transmitter, \( x^t \), and the receiver, \( x \), in the channel.

The resulting fading envelope, \( |E_S(x, x^t)| = \sqrt{E_{Sr}^2 + E_{Si}^2} \) of the empty room is shown in Fig. 3-9. The x-axis is the location from the position \( R1 \) in meters and y axis is normalized magnitude in decibels. \( E_{Sr} \) is real component of received signal and \( E_{Si} \) is the imaginary part. As shown in the picture, the received signal is changes very rapidly with small amount of mobile movement. During the fade times, loss of data and
communication can take place. If the speed of the mobile is 1.8 km/hr then it takes 4 seconds to move 2 m. With this mobility condition, if the fading threshold is assumed to be less than -2 dB, the system experiences on average a fade every two seconds. If the fade threshold is raised to –1 dB, then the system encounters 1 fade every second. If the speed of the mobile is doubled, then the fade rate is doubled too.

![Figure 3-9. The fading envelope of the empty room.](image)

The received signal is calculated based on the image method and the fading statistics are calculated using the received signal variation when the receiver position changes. To model the envelope fading, the signal is separated into the real and imaginary parts. The real and the imaginary parts are averaged by a moving average filter using a sample size of 12 points. The resultant diagram is shown in Fig. 3-10.
In the picture, the dotted line is the real and imaginary parts of the signal and the solid line is its moving average result. The inphase and the quadrature components are composed of two processes, one is the mean trend, $\bar{E}_S$ and the other is the residual process, $E_S - \bar{E}_S$. The mean process is relatively slowly changing with distance, but the residual process changes faster. In Fig. 3-10, the solid line is the mean process and the difference between solid and dotted line is the residual process.

![Figure 3-10. The real and the imaginary component of received signal.](image)

(a) Real part  
(b) Imaginary part

3.4 Summary

In this chapter, the indoor wireless channel is analyzed considering two computational models. A 1-D lossless model is analyzed to evaluate the dependence of poles and zeros of transfer function with channel parameters. The other model is the image source based computational model that generates the CIRs for a three dimensional empty rectangular room. Based on the results of 1-D lossless channel, the
channel impulse response is calculated and the dynamics of pole and zero are anticipated. The number of poles is proportional to the channel dimension and reflection coefficients. The locations of the poles are also affected by the size of the sample space and the reflection coefficients of the boundary. The locations of the zero are influenced additionally by the geometrical configuration of transmitter and receiver. Both the locations of pole and zero move outward by the increase of reflection coefficients of the boundary. The poles and zeros transition from the real axis to imaginary axis along the circle with decrease in the channel dimension. In this case, the number of poles and zeros are reduced. As mentioned earlier, the poles are not affected by the movement of the receiver and transmitter. Under the receiver movements, some of the zeros are found to be invariant.

The channel impulse response of a rectangular room was calculated using the method of images. The size of the room and the geometrical configuration of transmitter and receiver are kept the same with the measurement scenario of the empty classroom case. The initial region of the CIR consisting of the LOS peak and the second peak, has nearly the same structure and arrival time as the measurements, but significant differences exist in other regions. The difference may come from non-ideal conditions of classroom and transmitter and receiver such as radiation pattern of transmitting and receiving antenna, roughness of the enclosure surfaces and the volume of transmitter and receiver, all of which are not modeled in the computational results.

The spatial variation of received energy is calculated based on the channel
impulse response which is calculated using the image method. The received energy is separated into the real and imaginary parts. The local trend effect is separated out by a moving average filter. The variation of received energy is composed of two processes, the mean process and residual process. The mean process is highly location dependent. The residual process describes the change of received energy within a limited area.
4. Numerical Analysis of Wideband Channel Impulse Response

4.1 Introduction

The wideband channel characteristics are investigated using the channel measurements made in the classroom and in CACT. The wideband channel impulse responses that are shown in Chapter 2 are used to analyze and characterize the indoor wireless channel. Two different approaches are used to analyze the affect on wireless communication due to the fading effects of the channel. One is the delay spread characteristic of indoor CIR and the other describes the channel using the pole-zero model.

The indoor wireless channel is inherently a multipath channel. If the change or motion of objects inside the channel is slow relative to the data rate, then the channel can be assumed nearly static. For such cases the time of arrival of the dominant peak is nearly fixed during the transmission of several symbols. The chance of communication error increases due to the increase of the power in the multipath components. The multipath components then contribute to intersymbol interference (ISI). There are a number of different measures to quantify the effects of communication errors. The mean excess delay\textsuperscript{[49]}, root mean square (RMS) delay spread\textsuperscript{[50]}, maximum excess delay\textsuperscript{[2]},
coherence bandwidth\textsuperscript{[51]} and maximum reliable data rate\textsuperscript{[28]} are some of the metrics that have been proposed. The measured wireless channel is characterized using some of the aforementioned parameters and their effect on data rate is analyzed.

From the results of 1-D lossless space, one can estimate that a total of $8 \times (n_L \times n_W \times n_H)$ poles are needed to perfectly describe the 3-D indoor channel. Here $n_L$, $n_W$, $n_H$ are the length, width and height of the channel space in sampling units. A limited number of dominant poles and zeros are therefore used to describe the indoor channel. The measured delay parameters are determined in the next section and the pole-zero model is given in Section 4-3.

\section*{4.2 Channel Delay Parameters}

The channel impulse response $h(n)$ obtained by IFFT of the measured channel transfer function is considered here. The interval between discrete pulses, 0.724 nsec depends on the bandwidth of the window 1.382 GHz, which is used in the IFFT. The discrete multipath CIR is expressed by Equation (2-2). Using the channel impulse response, the power delay profile (PDP) is defined by\textsuperscript{[52]}\textsuperscript{[49]}

$$P(\tau, t) \equiv | h(\tau, t) |^2$$ (4-1)

This relationship between power delay profile and channel impulse response is valid if the transmitted pulse is much smaller than the CIR\textsuperscript{[51]}\textsuperscript{[30]}.

The mean excess delay (MED) of power delay profile is defined by
\[ m_\tau = \frac{\sum_n P(\tau_n) \cdot \tau_n}{\sum_n P(\tau_n)} \]  

(4-2)

The mean excess delay is the weighted center or mean delay of the power delay profile. The received signal is synchronized at the instant of maximum peak arrival time for detection. If the difference between maximum peak arrival time and mean excess delay is increased then the probability of synchronization error is also increased.

The RMS delay spread (RDS) is the second central moment of the power delay profile and expressed by

\[ \sigma_\tau = \sqrt{\frac{\sum_n (\tau_n - m_\tau)^2 \cdot P(\tau_n)}{\sum_n P(\tau_n)}} \]  

(4-3)

The RMS delay spread gives information about the dispersiveness of the wireless channel and it highly influences the channel transmission error.

The maximum excess delay is the time period for the level of power to go down a certain level from the peak value. The received power which is smaller than this level is considered as noise in the communication system. The maximum excess delay also gives characteristics of the channel which relate to the communication error. The size of the signal detector such as an equalizer maybe determined based on the maximum excess delay.

The coherence bandwidth is the statistical average bandwidth of the channel
based on the channel correlation function and is defined by

\[
B_c = \frac{1}{\alpha \sigma_{\tau}}
\]  

(4-4)

where \(\alpha\) is a constant. If the PDP is of exponential form then \(\alpha = 2\pi^{[53]}\). Empirical work on coherence bandwidth is presented by Howard and Pahlavan\([32]\) who evaluate \(\alpha\) for various channel environments. The coherence bandwidth is a function of both the decay rate of transmitted signal with respect to distance and the channel RMS delay spread. The maximum reliable data rate is given by \(D_m = \frac{1}{4\sigma_{\tau}}\). The constant factor 4 is derived by Glance and Greenstein\([28]\). This is a rough indication of the maximum data rate supported by the channel without equalizer or diversity techniques. The physical meaning of maximum reliable data rate is nearly the same as coherence bandwidth.

4.2.1 Measurement Environments

The delay parameters are acquired for the four different receiver locations in a classroom and in CACT using the channel measurements. The schematics of the classroom and CACT are shown again in Fig. 4-1. The locations R1 to R4 denote the classroom and RC1 to RC4 denote CACT. In the case of the classroom, the data is collected and calculated also for four different channel environments at each receiver location.

The characteristics of 4 receiver locations and the distance between the
transmitter and each receiver in the classroom are as follows:

![Diagram showing the schematics of channel in Classroom (Ball 313) and CACT.](image)

Figure 4-1. The schematics of channel.
• R1: In front of student section, 5.1 m
• R2: In the middle of student side, 2.9 m
• R3: beside the window, 5.2 m
• R4: On the instructor’s desk, 2.3 m

And the channel environments of classroom with different scenarios are:

• S1: A completely empty classroom to simulate an empty rectangular room, LOS
• S2: An empty classroom with furniture, LOS
• S3: A classroom during the session, LOS
• S4: A classroom in the middle of time to rest, NL0S

The measurement environments of 4 receiver locations in CACT and its distance from the transmitter are summarized in:

• RC1: NL0S, 4.2 m, strong single reflection by metal cabinet
• RC2: NL0S, 2.2 m, LOS blocked by fabric partition
• RC3: LOS, 5.5 m, longest T-R separation
• RC4: NL0S, 3.5 m, most of the dominant reflections are blocked by the metal plate partitions

4.2.2 Calculation and Analysis of Delay Parameters

The delay parameters discussed above were estimated from the measurements and are given in Tables 4-1 to 4-5. The mean excess delay (MED), \( m_e \) is given in Table
In the case of the classroom, there are no big differences in the three line of sight (LOS) cases, a completely empty classroom, an empty classroom with furniture and class in session for a fixed receiver position. The magnitude of MED is proportional to the transmitter and receiver separation when there exists a LOS component. The LOS component and some other dominant peaks are blocked by the students in the rest time scenario and the mean excess delay increases. Although, the LOS component for CACT results are blocked by the partition in location RC2, a transmitted component exist that is large enough in magnitude to influence the delay spreads. The location RC4 in CACT has a large MED because the location is enclosed by conducting material partitions.

Table 4-1. Measured mean excess delay (MED) [nsec].

<table>
<thead>
<tr>
<th>Environment</th>
<th>Location</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>R(C)1</td>
</tr>
<tr>
<td>A completely empty classroom</td>
<td>24.44</td>
</tr>
<tr>
<td>An empty classroom with furniture</td>
<td>21.16</td>
</tr>
<tr>
<td>Class in session</td>
<td>23.24</td>
</tr>
<tr>
<td>Rest time in classroom</td>
<td>42.15</td>
</tr>
<tr>
<td>In CACT</td>
<td>44.31</td>
</tr>
</tbody>
</table>

The difference between mean excess delay and the location of maximum peak (DMM) is shown in Table 4-2. These results give us the problems on synchronization issues at the receiver. The DMM is a location dependent parameter. The DMM of a completely empty classroom at location R1 is bigger than an empty classroom with
furniture. The location R1 is in front of student side and a number of strong reflected component are scattered by the furniture. It is reversed at R2. The location R2 is in the middle of student side and some reflected component by the furniture arrive at the receiver. In the classroom case, the NLOS conditions have larger DMM than LOS cases. The DMMs exhibit a similar pattern with respect to the MED in CACT. The DMM depends on both the existence of LOS component and the locations of transmitter and receiver.

Table 4-2. The difference between MED and location of maximum peak (DMM) [nsec].

<table>
<thead>
<tr>
<th>Environment</th>
<th>Location</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>R(C)1</td>
</tr>
<tr>
<td>A completely empty classroom</td>
<td>7.77</td>
</tr>
<tr>
<td>An empty classroom with furniture</td>
<td>4.49</td>
</tr>
<tr>
<td>Class in session</td>
<td>6.57</td>
</tr>
<tr>
<td>Rest time in classroom</td>
<td>13.76</td>
</tr>
<tr>
<td>In CACT</td>
<td>24.52</td>
</tr>
</tbody>
</table>

Table 4-3. Measured RMS delay spread [nsec].

<table>
<thead>
<tr>
<th>Environment</th>
<th>Location</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>R(C)1</td>
</tr>
<tr>
<td>A completely empty classroom</td>
<td>13.11</td>
</tr>
<tr>
<td>An empty classroom with furniture</td>
<td>11.18</td>
</tr>
<tr>
<td>Class in session</td>
<td>14.88</td>
</tr>
<tr>
<td>Rest time in classroom</td>
<td>40.12</td>
</tr>
<tr>
<td>In CACT</td>
<td>42.30</td>
</tr>
</tbody>
</table>
The RMS delay spread (RDS), $\sigma_r$ is given in Table 4-3. The RDS is the measure of dispersiveness of the wireless channel and exhibits a similar trend as the DMM of Table 4-2. It exhibits the same pattern at R1 and R2 in the classroom for the two different scenarios as explained earlier. The magnitude of RDS is nearly double of DMM except at RC4 of CACT. The physical bandwidth of channel is calculated based on the RDS.

Table 4-4. Measured 10dB maximum excess delay [nsec].

<table>
<thead>
<tr>
<th>Environment</th>
<th>R(C)1</th>
<th>R(C)2</th>
<th>R(C)3</th>
<th>R(C)4</th>
</tr>
</thead>
<tbody>
<tr>
<td>A completely empty classroom</td>
<td>20.83</td>
<td>19.01</td>
<td>11.20</td>
<td>7.55</td>
</tr>
<tr>
<td>An empty classroom with furniture</td>
<td>5.47</td>
<td>18.49</td>
<td>8.85</td>
<td>7.55</td>
</tr>
<tr>
<td>Class in session</td>
<td>4.69</td>
<td>18.23</td>
<td>9.12</td>
<td>5.47</td>
</tr>
<tr>
<td>Rest time in classroom</td>
<td>38.28</td>
<td>18.75</td>
<td>35.68</td>
<td>25.52</td>
</tr>
<tr>
<td>In CACT</td>
<td>30.99</td>
<td>16.93</td>
<td>9.38</td>
<td>110.42</td>
</tr>
</tbody>
</table>

In dispersive communications channels, an equalizer is adopted to overcome the effect of multipath. The length of the equalizer is determined by the duration of multipath existence above a certain power level. The maximum excess delay at a certain power level (MDP) gives information about the persistence of multipath which affects the communication. The result of 10dB maximum excess delay is shown in Table 4-4. MDP highly depends on the strength of each multipath component and is different for different threshold levels. From the results, $-10$ dB multipath component with respect to maximum peak exists around 20 nsec in a classroom with LOS.
component. A 40 nsec interval contributes to interference for the NLOS cases in a classroom. In CACT, the MDP is changed based on the T-R separation distance and also their independent locations.

The coherence bandwidth, $B_c$, is calculated based on the RMS delay spread, $\sigma_r$, and decay rate, $\alpha$ of the wireless channel. The $B_c$ is the measure of reliable channel data rate or channel capacity based on the empirical result of Howard et al.\textsuperscript{[32]}. The result of coherence bandwidth is given in Table 4-5 with the assumption of $\alpha = 2\pi$. The coherence bandwidth of the classroom with LOS component is more than 10 MHz but it’s spread out from 10.4 MHz to 3.9 MHz for the NLOS case. In the case of CACT, it ranges from 6.8 MHZ down to 2.4 MHz.

<table>
<thead>
<tr>
<th>Environment</th>
<th>Location</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>R(C)1</td>
</tr>
<tr>
<td>A completely empty classroom</td>
<td>12.14</td>
</tr>
<tr>
<td>An empty classroom with furniture</td>
<td>14.24</td>
</tr>
<tr>
<td>Class in session</td>
<td>10.70</td>
</tr>
<tr>
<td>Rest time in classroom</td>
<td>3.97</td>
</tr>
<tr>
<td>In CACT</td>
<td>3.76</td>
</tr>
</tbody>
</table>

The delay parameters are calculated based on the measured power delay profile (PDP) and used as a measure of reliability criteria of the indoor channel for wireless communications. The DMM and RDS have similar pattern and give information about dispersiveness of the indoor wireless multipath channel. MDP denotes the existance
duration of effective multipath components. The reliable data rate is estimated using the coherence bandwidth, $B_c$.

### 4.3 Pole-Zero Calculation using Singular Value Decomposition

For discrete-time signals, the transfer function of the channel can be represented as the ratio of the Z-transforms of the output ($Y(z)$) and input $X(z)$ signal. The transfer function, $H(z)$ in proper rational function form is represented by

$$H(z) = \frac{Y(z)}{X(z)} = b_0 \frac{z^{N-M}(z - Z_1)(z - Z_2) \cdots (z - Z_M)}{(z - P_1)(z - P_2) \cdots (z - P_N)} \quad (4-5)$$

where $M$ is the number of nonzero zeros and $N$ is the number of poles with the assumption that there are no pole-zero cancellations. The nonzero zeros are $Z_1, \ldots, Z_M$ and the poles are $P_1, \ldots, P_N$. The transfer function $H(z)$ can be represented as a function of $z^{-1}$ and both $H(z)$ and $H'(z^{-1})$ are identical.

$$H(z) = H'(z^{-1}) = \frac{b_0 + b_1 z^{-1} + \cdots + b_M z^{-M}}{1 + a_1 z^{-1} + \cdots + a_N z^{-N}} \quad (4-6)$$

If time domain channel input, $x(n)$ is a unit impulse then the channel impulse response, $h(n)$ is as follows

$$h(n) = -a_1 h(n-1) - \cdots - a_N h(n-N) + b_0 \delta(n) + b_1 \delta(n-1) + \cdots + b_M \delta(n-M) \quad (4-7)$$

Using the recursive difference equation with the assumption of causality, channel
impulse response of each time instant is expressed in matrix form.

\[
\begin{bmatrix}
  h(0) \\
  h(1) \\
  \vdots \\
  h(M) \\
  h(M+1) \\
  \vdots \\
  h(M+N) \\
  h(M+N+1) \\
  \vdots \\
  h(np)
\end{bmatrix}
= 
\begin{bmatrix}
  0 & 0 & \cdots & 1 & 0 & \cdots & 0 \\
  -h(0) & 0 & \cdots & 0 & 1 & \cdots & 0 \\
  \vdots & \vdots & \ddots & \vdots & \vdots & \ddots & \vdots \\
  -h(M-1) & -h(M-2) & \cdots & 0 & 0 & \cdots & 1 \\
  -h(M) & -h(M-1) & \cdots & 0 & 0 & \cdots & 0 \\
  \vdots & \vdots & \ddots & \vdots & \vdots & \ddots & \vdots \\
  -h(M+N-1) & -h(M+N-2) & \cdots & 0 & 0 & \cdots & 0 \\
  -h(M+N) & -h(M+N-1) & \cdots & 0 & 0 & \cdots & 0 \\
  \vdots & \vdots & \ddots & \vdots & \vdots & \ddots & \vdots \\
  h(np-1) & h(np-2) & \cdots & 0 & 0 & \cdots & 0
\end{bmatrix}
\begin{bmatrix}
  a_1 \\
  a_2 \\
  \vdots \\
  a_N \\
  b_0 \\
  \vdots \\
  b_M
\end{bmatrix}
\]

Where \( np + 1 \) is the number of data points of the channel impulse response. Equation (4-8) is represented by Equation (4-9) using the vector matrix notation.

\[
[h]_{1\times(np+1)} = [A]_{(N+M+1)\times(np+1)} [C]_{1\times(N+M+1)}
\]

The system is an overdetermined case when \( np \) is greater than \( M + N \). The coefficients of the transfer function are calculated using matrix inverse for the case where \( M + N = NP \) and \( LU \) decomposition or singular value decomposition (SVD) is applied for the case of overdetermined system. The number of poles and zeros required for the model are generally unknown. From the results of 1-D space analysis, the number of poles required is of the order of the discrete sample space dimension.

In this work, the matrix equation is solved by SVD when the equation is overdetermined\(^{[54]}\). The matrix is decomposed using singular values as \( A = U W V^T \),
where $\mathbf{U}_{(N+M+1)\times(N+M+1)}$ and $\mathbf{V}_{(np+1)\times(np+1)}$ are orthogonal matrices. When $\mathbf{A}$ is of rank $r > 0$, $\mathbf{W}$ will have exactly $r$ positive singular values, $s_k$ along the main diagonal extending from upper left corner. The remaining components of the matrix $\mathbf{W}$ are zero. The pseudoinverse of matrix $\mathbf{A}$ is $\mathbf{A}^i = (\mathbf{UWV}^T)^i = (\mathbf{V}^T)^i \mathbf{W}^i \mathbf{U}^i$. Using the orthogonality of the matrix $\mathbf{U}$ and $\mathbf{V}$, the pseudoinverse becomes

$$\mathbf{A}^i = \mathbf{V} \mathbf{W}^i \mathbf{U}^T$$  \hfill (4-10)

The components, $s_k^i$ of pseudoinverse of $\mathbf{W}$ are given by $1 / s_k$ when $s_k \neq 0$ and are 0 when $s_k = 0$. The coefficient vector is given by

$$\mathbf{C} = \mathbf{A}^i \mathbf{h}$$  \hfill (4-11)

The coherent and diffuse parts of the channel impulse response are used to find out the poles and zeros that characterize the indoor wireless channel. The number of zeros is determined based on the duration of the coherence region and the number of poles is chosen to minimize the error which is defined by

$$\varepsilon = \frac{1}{np + 1} \sum_{n=0}^{np} \left( h(n) - \hat{h}(n) \right)^2$$  \hfill (4-12)

where $h(n)$ is the original channel impulse response and $\hat{h}(n)$ is the reconstructed channel impulse response using the calculated poles and zeros. If the distance between a pole and zero are less than a chosen error tolerance, they are assumed to cancel each
other and are removed as parameters. The characteristics of pole-zero locations are compared with results of the 1-D lossless channel.

### 4.3.1 Poles and Zeros

The 100 nsec duration of the channel impulse response from the maximum peak is used to calculate poles and zeros. This is typically the region where the CIR amplitude is above the noise floor of −50 dB. The coefficients of numerator and denominator of the transfer function are solved using singular value decomposition (SVD). The number of zeros is determined based on the duration of the coherence region in the CIR. In general, a 40 nsec duration from the maximum peak has different location dependent features and this is approximated as the length of the coherent component. Considering the 2.63 m mean free path of the classroom, the coherence region contains the LOS and an average number of 4.5 reflected waves. The number of poles are arbitrarily chosen to minimize the error, ε with the consideration of calculation complexity. In this work approximately 15 poles were found to be sufficient to capture the diffuse component trend of the CIR.

The poles and zeros of a completely empty classroom for the four different receiver locations are given in Fig. 4-2 (a-b). From the result of 1-D lossless channel, the locations of poles should be invariant by the movement of the mobile, affected only by the size of the channel space and the reflection coefficients of the boundaries.
The pole diagram given in Fig. 4-2 (a) shows clusters over which the common poles are arranged. Although they are not exactly equal, they follow the same trend in angle and we may assume that there is a certain degree of location independent invariance in the
distribution of poles. The dispersiveness of the aggregated poles of the different locations of receiver may come from nonideal conditions of the channel and the communication system, as mentioned earlier.

Figure 4-3. The detailed local plot of zero of a completely empty classroom.

The majority of zeros are located at a fixed radius shown by the dark circle, and a small number of zeros are inside and outside the circle. The location of zeros are affected by the size of the channel and the reflection coefficients and the geometrical configuration of transmitter and receiver. When the mobile is moved in the 1-D channel, a number of zeros change their location and some were found to be invariant. In this case, the zeros on the circle may represent the invariant components, but the zeros inside and outside of the circle change their locations with change in placement of the receiver. More detailed localized plot of zero diagram is given in Fig. 4-3. From the
figure, the location of zeros for different placements of the receiver form fairly close clusters, motivating their characterization as invariant parameters.

### 4.3.2 Reconstruction of CIR based on Pole-Zero Model

The channel impulse responses are reconstructed using the limited number of poles and zeros which are calculated in the previous section. When reconstructing the channel impulse response, if the distance between pole and zero is less than 0.01, then the pole and the zero are cancelled out [55].

The reconstructed channel impulse responses of the empty classroom are given in Fig. 4-4 (a-d). The $x$ axis is time in seconds from the maximum peak and $y$ axis is the relative magnitude in decibels. In the picture, solid line is the original channel impulse response which is calculated from the measured complex channel transfer function and dotted line is reconstructed CIR based on the limited number of poles and zeros. The error, $\varepsilon$, between original CIR and reconstructed CIR is $2.272 \times 10^{-4}$ for R1, $4.791 \times 10^{-4}$ for R2, $1.096 \times 10^{-4}$ for R3 and $8.773 \times 10^{-5}$ for R4.

The reconstructed channel impulse responses based on the pole-zero model are least square approximations and the energy level is decaying by the time. The reconstructed signal is fitted well at the beginning part where contain most of channel energy by enough number of zeros. The location R4 has least error because more fit than other places in high energy region. The tail region of reconstructed CIR is governed by both the pole and the zero. The slope of the tail region is highly affected
by the locations of poles. In the case of the single pole system, the decay rate is inversely proportional to the magnitude of pole. In the case of the multiple pole system, the slope is governed by the sum of magnitude of all pole. In location R2, more poles are located near the origin than any other location and it has the steepest slope.

Figure 4-4. The measured and reconstructed channel impulse response.
4.4 Summary

The channel impulse response is the delayed and attenuated multipath received signal when a single pulse is transmitted. The delayed and attenuated pattern depends on both the communication system and the wireless channel itself. The power delay profile (PDP) has been used to describe the delay and attenuation pattern in an indoor wireless channel. PDP delay parameters such as mean excess delay, RMS delay spread, 10 dB maximum excess delay and coherence bandwidth are calculated using the measured CIRs in the classroom and research laboratory.

The mean excess delay gives information about the possible errors in synchronization between transmitter and receiver. Both the difference between mean excess delay and location of maximum peak and the RMS delay spread are measure of the wireless channel dispersiveness and the values of RMS delay spread are roughly double in magnitude of the difference between the mean excess delay and the location of maximum peak. To determine the memory component size of a detector such as an equalizer, the maximum delay based on a power threshold is useful. The coherence bandwidth or maximum reliable data rate is calculated based on the RMS delay spread. The coherence bandwidth of a classroom with LOS component has a minimum value of 10 MHz. It ranges from 3.9 MHz to 10.4 MHz for the NLOS case.

The coefficients of the transfer function are solved by SVD from the measured channel impulse response and are the least square error solutions for a fixed order of
denominator and numerator. The poles and the zeros are calculated from these 
denominator and numerator polynomials. The number of zeros are determined based on 
the duration of the coherence region and the number of poles are arbitrarily chosen to 
minimize the reconstruction error. The number of zeros and poles are 150 and 15 
respectively. To maintain stability of the solution, it was found that the number of zeros 
must be greater than the number of zeros and compensate the location dependent 
variations in the coherent region. The solutions for the poles at different locations show 
some degree of invariance. Similarly a large number of invariant zeros were identified in 
the solution, all of which had a constant radius. Other zeros inside and outside this 
radius exhibited changes as the receiver location changed. The reconstructed CIRs are 
fit well at the coherence part by enough number of zeros. The location R4 has least 
error than other receiver locations because the T-R separation distance is short and most 
of the high energy peaks are gone after the LOS component.
5. Analysis of Narrow Band Channel Impulse Response

5.1 Introduction

The bandwidth of communication systems in current standards is limited to range from typically 30KHz to 140 MHz. The effects on system performance due to the change of bandwidth and carrier frequency are investigated in this chapter. The narrow band channel impulse responses are acquired by appropriate bandpass filtering of the measured channel transfer functions.

As mentioned in Chapter 2, the super Gaussian filter is used as a band pass filter to calculate the narrow band channel impulse response. Three different carrier frequencies are selected to characterize the narrow band indoor wireless channel. They are, 1.9 GHz, 2.35 GHz and 2.45 GHz carriers. These carrier frequencies represent current or upcoming wireless communication system standards. The 1.9 GHz frequency is used in 3rd and 4th generation cellular systems, the 2.35 GHz is applied for wireless local loops and 2.45 GHz is the standard for IEEE wireless local area networks (WLANs). The bandwidth of each system is chosen based on both assigned and available spectrum and the data rate requirement of the applications supported. The channel measurement resolution of 1875 KHz is also a consideration in this analysis.
The orthogonal frequency division multiplexing (OFDM) scheme has been adopted in the European digital audio broadcasting (DAB) standard and it is a candidate modulation scheme for digital television broadcasting in North America [6] and also for the next generation wireless LAN referred to as IEEE 802.11g. The OFDM technique was developed to transmit a message in parallel channels, utilizing a set of carriers such that the signals are orthogonal to each other. This approach allows transmission at a maximum data rate on a linear band limited channel while minimizing the effects of interchannel interference (ICI) and intersymbol interference (ISI). Chang [56] proposed the use of sub-channel spacing that was equal to the symbol rate, so that the modulated signals are orthogonal and amenable for separation at the receiver using matched filter or correlation based techniques. This also leads to increased bandwidth efficiency, by avoiding placement of guard bands etc.. The discrete Fourier transform (DFT) was proposed by Weinstein and Ebert to perform baseband modulation and demodulation and Peled and Ruiz adopt a cyclic prefix (CP) to solve orthogonality problem, instead of using an empty guard space.

The effects of multipath communication channel on OFDM communication systems is investigated as a function of the number of carriers and the power in the channel noise, using measured narrow band complex channel impulse responses. The frequency sweep method which is used in the measurement component of this thesis is one of best channel measurement techniques for evaluating OFDM systems because both the modulation scheme and the channel measurements operate on the basis of
discrete Fourier frequencies.

The narrow band channel impulse responses obtained from the channel measurements are described in Section 5.2, the basic principles of OFDM and channel simulation based on the narrow band CIR is described in Section 5.3 and finally the chapter summary is given.

![Figure 5-1. The super Gaussian band pass filter, $S(f; w, 2)$.

5.2 Narrow Band CIR and its Intersymbol Interference (ISI) Power

The narrow band complex CIR is calculated by taking the inverse Fourier transform of the measured complex transfer function. The transfer function is filtered by the second order super Gaussian filter, $S(f; w, 2)$. The shape of this band pass filter is given in Fig. 5-1. The horizontal axis is frequency in GHz and the vertical axis is the
magnitude. The center frequency is set to 2.45 GHz and the bandwidths are 10, 20, 50, 100 and 200 MHz each.

Both the real and the imaginary components of the transfer function are filtered by $S(f; \omega, 2)$. An example of the narrow band complex CIR characterizing the channel at location R1 in the empty classroom is shown in Fig. 5-2. The bandwidth of the filter used for calculating the CIR shown in Fig 5-2 is 10 MHz and the center frequency is 2.45 GHz. The x axis is time in seconds and the y axis is the magnitude.

![Figure 5-2. Narrowband complex channel impulse response.](image)

**5.2.1 The Effect of Transmission Bandwidth**

The effect on the communication system due to the change in channel bandwidth is investigated. The magnitude of the CIR for spectral bandwidths ranging from 10-200 MHz are given in Fig. 5-3 for the empty classroom at location R1. The carrier
frequency of 2.45 GHz has been considered for these results. In the figure, the $x$ axis is time in seconds and the $y$ axis is magnitude in decibels, normalized by the maximum $h_{\text{max}}$ which occurs at index $N_{\text{max}}$. As bandwidth increases, one can resolve an increased number of peaks and valleys in the CIR. For the case of 10 MHz bandwidth, the CIR appears as a smooth, slowly decaying function in time. The number of peaks or maxima increase to approximately 4 peaks for 20 MHz, 10 peaks for 50 MHz and so on. The result of examining the narrowband CIR in various channel measurement environments in the classroom and university laboratory at the four different locations show a similar pattern.

![Figure 5-3. The normalized channel impulse responses with different bandwidth.](image)

To characterize the interference noise of the CIR as a function of bandwidth, the data rate has to be taken into consideration. For a data rate $R$ symbols per second, the
channel is sampled at time locations \((i \cdot N_s), i = 0,1\cdots\) and \(N_s = \frac{1}{Rt_s}\), where \(t_s\) is the measurement sampling duration. If the index \(i_{\text{max}}\) corresponds to the location where the relative ISI noise power arising from the interfering components of the channel may be calculated as \(P_{\text{ISI}}\),

\[
P_{\text{ISI}} = \sum_{k_1 \neq i_{\text{max}}} \frac{h_{NB_1}^2}{h_{\text{max}}^2}
\]

The ISI power for the empty classroom for different bandwidths is given in Fig. 5-4.

![Figure 5-4. The normalized ISI power of narrowband CIRs.](image)

The ISI power is calculated using the bandlimited CIRs shown in Fig. 5-3. In Fig. 5-4, the relative ISI power is plotted on the \(y\) axis as a function of data rate in Mbps. Both axes are in logarithmic scale. Typically, to accommodate increased data transmission
rate requirements, the channel bandwidth is increased. From the figure, one can see that at a fixed data rate, for example at 10 Mbps, the ISI decreases as the bandwidth is increased from 10 to 200 MHz. However, the magnitude of ISI power will also increase with increasing data rate. One can see from the figure that at the higher data rates, the ISI powers are nearly comparable for all channel bandwidths considered. These features imply that since ISI power increases nonlinearly with increase in data rate, increasing the channel bandwidth in proportion to data rate will not be sufficient to maintain the desired error performance. Modulation schemes such as OFDM are designed to minimize the effects of ISI with increasing data rate. Their performance is studied in Section 5.3.

5.2.2 Impact of Carrier Frequencies and Bandwidth

A number of different wireless standards are either in place or under study to support a diverse set of applications on wireless systems. The selection of a specific set of carrier frequencies may be determined based on available spectrum, FCC regulations etc.. Three carriers that are of interest are:

i. 1.9 GHz which exists in the current commercial cellular phone system, operating with bandwidths of 10 and 20 MHz; The maximum chip rate of CDMA2000 1xEV-DO is 2.4 Mbps in indoor fixed mobile environment and 0.144 Mbps for outdoor systems with mobility. The data rate has been expanded to 3.09 Mbps in CDMA2000 1xEV-DV standard.
ii. 2.35 GHz carrier has been selected for wireless local loop (WLL) systems in North America and the wireless broadband system (WBS) in Korea\textsuperscript{[57]}. Currently field tests are being undertaken in Korea by the cooperation between Flarion Technologies, the architect of flash-OFDM systems and major cellular and broadband carriers such as SK Telecom, KT and Hanaro communication\textsuperscript{[58]}. The 50 and 100 MHz bandwidths are considered for broadband data communications under the conditions of reduced mobility and power relative to cellular standards. The WLL and WBS support low speed movement of mobile terminals, and devices such as PDA, notebook and tablet computers, and the size of the cell is nearly ten times bigger than that of the WiFi WLAN.

iii. Thirdly, the 2.45 GHz industrial, scientific and medical (ISM) band is the standard for wireless local area networks, such as IEEE 802.11b to support broadband wireless applications with pico cell frequency reuse patterns and 200 MHz bandwidths.

The aforementioned carrier frequencies and associated bandwidth parameters are applied to calculate the narrow band CIR from the CACT research laboratory channel transfer function measurements. The impact on receiver placement at four different locations is also examined. The shape of the band pass filter is the same as shown in Fig. 5-1, but with different center frequencies.
5.2.2.1 Carriers in 1.8 - 2.2 GHz Band

In this section, the CIRs are computed from measurements at carrier frequency \( f_c = 1.9 \ \text{GHz} \) for bandwidths 10 and 20 MHz. The data correspond to the four measurement locations RC1, RC2, RC3 and RC4 in CACT. The CIRs are shown in Fig. 5-5(a,b) for the 10 and 20 MHz cases respectively. The \( x \) axis is time in seconds and \( y \) axis is the normalized relative magnitude in decibels. The ISI powers calculated using Eq. (5-1) are given in Table 5-1. The same wireless channel is seen to be differently affected during data exchange by the channel bandwidth and data rate. The location RC3, which corresponds to the LOS case, has the smallest ISI power for the 10 MHz case, but it has the second largest ISI when bandwidth is doubled. There is -78.4 dBm deep fading at 1,910.6 MHz in the transfer function of the RC3. By the effect of this frequency null, the main peak in narrow band CIR in RC3 is wide and it generates high ISI power in the 20 MHz bandwidth.

![Figure 5-5. The narrow band CIR at 1.9 GHz.](image)
Table 5-1 ISI powers for $f_c = 1.9$ GHz

<table>
<thead>
<tr>
<th>Location</th>
<th>10 Mbps 10 MHz</th>
<th>20 Mbps 20 MHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>RC1</td>
<td>0.0647</td>
<td>0.0577</td>
</tr>
<tr>
<td>RC2</td>
<td>0.062</td>
<td>0.077</td>
</tr>
<tr>
<td>RC3</td>
<td>0.05</td>
<td>0.23</td>
</tr>
<tr>
<td>RC4</td>
<td>0.43</td>
<td>0.61</td>
</tr>
</tbody>
</table>

5.2.2.2 Carriers in 2.3 GHz Region

The 2.35 GHz carrier frequency with 50 and 100 MHz bandwidths are examined in this section. The calculated narrow band CIRs are given in Figs. 5-6 (a-b). The ISI powers for 40 and 80 Mbps data rate are given in Table 5-2. A distinctive feature of this region is that the ISI power of RC1 is abnormally high for both bandwidth cases. It arises due to the deep fading and spectral nulls that are seen around the 2.32 GHz in the channel transfer function for location RC1. Therefore frequency selective fading characteristics can directly influence the narrow band communication.

![Figure 5-6](image-url)

(a) 50 MHz bandwidth  
(b) 100 MHz bandwidth

Figure 5-6. The narrow band CIR at $f_c = 2.35$ GHz.
Table 5-2. ISI powers for $f_c = 2.35$ GHz

<table>
<thead>
<tr>
<th>Location</th>
<th>40 Mbps 50 MHz</th>
<th>80 Mbps 100 MHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>RC1</td>
<td>0.73</td>
<td>1.15</td>
</tr>
<tr>
<td>RC2</td>
<td>0.25</td>
<td>0.58</td>
</tr>
<tr>
<td>RC3</td>
<td>0.37</td>
<td>0.31</td>
</tr>
<tr>
<td>RC4</td>
<td>0.82</td>
<td>0.92</td>
</tr>
</tbody>
</table>

5.2.2.3 Carrier in 2.4 GHz Region

Figure 5-7. The narrow band CIR at 2.45 GHz with 200 MHz bandwidth.

The maximum data rate of IEEE 802.11b WiFi wireless LAN, currently available in the market is 11 Mbps and its spectral channel spacing is 25 MHz which include a guard gap between channels[^59]. The symbol rate of IEEE 802.11b is 1.375 Mega-symbols per second with 8 bit code length using complementary code keying (CCK) and quadrature phase shift keying (QPSK) modulation scheme. The maximum data rate
of IEEE 802.11a is 54 Mbps but the physical layer of this standard is 5.8 GHz band. The upgrade version of 802.11b is 802.11g which uses the same physical and MAC layer, 2.4 GHz ISM band and maximum data rate is 54 Mbps\textsuperscript{[60]}. Here, the CIR is calculated for 200 MHz bandwidth at $f_c = 2.45$ GHz and the results are given in Fig. 5-7. The ISI powers are tabulated in Table 5-3 for data rate of 160 Mbps. The ISI power in location RC2 is largest because there is broad peak which includes the maximum and this peak contributes to the majority of ISI power.

<table>
<thead>
<tr>
<th>Location</th>
<th>160 Mbps</th>
<th>200 MHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>RC1</td>
<td>0.66</td>
<td></td>
</tr>
<tr>
<td>RC2</td>
<td>0.94</td>
<td></td>
</tr>
<tr>
<td>RC3</td>
<td>0.34</td>
<td></td>
</tr>
<tr>
<td>RC4</td>
<td>0.85</td>
<td></td>
</tr>
</tbody>
</table>

5.3 The Influence of Channel Features on an OFDM System

The orthogonal frequency division multiplexing (OFDM) scheme is applied in this section to evaluate its resilience to ISI features in the channel. OFDM is being considered as a promising multicarrier modulation scheme for supporting high data rates on linear band limited channels. OFDM schemes are specifically chosen for their potential to minimize effects of interchannel interference (ICI) and ISI. To this end, the channel transfer functions measured in multipath environments are applied for analysis of OFDM performance. Since various standards operate in different frequency bands,
the performance will also be examined as a function of carrier frequency and channel bandwidth. The quadrature phase shift keying (QPSK) scheme is selected for modulation of data.

### 5.3.1 Principle of OFDM

OFDM is a process where the data to be transmitted at a rate of $R$ symbols per second over a channel bandwidth $W$ Hz is split into a set of $N$ parallel streams and each stream is transmitted on a different carrier signal. The carrier spacing $\Delta f$ is chosen such that the signals in the subchannels are orthogonal to each other. This is typically achieved by considering the complex Fourier exponentials as carrier signals. The bandwidth of each subchannel $\Delta f = W/N$. The symbol extends over a duration $T = \frac{1}{\Delta f}$. Typically, $T$ is selected to be larger than the delay spread of the channel, to minimize the effects of ISI on the subchannels. The set of $N$ information symbols generated from a chosen modulation scheme are represented as $X_n(i)$, $n = 0, \ldots, N-1$, where $i$ is the block index. The carriers are chosen to be the set $\{e^{j2\pi f_n t}\}$, where $f_n = n\Delta f$. Neglecting the index $i$, the OFDM signal transmitted during a block may be represented by,

$$ x(t) = \sum_{n=0}^{N-1} A X_n e^{j2\pi ft/T} dt, \quad 0 \leq t \leq T $$

The constant $A = \sqrt{\frac{2E}{T}}$, where $E$ is the average symbol energy. The QPSK
modulation scheme chosen here results in symbol alphabets \{(1+j1), (-1+j1), (-1-j1), (1-j1)\}, that are assumed by $X_i$. In practice, the inverse fast Fourier transform (IFFT) may be implemented at the transmitter to generate samples of $x_k = x(kT/N)$, $k = 0, 1, \ldots, N-1$. These samples are transmitted successively over the channel using signalling intervals of length $T/N$.

The channel is represented in terms of its impulse response $h_k$, $k = 0, \ldots, M-1$, sampled at the symbol spacing $T/N$. If the CIR duration $M$ is smaller than $N$, then intrablock ISI can be avoided by adding a cyclic prefix of length $M-1$ [61] to each block of size $N$. The prefix serves to absorb the channel ISI resulting from the transmissions at the tail of the previous block. At the receiver, the first $M-1$ samples corresponding to the prefix are discarded. The received signal samples $r_n$ are corrupted by additive Gaussian noise. The received samples are processed at the receiver to recover the information symbols $X_k$.

A schematic diagram of the discrete time OFDM system implemented in the simulation is given in Fig. 5-8. Where $FFT$ and $IFFT$ represent the fast Fourier transform and inverse fast Fourier transform operations respectively, $P/S$ and $S/P$ indicate parallel to serial and serial to parallel conversions, $CP$ and $DCP$ are operations of attaching a cyclic prefix (CP) and detaching the CP respectively. $AGC$ stands for automatic gain control and $PD$ is a parallel signal detector. The block between channel input, $x_n$ and channel output, $r_n$ represents the bandlimited wireless channel, $h_n$, $n = 0, 1, \ldots, M-1$. The additive white Gaussian noise is $g_n$. 
In the simulation of OFDM transmission, the sequence of operations begins by generating $N$ complex information symbols $X_k, j = 0, 1, \ldots, N - 1$ in successive frames and taking the inverse Fourier transform of this sequence. The time-domain signal consists of $N$ samples $x_n, n = 1, 2, \ldots, N$. To the front end of this sequence, a cyclic prefix of length $L_c$ is appended. This is chosen to be a repetition of the last $L_c$ information symbols, to achieve a circular convolution with the CIR and exactly cancel out interblock ISI. The transmitted complex signal sequence $x_n$ is convolved with the complex channel impulse response. The samples of the received signal are,

$$r_n = (x_n \ast h_n) + g_n$$  \hspace{1cm} (5-3)

After discarding the first $L_c$ samples of the received frame, the received sequence is processed through a FFT algorithm to yield,
\[ R_k = X_k H_k + G_k, \quad k = 0, 1, \ldots, N - 1 \]  
\[ (5-4) \]

where \( H_k \), \( k = 0, 1, \ldots, N - 1 \) are the samples of the channel transfer function at the carrier frequencies of the \( N \) subchannels and \( G_k \) is a complex Gaussian noise. Eq. (5-4) shows that the performance of the OFDM system is limited by the channel characteristics in the frequency domain. Each subchannel has an effective complex attenuation factor \( H_k \), that determines the overall bit error rate performance of the channel. A set of performance metrics may be defined in terms of the samples \( H_k \).

The signal to noise ratio may be defined as a function of subchannel \( k \) \[^{62}\].

\[ SNR_k = \frac{T P_k |H_k|^2}{\sigma_G^2} \]  
\[ (5-5) \]

where \( P_k \) is the average power associated with the \( k \)th subchannel. The conditional expectation of the received signal,

\[ E[R_k|X_k] = X_k E[H_k], \quad k = 0, 1, \ldots, N - 1 \]  
\[ (5-6) \]

If the channel is time-invariant or slowly varying, the conditional expectation may be calculated for each of the unique information symbols and their location in the constellation diagram predicted. The symbols are then scaled and rotated by a constant factor \( E[H_k] = \frac{1}{N} \sum_{i=0}^{N-1} H_k = h(0). \) Clearly, apriori channel estimates must be available or obtained through the transmission of pilot symbols to specify expected location of the
symbol in the constellation space and the decision boundaries for error detection.

The constellation space resulting from the QPSK modulation scheme considered here is shown in Fig. 5-9.

The four symbols may be denoted as $S_i$, $i = 0, 1, 2, 3$. The distance between two adjacent symbols is $d = \sqrt{2E}$. For a case where $H_k$ may be assumed to be a uniform factor $\hat{H}$ across the subchannels, the received symbols $R_k$ are simply scaled and rotated by this value and corrupted by AWGN. Assuming complete knowledge of the constant factor, the decision on classifying the symbol received is based on the distances $\hat{d}_i = \|R_k - \hat{H} S_i\|$, $i = 0, 1, 2, 3$. The detected symbol is the $S_i$ that minimizes this distance criteria. These decisions are derived by maximizing the aposteriori probabilities (MAP) of the received symbols. This constraint minimizes the average probability of error. For QPSK signals with AWGN interference only, the average probability of symbol error

![Figure 5-9. The constellation of QPSK signal.](image)
can be shown to be \[ P_s = 2P_1 - P_1^2 \] (5-7)

where \( P_1 = Q \left( \sqrt{\frac{E_s}{N_0}} \right) \) and \( E_s \) is the symbol energy. From Fig. 5-1, assuming a signalling duration \( T_s \), the symbol energy may be calculated as \( E_s = 2E^2T_s \).

![Figure 5-10. The band limited complex channel impulse response.](image)

The performance of QPSK-OFDM scheme is examined for the channel representative of location RC1 in CACT. Fig. 5-10 depicts the CIR computed by selecting a center frequency of 2.45 GHz and bandwidth of 200 MHz. The channel shown is sampled at a rate of 20 Mbps, corresponding to time samples of 0.05
microseconds. Note, the wideband channel measurements yield a total CIR of length 250 nanoseconds. The CIR for this bit rate is therefore five samples long. The real and imaginary components of the CIR and its magnitude are also shown.

The simulation of QPSK-OFDM was carried out by transmitting $10^6$ symbols over a bandlimited channel, considering a signal to noise ratio range $\text{SNR} = 1:15$ decibels.

First, some basic comparisons are made to understand the effect of the channel on OFDM transmission. Fig. 5-11 (a-c) depicts for $\text{SNR} = 10$ dB four constellations observed at the receiver. The number of subchannels is selected to be $N = 1024$. In (a), the channel is ideal, containing a single delay component $h(0) = \alpha$ and is corrupted only by AWGN. Based on whether $\alpha$ is greater or less than one, the signal vectors are scaled and the distance between the signals increases or decreases. If $\alpha$ is exactly known and applied in the detector, the resulting symbol error rate is $9.3 \times 10^{-6}$. In (b), the constellation shown is for the case $h(0) = \alpha + j\beta$. The presence of the imaginary component induces a rotation of the signal vectors by an angle $\theta = \tan^{-1}(\beta/\alpha)$ and scaling by a factor $\sqrt{\alpha^2 + \beta^2}$. The effect in the signal space can be clearly observed. In (c) the CIR $h(n) = h_0\delta[n] + h_1\delta[n-1]$, creating interference from a single multipath component. We assume the delay spread is known and use a cyclic prefix of length one to cancel the intrablock interference. The channel transfer function may be represented as $H_k = h_0 + h_1 z^{-1}$. Also indicated in the figure are the expected locations of the symbols, calculated using the conditional expectation given in Eq. (5-6), assuming
From Fig. 5-11 of the detected signal constellation of known channel, the expected constellation of the channel is scaled by the factor of 1.3 when the CIR is $h_0 = 1.3$, figure (a), and their locations are in the middle of the signal cluster with 10 dB Gaussian noise. Fig. 5-11 (b) shows the result of complex channel effect. The expected detected signal was rotated by the angle of $\tan^{-1}(0.5/1.3) = 21^\circ$ and scaled by factor of
\[ |(1.3 + j0.5)(1 + j1)| = 1.97 \] and the expected location of the symbol \( S_1 \) is \( 0.8 + j1.8 \).

By the presence of the ISI component, \( h_1 = 0.3 \), in channel, figure (c), the detected signal form the circle and the radius of the circle is ISI magnitude times scaling factor and is \( 0.3 \times 1.3 = 0.39 \) when real and single tap ISI case. The detected signal are scattered around the circle, ISI only case, with additional Gaussian noise. In QPSK-OFDM system, there are two error source, one is channel noise and the other is ISI. When radius of the ISI circle is less than scaling factor of the channel, the channel noise is dominantly factor of the transmission error.

![Signal constellation](image)

Figure 5-12. The constellation of detected signal.

The constellation of the detected signal is given in Fig. 5-12 with the channel shown in Fig. 5-10. In the figure, ‘+’ constellation point is ISI only channel and
expected constellation of the channel can be determined using the $h_0 = -0.94 + j0.35$ of the CIR and the expected location of signal $S_0 = -1.29 - j0.59$. The shape of the closed loop of the ISI only case is related with $H_k$ and the radius is related with the magnitude of ISI power and scaling factor which determined by $h_0$. The closed loop also rotate by the $90^\circ$ with adjacent signal constellation.

5.3.2 The Channel Effect on OFDM System

The effect on OFDM communication system is analyzed by the different kinds of standards and the channel parameters based on the narrow band CIR and the result are compared with the results of delay parameters and ISI power of channel. The simulation is conducted under the assumption of known $h_0$. This mean that there is no detection error.

![Figure 5-13. The BER by the change of number of multicarrier.](image-url)
The number of multicarrier is changed and is pass through the CIR at RC1 in CACT with the filter described in Fig. 5-1 and 10 dB Gaussian noise is added. The BER by the change of number of multicarrier is shown in Fig. 5-13. The $x$ axis is data rate in Mbps and $y$ axis is BER. There is no big difference by the change of subchannel or number of multicarrier. No error was detected when data rate is less than 40 Mbps. And there is no change in BER after 80 Mbps.

![Figure 5-14. The BER by the change of SNR.](image)

The effects of channel noise on BER at RC1 with the same conditions are given in Fig. 5-14. By the increase of the data rate the ISI power is increased as shown in Fig. 5-4. The radius of the ISI contour is less than the signal power then the BER is dominantly affected by the channel noise. From the figure, data rate is less than 40
Mbps then the BER is increased by the increase of channel noise power. In this region, the relationship between BER, $\varepsilon_{BER}$ and noise power, $\sigma_n^2$ is given by

$$
\varepsilon_{BER} = K(d_i) \sigma_n^2
$$

(5-3)

Where the coefficient $K$ is function of data rate. This relationship is valid when the SNR is between 1 dB and 10 dB. The BER bound of QPSK modulation system also linearly proportion in this range. The radius of ISI circle is bigger than signal power then the signal power must be increased to communicate.

From the Eq. (5-4), the variance of the each constellations, $R_k - X_k h_0$, can be drived using Parseval’s theorem and is expressed by

$$
\text{Var}(R_k - X_k h_0) = E_s P_{ISI} + \sigma_G^2
$$

(5-4)

Where $E_s$ is symbol power of QPSK and $\sigma_G^2$ is channel noise power in message domain. The variance of each constellations are proportional to both ISI power and channel noise. As shown in Fig. 5-11 and 5-12, the distribution of the each constellation are not Gaussian distributed.

The BER is calculated under the various carrier frequencies and channel bandwidth which are shown in previous section. These results are compared with the results of channel delay parameters and ISI power.
5.3.2.1 Carriers in 1.8 - 2.2 GHz Band

The BER in CACT at 4 different locations are calculated using narrow band CIRs of 1.9 GHz carrier frequency with 10 MHz and 20 MHz bandwidth as shown in Fig. 5-15 both with added Gaussian noise, 10 dB SNR, and without noise. The BER of 10 MHz channel is 0 when data rate is less than or equal to 10 Mbps with and without noise at RC1, RC2 and RC3 but RC4 has transmission error. It’s similar trend with ISI power but data rate is higher than coherence bandwidth, $B_c$ of wide band analysis with the assumption of 2.5 Hz per 1 bit transmission efficiency as CDMA2000 chip transfer efficiency.

Figure 5-15. The BER at 1.9 GHz carrier frequency.

In the case of 20 MHz bandwidth, the location RC1, RC2 and RC3 are no error when the data rate is less than or equal to 20 Mbps without additional noise. It’s very similar to the case of 10 MHz. By the increase of channel bandwidth to 20 MHz the
location RC4 has no error with the 10 Mbps data rate. But with the channel noise RC3 has error at 20 Mbps and is same trend of ISI power which given in Table 5-1. The delay parameters only gives general characteristics of wireless channel and is impossible to estimate BER of specific system.

5.3.2.2 Carriers in 2.3 GHz Region

The BER of 2.3 GHz carrier frequency with 50 MHz and 100 MHz bandwidth is given in Fig. 5-16. In the case of no additional noise at 50 MHz channel bandwidth, RC1 and RC4 have no error with 10 Mbps data rate and RC2 and RC3 have no error until 40 Mbps. The descending order of ISI power at 40 Mbps is RC4, RC1, RC3 and RC2 and descending order of BER is RC4, RC1, RC3 and RC2. The order of BER is exact matched with the ISI power. By the effect of the deep fading at 2.32 GHz in RC1, the ISI power of this receiver location is abnormally big. The effect of deep fading in
complex transfer function affect on narrow band data transmission.

The highest no error data rate is increased when the increase of the channel spectral bandwidth from 50 MHz to 100 MHz, the highest no error data rate are changed at RC1 and RC4 to 20 Mbps, RC2 and RC3 to 40 Mbps. These results are matched with the ISI power. The ISI power approach around 0.5 then BER is highly affected by ISI power.

5.3.2.3 Carrier in 2.4 GHz Region

![Figure 5-17. The BER at 2.45 GHz center frequency with 200 MHz bandwidth.](image)

The BER of 2.4 GHz carrier frequency with 200 MHz bandwidth is shown in Fig. 5-17. The maximum no error data rate are 40 Mbps for RC1, RC2 and RC3 and 10 Mbps for RC4. But the ISI power of RC2 is the largest and RC3 has the smallest value.
In this bandwidth the trend of highest no error data rate is different with ISI power. In this bandwidth, the ISI power of all the location is big enough. The channel doesn’t support theoretical full data rate of channel bandwidth by the increase of data rate and channel spectral bandwith with the same increase ratio. The same physical channel is differently affect by the carrier frequency and bandwidth of channel.

5.4 Summary

The narrow band channel impulse response is used to analyze the effect on future high data rate communication systems and standards. The 3 different carrier frequencies with various bandwidth are selected to characterize next generation high speed cellular system, wireless broadband system and wireless LAN. The ISI power is calculated using the band limited channel impulse response and its bit error rate is counted using QPSK-OFDM system which are known as effectively support multicarrier high data rate communication system.

The narrow band CIR is calculated by inverse Fourier transform of band pass filtered complex channel transfer function. The 10 MHz and 20 MHz band limited CIRs with 1.9 GHz carrier frequency are used to characterize high data rate cellular system. From the result, the same physical wireless channel is differently affects on communication by the change of spectral bandwidth. The CIR of 2.35 GHz carrier frequency with 50 MHz and 100 MHz bandwidth are calculated for the WLL and WBS. The narrow band CIR is influenced by the frequency selective fading characteristics of
the pass band. The CIR of 2.45 GHz ISM band with 200 MHz bandwidth is calculated to simulate upcoming wide band wireless LAN. The wider spectral channel bandwidth is needed to support high data rate with the same ratio ISI power. The increase rate of ISI power is faster than the increase rate of bandwidth and data rate. The wireless communication is more and more affected and limited by the wireless channel itself.

The QPSK-OFDM scheme is used to simulate data transmission through the complex band limited channel which based on the channel measurement. The OFDM is selected for European standard of digital audio broadcasting (DAB) and digital video broadcasting (DVB) and strong candidate modulation scheme of WLL in North America and WBS in Korea. The information of channel is critical to determine decision line for signal detector. The ISI power is less 0.5 the BER is more affected by the channel noise. In this region the BER of OFDM is direct proportional to the power of channel noise. Through the transmission simulation, it is proved that ISI power is useful measure of wireless channel with QPSK-OFDM. The maximum no error data rate is increased by the increase of channel bandwidth when considering ISI only case. The BER increase rate is faster than the increase rate of channel bandwidth and data rate.
6. Conclusions and Future Work

6.1 Conclusions

In this work, measurements of indoor wireless channel transfer function are analyzed and their features interpreted relative to changes induced in the channel. The channel transfer function is measured using the frequency sweep method using a vector network analyzer and the channel impulse response (CIR) is calculated by filtering the transfer function and taking an inverse Fourier transform. The time-domain features of the CIR are analyzed and effect of channel interferers on multipath response discussed. In this context, the effect of furniture such as desks and chairs in a classroom, the impact of carpeted floors, windows and curtains and presence of stationary or moving human beings are identified in the CIR. The CIR is parametrically characterized using the conventional delay parameters such as the mean excess delay and the RMS delay spread values. A pole-zero characterization of the transfer function is also undertaken to differentiate the channel environments. The image source based computational model for the CIR is applied to examine the temporal and spatial features of the fading signal. Finally, a narrowband analysis of the channel transfer functions is carried out considering a set of standard center frequencies and bandwidths that are either being used currently or proposed for upcoming standards. The QPSK-OFDM scheme is
analyzed to evaluate its performance in frequency selective channels. The effect of uncertainty in the channel on OFDM performance is identified to compare its utility relative to single-carrier modulation schemes. A brief description of the conclusions for each of the aforementioned components is presented below.

6.1.1 Channel Measurements

The indoor wireless channel is measured using an Agilent 8753ES vector network analyzer and Astron AXQ24SM-A quarter wave omni-directional monopole antenna as a transmitter and a receiver. The measurements were conducted in a classroom environment and in the CACT research laboratory. For each of these environments, measurements were taken at four different receiver locations. In the classroom case, four different scenarios were analyzed that successively increased the level of interference in the channel. These scenarios ranged from the empty classroom, classroom with student desks, classroom in session with students seated and the same classroom considered when people in the room were in motion. For the last case, the LOS was blocked. The LOS paths that existed in the first three scenarios were closely identified in the CIR. A number of dominant paths that followed the LOS and in some cases, preceded the LOS were also identified. In summary, it can be concluded that for a typical empty room, that is, one with windows, doors and metal covered heaters, the singly reflected paths from the metal elements and the six walls were fairly well resolved for all receiver positions. For the case where the T-R separation was small,
these singly reflected paths were well separated from the regime formed by double and triply reflected paths by a duration of no arrivals. For T-R separations that almost extended to one of the room dimensions and also for locations at the middle of the room, the separation between singly and doubly reflected regions was not clearly evident. For the case where the receiver was placed near the wall, this initial region was characterized by a series of almost equal amplitude echoes arriving in quick succession forming an extended group of singly reflected paths. When chairs were added to the scenario, the region comprised of LOS and first order reflections were generally perturbed by destructive interference effects. The second and third order reflections were perturbed by constructive interference, increasing the multipath energy in this regime. This region typically contributed in the range of $-10$ to $-20$ decibels and resulted in the removal of time durations where the CIR had effectively no arrivals in the empty room case. In the third scenario, the presence of stationary human beings in the channel affected the CIR differently based on the location of the receiver. For the case where the receiver was in front of the seated students and close to the transmitter, some of the distinct first order reflections were smeared out by the effect of interference with human bodies in the return path. For positions where the path delay from these reflections extended to the doubly and triply reflected regimes of the previous scenarios, constructive interference effects were observed. Finally, in the last scenario, where students were in motion in the room and LOS paths were blocked, the primary effect was the merging of single, double and triple reflected regimes into a contiguous regime.
that continued into the diffuse regime comprised of higher order reflections. Here little
distinction could be made on position dependent features when the receivers were
located in the middle of the room. For the extreme cases, of receiver placement close to
the transmitter and close to the wall and corners, one could still distinguish a separation
between the first, second, third order regime and higher order reflections forming the
diffuse region of the CIR. The diffuse component forms the tail of the CIR and its decay
rate was comparable for all cases and positions of classroom measurements made in this
study. The coherent region typically extended from 25–50nseconds and the diffuse
component was captured until about 100 nsecs, beyond which the noise floor was
reached at −50:−60 dB.

The measurements made in CACT captured the intensive multipath nature of a
working environment that consisted of a people moving about, seated and several
obstructions arising from metal furniture and partitions. The key feature observed here
was that the rate of decay of the CIR in the diffuse regime was not invariant with
receiver position as observed in classroom measurements. The presence of partitioned
regions within the outer environment, coupled with the strong reflectors present in the
room creates specific regimes of spectral nulls that strongly influence the CIR patterns.

6.1.2 Models and Characterization of CIR: Wideband analysis

In this section, the wideband channel transfer function derived from the
measurements was characterized using a pole-zero model. A preliminary analysis of a
one-dimensional free space channel was carried out to derive the functional form of the channel impulse response as a function of channel dimension, T-R separation and reflection coefficients. It was shown that for such a channel, the number of poles were determined by the channel dimension alone and the magnitude was a function of the magnitude of the reflection coefficient. The zeros however were in general dependent on the T-R separation distance as well as the unique positions of receiver and transmitter. Therefore in general, the features of the coherent region may not be easily tabulated. Using these insights a pole-zero characterization of the measurements obtained for the empty classroom scenario were obtained. Matching the coherent region with sufficient number of zeros, the order of the poles was selected based on a minimum mean square error criteria. The distribution of poles in the Z-plane was examined as a function of receiver positions. A cluster of poles were formed in the plane that may be construed to be the invariant set of poles in the ideal case. On examination of the distribution of zeros, a subset of invariant zeros were also determined as the transmitter position changed. These typically resided on a circular trajectory in the Z-plane. A number of zeros however occupied random locations in the Z-plane that changed when the receiver was moved.

The wideband CIR was also described using characteristic delay parameters of power delay profile. These parameters include the mean excess delay, RMS delay spread, maximum excess delay calculated based on a threshold power level and coherence bandwidth. The results showed that when a LOS signal was present, the
mean excess delay exhibited the direct proportionality expected with T-R separation distance. The mean excess delay ranged from 10 – 25 nsecs for the classroom scenarios. For CACT, the results are not directly predictable using the T-R separation. They exhibit significantly higher values ranging from 20 – 67 nsecs. If the receiver is designed to synchronize its signal based on the delay where the CIR has a maximum value, the distance from this position to the mean excess delay signifies the variance or spread of the multipath components. It is found to be the smallest for cases where the T-R separation is small or when the receiver is located near the reflecting surfaces. For locations where all of the interference phenomenon are dominant, this parameter increases in size. In general, this parameter almost doubles in size when LOS is absent and channel is time-varying. These same features may also be obtained by calculating the RMS delay spread parameter.

### 6.1.3 Narrowband Analysis

The third component of this thesis examined the performance of the channel under various narrowband conditions, where the bandwidth and center frequency were varied to match existing and upcoming standards specifications. The 1.9 GHz, 2.35 GHz and 2.45 GHz carrier frequencies were selected to characterize wideband next generation cellular system, wireless broadband system and wireless LANs respectively. The narrowband CIRs were obtained by bandpass filtering the channel transfer function measurements. Under the assumption that the ISI power is to be maintained below a
certain threshold to guarantee a bit error rate (BER) performance, the required bandwidth increased nonlinearly with the transmission data rate. This feature suggests that wideband systems would have to implement transmission schemes that would be robust to increased multipath at higher bandwidths. The orthogonal frequency division multiplexing (OFDM) scheme has been proposed as a scheme that can effectively counter the effects of multipath transmission.

A simulation of the BER performance was carried out using QPSK modulation and OFDM. Although, interblock ISI can be effectively cancelled with the knowledge of the delay spread and by using a cyclic prefix, the effects of intrablock performance can impact the BER significantly if ISI power is high. In general, the transmitted symbols exhibit a scaling and rotation in the signal space, both of which are time-varying phenomenon when the channel changes.

6.2 Future Work

This dissertation work was done based on the channel measurements using a frequency sweep range of 30 KHz to 3 GHz. The candidate physical layer of IEEE 802.11a wireless LAN is 5.8 GHZ ISM band to support a maximum of 54 Mbps data rate. The indoor channels have been shown to be highly selective based on the frequency range considered. Therefore, the analysis of the 5.8 GHz band must be carried out to identify the potential shortcomings of this band in relation to selected channel environments. Most of the work done in this thesis is under the condition of a
stationary channel, with the receiver also being fixed in space. The effects of Doppler frequencies must also be considered, since they can even at slow speeds have significant impact if the transmission data rate is high. For this the continuous time measurements of channel must be designed to evaluate the spatial and temporal characteristics of the channel.
REFERENCES


BIOGRAPHY

Hark-sang Kim was born on 12 September 1966, in Sangju, Korea. He received his Bachelor of Engineering degree from Kyungpook National University, Taugu, Korea in February 1987, and the Master of Engineering degree from the alma mater in February 1989, both in Electronics. After that he joined Agency for Defense Development of Korea, national military laboratory. At there he participated in a number of project on ground weapon system, chemical sense system and laser remote detection technology as a researcher and a senior researcher until July 1999. He is currently a candidate for the Doctor of Engineering in Electrical Engineering, at the Center for Advanced Computation and Telecommunications, University of Massachusetts Lowell.

His current research interests are in the area of measurement, modeling and analysis of indoor wireless wave propagation for high speed cellular system, wireless broadband system and wireless LAN. He coauthored a number of papers, patents, program and technical reports. He was awarded silver grade of national defense science award in 1994, merit scholarships from 1983 to 1988 and Engineering Dean award in 2002. He is a member of IEEK and associate member of Sigma Xi.