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**Characterization of
29 GHz Indoor Radio Channels
Using A
2-Channel Angle Diversity System**

by

Udo Licht, B. Eng. Mgmt.

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

Master of Engineering

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

May 16, 1996

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Using a 2-Channel Angle Diversity System"**

submitted by

Udo Licht

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

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Abstract

One of the possible designs for an indoor mm-wave communication system involves using angle diversity with sectorized horns on both the transmitter and receiver. Since an obstruction will rarely block all scattered components of a signal, this scheme may be quite robust against shadow fading as well as providing micro diversity gains. This paper examines a measurement system with one transmit and two receive antennas. The purpose of the research project is to design a viable broadband measurement system and examine channel fading characteristics including cross-correlation between receive channels as well as CDF's of received signal power. A 200 Mbps PN sequence is modulated onto a 29.125 GHz carrier, transmitted over a 20° horn and received by 30° horns pointed 30° apart. The signal power is sampled. The receive channel is found to be quiescent 40-60% of the time. Cross-correlation coefficients for fast fades on the two antennas are in the 0.1-0.2 range, but when shadow fading is added, many tests show correlations greater than 0.5. Fade margins at the 1% level are in the 13 dB range. Gains achieved by selection combining of the two receive channels are examined, and found to be low (max 3 dB gain at 1%). Various combinations of receiver horn pointing angles are tested. As well, characteristics of the received signal power are studied for measurements with an omnidirectional antenna on the transmitter, and correlations and fade depths are found to be shallower than when a horn is used.

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List of Acronyms and Abbreviations

ADC - analog to digital converter

A/D - analog to digital

BER - bit error rate

CDF - cumulative distribution function

cm - centimetre

CW - continuous wave

dB - decibel

dBi - dB gain of an antenna referenced to isotropic gain

dBm - decibels relative to one milliwatt of power

EHF - extremely high frequency

GHz - gigahertz = 1 billion cycles per second

GWSSUS - Gaussian wide sense stationary uncorrelated scattering

Hz - hertz (cycles per second)

IF - intermediate frequency

ISI - intersymbol interference

kbps - kilo-bits per second

kHz - kilohertz = one thousand cycles per second

LAN - local area network

LNA - low noise amplifier

LO - local oscillator

LOS - line of sight

Mbps - mega-bits per second

m - metre

m/s - metres per second

MHz - megahertz = one million cycles per second

mm - millimetre

msec - milliseconds = one thousandth of a second

mW - milliwatt = one thousandth of a watt
nsec - nanosecond
PC - personal computer
PCEG - programmable complex envelope generator
PDF - probability density function
PN - pseudorandom
PRAT - programmable step attenuator
PSK - phase shift keying
QPSK - quadrature phase shift keying
RF - radio frequency
RMS - root mean square
Rx - receiver
s - second
SNR - signal to noise ratio
SONET - synchronous optical network
Tx - transmitter
UHF - ultra high frequency

List of Mathematical Symbols and Notation

D - antenna directivity

e - 2.71828183

F - noise figure

$F_{\max}$ - maximum Doppler frequency

G_x - gain of component 'x'

I_0 - zeroth order modified Bessel function of the first kind

K - Boltzman constant = $1.37 \cdot 10^{-23}$ Joules/degree

k - ratio of diffuse to steady components for a Rician distribution

N - distance / power law exponent

P_{inst} - instantaneous power

$p_r(r)$ - PDF with respect to r

$P(r \leq R)$ - CDF of the signal envelope, r

$r(t)$ - equivalent lowpass received signal

V - velocity

$\bar{x}$ - mean signal level

Γ_d - root mean square delay spread

Π - 3.14159

η - antenna efficiency

λ - wavelength

θ_k - phase shift for the k^{th} path

σ - standard deviation

$\sum_a^b$ - summation from a to b

1. INTRODUCTION

1.1 Background

1.1.1 Rationale For Indoor Wireless Systems

In the last ten years or so, a revolution has taken place in the telecommunications industry. The explosive expansion of the computer market coupled with the rapidly increasing capabilities of these machines has led to an ever-growing demand for better digital communications networks. A few years ago, people were content to connect to low bit-rate office-wide LAN's that offered services such as e-mail and allowed basic client-server systems to be set up. The rapid growth of the Internet means that these LAN's are now connected in a worldwide network and, with the introduction of interactive multimedia services, demand for broadband (10+ Mbps) systems is rising. Although technological innovations in cable transmission systems, such as fiber-optic lines, can be employed to meet this demand, there are several practical problems associated with their implementation. While it is highly cost-effective to replace inter-city trunk lines with fiber, re-wiring of office buildings and private homes would be prohibitively expensive at the moment. Coax lines would be cheaper to install than fiber, but would limit the bandwidth and would still require a laborious and disruptive re-wiring of each building. Broadband indoor wireless systems may be the answer to this problem, since little re-wiring is required and available bandwidth is high.

1.1.2 Broadband MM-Wave Systems

Basic narrowband (voice-only) wireless systems have been in the marketplace for a while, in the form of cellular phones, and the technology is mature. The characteristics of mobile radio channels in the 900 MHz band have been studied extensively and problem areas, such as impairments due to obstructions, are well defined. It is now widely proposed to apply these concepts to a broadband system in the mm-wave (20-60 GHz)

range and design an indoor network capable of competing with fiber-optic cables in terms of capacity. Wireless implementations have several advantages over fixed link designs:

- 1) New wiring is only needed as far as the base stations, lowering installation costs and complexity, and minimizing disruptions.
- 2) The end-stations are tetherless. Thus, no re-wiring is needed if they have to be moved.
- 3) The same technology can be applied to a mobile setting: The end-station may be a robot on a factory floor or instruments attached to a hospital bed.
- 4) Due to the high carrier frequency, achievable bandwidths are much greater than for alternatives that use existing cables, such as twisted-pair (telephone cable) or the power grid.

Wireless systems in general also have some unique problems:

- 1) Licensing requirements. Most frequency bands are strictly regulated and considerable paper work and cost may be involved to get regional permission to transmit. Co-channel interference is the problem here. All frequencies below about 1 GHz are heavily used, so only very narrowband systems would be viable at these frequencies.
- 2) Power output: Strong RF signals, especially in the microwave frequencies, can present a health hazard.
- 3) Fluctuations in signal strength due to changes in the environment, obstructions blocking an antenna (shadowing), or vector addition of multiple components of a signal at the receiver antenna (multipath fading).
- 4) ISI at high bit rates due to echoes of a signal arriving at the receiver at later times than the earliest received signal. (See section 2.1.1.1.2)
- 5) Interference from other systems using the same frequency.

MM-Wave broadband transmission alleviates some of these problems:

- 1) The frequency range is relatively unused, and Canadian licensing requirements for the increasingly popular 28 GHz range are only now being drawn up by Industry Canada.

- 2) Attenuation of materials such as concrete is much higher at these frequencies, lowering the chance of co-channel interference with systems in different rooms. Other systems operating in the same frequency range and across the same bandwidth can therefore be placed closer together, allowing each system to use a greater bandwidth.
- 3) In narrowband systems, multipath fading causes the signal to fade across the entire frequency spectrum, while a broadband system will exhibit 'frequency-selective' fading - some frequency components will be affected less than others at a given time (see section 2.1.1.1.2). As a result, fades caused by multipath scattering will not be as deep for a broadband system.
- 4) Mm-wave systems are ideal for broadband indoor wireless because the high carrier frequency allows bandwidths for each portable to be in the GHz range.

1.2 Thesis Outline

1.2.1 Background

The probability that broadband mm-wave systems might become serious competitors to fiber distribution in the near future prompted the Canadian Institute for Telecommunications Research (CITR) to launch a major initiative to examine the characteristics of EHF indoor radio channels. One of the ideas, to be tested at Carleton University, was whether a system of sectorized horn antennas, pointing in different directions on both the transmitter and receiver would result in a robust design, since the usual high attenuation caused by an obstruction in the LOS path might be mitigated by being able to choose one of multiple reflections. This type of angle diversity might also be used to lessen multipath fades. The high gain of the horns could then be exploited to lower the required transmit signal power, while their high directivity would lower co-channel interference.

The final goal of the CITR project is to design, build and test a viable indoor wireless system capable of supporting a large number of portable stations operating in an

office environment at 155 Mbps (SONET-OC3 rate). The base stations would be set up in a grid in an office and portables coming on-line would test the signal strength from each base station and would create a connection with the strongest one. Occasionally, each portable would test the signal strength in each of its sectors, as the base switched between its horns. The base-portable horn combination producing the highest signal power would then be used for data transmission. Moving portables could also be handed off to other base stations.

1.2.2 Purpose of Thesis and Expected Thesis Contributions

This thesis investigates several aspects of mm-wave indoor radio propagation that are relevant to the CITR Broadband Indoor Wireless Project. Several universities are participating in this project, and Carleton was responsible for solving several of the problems in the initial phase. This thesis proposes solutions and resolves the issues detailed below:

- 1) To study the effects of theoretical models in practice, it was necessary to develop a measurement system consisting of a transmitter and receiver that was sensitive enough to allow accurate measurements across typical indoor distances (30 ft) and fast enough to allow transmission at 155 Mbps, at a minimum.
- 2) Little data is available on broadband fading characteristics at 29 GHz, especially when directional antennas are used. This information is essential for the construction of a link budget which must take outage probabilities into account. The characteristics of the broadband received signal power, including multipath and shadow fading, were therefore studied. While the use of directional horn antennas had priority, channel fading characteristics using omni-directional antennas were also examined.
- 3) An initial study of channel characteristics was required for a proposed system using sectored-horn antennas. (See section 1.2.3 below). Initially, it was necessary to know what gains could be achieved using 2-antenna angle diversity on the receiver, and thus the cross-correlation of the signal power on the two receive channels was examined, as well as the effects of selection combining.

1.2.3 Overview of The Experiments

The measurement system for this experiment included a transmitter that transmitted a PN sequence as a 200 Mbps/channel QPSK signal modulated on a carrier frequency of 29.125 GHz. The receiver filtered the incoming broadband signal to a bandwidth of 250 MHz. Since the objective was to analyze the fading envelope, a crystal detector connected to an A/D board in a computer was used to sample the output power of the channel. The computer sampled each channel at 1-5 KHz and the channel characteristics were analyzed using MATLAB.

Since the effects of using directional antennas and angle diversity was to be examined, the receiver had two 30° horns on its front-end, each pointing at a different area of the room. The data acquisition system cycled between the two channels, taking 1 sample from each. The transmitter alternately had either a horn or an omni-directional antenna attached to it.

The effects of movement were studied by either having the receiver or people in motion. Several different areas in the Minto Centre and the Mackenzie Engineering Building at Carleton University were used for taking measurements.

1.2.4 Thesis Organization

This thesis is organized into three main sections: the first describes the theoretical background to the experiment, the second describes the design of the transmitter and receiver, and the third describes the characteristics of the data collected and draws conclusions from the results.

Specifically, Chapter Two describes the theoretical concepts related to wireless communication in general and mm-wave frequencies in particular. The effects of multipath propagation on fades are examined, as is path loss from free-space propagation and attenuation due to obstructions. Various types of diversity are studied, as are methods of combining diversity channels to lower combined signal fade depths.

Chapter Three describes the current knowledge as it relates to 29 GHz indoor radio channels, including the results of impulse response measurements at various

locations as well as path loss and preliminary fading measurements. It also outlines the thesis problem in detail.

Chapter Four details the design, building and testing of the transmitter and receiver, and describes in detail the types of measurements that were taken with the setup.

Results from the data collection portion of the thesis are studied in Chapter Five. In particular, CDF's, cross-correlation of channels, and gains from selection combining are examined.

Conclusions and resolutions to the problems to be solved as part of the thesis objectives are presented in Chapter Six.

2. RELEVANT COMMUNICATIONS THEORY

2.1 Fading

2.1.1 Causes and Types

Signal fades on wireless channels can be grouped into two categories: 'shadow' fading and 'multipath' fading. Shadow fading is due to obstructions moving between the transmitter and receiver, while multipath fading is due to time varying vector additions of signals that take different paths between the transmitter and receiver. Fading is of great concern on wireless channels since overcoming them requires the transmission of much more power than would be necessary on a quiescent channel.

2.1.1.1 Multipath Effects

2.1.1.1.1 Rayleigh and Rician Fading

Objects in the vicinity of a wireless system can cause two types of phenomena: reflections and perhaps scattering cause multipath fading and ISI. Unless the transmitted beam is very narrow, and a line-of-sight (LOS) path exists, the transmitted signal will reflect from objects in the room and several replicas of the original signal will arrive at the receiver. The time of their arrival is determined by the lengths of the paths they take.

The beamwidth of each replica is large enough that different components of the beam strike different objects near the receiver and are reflected into the receive antenna along slightly different paths. Since the path length of the reflected signals is slightly different, they will arrive with different phases at the receiver. When these signals are now added, constructive or destructive interference occurs. The addition of many such signals results in 'multipath' fading. The resultant signal fades in time and can sometimes be represented by quadrature Gaussian components [4]. This results in a Rayleigh distribution of the signal envelope at the receiver. In mathematical terms, if quadrature

components 'x' and 'y', each a Gaussian random variable with zero mean and variance σ^2 , are added to form signal u:

$$u=x+jy.$$

Then the resulting probability density function of the envelope of the signal, $r=|u|$, is given by [4]:

$$p_r(r) = \frac{r}{\sigma^2} \cdot e^{-r^2/2\sigma^2}, \quad r \geq 0. \quad (1)$$

The cumulative distribution function (CDF) is then given by:

$$P(r \leq R)=1-e^{-r^2/2\sigma^2}, \quad r \geq 0. \quad (2)$$

The instantaneous received signal power is $P_{\text{inst}}=r^2/2$. Therefore, the probability that the instantaneous received power is less than a threshold P_o is given by:

$$P(P_{\text{inst}} \leq P_o)=1-e^{-P_o/\sigma^2} \quad (3)$$

The above equations only hold true if all signal components - with phases varying uniformly from 0 to 2π - have the same strength and there is no dominant component such as a line-of-sight path. In the case of a dominant coherent component, the probability density function of the envelope becomes Rician:

$$p_r(r)=\frac{r}{\sigma^2} \cdot e^{\frac{-r^2+2kr^2}{2\sigma^2}} I_0\left(\frac{r\sqrt{2k}}{\sigma}\right), \quad 0 \leq r \leq \infty. \quad (4)$$

where I_0 is the zeroth-order Bessel function of the first kind and k is the ratio of power in the specular signal component to the power in random multipath.

$$k = \frac{A^2}{2\sigma^2}, \quad A \geq 0. \quad (5)$$

where A is the peak value of the specular component.

If the dominant component tends to zero, the k factor approaches zero and the above distribution becomes Rayleigh .

In all cases, the signal strength varies rapidly compared with shadow fading as the antenna or, or an object near the wireless system moves, since a movement of even $1/4$ wavelength (0.25 cm at 29 GHz) is enough to cause significant phase changes in reflected signal components based on Doppler considerations. Fading speeds are often in the range of 10-100 Hz for indoor channels.

2.1.1.1.2 Frequency Selective Fading

As indicated at the beginning of the previous section, multipath effects occur not only from signal components reflected from surfaces near the receiver, but also because signal components reflect from objects anywhere in the beam of the transmit antenna. Since these objects may be spaced far apart, the path length of each component may be quite different, and replicas of the same signal may arrive many nsec apart. If the symbol rate of the system is low enough, all replicas of one symbol will arrive before the next symbol does. In this case, multipath causes no inter-symbol interference and the channel is deemed to be frequency non-selective. With a high symbol rate, replicas of the first symbol will add constructively and destructively to replicas of the next one. If they interact with the main lobe of the second symbol, the waveform is distorted and ISI results. If the Fourier transform of the new waveform is examined, components of the frequency spectrum have been altered to different - and sometimes independent - extents. This is known as a frequency-selective channel. Whether a channel is frequency selective depends on both the symbol rate and the variation in path lengths that signal components take. The power weighted variance of time delays at which signal components arrive is

known as the RMS delay spread of the channel, as explained in section 2.2 below. A channel is defined to be frequency selective if frequency components in its bandwidth fade selectively. The bandwidth over which fading of frequency components shows a higher correlation is known as the 'coherence bandwidth', of the channel. A simple approximation for its value under GWSJS (Gaussian Wide Sense Stationary Uncorrelated Scattering) conditions is the reciprocal of the RMS delay spread. For such channels, it has been found that the influence of RMS delay spread is negligible so long as it is less than approximately 1/100th of the symbol duration [1].

The classical models assume that correlations between frequency components will not change much for different fading intervals. This is known as 'uncorrelated scattering'. Measurements by Bultitude and Imbeau [1] cast some doubt as to the applicability of this method to indoor channels, however.

2.1.1.2 Doppler Spread and Slowly Fading Channels

When the environment causing multipath effects to occur in the receiver changes slowly enough that more than one symbol period elapses before significant changes occur, the channel is 'slowly fading'. The time period over which the fading environment is constant is known as the 'coherence time' of the channel. Movement of either the equipment or a nearby object over at least 0.25 of a wavelength is generally required to considerably change the depth of a fade.

The Doppler spread of the channel can be used to estimate the frequency of fade occurrences. The Doppler phenomenon occurs because a moving receiver will see signal components that arrive from directly in front of it as being higher in frequency. At the same time, signal components arriving at a 90° angle to the direction of movement will appear at the proper frequency, and signals arriving from behind will appear to be lower in frequency. The peak Doppler shift is expressed as:

$$F_{\max} = V/\lambda \quad (6)$$

where V is the velocity of the receiver, and λ is the wavelength of the carrier frequency.

It is clear that equation 6 can express the number of wavelengths traversed per second. Under GWSSUS conditions it can be shown that the coherence time is approximately $1/F_{\max}$.

At a carrier frequency of 29 GHz, and a receiver speed of 1 m/sec, $F_{\max}=100$ Hz and the coherence time is approximately 2.5 msec. In digital systems, fading is classified as slow so long as symbol durations are much lower than the coherence time on the channel [4]. Since the symbol rate for the broadband EHF system is 200 Msymbol/sec, with a symbol time of 5 nsec, it is clear that all channels in this study will be slowly fading. The Doppler effect is the same if objects moving in the vicinity of a stationary receiver are causing the multipath fading environment to change.

2.1.1.3 Shadow Fading

'Shadow' fading is caused by moving obstructions between the transmitter and receiver that attenuate the signal. In a mobile setting, these obstructions can be buildings, trees, cars, or the ground itself if the mobile unit enters a depressed area. In an indoor setting, the obstructions are usually people. In contrast to an outdoor, mobile setting, the portables for broadband indoor communication are usually stationary (attached to a workstation, for example), and the obstructions move. In an outdoor channel, there are a great many types of obstacles that will cause these fades.

If such fading results in a normal distribution of fade depths, on a decibel scale, the distribution of the result will have what is referred to as a 'log-normal' distribution:

$$p(x) = \frac{1}{\sqrt{2\pi}\sigma} * e^{\frac{-(x-\bar{x})^2}{2\sigma^2}}, \quad (7)$$

where x is in dB, $\bar{x}$ is the mean value of the signal (averaged over all local fluctuations) and σ is the standard deviation.

A mobile can move a distance of many wavelengths before this fade value changes. Similarly, a person moving through the path of an indoor channel can move a reasonable distance (several cm to 1 m) before the receiver sees a significant change in signal power due to enhanced or reduced shadow fading. These fades therefore tend to vary slowly - on the order of seconds - and are almost independent of symbol rates, since material attenuation is not much different for a narrowband signal than a broadband one (which has only slightly higher frequency components).

Since material attenuation tends to increase as frequency increases, signals with a carrier in the 20-60 GHz range are affected much more by shadow fading than conventional signals below 2 GHz.

2.2 Multipath Spread

Multipath delay occurs when reflections and scattering cause a transmitted signal to be split into various components that follow different paths and therefore arrive at different times at the receiver. The amount of time elapsed between the first and last replicas of the signal is known as the multipath spread. A form in which this variable is sometimes quantified is known as the root-mean-square delay spread, τ_d , calculated as the square root of the ratio of the power in the second term to that in the first term of a Taylor series expansion for the transfer function $H(f)$ of the channel [22]. This is shown in equation 8 below:

$$\tau_d = \sqrt{\tau^2 - (\bar{\tau})^2} . \quad (8)$$

where

$$\tau^n = \frac{\sum_i \tau_i^n |h(\tau_i)|^2}{\sum_i |h(\tau_i)|^2}, \quad n=1,2. \quad (9)$$

The τ_i represent multipath signal arrival times in a discrete measurement system [1].

Since low-power noise and equipment-related impulse response aberrations can cause incorrect readings when incoming components have degraded to very low levels, a cut-off point is implemented in practice. A cut-off level between 20 and 30 dB is typical [2] [3].

As explained in section 2.1.1.1.2 above, this statistic is vital because it determines whether a channel will experience severe ISI. It has been shown that, on a GWSSUS channel, a symbol rate of approximately 1/100 of the reciprocal of the RMS delay spread at the 90th percentile is the maximum rate that can be used without degradation [1].

2.3 Path Loss

Attenuation of a transmitted signal due to path loss is determined by two factors:

- 1) Free-space loss, which depends on the frequency of the transmitted signal and the distance to the receiver,
- 2) Attenuation of materials through which the signal must pass.

Free space loss is governed by the equation:

$$\text{Loss (in dB)} = 10 \log_{10} \left[\frac{4\pi d}{\lambda} \right]^2, \quad (10)$$

where d is the distance from transmitter to receiver and λ is the wavelength.

A comprehensive model of path loss is either a summation of the free space loss and the effects of all obstructions, or obstructions and reflections are integrated into the path loss model by expressing the path loss as a function of distance raised to a variable exponent 'n':

$$\text{Total Path loss (dB)} = 10\log_{10} \left[\frac{4\pi}{\lambda} \right]^2 + 10\log_{10} (d^n), \quad (11)$$

$$= (\text{free space loss at 1m}) + (\text{effects of distance, reflections, wall losses})$$

The first model is more accurate for calculating losses since the effects of each wall are considered separately, however the mitigating effects of reflections are not included. The second model is more tractable since it allows easy comparison of results from different studies.

2.4 Effects of Antenna Design

Two types of antennas are considered in this thesis. The omni-directional antenna radiates (or receives) energy uniformly in all directions in one plane, and usually over a broad range in the orthogonal plane. A horn antenna radiates all energy into a specific sector, and the gain of such an antenna with a rectangular aperture is given as follows:

$$G = \eta D, \quad (12)$$

where η is the efficiency of the antenna (usually between 0.6 to 0.95 depending on the field type- a tapered field reduces it to 0.75) and D is the directivity, given by:

$$D = 4\pi(a*b)/\lambda^2, \quad (13)$$

where a and b are the dimensions of the horn opening.

Directional receiver antennas have several advantages over omni-directional antennas when multipath effects are a problem. Apart from the advantage that a directional antenna has a higher gain, the narrow coverage allows the receiver to eliminate many of the scattered signal components, reducing multipath fade depths. All incoming signal components in the narrow area covered by the beam will have approximately the same path length, so the channel RMS time delay spread and resultant tendency for severe ISI to occur, is reduced.

While effective against multipath, directional antennas are at a disadvantage when shadow fading is a factor. Since the number of signal components is small and their arrival angles are almost the same, an obstruction can easily block all components at once, producing a deep fade. An omnidirectional receive antenna would see scattered components from other areas of the room that the obstruction would not block. The effect is compounded if the transmit antenna is also directional, since the narrow transmit beam reflects from few objects and is only weakly dispersed.

2.5 Micro Diversity

2.5.1 Space Diversity

To combat the effects of multipath fading due to reflections from nearby surfaces, microdiversity receiver designs can be used. Taking advantage of the fact that changes in receiver antenna position of $1/4$ wavelength can significantly alter the received signal strength, a system of two or more receiver antennas spaced more than $1/4$ wavelength apart will often see a significant difference in signal powers between antennas. This is known as a space diversity system. By selecting the best signal from the available array or by combining the signals from each antenna, a large reduction in deep fades can be achieved over non-diversity systems if multipath fading is strong. The amount of gain depends heavily on the strength of cross-correlation between the channels. Cross-

correlation strength is a function of how far apart the antennas are located, the distance to nearby scattering objects, and the range of angles that replicas of the signal arrive at.

2.5.2 Frequency Diversity

If a signal is transmitted over a channel at centre frequencies that are spaced apart by more than the coherence bandwidth, then the fact that the signals for the two frequencies will fade in an uncorrelated fashion. Again, these gains depend on the amount of correlation between the two frequencies. Due to the large cost of stable oscillators for mm-wave channels and the large bandwidth that each channel uses, this type of diversity is not practical for our purposes.

2.5.3 Time Diversity (Spread Spectrum)

If significantly delayed multipath components of a signal are present in a channel, significant gains may be realized by receiving each component separately as it arrives, and combining the result. Coding a narrow-bandwidth signal and spreading it over a wider bandwidth allows more components of the signal to be identified, since the wider bandwidth of the filter doesn't broaden the individual pulses as much. While practical for low bit-rate systems, this type of diversity has limitations for broadband systems because the initial bandwidth of the system is so high that it cannot be spread by the required factor of 10 or 100 without interfering with systems using adjacent frequency bands.

2.5.4 Angle Diversity

Angle diversity takes advantage of the fact that replicas of the signal arrive from various angles at the receiver. Several directional antennas, each covering a specific sector, are set up to receive only those replicas of the signal that fall within their narrow beamwidth. Much of the multipath scattering from nearby objects doesn't enter a directional antenna, so multipath fading is reduced, and an obstruction blocking one of the signal components is unlikely to block all of them, hopefully allowing one or more antennas to continue receiving full signal strength. Unless integrated antennas are

employed, diversity antennas are likely to be more a wavelength apart in a mm-wave system, so space diversity against multipath fading is automatically introduced. Angle diversity is therefore effective as both a micro and macro diversity scheme.

2.6 Macro Diversity

Macro diversity systems utilizing space diversity are designed to combat shadow fading. Since an obstruction that causes shadow fading will block the signal in a large area around the receiver, receiver antennas must be spaced far apart to increase the probability that if one antenna is blocked, another will still see an unobstructed path.

2.7 Combining Methods

To realize the gains in SNR associated with various types of diversity, the receiver must combine the input from each branch of a diversity system in such a way that the branches which are performing the best are favored and the others are either discarded or have their impact on the decision-making process minimized. Expressed mathematically, the receiver combines branches to form a decision variable as follows:

$$r(t) = \sum_{k=1}^L c_k y_k(t), \quad (14)$$

where:

L denotes the number of branches,

$r(t)$ denotes the decision variable,

c_k denotes the weighting coefficient for the k 'th branch, and

$y_k(t)$ denotes the received signal on the k 'th branch at time t .

This signal is described as the summation of a noise term and the original signal. The original signal is weakened by attenuation and has a phase offset:

$$y_k(t) = \alpha_k e^{-j\theta_k} u_k(t) + z_k(t), \quad k=1,2,\dots,L. \quad (15)$$

where

α_k denotes the channel attenuation,

θ denotes the phase offset,

u_k denotes the original signal, and

$z_k(t)$ denotes the additive white gaussian noise on the channel.

[4]

2.7.1 Maximal Ratio and Equal Gain Combining

Maximal ratio combining will theoretically yield the highest gains of any combining method since it employs all possible information about a channel. A knowledge of both the channel attenuation and phase shift of each receiver branch is necessary. The signal on each branch is weighted by a factor proportional to its (noiseless) signal strength and the phase shift is compensated for, before summation of the signals from all branches takes place. In equation 14 given in section 2.7, this would amount to setting the weighting factor c_k to:

$$c_k = \alpha_k e^{j\theta_k}. \quad (16)$$

This technique has the drawback of being very complex. As objects move and multipath components change, the phase of the signal on each branch will vary quickly, making it difficult to achieve a lock. It is also unlikely that the attenuation of each signal path will be known.

For equal-gain combining, phase information from each channel is still employed, but no weighting of each branch according to signal strength is done. The signals from each branch are simply co-phased and summed. Improvement in SNR is lower, and the complexity associated with co-phasing each branch is still a problem.

2.7.2 Selection Combining

Selection combining is a much more simple technique than the previous two, although it also shows the least improvement in SNR. At any given time, the branch with the highest received power is chosen for input. All other branches are ignored, and no attempt is made to determine the noise power on each branch before choosing. Since no signal summation is necessary, no phase information is needed. Expressed mathematically, at any time 't', the branch 'y_n' is chosen such that:

$$y_n > y_k, \quad k=1,2,\dots,L. \quad (17)$$

Therefore, the expression for the result of the combiner, at any time 't' is:

$$r = \sum_{k=1}^L c_k y_k \delta_{k,n}, \quad (18)$$

A variation on this scheme is switch-and-stay combining where a threshold is set, such that when a branch is chosen by the receiver, no switching is done until the signal level on that branch falls below the threshold. If the signal strength does fall far enough, switching to another branch occurs regardless of whether the signal level on the new branch is also below the threshold.

2.8 Channel Cross-Correlations

Evaluating the degree of cross-correlation between two branches of a diversity system can often help in determining the amount of gain that can be expected by combining them. In general, the lower the degree of correlation, the higher the expected gain. The amount of correlation between two sets of data, 'x' and 'y', each with 'n' data points, is defined as:

$$\text{Correlation Value} = \frac{\text{Covariance}(x, y)}{\text{Variance}(x) * \text{Variance}(y)}, \quad (19)$$

where

$$\text{Variance}(x) = \frac{\sum_{i=1}^n (x_i - \bar{x})^2}{n} \quad \text{and} \quad \text{Covariance}(x, y) = \frac{\sum_{i=1}^n (x_i - \bar{x})(y_i - \bar{y})}{n}. \quad (20)$$

As can be seen from the above equations, the only factor taken into account by the correlation value is the degree that each data point of a set is offset from the average value *for that data set*. When the mean signal powers on two diversity channels are different, this equation has limited value for predicting the degree of gain that can be achieved by combining, since it compares only the trend of signal strength changes and not the absolute signal strength on the two channels. This constraint holds especially true for angle diversity measurements, as shown in Chapter Five.

As a practical matter, the mean signal power of each channel will change over time (due to shadow fading, for example). It would therefore be erroneous to find the values of $\bar{x}$ and $\bar{y}$ by averaging over the entire data set. The data must be segmented and processed in stages, such that averaging only occurs over small sections of time, where the local mean signal power will not vary. A certain size of time 'window' is therefore selected for a data file, and is slid through the data. The window size depends on the amount of time that the mean signal power for the channels is expected to be invariant, and the number of data points that the window is slid past between two computations

depends on the amount of time that the instantaneous signal power on a channel is expected to be relatively invariant.

2.9 Communication System Design Considerations

2.9.1 Receiver Noise Figure

One of the most important aspects to consider in the design of any wireless system is the SNR that will be required to achieve a certain Bit Error Rate (BER). The amount of noise introduced by the receiver can be a significant factor. For the purpose of this thesis, it was important to determine this factor to be assured that for all likely depths of fading, the signals would not be lost in the noise floor.

The noise at a component output consists of the addition of thermal noise introduced into the system at the component input to the noise introduced by the component itself. A measure of this additional noise due to the component is the *noise figure*, F , calculated as the degradation of SNR between the input and output of the component. As outlined in [5], the equation for calculating the overall noise figure for a multi-component receiver is:

$$F = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \frac{F_4 - 1}{G_1 G_2 G_3} \dots + \frac{F_x - 1}{G_1 G_2 \dots G_{(x-1)}}, \quad (21)$$

where F_x is the noise figure for component number 'x' in the chain, starting at the front end, and G_x is the gain for component number 'x' in the chain.

3. REVIEW OF CURRENT RESEARCH AND REASON FOR THESIS

3.1 UHF and EHF Channel Characterization

3.1.1 Research Objectives and Achievements To Date

The underlying purpose of research on channel characterization for wireless systems is to examine the channel impairments that decrease system BER performance. Much work has been done on outdoor mobile environments in the 900 MHz and 1.75 GHz range, while comprehensive work in the mm-wave range has generally been restricted to impulse response measurements and path loss characterization. Models for dispersion and fading characteristics of UHF outdoor mobile channels have been fairly well developed, but are lacking for indoor mm-wave systems. To some degree, they are applicable to the EHF band as well, so they are presented as background information below. Much of the theoretical results for mm-wave channels are based on models developed for lower-frequency bands. Some measurements from the UHF band are discussed below, together with measurements from the mm-wave channels for comparison.

3.1.2 Multipath Spread

3.1.2.1 Models

A model for the theoretical representation of multipath spread was proposed by Saleh and Valenzuela in 1987, based on their work at 1.5 GHz [6]. The model was based on the exhibition of ray-like behavior by distinct multipath components. It was proposed that components arrived in clusters at the receiver, due to the widely varying path lengths of each scattered ray. The varying arrival times of each cluster cause 'multipath spread'. Local scattering caused the varying component strengths within each cluster. Arrival

times of clusters were modeled as a Poisson process and the signal strength of clusters was modeled as a Rayleigh probability density function, as was the signal strength of components within a cluster. A good, non-mathematical overview of a ray-based model for multipath scattering is also given by Mitzlaff [7].

3.1.2.2 Measurements

A paper by Davies et al. [8] in 1991 was one of the first to report RMS delay spreads in the mm-wave band. At 60 GHz, static RMS delay spreads were found to be around 30 ns at the 90th percentile for large rooms but only 10 ns for smaller rooms. Horn antennas were used for these measurements, so RMS delay spreads were somewhat lower than if omni-directional antennas had been employed.

A later paper by Bultitude and Imbeau [1] found RMS delay spreads at 40 GHz to be 28 ns at the 90th percentile for 8 m long links in moderately sized rooms. Averaging over various environments, median RMS delay spreads were in the 20-35 ns range. Among a variety of areas studied by Bultitude and Hahn [9], measurements in a small (22 by 16.5 ft) room showed that static RMS delay spreads at the 90th percentile were 16 ns for both 40 GHz and 60 GHz measurements when the room was empty, and 20ns when it was furnished. Omni-directional antennas were used. Multipath conditions were the most severe in large measurement areas, with 33 ns RMS delay spreads at the 90th percentile. Results from the 950 MHz band for the same areas showed that RMS delay spreads were usually 15-20% higher. It was also found that although RMS delay spreads increased as a function of distance in an empty room, this effect was mitigated by furnishings. Kalivas et al. [10], in experiments where several larger rooms at Carleton university were examined, found that at 21.5 GHz, average RMS delay spreads for an omni-omni setup with an LOS path were 25.9 ns, while a setup with a horn on the transmitter resulted in a value of 26.6 ns. As corroboration, Smulders et. al [11] found that median RMS delay spreads at 58 GHz were in the 13-98 ns range using omni-directional antennas. All RMS delay spreads over 40 nsec were found in long hallways, very large rooms, or large rooms with concrete or metal walls.

As mentioned above, RMS delay spreads at lower frequencies are usually slightly higher. Bultitude et. al [2] found that at 900 MHz and 1.75 GHz, median RMS delay spreads were 26-30 ns and 28-29 ns respectively, with the larger spreads found in larger buildings. Saleh and Valenzuela [1], also using an omni antenna, found median spreads of 25 ns, with maximum spreads of 50 ns. These spreads were however calculated in rooms only, and with non-LOS paths. Hallway measurements with LOS paths and long links found RMS delay spreads up to 150 ns.

3.1.2.3 Conclusions Regarding RMS Delay Spreads

RMS delay-spread characterization for the EHF band is fairly well advanced. Although no models have been created for this frequency in particular, a comparison of EHF RMS delay spreads to UHF RMS delay spreads shows no reason why the model developed by Saleh and Valenzuela would not be valid. It is clear from the above results that any channel operating in the 100 Mbps range will experience severe ISI since even the smallest RMS delay spreads found (10 ns) would still limit a GWSSUS channel to 1 Msymbol/sec before degradation occurred (see section 2.1.1.1.2).

3.1.3 Envelope Fading Statistics

The standard models for envelope fading characteristics are the same as outlined in Chapter 2. No new models have been proposed for the indoor channel or for EHF measurements in particular.

Few measurements of envelope fading have been done in the 20-60 GHz band. The ones that were done concentrate mainly on CW measurements, and the only wideband measurements performed to date were limited to signal bandwidths of 40 MHz.

A study in 1992 by Liu and Delisle at Laval University [12] characterized fading at 21.6 GHz, 37.2 GHz and 59.6 GHz using horn antennas with 10°-30° beamwidths on both the transmitter and receiver, and employing CW signals. The receiver was moved during the measurements, and no other movements were present. Long-term fading was found to follow a log-normal distribution with a standard deviation of -4 dB, -4.3 dB and

-3.8 dB at 21.6 GHz, 37.2 GHz and 59.6 GHz respectively. Short term fading was Rician, with K ($10 \cdot \log_{10} k$) values of 1.64 dB, 4.3 dB and 6 dB.

Kalivas et al. studied fading at 21.6 GHz and 37.2 GHz using CW signals [13][10][14]. Measurement configurations included LOS and non-LOS paths and a two-channel diversity receiver was employed. The transmitter was moved during sampling, and interference from the movement of people was allowed. When both the receiver and transmitter were outfitted with omni-directional antennas and a 21.6 GHz signal was employed, average CDF's for LOS and non-LOS paths were found to be close to Rayleigh, although outage rates at certain fading levels of individual measurements departed up to 9 dB from the theoretical values. Results from 37.2 GHz were all close to Rayleigh. LOS and non-LOS measurements with horns on the receiver showed mixed results, with many departures from Rayleigh, even for the non-LOS cases. In four cases, a Rician distribution with a K factor of 9 was found. Deep fades were consistently smaller at 37.2 GHz than at 21.6 GHz. People moving at more than 5 m from the receiver produced fades less than 10 dB deep, but closer approaches yielded fade depths of more than 25 dB.

Some measurements were made by Bultitude and Imbeau [1] using CW signals as well as signals with a bandwidth up to 40 MHz modulated onto a 40 GHz carrier. Fading was caused by the movement of personnel. It was found that half the measurements could not be modeled as either Rician or Rayleigh, while others showed Rician distributions with K values of +2 dB to -3 dB at the 90th percentile, with the lower K values found at higher signal bandwidths. Fade depths at the first percentile decreased from -18 dB to -15 dB as the signal bandwidth was changed from 1 MHz to 40 MHz. This decrease is lower than expected, and the prediction that decreases in fade depths would be linear did not hold.

3.1.4 Path Loss and Material Attenuation

There are several ways of characterizing path loss, as described in Chapter 2. The most common method was to integrate the effects of obstructions and reflections into the

path loss exponent, N . Experiments performed by Kalivas et al. at Carleton University [13] [10] indicate that in corridors, the value for N is 1.2 at 21.6 GHz and 1.65 at 37.2 GHz. Non-LOS measurements with 1 to 4 walls between the transmitter and receiver yielded $N=2.95$ for 21.6 GHz and $N=3.3$ for 37.2 GHz. On average, each wall caused 3.7 dB of attenuation at 21.6 GHz versus 5.4 dB at 37.2 GHz.

Measurements by Todd et. al [15] corroborate these results. At 1.7 GHz, non-LOS environments produced results of $N=2.7$ versus $N=3.5$ at 37.2 GHz. LOS experiments indicated that $N=1.55$ at 1.7 GHz and $N=2.15$ at 37.2 GHz. The values for 37.2 GHz were not all taken in corridors, and the ones that were had $N < 2$.

A third study by Liu and Delisle [12] used directive antennas for their measurements at 21.6 GHz, 37.2 GHz and 59.6 GHz and found quite different results. In LOS environments, values of $N=2.01$, 2.05 and 2.04 were found for the three frequencies, suggesting that the directive antennas eliminated many of the reflected multipath components.

3.1.5 Wall Reflectivity

An important factor to consider when examining possible gains due to angle diversity is the number and strength of reflections in an area. Although the delay spread gives some measure of this, it does not determine the number of times that a ray was reflected. If a horn is used on the transmitter, then it is likely that a ray needs to be reflected several times before it experiences a large enough change in direction to enter a receive horn that is misaligned with the transmit horn. D.D. Falconer et al. [16] present results for wall reflectivity at 60 GHz as shown in Table 3-1:

Table 3-1: Wall Reflectivity at 60 GHz

Material	% Reflectivity
Plasterboard (1 cm)	2
Concrete	16
Aluminum	>99
Wood (20 mm)	2
Glass (3 mm)	16

This indicates that reflected rays will be substantially weakened after even one or two reflections from all materials except for aluminum.

3.1.6 Diversity Systems

The most common form of diversity proposed for combating fading in the EHF band is space diversity, and the performance of one such system was examined by Kalivas et. al [10][13] at 21.6 GHz and 37.2 GHz. Omni-directional antennas were used on the transmitter, and either horns or omnis were used on a two-branch receiver. Distances between antennas were about 10 wavelengths. Selection combining was employed. Both LOS and non-LOS measurements with omni antennas on the receiver showed that gains were close to what would be expected of Rayleigh channels for both frequencies. When horns were used in non-LOS measurements, gains were 2 dB higher than Rayleigh at an outage rate of 10^{-2} .

The theory behind a method to combat multipath-induced ISI, called Commutation Signaling, which uses bandwidth spreading modulation has been proposed by Turin [17] and examined extensively by Leib and Cheung [18]. It is based on the cyclical use of quasi-shift orthogonal signal sets so that the effective time between the reuse of identical symbols becomes large enough that severe ISI is avoided. The motivation for the use of this scheme is that the number of required signal sets increases linearly with the multipath spread, while the use of signal sets in a straightforward M-ary modulation scheme would increase exponentially, or conversely, the required bandwidth

would increase exponentially with M-ary orthogonal signaling. It was found that the use of commutation signaling with a square-law equal weight RAKE combiner would be ideal for the indoor channel. This type of system could be combined with antenna sectorization to minimize ISI.

Although no measurement results have been reported, Motorola's ALTAIR system is an example of a high-frequency microcellular architecture utilizing directive antennas and six-way sectorization on both the base station and portable to lessen the impact of fading and interference from other portables using the same frequency [7][19]. Six-way sectorization means that 36 possible paths exist for the system to choose from, allowing it to select the strongest combination at any given time. This 18 GHz system serves as a base model for the proposed CITR 28 GHz system.

3.2 Analysis of Recent Research

Research on EHF channel characterization is relatively new. Most of the work that was done has concentrated on path loss measurements and estimation of RMS delay spreads. Although some information is now available on channel fading, almost no wideband measurements have been done, and the maximum signal bandwidth investigated was 40 MHz. Few measurements have been done with horns on either the receiver or transmitter.

Little data has been gathered on diversity systems. Only space diversity was investigated, but it remains unclear as to how much of the fading was attributable to shadow and multipath fading from moving objects and how much was due to pure multipath from the moving antenna. It is unlikely that many of the results reflect the movement of people since a receive antenna spacing of 10 wavelengths would ensure that shadow fading would block both channels, allowing little gain from selection combining. It was not mentioned whether quiescent periods of the channel were removed before performance was examined. Also, cross-correlations between the receive channels were not examined [10][13].

3.3 Thesis Problem

3.3.1 Issues Requiring Resolution

To successfully design and implement a wideband MM-wave indoor wireless system, several issues clearly required resolution.

- 1) No wideband signal envelope measurements in the 100-200 MHz range had been taken, as needed for an ATM system running at the SONET OC-3 rate. The result is that no information about fade depths was available for such a system, and data had to be extrapolated from narrowband measurements.

- 2) No data were collected that reflected fade depths due to movement of people alone. All data either examined fade depths from multipath due to the movement of antennas in a local area, or added the infrequent normal movement of objects to the movement of the antenna. The result is that fade depths due to shadowing have not been adequately determined, even for a narrowband system. Also, the impact on multipath fading due to movement of objects alone had not been studied. Thus, it was not known whether multipath fades are different if objects move instead of the portable. These are important areas to examine since in most cases, both the receiver and transmitter will be stationary in an indoor system, while fading comes from objects moving in the room. This is in direct contrast to a cellular mobile system. Not many measurements had been done in the area of envelope fading to begin with, not even for pure multipath fades.

- 3) No data were available on the impact of angle diversity on the depth of fades, or on the degree of correlation between channels in an angle diversity setup.

- 4) A large concern with a microcellular architecture is co-channel interference due to several portables transmitting simultaneously. Directional antennas should reduce this problem, but no studies had been done to see how much signal spillover will occur from the properly aligned antenna to a neighbouring one. So far, it was assumed that such spillover would be negligible.
- 5) Differences between fading characteristics from systems utilizing omni-directional or directional antennas were not adequately examined. Specifically, it was not known how large the effects would be on the depth of shadow fades.

3.3.2 Thesis Purpose and Design

The purpose of this thesis is to provide answers to the above problems by designing and constructing a measurement system and performing a series of experiments that measure variations in the wideband signal envelope as environmental conditions change. The project was divided into four major sections:

- 1) Design, construction and testing of the transmitter and receiver systems, which included the selection and testing of individual components as well as designing and testing the layout of functional blocks within the systems.
- 2) Setup of the system for data acquisition. This included equipment calibration and design of a test plan for data collection, which specified, among other things, location of tests, type of antennas to use, sampling rate (determined from test runs), types of movement to examine and antenna alignment.
- 3) Data collection as per the test plan described above.
- 4) Data analysis including the creation and study of CDF's, examination of channel cross-correlations, and impacts of selection combining.

3.3.3 Importance of Expected Results

The design of any wireless system requires a fundamental knowledge of the channel characteristics. Without this, the severity of ISI cannot be determined and a link budget cannot be created. The major project on broadband wireless systems in CITR was set up in part to examine these issues for a broadband indoor MM-wave system operating in a workplace environment.

The completion of the work performed for this thesis will allow the designers of broadband indoor MM-wave systems to predict required fade margins with much greater accuracy, since both multipath and shadow fading were studied in a variety of environments. In particular, this thesis describes the fading characteristics of a system operating in an office or lab environment and concentrates on the use of directional antennas in an angle diversity setup. The creation of a sectored antenna system such as Motorola's 18 GHz ALTAIR system, is a major goal of the CITR project, so this thesis contributes significantly to the de-risking of such a design. Designers will have concrete results for both fading and channel cross-correlations to work with, allowing the creation of a link budget with a high level of confidence in its accuracy. Conclusions about the impact of using selection diversity and the amount of power spillover into adjacent antennas will also be valuable in determining parameters such as the degree of sectorization and antenna directivity.

4. DESIGN OF EXPERIMENTAL SETUP AND MEASUREMENT PROCEDURES

4.1 Introduction

This chapter describes both the design of the measurement system that was used to carry out all experiments for the thesis as well as the data collection procedure. Since the measurement system design and construction formed a major part of this work, this chapter will describe the experimental design in detail, including the various tests and calculations that were made to ensure accurate results. Functional diagrams of the transmitter and receiver as well as individual components are included in this description. All test results described below were performed in the lab, with the exception of component noise figures which were taken from manufacturer specifications.

The description of the data collection procedure encompasses all stages of the equipment setup including parameters such as calibration, antenna alignment and types of movement studied, determination of sampling rate, and locations tested.

4.2 Experimental Setup

4.2.1 Overview of The Experimental Setup

The setup of the measurement system can be described as follows:

A 200 Mbps signal was generated from both channels of a complex envelope generator (PCEG) and combined in a QPSK modulator to form an IF signal with a bandwidth of 400 MHz and a center frequency of 2.5 GHz. After upconversion to 29.125 GHz, this signal was transmitted over an omni-directional antenna or a horn antenna with a 20° beamwidth. A distance of 3 to 20 metres separated the receiver from the transmitter. Two horn antennas, each with a beamwidth of 30° and pointed 30° apart,

intercepted the signal. Only a single receive chain was used to sample both channels, so a high-speed computer-controlled RF switch selected the antenna input to sample from. After downconversion to 1.125 GHz, the signal was bandlimited to 250 MHz and passed through a crystal detector. The resulting signal was passed through to an A/D converter and sampled under PC control at a rate of 1 KHz. Sampling was alternated between the two channels, for an aggregate sampling rate of 2 kHz. Dynamic range at the receiver was 37 dB, limited by the noise floor, maximum input power of the receiver, and the range of the ADC.

Figure 4-1: Block Diagram of the Transmitter

4.2.2 Transmitter Design

4.2.2.1 Overview

The transmitter portion of the measurement system consists of five major blocks, each of which are examined separately below:

- 1) The baseband stage, including signal generation
- 2) The IF stage
- 3) The RF stage at 29.125 GHz, including the antenna
- 4) The LO setup for the IF stage
- 5) The LO setup for the RF stage

The connection of individual blocks and their components is shown in Figure 4-1.

4.2.2.2 Baseband

The baseband portion of the transmitter encompasses all components which perform the signal generation function. This includes

- 1) The complex envelope generator, or PCEG, which creates the signal
- 2) The HP 8675A signal generator which provides timing for the PCEG
- 3) DC blocks mounted on the dual output of the PCEG
- 4) Attenuators on the PCEG output

The connection of these components can be seen in Figure 4-1.

4.2.2.2.1 PCEG

The Programmable Complex Envelope Generator, or PCEG, was developed at Carleton University to handle a wide variety of possible output sequences. It was equipped with two separate output channels so that complex constellations could be generated with the aid of an I-Q modulator. For the purposes of this experiment, identical signals were generated on each channel, since only the signal amplitude was of interest. The use of both channels was therefore required only to boost the output power.

The function of the PCEG was to output two sets of digital signals, each of which had a bandwidth of 200 MHz (baseband). The PCEG has a high-speed 64 kbyte memory bank for each of its two channels, and a Gold Code PN sequence of length $n=2047$ was programmed into each bank. A PN sequence was used because the frequency spectrum of such a sequence approximates a Sinc function, with a mainlobe baseband bandwidth equal to the bit rate at which the sequence is generated.

The PCEG channels can output a range of 1024 possible signal levels. To minimize distortion, the high output was kept to level 511. The clock was supplied by using the RF output of an HP 8675A Signal Generator running at 200 MHz, with an output voltage level of 254 mV p-p. It was fed into the timing circuit of each channel. At startup, each channel of the PCEG cycled through its data bank in an endless loop, generating digital output at 200 Mbps. Although the sequence in each channel was identical, the clocks were not synchronized, resulting in an asynchronous output. However, since a PN sequence was being used, the combination of the two channels in the I-Q modulator created acceptable results.

For testing purposes, the output of each channel was fed into a digitizing scope (HP 54542A), and it was found that a maximum bit rate of 200 Mbps could be sustained without serious distortion.

4.2.2.2.2 DC block, Baseband Filter and Attenuators

The purpose of the attenuator was to provide a pad between the PCEG channel output and the I-Q modulator to reduce reflections. The DC block (SMC BLK-721N) was needed because the I-Q modulator could not accept DC components. The filter (MCL SBLP-300) was used to bandlimit the baseband signal.

Equipment Characteristics

Table 4-1: Baseband Filter and Pad Characteristics

Channel	Pad Attenuation	Filter Type
Channel A	5.8 dB	300 MHz lowpass
Channel B	6.05 dB	300 MHz lowpass

Figures B-1 and B-2 in Appendix B contain frequency response measurements of the attenuators.

4.2.2.2.3 Overall Characteristics:

Table 4-2: Signal Characteristics at DC Block Output

Channel	Bandwidth	Power
Channel A	200 MHz	-5.24 dBm
Channel B	200 MHz	-5.48 dBm

Please see Figures B-3 to B-11 in Appendix B for frequency response measurements of both channels, and spectral components of the output signals.

4.2.2.3 IF Stage

The IF stage includes all components which upconvert the signal to the IF frequency and prepare it for upconversion to the RF spectrum. The components are:

- 1) An I-Q modulator for upconversion
- 2) A filter to eliminate spurious signals
- 3) An amplifier
- 4) An attenuator

4.2.2.3.1 I-Q Modulator

The intent of future projects is to provide quadrature modulation capability, however for the purposes of this thesis, the I-Q modulator (Merrimac VMM-2B-

2500/71427) was used only to upconvert the signal from baseband to 2.5 GHz. For this reason, both input channels (I and Q) carried the same information - the PN sequence generated by the baseband stage. The I input was connected to channel B of the baseband stage, and the Q input was connected to channel A.

Input Signal:

Frequency: Baseband with DC removed - 200 MHz bandwidth (mainlobe)

Power: I channel: -5.48 dBm

Q channel: -5.24 dBm

LO:

Frequency: 2.5 GHz

Power: 8.6 dBm

Output Signal:

Frequency: Centered at 2.5 GHz

Bandwidth: 400 MHz (mainlobe)

Power: -9.66 dBm

4.2.2.3.2 Filter

The function of the filter (Anthony ABP 2500-400-4SS) was to remove unwanted out-of-band components generated by the I-Q modulator. It bandlimited the signal to 500 MHz.

Filter Characteristics

Power loss (filter and connecting cables): 0.9 dB

3 dB bandwidth: Approx. 500 MHz

Passband ripple: Approx. 0.6 dB (across 380 MHz)

Please see Figure B-12 in Appendix B for frequency response measurements of the IF filter.

4.2.2.3.3 Amplifier and Attenuator:

Amplification with a power amp (MCL ZHL-1042J) was needed to prepare the signal for upconversion to the RF frequency and the attenuator was used to fine-tune the output power of the IF stage.

Amplifier Characteristics:

Gain (uncompressed, at 2.5 GHz): 26.4 dB (+/- 0.3 dB)

Signal Output: 16 dBm

1 dB Compression Point: 17.6 dBm

Please see Figures B-13 and B-14 in Appendix B for a frequency response graph and a gain sweep of the IF amplifier.

Attenuation:

Attenuator Loss: 10.15 dB

Cable Losses (between amplifier and upconverter): 0.4 dB

Please see Figure B-15 in Appendix B for the frequency response of the attenuator.

4.2.2.4 Overall IF Stage Characteristics:

Output Power To RF Stage: 5.52 dBm

Gain of IF stage: 13.76 dB

Input frequency: baseband

Input signal: Two identical Gold code PN sequences of length $n=2047$.

Output frequency: 2.5 GHz

Output signal: Asynchronous addition of both input sequences.

Please see Figure B-16 in Appendix B for a graph of the IF stage output spectrum.

4.2.2.5 RF Stage and Antenna

The RF stage includes all components in the front end of the transmitter:

- 1) An upconverter and wideband filter
- 2) An RF amplifier and isolator
- 3) A variable attenuator for power control
- 4) A directional coupler for power monitoring
- 5) An omni-directional or horn antenna
- 6) Associated cables and waveguide terminations

The connection of all components is illustrated in Figure 4-1.

4.2.2.5.1 Upconverter

The Millitech 47471H-2130 upconverter used a high-frequency LO input to upconvert the IF signal to the RF band. An attached filter removed unwanted components generated by the upconverter. The final frequency of 29.125 GHz was reached by the addition of the LO frequency (26.625 GHz) to that of the IF signal frequency (2.5 GHz) inside the upconverter.

IF input: 5.5 dBm centered at 2.5 GHz

LO input: 9.2 dBm at 26.625 GHz

RF output: -7.8 dBm centered at 29.125 GHz

4.2.2.5.2 RF Amplifier and Isolator

This sensitive amplifier (Miteq AMF-6B-275300-19P) was needed to increase signal strength for transmission over the distances required by the project. The HP R365A isolator prevented reflections from unterminated or mismatched components in the RF front-end from damaging the amplifier.

Amplifier and Signal Characteristics:

Input Power: -7.8 dBm

Amplifier Gain (Uncompressed, at 29.0 GHz): 26.2 dB (+/- 0.4 dB)

1 dB Compression Point at 29 GHz: approximately 20.6 dBm (only tested to 0.8 dB compression to prevent possibility of damage)

Gain Flatness from 28.5 GHz to 29.5 GHz: 0.9 dB

Noise Figure at 29 GHz: 7.81 dB

Signal power after isolator: 18.2 dBm (+/- 0.6 dB)

Please see Figure B-17 in Appendix B for a gain sweep of the RF amplifier.

4.2.2.5.3 Variable Attenuator and Directional Coupler

An important function of the RF front end was power control, both for calibration and for allowing the receiver to be placed at various distances from the transmitter during measurements. A variable attenuator (HP R382A) was employed for this purpose. It was manually controlled and allowed a sweep from 0 dB to 50 dB attenuation in small steps. During measurements, attenuation rarely exceeded 16 dB, and the instrument's accuracy in this range was better than 0.2 dB.

The directional coupler (HP R752D) was used for the continuous monitoring of output power by connecting it to an HP 437B power meter via an HP R8486A head (20.5-40 GHz range). This allowed the use of the transmitting equipment over a long period of time with a high degree of confidence in its power output, since possible power fluctuations could be readily observed.

Directional Coupler and Signal Characteristics

Secondary Output: 20 dB below primary

Power Loss (Variable Attenuator and Coupler): 1 dB

Average Signal Power (from secondary output, incl. waveguide termination): 13.9 dBm

Max Power into Antenna: 12.9 dBm (+/- 0.7 dB)

4.2.2.5.4 Antennas

Both horn and omni-directional antennas were used on the transmitter. The horn antenna had a straight flare with a 3 dB beamwidth of 19.3° and a maximum gain of 19 dBi. The omni-directional antenna was a biconical, with a gain of approximately 2.5 dBi

along its azimuth. The radiation patterns of both antennas were measured at the anechoic chamber of the CRC, whose assistance is gratefully acknowledged. Please see Appendix F for selected patterns.

4.2.2.5.5 RF Cables and Waveguide Terminations

Coaxial cables were used in the front end to connect the output of the isolator to the input of the variable attenuator as well as the output of the directional coupler to the antenna. All other connections were either waveguides (WR-28) or direct couplings using K-type connectors. The cables were Suhner Sucoflex 104, with K-type (3.5 mm) connectors (approx. 0.2 dB loss each).

Loss of each cable (50 cm): approx. 1 dB

Waveguide terminations were required at the ends of each cable. Two types of terminations were used, with approximate losses as follows:

- 1) HP R281A (at upconverter and isolator outputs): Loss: approx. 0.1 dB
- 2) Generic (lower performance- all other locations): Loss: 0.3 dB

4.2.2.6 IF LO

This block created the 2.5 GHz clock signal that was needed for driving the I-Q modulator. It consisted of 6 components, as can be seen from Figure 4-2.

- 1) HP8673D synthesizer time base output (10 MHz)
- 2) Qualcomm Q 0410-1 PLO synthesizer
- 3) Amplifier (Mini-circuits MCL ZFL-1000)
- 4) Frequency doubler
- 5) Bandpass filter
- 6) Two amplifiers (ANZAC AMC-180)

Figure 4-2: Block Diagram of IF LO **4.2.2.6.1 HP 8673D Synthesizer**

The 10 MHz time base output was connected to the input of the Qualcomm synthesizer to provide timing signals that were synchronized with the RF LO and the receiver IF LO timing circuits.

4.2.2.6.2 Qualcomm Synthesizer

The Qualcomm QO410-1 PLO synthesizer used the 10 MHz input to create a 1.25 GHz output signal.

Output Power (at 'synthesizer out' port): 2.53 dBm

4.2.2.6.3 Doubler:

Type: PMC XK-A03 8837

Input Frequency: 1.25 GHz

Output Frequency: 2.5 GHz

Loss: 12.4 dB

4.2.2.6.4 Tunable Bandpass Filter

Type: K&L 5BT-1500/3000-2N

Bandwidth (3 dB): 25 MHz

Center Frequency: 2.5 GHz

Loss: 0.6 dB

4.2.2.6.5 Overall Specifications of IF LO:

Table 4-3: Power Out of IF LO Components (dBm)

Qualcomm Synthesizer	2.53
MCL-ZFL 1000 Amplifier	8.55
Frequency Doubler	-3.7
Bandpass Filter	-4.4
Anzac AMC-180 Amplifier	2.1
Anzac AMC-180 Amplifier	8.6

Please see Figure B-18 in Appendix B for a graph of the output power spectrum of the IF LO stage

4.2.2.7 RF LO

The RF local oscillator provides a signal centered at 26.625 GHz to the RF upconverter. It consists of three components:

- 1) The HP8673D Synthesizer
- 2) A microwave amplifier
- 3) A frequency doubler

A block diagram of the connections between the above components is shown in Figure 4-3.

Figure 4-3: Block Diagram of RF LO

4.2.2.7.1 HP 8673D Synthesizer and Microwave Amplifier

The synthesizer created a 13.3125 GHz signal and provided synchronization to both the IF LO and the receiver timing circuits. The HP8349B microwave amplifier amplified the signal to the required level for LO output. The synthesizer, amplifier and mm wave source module were integrated as one system, with the amplifier providing automatic level control to the synthesizer. Signals were carried between the synthesizer and amplifier via a Sucoflex 104 cable, with SMA connectors. A specialized cable (MFR 65512) with N-type connectors was used between the amplifier and the source module.

4.2.2.7.2 Frequency Doubler

The signal was doubled to the required 26.625 GHz in the mm-wave source module (HP83554). A WR-28 waveguide output to the upconverter was provided.

4.2.2.7.3 Overall Characteristics of RF LO Block

Output Power: 9 dBm

Output Frequency: 26.625 GHz

4.2.2.8 Overall Tx Specifications

Maximum Output Power (excluding antenna gain): 12.9 dBm

Output Bandwidth: 400 MHz

4.2.3 Receiver Design

The receiver can be broken into 3 stages:

- 1) The RF stage, including the horn antennas
- 2) The IF stage, which includes the IF amplifier chain
- 3) The baseband stage, including the crystal detector and data acquisition system.

The various stages and components of the receiver are outlined in Figure 4-4.

4.2.3.1 RF Stage

The RF stage of the receiver prepares the received signal for downconversion. It also performs switching between the two receive antennas, creating the equivalent of two input channels. It consists of the following components:

- 1) Receive antennas (horns)
- 2) High-speed PIN switch
- 3) Cascaded low-noise amplifiers
- 4) Associated cables, flexible waveguide and waveguide terminations

The connection of components is illustrated in Figure 4-4.

Figure 4-4: Block Diagram of the Receiver

4.2.3.1.1 Receive Antennas

The receive antennas were two NARDA horns - one for each channel. They were mounted closely above each other, as described in section 4.3.1.1. The waveguide output of the antenna for channel 2 was connected directly to the PIN switch input, while a flexible waveguide was used to connect the antenna for channel 1 to the switch. This was done to allow the pointing direction of the antenna on channel 2 to be adjusted independently.

Table 4-4: Antenna Characteristics for both Receive Channels

	3 dB Beamwidth	Ma. Gain	Horn Type
Channel 1	32°	15.3 dBi	NARDA 48127 S/N 1
Channel 2	32°	15.2 dBi	NARDA 48127 S/N 40

These results fit very well with the theoretical gain of 15.4 dBi for a flared horn with a physical aperture of 2.6 by 1.7 cm. (see Chapter 2 for equations). Please see Figures F-6 to F-9 in Appendix F for detailed radiation patterns of both antennas.

4.2.3.1.2 PIN Switch

The purpose of the broadband SPDT PIN switch (Hughes 47971H 2184) was to allow the multiplexing of the two receive antennas onto a single receive chain, creating the equivalent of two channels without the need for balancing two channels and duplicating components. The switch was controlled by an integral driver that utilized TTL inputs from the same timing circuit that controlled the sampling instants, ensuring synchronization. See section 4.2.3.4 for details on timing.

PIN Switch Characteristics

Table 4-5: PIN Switch Insertion Loss (29.0-29.5 GHz)

	Minimum	Maximum
Channel 1	1.23 dB at 29.50 GHz	1.38 dB at 29.15 GHz
Channel 2	1.09 dB at 29.48GHz	1.21 dB at 29.10 GHz

Table 4-6: PIN Switch Isolation (29.0-29.5 GHz)

	Minimum	Maximum
Channel 1	27.34 dB at 29.00 GHz	28.63 at 29.37 GHz
Channel 2	25.54 dB at 29.01 GHz	26.70 at 29.41 GHz

See Tables C-1 to C-4 in Appendix C for detailed levels at specific frequencies.

For test purposes, a signal was fed directly from the transmitter into one channel of the switch, and the output of the receiver was sampled as switching was performed at 1 KHz. No aberrations were found in the waveform.

4.2.3.1.3 Low-Noise Amplifiers

The purpose of the LNA's was to amplify the low-power received signal in preparation for downconversion by the lossy harmonic mixer, while introducing as little noise as possible. This required a cascade of two LNA's.

Table 4-7: LNA Characteristics (from Manufacturer Specifications)

	Gain (29 GHz)	1 dB Compression Point	Noise Figure
LNA #1 (ser #318062)	25.1 dB	7.4 dB	3.01 dB
LNA#2 (ser # 318061)	24.4 dB	8.8 dB	3.43 dB

Accurate tests of the LNA gain were done at 26 GHz and compared to manufacturer specifications for those levels. Differences were found to be less than 0.5

dB for LNA#1 (23.5 dB, vs 24.0 dB specified) and less than 0.2 dB for LNA#2 (24.3 dB, vs 24.5 dB specified). Using the broadband signal from the transmitter, somewhat less accurate tests (+/- 0.7 dB) at 29.125 GHz showed an average uncompressed gain of 24.4 dB for LNA#1 and 24.6 dB for LNA#2. Since both lab tests showed somewhat lower gains for LNA#1 than the manufacturer specified, a more conservative estimate of 24.5 dB was assumed for receiver sensitivity calculations. Individual compression points were not tested in the lab, since a comprehensive gain sweep of the RF stage was completed. Please see Figures C-1 to C-3 in Appendix C for a gain sweep of both LNA's, and a gain sweep of the RF stage.

4.2.3.1.4 Cables, Waveguides and Waveguide Terminations

A flexible waveguide was used to connect the antenna on channel 2 to the PIN switch input, while the antenna on channel 1 was connected directly, via its waveguide output. A waveguide termination with an approximate loss of 0.3 dB connected the switch output to LNA #1. A Suhner Sucoflex cable was used to connect the output from LNA#1 to the input of LNA#2, and an HP R281A waveguide termination with an approximate loss of 0.1 dB connected the second LNA to the harmonic mixer.

Loss of Flexible Waveguide: approx. 0.3 dB

Loss of Cable: 1 dB

4.2.3.1.5 Overall RF Stage Characteristics

Uncompressed Gain at 29 GHz: 47.3 dB (+/- 0.7 dB, depending on input power)

Noise Figure: 4.58 (Channel 1) 4.82 (Channel 2)

Output at 1 dB Compression Point: 4.9 dBm

4.2.3.2 IF Stage

The main purpose of the IF stage was signal amplification. It consisted of 7 components: A harmonic mixer, two bandpass filters, a variable attenuator, 3 amplifiers, and associated cables.

4.2.3.2.1 Harmonic Mixer

Downconversion was accomplished by using the 8th harmonic from a harmonic mixer. To get the required output frequency of 1.125 GHz, the mixer was driven by a 3.5 GHz LO. The LO was supplied by driving an HP 11975A amplifier with the signal generated from the LO output of an external mixer interface module (HP 70907B) attached to an HP 70004A Spectrum analyzer (see Figure). This mixer module was designed specifically for the purpose of driving the harmonic mixer through the HP amplifier.

Since signal powers at the mixer output were too low to be measured directly for most cases, measurements were taken at the output of the first IF amplifier (ZHL 1042J), whose characteristics had been previously tested, and translated back.

Mixer Characteristics

Uncompressed Mixer Loss: 25.6 dB (Manufacturer Specification: 26 dB max)

1 dB Compression Point: 2.1 dBm at input

LO Input Power: 16 dBm

See Figure C-4 in Appendix C for a gain sweep of the mixer.

4.2.3.2.2 Bandpass filters

The first K&L filter was connected directly after the mixer, to remove unwanted out-of-band components picked up by the antennas or generated by the mixer. The second filter was connected at the end of the IF stage, after the last amplifier, to produce sharper rolloff and remove any spurious emissions generated by active components or from possible RF interference.

The bandwidth of the filters was chosen so that only the centre section of the mainlobe of the 400 MHz incoming signal would pass through (2.5 GHz +/- 125 MHz). This ensured that the frequency spectrum across the entire bandwidth seen by the amplifiers and crystal detector was relatively flat.

Table 4-8: Characteristics of Bandpass Filters in Receiver

	Loss	3 dB Bandwidth
Filter # 1 (Type 4B121-1125/T250-0/0 S/N 2)	0.37 dB	260 MHz
Filter #2 (Type 4B121-1125/T250-0/0 S/N 1)	0.38 dB	268 MHz

See Figures C-5 and C-6 in Appendix C for frequency response measurements of the filters.

4.2.3.2.3 Amplifier #1 and Variable Attenuator

The first amplifier was an HP 1042J, whose function was to raise the IF signal level to allow adjustment of the output power level with the variable attenuator, both for the final setup and for intermediate measurements. The variable attenuator was needed to fine-tune the IF power output such that all values would be in the range of the A/D converter.

Amplifier Characteristics:

1 dB Compression Point: 20 dBm

Uncompressed Gain: 25 dB

Noise Figure: 4.5 dB

4.2.3.2.4 Amplifiers #2 and #3

These amplifiers were cascaded and used to amplify the signal to the proper output level. The first was an Mini-Circuits MCL ZFL-2000 medium power amp, and the second was an HP 8347A high-power amplifier with a power level readout that was used as a basis for adjusting the transmitted power level during measurements. Note that the medium power amplifier was not tested to its compression point because its maximum output signal in this setup was in the -12 dB range, and tests up to the compression point specified by the manufacturer showed no evidence of a non-linearity.

Table 4-9: Characteristics For Receive Amplifiers #2 and #3

	1 dB Compression Point	Uncompressed Gain	Noise Figure
Amplifier # 1 (MCL ZFL-2000)	16 dBm (specs)	20.47 dB	7.0 dB
Amplifier #2 (HP 8347A)	20 dBm	29.22 dB	15.0 dB

See Figures C-7 and C-8 in Appendix C for gain sweep and frequency response measurements of the HP 8347A amplifier, and Figure C-9 in Appendix C for a gain sweep of the ZFL-2000 amplifier.

4.2.3.2.5 Overall IF Stage Characteristics

IF stage gain (uncompressed, including attenuator): 14.2 dB

1 dB Compression Point: 13.63 dBm

Please see Figure C-10 in Appendix C for a gain sweep of the IF stage.

Maximum output signal used during measurements: 17.2 dBm

4.2.3.3 Envelope Detection

The baseband stage of the receiver was designed to sample the envelope of the signal and save the results in digital form. A crystal detector was used to detect the envelope, and after amplification, a data acquisition card by Data Translation (DT2828) in a 386 PC was used to sample and convert it into digital form.

4.2.3.3.1 Crystal Detector

The crystal detector (HP423B) converts RF power values into proportional (square-law) values of DC voltage. The use of a matched load resistor allows the detector to be accurate to within 0.5 dB across a 30 dB range (manufacturer's specifications). The detector has an 8 GHz bandwidth.

Please see Figure C-11 in Appendix C for a graph of output voltages vs input power for the crystal detector.

4.2.3.3.2 Baseband Amplifier

The baseband amplifier converts the input voltage (0-1V) into a voltage that utilizes the full scale of the A/D converter (0-5V).

4.2.3.3.3 A/D Converter

The DT2828 data acquisition card and PC combination can sample at speeds up to 50 Kbps. The card has a 12-bit A/D converter resulting in 4096 discrete levels for a -5V to +5V input, but only the positive range was used. Also, the last bit of the 11-bit conversion is erroneous, resulting in a practical scale of 1024 levels for the 0-5V input.

4.2.3.4 Timing

4.2.3.4.1 Transmit/Receive LO Synchronization

To synchronize the transmitter and receiver setup, the 10 MHz frequency reference on the HP 8673D synthesizer from the transmit RF LO was used as a master, with the HP 8567A signal generator on the PCEG clock and the HP 7000 spectrum analyzer unit on the receiver IF LO driven by it.

4.2.3.4.2 Sampling Control

An HP 3312A function generator was used to provide the master clock for control of the sampling rate. Since the two receive channels were sampled in sequence, the function generator was used to drive a frequency divider circuit based on the 74163N chip, that synchronized the switching of channels at the PIN switch with the sampling instants of the A/D converter. Control over the timing of measurements was provided at the timing chip, since the chip output could be interrupted manually. The output of only one of the two channels could also be examined from here by activating a switch that caused the chip output to the PIN switch to be tied high or low during sampling. When the chip was deactivated, both outputs were automatically tied low. This ensured that sampling would start on the same channel at the start of each measurement.

Figure 4-5: Synchronization of PIN Switch and Data Acquisition System

4.2.3.5 Receiver Specifications:

4.2.3.5.1 General:

Gain of IF and RF stage (uncompressed): 61.6 dB (measured)

(theoretical: 60.2 dB by adding component gains)

1 dB Compression Point: 13.6 dBm

Bandwidth: 250 MHz

Please see Figure C-12 in Appendix C for a gain sweep of the receiver.

4.2.3.5.2 Noise Figure and SNR

Table 4-10: Receiver Noise Figure and SNR

	Channel 1	Channel 2
Receiver Noise Figure	5.22 dB	5.46 dB
Receiver SNR (17.2 dBm max output)	41.8 dB theoretical 37-38 dB measured	41.7 dB theoretical 37-38 dB measured

The theoretical receiver noise figure is calculated as described in section 2.9.1. Please see Appendix D for calculations.

The above noise figure was verified by examining the receiver output signal power on a spectrum analyzer when no input signal was present. The noise floor was found to be at approximately -104 dBmW/Hz, resulting in a maximum SNR of 37.2 dB, however during these measurements the antennas were connected at the front-end and may have picked up small amounts of spurious noise from the transmitter RF LO which was active nearby. The room temperature was also somewhat higher than 290°K, and amplifier temperatures were elevated well above that level (LNA temperatures were in the 50°C - 60°C range). Manufacturer specifications for device noise figures may also have been optimistic.

4.3 Measurement Procedure

4.3.1 Transmitter/Receiver Setup

4.3.1.1 Physical Setup

For measurement purposes, the transmitter was set up on a 83 cm high wooden cart, with the transmit antenna elevated to a height of 185 cm. The antenna was raised to simulate the placement of a base station antenna. The tripod on which the transmit

antenna was secured had a mounting plate on top that allowed the antenna to be rotated 20° vertically and 90° horizontally, for alignment with the receive antennas.

The receiver was set up on a separate cart, with the receiver antennas mounted above each other on a wooden platform. The top antenna, for channel 2, was elevated 123 cm (centre to centre) from the floor, and the bottom one, for channel 1, was placed 5 cm (antenna centre) lower. The size of the PIN switch dictated the spacing between the two antennas. Each antenna could move independently across a 100° arc in the horizontal plane and graduations above and below the antennas allowed each to be accurately aligned, in 5° increments. The entire setup could be steered in a 300° arc across the horizontal plane and 60° in the vertical plane. When the two antennas were pointed 30° apart, the straight-line horizontal distance between the two centres was 2.5 cm, resulting in a total separation between their centres of 5.6 cm, or 5.4 wavelengths at 29.125 GHz. Please see Appendix A for a photograph of the receive antennas.

The PC with the A/D board in it, and the spectrum analyzer used for the receiver LO were put on a third cart that was placed several metres behind the receiver cart during measurements. Please see Appendix A for photographs of the receiver and transmitter.

4.3.1.2 Antenna Setup

During measurements, 3 antenna alignments were studied at each location:

- 1) Channel 2 antenna aligned with the transmit antenna, channel 1 antenna pointed 30° to the left ('misaligned').
- 2) Same as 1) except that the antenna on channel 1 was pointed to the right.
- 3) Channel 1 antenna pointed to the left by 15° and channel 2 antenna pointed to the right by 15°.

For analysis, the first two categories were grouped together since the objective was to examine the differences between aligned and misaligned cases and in a typical indoor environment, there would be no pattern of differences between measurements with right or left misalignment.

Figure 4-6: Illustration of Horn Configurations During Measurements

When a horn antenna was on the transmitter, it was always directed at the receiver, and elevation differences between the transmit and receive antennas were corrected by aligning the antennas in the vertical plane.

4.3.2 Calibration

4.3.2.1 Tuning Receiver Output Power:

The maximum receiver input power level was limited by the compression point of the harmonic mixer, and the step attenuator in the receiver was used to fine-tune the receiver output until a level of 17.2 dBm was achieved at the end of the IF stage when the RF front-end was fed the maximum input level and the mixer was slightly compressed. At 17.2 dBm, the output from the crystal detector was just less than 1 volt, and the data acquisition system registered this voltage as the highest possible level (2048). This fine-tuning therefore allowed the data acquisition system to swing across most of its range, from approximately 9 (no signal and pure noise: 8) to 2048 (corresponding to power levels of -32.8 to +17.2 dBm at the detector). This maximum receiver input level was set as the reference level. All data that were subsequently sampled and calibrated had this level set as the 0 dB mark. During measurements, the transmitter signal power was adjusted so that the average signal power on the strongest received channel was approximately 5 dB below this level, ensuring a wide margin for fades yet also allowing an adequate margin for enhancements. This also kept power levels below the receiver compression point (13.6 dBm) for the majority of the time.

4.3.2.2 Data Acquisition System Calibration:

To achieve accurate results across the entire dynamic range of the receiver, the system was precisely calibrated before measurements were started. The transmitter output at the antenna was connected to each channel of the receiver (at the antenna input) via an extra cable, a programmable step attenuator (PRAT) and a waveguide termination. Separate calibrations were done for each receive channel. The PRAT was used to attenuate the transmitter output to the level where the maximum input to the receiver would equal the maximum input seen under measurement conditions (see above). The variable attenuator was used to decrease power in precise increments during calibration. These increments were as follows:

Table 4-11: Power Increments Used During Calibration

Attenuation Level	0-10 dB	10-20 dB	20-30 dB	30-45 dB	45-50 dB
Increments Used	0.1 dB	0.2 dB	0.5 dB	1 dB	5 dB

As attenuation was increased, 1000 samples were taken by the A/D board at each level, at 100 samples/second. A computer program was used to take an average of the samples at each level, and a calibration table was formed, listing received power levels against A/D outputs. Please see Table D-1 and Figures D-1 and D-2 in Appendix D for detailed listings and graphs of the calibration curves for both channels. All measured data was processed with these calibration curves.

The last bit of the A/D conversion was not reliable, however at power levels that were no more than 33 dB below the reference level, moving from one calibration step to an adjacent one involved the change of at least one bit. This resulted in an error margin of 1 dB (the coarsest calibration step) at low power levels, with considerably smaller error margins at higher power levels. For signals less than 33 dB below the reference level, changes in the average calibration value between adjacent 1 dB steps were not as pronounced. A 1 bit change in the calibration value covered 2 calibration steps, so errors in that region were as high as 2 dB.

Note that no interpolation was done when the measured data was matched with the calibration curve, to preserve accuracy. This resulted in discrete steps in final data logs.

To ensure stable results that would be accurate for later measurements, the transmitter and receiver equipment was powered up for one hour prior to the start of calibration, allowing amplifiers to warm up. The same warmup time was used before regular measurements as well, since tests showed that power levels were within 0.2 dB of their long-term value at that point.

4.3.3 Parameters of the Experiments

To achieve the goals of the thesis, a wide variety of measurements was done at a number of locations throughout the engineering areas of Carleton University. Most were done

during the day, when the movement of people in rooms and corridors allowed a large amount of shadow fading to occur. Two main types of movement were studied at each location:

- 1) Both receiver and transmitter stationary, with people walking in the room.
- 2) Transmitter stationary, and no movement of objects in the area, but receiver moving

The first case simulated a typical environment for an indoor setup, and most of the analysis was done on these measurements. These measurements were done for 5 to 20 minutes at a time, to capture a representative number of fades. The length of time depended on the amount of activity in the room. If not enough normal movement occurred, people doing the measurements walked through the room in as natural a fashion as possible. This did not distort results since the frequency of movement and traffic patterns were not factors to be analyzed.

In the second case, the receiver was moved in a 34 cm sphere at approximately 1m/sec, to simulate a setup where the receiver would be portable. These measurements were used to determine the average power level at the receiver location, as well as to examine fade depths due to pure multipath effects. Fades due to multipath were frequent, and at a speed of 1m/sec, and a wavelength of 1.03 cm, the receiver covered approximately 100 wavelengths per second. A 30 second measurement time was therefore long enough.

Two types of antennas were used on the transmitter—a biconical antenna with an omnidirectional radiation pattern, and a horn antenna with a 20° beamwidth in the H (azimuthal) plane. The latter was used in the majority of cases.

Distances between the transmitter and receiver were varied between 3 and 20 metres, to simulate a variety of possible scenarios. In all LOS cases, both the transmitter and receiver were placed so that walls would not obstruct any receive antenna or severely restrict the transmitted radiation pattern. This meant that all antennas were placed at least 1.5 metres from the nearest wall behind or beside them. Four possible environments were

covered:

- 1) Computer labs and general-purpose rooms
- 2) Hallways
- 3) Large open pedestrian areas (Minto Centre lobby)
- 4) Factory floor environment (Civil Engineering lab)

The three computer and equipment labs in which measurements were made all had large wooden tables in them, as well as a variety of equipment with metallic covers. This included monitors as well as portable test equipment. The walls were mostly made of concrete blocks and glass windows. All three rooms were relatively large, being at least 10 by 5 metres. In two of the rooms, large numbers of people were present during tests. In the undergrad lab, this resulted in a significant amount of motion, however in the graduate lab, the 10 to 15 people remained mostly seated and did not add many fades to the experiment.

The two hallways that were used were both more than 20 metres long and had concrete walls, with metal doors leading to various rooms. Movement in both corridors was sparse.

The open area that was tested was the Minto Centre lobby, an irregularly shaped area with benches, overhead metallic pipes and large numbers of people. The people moved both singly and in groups, and tended to move faster than people in the labs. The walls were made of concrete and all doors were metal. An atrium at the center of the area rose to a height of 3 stories.

The Civil Engineering lab in the Minto Centre was chosen to represent a factory environment. It was a large, 2 story high area filled with a variety of metallic equipment. The walls were made of concrete. Movement was infrequent, and machinery blocked much of the space. Large metal beams were present, both overhead and rising vertically.

The six standard measurements (3 antenna alignments and static/moving receiver for each) were done for at least one location in each environment. Non-LOS tests were done around corners in hallways in the Engineering building and through walls between a

hallway and a lab, as well as through an obstructing section of wall in the Minto Centre lobby. For all non-LOS tests, the receiver was kept static.

A total of 101 measurements of various kinds were included in the final data analysis, with approximately 500 Megabytes of raw data processed for this purpose.

Please see Appendix G for floor plans of the various locations tested.

5. DATA ANALYSIS

5.1 Introduction

The purpose of this section is to examine the characteristics of the received signal power during the times that the radio link is affected by various types of movement in several different environments, and to analyze the gains in SNR that could be achieved by using multiple receiver antennas in an angle diversity setup. The bulk of the data analysis is therefore focused on three objectives:

- 1) To examine the CDF for each type of measurement and calculate required fade margins on both channels at various probability levels.
- 2) To examine correlations between the two receive channels for two cases: multipath-only and combined multipath and shadow fading.
- 3) To calculate a new CDF for each measurement based on selection combining of the two receive channels, and analyze possible improvements to the required fade margin at various probability levels.

Altogether, four different types of locations and six different antenna setups were used to gather information on the above three points for two different scenarios:

- 1) A multiple-horn setup on the receiver, with only one horn transmitting.
Most of the setup configurations and measurements were done for this scenario.
- 2) An alternative design with an omni-directional antenna on the transmitter.

The following section describes the parameters that were used to set up the data collection program as well as the parameters needed to calculate results. The three sections after that each examine one of the above objectives in detail.

The programs needed to analyze the data were written in a language called 'MATLAB' and run on a 486 50 MHz PC. Since only 8 Megabytes of memory were

available, the programs were set up to handle a maximum of 500,000 points per data file. For more information on the programs and their capabilities, please see Appendix H.

5.2 Collection Parameters and Typical Results

5.2.1 Collection Parameters

To obtain good results, several important data collection parameters had to be set. Finding the proper data acquisition speed was a primary goal, since too slow a speed would mean that aliasing of high frequency components in the received signal envelope would occur and deep fades might be missed. The data acquisition system could handle a maximum of 25k samples/sec per channel, providing an upper limit of 12.5 kHz on the frequency spectrum of the signal envelope. To test the maximum speed necessary, a test was run in the radio lab where each channel was sampled at 5k samples/sec while people were moving around. Since the maximum rate of movement was approximately 1 m/sec, this sample rate could safely be assumed to be an upper bound on the required rate (see section 2.1.1.2). For this test, channel 1 was misaligned by 30° away from the transmitter direction to ensure that fading characteristics on misaligned antennas would also be fully recorded. The resulting data was then decimated so that CDF's were obtained for lower sample speeds. The deviation of CDF results at a 1% probability level is shown in Table 5-1 below:

Table 5-1: Deviation of CDF's Using Decimated Data at the 1% Level

Sampling Speed	5000/sec	2500/sec	1667/sec	1250/sec	625/sec
Channel 1	0 dB	-0.08 dB	-0.03 dB	-0.12 dB	-0.12 dB
Channel 2	0 dB	-0.06 dB	0.53 dB	0.01 dB	0.02 dB

Note: A positive deviation means that the power level at the 1% probability level was higher in the subset of decimated data than levels in the full data set (5000 samples/sec).

Even though deviations were less than 0.5 dB in all cases, a minimum speed of 1000 samples/sec/channel was chosen to ensure data integrity. According to the Nyquist theorem, this limits the frequency spectrum of the signal envelope to 500 Hz. Since movement was not expected to exceed 1m/sec, and the wavelength was approximately 1 cm, a maximum of 100 fades/sec could be expected to occur, making a cutoff of 500 Hz a reasonable figure for capturing almost all fades (see section 2.1.1.2).

To determine the length of time required to get an accurate measurement at one location that would be representative of long-term results, several measurements, each 10 minutes in length, were done with 1 person continuously walking around. The CDF of the total measurement was compared to CDF's of 2 minute subsets. It was considered that intervals shorter than 2 minutes might not allow the recording of a representative number of fades. Several such tests were run using different horn angles, and at 1% probability, the 2 minute subsets differed from the CDF of the full 10 minute set as shown in Table 5-2 below:

Table 5-2: Deviation of CDF Results For Short Measurements at the 1% Level

2 minute subset #	1	2	3	4	5
Aligned Horn	-0.18 dB	0.08 dB	-0.50 dB	-0.04 dB	0.61 dB
15° Misaligned	0.01 dB	0.40 dB	-0.20 dB	0.27 dB	-0.47 dB
30° Misaligned	-1.06 dB	2.56 dB	-1.2 dB	0.62 dB	-1.51 dB

Note: A positive deviation means that the power level at the 1% probability level was higher in the 2 minute subset of data than levels in the full data set.

The 30° misaligned horns had a lower average power level than the other measurements, and most of the deviations of these subsets are due to the higher measurement errors for this set of data (1 dB). Also, during testing, two measurements were made at each location for 30° misalignments (horn moved to the right, and then to the left). When combined for analysis, this would result in 4 minute sets, which reduced

the maximum deviation from long-term (10 minutes and longer) results to less than 1.2 dB. A minimum time of 2 minutes was therefore used for individual measurements, although many were in the 10 to 15 minute range to capture more fades if movement of people - and consequently fading - was infrequent.

At the receiver, one horn was positioned above, and parallel to the other, so it was important to test whether both receive antennas would receive the same average power. Differences in attenuation due to the inclusion of a flexible waveguide on channel 1 had already been taken care of through the calibration procedure, but horn input power might have been different since one horn would always be slightly misaligned with the transmit horn. A 30 sec test was done in which the receiving apparatus was moved in a 34 cm sphere to average over local fading, while data was collected at 1 ksample/sec. Both receive horns were pointed at the transmitter. The mean power of each channel was calculated, and the difference between the two was 0.09 dB - far less than the margin of error due to variables such as calibration inaccuracies (± 0.5 dB per measurement when fades lowered the received power level into the - 20 dB range) and equipment adjustments (e.g. the degree of bend in the flexible waveguide).

Similarly, it was important to determine whether it was necessary to change the transmit horn elevation angle when the receiver was moved closer to the transmitter. A test was run in the radio lab, where the transmit horn elevation angle was first aligned to point directly at the receive antennas (which were lower), and then pointed at a location several meters behind the receiver. In both cases, the receiver horns were pointed at the transmit horn from a distance of 5.9 m. Mean power dropped by 1.2 dB when the transmit horn was not aligned. To ensure that measurements at different distances would not be affected by the variable power loss associated with this type of misalignment, the transmit and receive antennas were always aligned in elevation.

Several parameters were also needed for data analysis, most notably for correlation calculations (see section 2.8) To find the behaviour of the two channels with unequal mean signal power, as is the case when one receiver horn is pointed directly at the transmitter and the other is pointed 30° away, it was necessary to normalize the data

by finding the local mean for each channel. Typically, averaging over 100 fades will give a reasonable local mean [personal communication with M. El-Tanany, Sept. '95] so with objects moving at 1m/sec and a wavelength of 1 cm, averaging over a 1 sec time interval is adequate. To determine the time interval (and number of data points) between adjacent correlation calculations, four sets of correlations were found for 9 measurements with people moving around in various locations, and 15 measurements with the receiver being moved in a local area. Each set corresponded to a different time interval: 20 msec, 10 msec, 5 msec and 2 msec. An average correlation coefficient was found for each of the four time intervals in each of the 24 files. The maximum deviations from the 2 msec values are shown in Table 5-3 below.

Table 5-3: Maximum Deviation Of Correlations As Window Slide Time Increases

Window Slide Time	5 msec	10 msec	20 msec
Maximum Deviation	4.0×10^{-5}	1.3×10^{-4}	3.0×10^{-4}

These are all insignificant amounts. Since all intervals showed the same result, 5 msec was chosen for computing all other correlations.

5.2.2 Typical Results

In most cases, having horns on the receiver and transmitter resulted in deep fading on the order of 15-25 dB whenever a person passed directly between the transmitter and receiver. This fading was more pronounced when the receive horn was aligned with the transmit horn. Typically, a receive horn that was misaligned by 30° would result in fades that were 2-7 dB shallower than an aligned horn when an object was moving directly in front of the receiver.

During the time a person passed close to or just inside the beam of the transmit horn (incomplete blockage), an aligned receive horn would result in fades with a depth of about 2-5 dB, while a receive horn that was misaligned by 30° would result in 4-7 dB fades. In both cases, the fades were sinusoidal in shape, with a frequency of about

30-50 Hz, and fade depths increased in an exponential fashion as the person approached the main beam of the transmit antenna. The fades are caused by the addition of multipath components, so they will have more of an effect on the misaligned horn, where the average received signal power is weaker. This explains the disparity of fade depths between aligned and misaligned horns during partial blockage of the antenna beam.

Having an omni on the transmitter reduced deep fades slightly, but extended the time over which small but regular fades were encountered. Deep fades occurred at almost exactly the same time on both channels, whatever the receiver setup, suggesting that much of the power seen by a misaligned horn came directly from the transmit horn, and not from an independent reflection.

When objects were moving only outside of the radiation pattern of the transmit horn, no fades at all were encountered. For the purpose of calculating CDF's and correlations, these sections were removed from the measurements. More information on the procedure for removing this data can be found in Appendix H.

To find the difference in mean signal powers between the two channels, measurements were done with the receiver moving in a small area. The average power of this measurement then gave the local mean for each channel. This type of measurement was done at almost all locations where horns were used on both the transmitter and receiver. Analysis showed that horns that were misaligned by 30° received 8.2 dB less power, on average, than aligned horns. When both horns were misaligned by 15° , channel 2 received only 0.49 dB more power, on average. Similar powers were expected since both receive horns should see approximately the same signal.

A typical measurement of signal power for both channels is displayed in Figure 5-1 and Figure 5-2. Channel 1 had its horn misaligned by 30° , and the horn for channel 2 was aligned. This 3.5 minute portion of a 10 minute measurement was taken in the Minto Centre lobby at a transmit/receive range of 19.5 m, with groups of people moving normally through the area. Sampling speed was 1 ksample/sec/channel. Figure 5-1 and Figure 5-2 show the received power with quiescent sections removed. Figure 5-3 shows the received power for channel 1, with quiescent sections included.

When one considers sample rates, the figures indicate that deep fades tend to be approximately 1 second long, as can be seen in Figure 5-1 and Figure 5-2. At 1 ksample/sec, Figure 5-1 and Figure 5-2 shows approximately 60 seconds of data (all instances of fades) out of the 3.3 minute measurement.

The fourth graph shows a typical measurement where receive and transmit antennas were aligned, and the receiver was moved. There was no other motion in the environment. This measurement was taken in the same area. Only a few seconds of data are displayed, to show the fades more clearly. When this data is compared to multipath fading due to the movement of people, several differences can be observed. When the receiver is moved, fades tend to be quite a bit deeper, and much more random. No specific fading cycles could be observed. A more in-depth evaluation of the fading characteristics for these measurements is given in the following section.

Figure 5-1: Received Power of Channel 1 (30° Misaligned), Quiescent Sections Removed: Fades Caused by Normal Movement of People

Figure 5-2: Received Power of Channel 2 (Aligned), Quiescent Sections Removed: Fades Caused by Normal Movement of People

Figure 5-3: Received Power of Channel 1 (30° Misaligned) - Fades Caused by Normal Movement of People

Figure 5-4: Received Power of Aligned Channel - Fades Caused by Receiver Moving at 1 m/sec

5.3 Fade Margins

To calculate the outage probability of a wireless system, it is important to know the probability that a signal will fade below a certain level. To calculate the required fade margin at a certain probability, a CDF of all data points in a measurement is set up. The fade margin can then be included in a system link budget to calculate the total amount of transmit power required. For example, in the case of a Rayleigh fading envelope, the average transmit power must be 18 dB above the minimum level required by the receiver to achieve an outage probability of 1% (if only the effects of Rayleigh fading are considered) since fades worse than 18 dB occur only 1% of the time.

CDF's for all measurements where people were walking around were calculated by including only the sections of a measurement where shadow or multipath fading occurred, and cutting the quiescent sections. Measurements where the receiver apparatus was moved were kept in full, since movement was constant and fade occurrences were continuous. Before the CDF was calculated, both channels were normalized to facilitate comparison of fade depths, and all results were expressed in dB relative to the mean power. To calculate the PDF for a measurement, a histogram was created with 471 bins. Bin values ranged from -40 dB to +7 dB relative to the mean, in 0.1 dB steps, and all data points in a measurement were added to their respective bins. A cumulative sum over all bins then gave the CDF. The cumulative sum was expressed as a percentage so that comparison between measurements of different length could be made. Fade margins for all groups of measurements were found at 10%, 1%, 0.5% and 0.1% probability levels, as shown in Table 5-4 below. Note that these fade margins were caused by a combination of multipath and shadow fading.

Table 5-4: Average Required Fade Margins (in dB) Computed From All Types Of Measurements

Measurement Type	Antenna Alignment	10.0%	1.0%	0.5%	0.1%
People Moved, Horn On Tx	Aligned	6.71	14.59	15.78	17.82
	15° Off	5.93	13.79	15.27	17.93
	30° Off	6.10	12.03	13.20	14.90
People Moved, Omni On Tx	Aligned	2.09	7.83	9.63	13.39
	15° Off	2.18	7.39	9.71	12.78
	30° Off	2.11	5.18	5.94	6.97
Receiver Moved, Horn On Tx	Aligned	2.06	3.73	4.04	4.48
	15° Off	3.43	5.95	6.34	6.95
	30° Off	5.43	8.56	9.13	9.85
People Moved, Omni On Tx, Non Line-Of-Sight	Aligned	3.72	9.22	10.87	13.23
	15° Off	3.47	7.39	8.16	9.87

CDF's for each measurement group were superimposed on one graph and the graph for each group is shown below. Non-LOS CDF's are not included since only a few Non-LOS measurements were done.

Figure 5-5: Superimposed CDF's For: Horn on Tx, People Moving, Rx Horn Aligned

Figure 5-6: Superimposed CDF's For: Horn on Tx, People Moving Rx Horn 15° Misaligned

Figure 5-7: Superimposed CDF's For: Horn on Tx, People Moving, Rx Horn 30° Misaligned

Figure 5-8: Superimposed CDF's For: Omni on Tx, People Moving, Rx Horn Aligned

2

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

PRECISIONSM RESOLUTION TARGETS

Figure 5-9: Superimposed CDF's For: Omni on Tx, People Moving, Rx Horn 15° Misaligned

Figure 5-10: Superimposed CDF's For: Omni on Tx, People Moving, Rx Horn 30° Misaligned

Figure 5-11: Superimposed CDF's For: Horn on Tx, Rx Moving, Rx Horn Aligned

Figure 5-12: Superimposed CDF's For: Horn on Tx, Rx Moving, Rx Horn 15° Misaligned

Figure 5-13: Superimposed CDF's For: Horn on Tx, Rx Moving, Rx Horn 30° Misaligned

As can be seen from the graphs, the range of CDF's was fairly wide for each group. This is likely due to the variety of locations that were used for the measurements. Also, fading tended to be deeper if people blocked the line of sight path as they walked close to the transmitter or receiver than it was if they blocked it while walking through the middle of the room. The most likely explanation is that a person walking close to the transmitter would block a larger percentage of the beam, allowing less overall signal power to propagate, while a person walking close to the receiver would block not only the direct line-of-sight but many of the reflected signals that arrived from another angle. The depth of fades therefore depends to some degree on how close people could get to one of the antennas as they moved around the room. A shorter distance between transmitter and receiver also tended to increase fade depths, since fewer reflected signal components were likely to be created in the narrower beam.

Typically, horn-horn measurements where people were moving and occasionally obstructing the LOS showed a 15 dB range at the 0.1% probability level. As the receive horn was misaligned, deep fades became less severe at a rate of approximately 3 dB per 15° . This result is likely due to the increasing influence of independently fading reflections during periods when the direct line of sight component is attenuated. Since the mean signal power drops at almost the same rate, actual signal powers during a deep fade are at approximately the same level, whether the receive horn is aligned or not. Having an omnidirectional antenna on the transmitter resulted in CDF's with only an 8 dB range at the 0.1% probability level. The reduction in CDF variations may be due to the fact that the radiation pattern of the omni antenna is less likely to be affected by the positions of people in the room, and the number of reflections from the antenna no longer depend on the range of the measurement. Deep fades exhibited the same tendency to become less severe as the receive antenna was misaligned, but on average, the deep fade depth was approximately 5 dB less for all cases, when compared to horn-horn measurements. This is likely due to the increased number of reflections, which enter the receiver from a wider angle and from various areas of the room and are therefore less likely to fade simultaneously with the direct line of sight ray. Even in this case, though, fading occurred

simultaneously on both channels, suggesting that the spillover from the direct line of sight path into the misaligned receive horn is still the dominant component for that channel.

In all of the above cases, a definite 'knee' can be seen on the graphs. This seems to represent a change in the way a fade is created: The upper portion of the 'knee' contains the majority of data points in the measurement and was created by frequent, shallow fades due to multipath as people partially obstructed and disturbed the radiation pattern of the antennas by walking near the edges of the main beam. The lower portion consists of the rare but deep fades created by shadow fading as people walked into the main beam of the antenna and created a complete obstruction. This explanation is reasonable because there is a distinct difference in fade depth and likelihood of occurrence between shadow and multipath fading, as can be seen in Figure 5-1 and Figure 5-2. When an omni-directional antenna was used on the transmitter, the knee was shifted farther down than when a horn was used. This reflects the fact that multipath-induced, shallow fades were more numerous. Note that no present theoretical formula can accurately model any of the above results in full. Although shadow fading is often modeled as having a log-normal distribution, this approach assumes that attenuation is caused by the addition of fades from a multitude of obstructions, with a uniform distribution of attenuation levels. On an indoor system with static antennas, shadow fading is generally caused by the movement of only a very few objects at a time.

When the receiver was moved and a horn was used on the transmitter, fades tended to increase in depth as the horn was misaligned. Fades were generally shallow, with a 6 to 8 dB spread at the 0.1% probability level and an average depth that was 6 to 13 dB less than when people were moving. Fades were not nearly as deep as Rayleigh fading models would suggest. This may be due to the significant line-of-sight component in received signals - which will theoretically cause the fading to become Rician [4] - as well as the broadband nature of the signal, which creates frequency-selective fading instead of fading across the entire band. Since no shadow fading was included in these measurements, no 'knee' is present in the CDFs. The increase in fade depth for a misaligned horn is due to the destructive addition of multipath components, which might

be equal to the direct line component power on the misaligned channel but would be far less powerful than the direct ray if the receive horn was aligned. As the receive horn is misaligned, therefore, fading tends to become more Rayleigh than Rician. This effect is not as evident when shadow fading due to obstructions causes the dominant deep fades in a measurement, since this type of fade is due to simple attenuation of the signal, and not the destructive addition of out-of-phase signal components.

In all cases, multipath-induced fades tended to be less severe when there was a person moving through the area, than when the receiver was moving. This result was to be expected since in the former case, multipath fades could only be observed when a person was moving at the edge of the antenna beam, where signal components that might be reflected into the receiver were weak and would have only a small effect on the dominant line-of-sight component.

5.4 Cross-Correlations Between Channels

Knowledge of the degree of correlation between two channels in a diversity system is an indicator of the amount of gain that can be expected from combining the envelopes of the channels. A cross-correlation of 1 would mean that fading on both channels occurs at the same instant and at the same rate, so diversity would bring no gains against fading. For independent fading on two diversity channels, Channel A and Channel B, a best-case combiner would create a new signal envelope whose fade margin at any probability level would be governed by the equation:

$$\Pr(\text{fade} < X \text{ dB for combined channels}) = \Pr(\text{fade} < X \text{ dB in Ch A}) * \Pr(\text{fade} < X \text{ dB in Ch B})$$

(22)

For example, a fade margin of 20 dB at the 1% probability level on either channel would translate to a fade margin of 20 dB at the 0.01% probability level with best-case combining. This equation only holds if both channels have equal mean signal power.

Cross correlations were found for 5 msec sections of a data file at a time and plotted in a histogram. The average cross-correlation for each file was found, and an overall average for each group of measurements was found. Computations were done as outlined in Chapter 2. The cross-correlation was defined to be 0 if the envelope on one channel showed no movement while the envelope on the other was changing. Similarly, it was defined to be 1 if neither envelope showed change during the 5 msec. However, all sections where both channels showed no movement were removed previous to the computations, so the latter definition was never required. For information on how the equations were implemented in code, please see Appendix H.

Two types of cross-correlations were calculated for measurements where people were moving around. The first type included all fades, but excluded sections where both channels had no fades at all. The second type included only fast fades. To remove slow fades, each data point was normalized with respect to the mean power of the points within 0.5 sec before and after it. Effectively, a square filter of length 1 sec was used. The filter response is therefore a sinc function with a mainlobe width of 1 Hz. The length of the filter was determined by the time required for 100 fades to be created.

A chart of average cross-correlation coefficients for each group of measurements is given in Table 5-5 below. In all LOS cases, enough measurements were done that meaningful histograms of the average correlation of each file in a group can be created, and these are also displayed below to show the range of correlations in each group.

Table 5-5: Power Cross-Correlation Coefficients For Fades Caused By People Moving

Transmit Antenna	Receiver Antenna Alignment	All Fades	Fast Fades
Horn	Aligned and 30° Off	0.466	0.232
	Each 15° Off	0.440	0.216
Omni (Line-Of-Sight)	Aligned and 30° Off	0.166	0.160
	Each 15° Off	0.162	0.156
Omni (Non LOS)	Each 15° Off	0.290	0.312

Table 5-6: Power Cross-Correlation Coefficients For Fades Caused By Receiver Moving

Transmit Antenna	Receiver Antenna Alignment	All Fades
Horn	Aligned and 30° Off	0.007
	Each 15° Off	-0.393

Figure 5-14: Cross-Correlation Histograms: People Moving, All Fades Included

Figure 5-15: Cross-Correlation Histograms: Receiver Moving, All Fades Included

Figure 5-16: Cross-Correlation Histograms: People Moving, Fast Fades Only

Omni Antenna On Tx: Rx Antennas Are 0° and 30° Misaligned

Omni Antenna On Tx: Each Rx Antenna 15° Misaligned

As can be seen from the graphs, correlations between channels for groups of files where people are moving with a horn on the transmitter are moderate to high. The results are the same whether each channel was misaligned by 15° or whether only one channel was misaligned by 30° . This suggests that in most such cases, much of the power entering both channels came from the same source, namely the direct ray from the transmit horn. If this direct LOS was blocked, much of the power to a misaligned receive horn was also blocked, and random reflections were not strong enough to overcome this. When only fast fades were considered, correlations were far weaker. The same results were seen from measurements where only the receiver was moved. This would suggest that the fast, multipath-induced fades on the two channels are more independent. Not all of these results can be attributed to angle diversity though, since the horn openings were several wavelengths apart and the receiver would therefore have space diversity as well.

Measurements with an omni-directional antenna on the transmitter showed much lower and more evenly spread correlation values. This is to be expected, since this configuration would produce more reflections, resulting in more instances of multipath fading. Since a larger proportion of the measurements would show multipath rather than shadow fades, the effect of strong correlations due to shadow fading would not be as obvious.

When the receiver was moved, only the multipath fades were measured, so correlations were consistently low. The surprising result for these measurements is that correlations were actually somewhat negative when both receive antennas were misaligned by 15° . This effect needs to be investigated further.

5.5 Gain From Selection Combining

Selection combining is one of the easiest methods of combining the outputs of several channels in a diversity system to achieve robustness against fading. The only information required by the algorithm is the magnitude of the received signal on each channel at a given point in time. The system examines the mean signal power on each

channel at given intervals and selects the channel with the highest power for its input. The maximum time between samples should be less than the time required for a fade to develop. In our case, the algorithm compared data points on the two channels one at a time and created a new envelope by selecting the greater of the two points in each case. Since sampling was done at speeds of 1000-5000 samples/sec, the time between samples ranged from 1 msec to 0.2 msec - more than fast enough to sample in each fade, as was shown in section 5.2.1 above.

Selection combining was done for all measurements where people were moving. Since correlations for these cases were quite high, the gain during deep fades was predictably small. When the horn on channel 1 was misaligned by 30° while the other was aligned, the unequal mean signal powers on the two channels further weakened the effectiveness of this method. In most of these cases, the small fades and enhancements created on both channels by people moving at the edge of the radiation pattern from the transmit antenna were less than the differences in mean signal strength (8 dB, on average - see section 5.2.2 above). In these cases, the aligned horn always had a stronger signal. These relatively shallow fades form the majority of the measurement, so the gain in mean signal power due to combining was negligible.

The only significant effect of selection combining was the mitigation of deep fades by 1-2 dB. Unequal mean signal powers on the channels when channel 1 was misaligned by 30° account for the small gains at the 90% level. A misaligned horn tended to have shallower fades than one that was aligned or almost aligned (15°), so selection combining had almost the same results in all cases for very deep fades. When an omni-directional antenna was used on the transmitter, gains were particularly low since fades tended to be less shallow.

For measurements where unequal mean signal powers clearly existed, gains were referenced against the channel with the highest mean signal power, since a non-diversity system would be set up to receive that channel. For measurements where antennas were each misaligned by 15° , no clear choice existed since the mean signal powers should theoretically be equal. Channel 2 was chosen arbitrarily in these cases. If one were to

choose the channel with the highest mean signal power as a reference in this case, one would have to assume that a non-diversity system would also always pick the best of the two orientations. This would only be possible if the non-diversity system could dynamically orient its antenna to pick the best channel. Without this assumption, this choice of reference would understate selection combining gains. Since the gains were calculated for many measurements in several environments, this method ensures that gains were referenced to the average signal that a non-diversity system would receive.

Table 5-7 details the results of the analysis and Figure 5-17 illustrates the effects that selection combining has on a typical envelope.

Table 5-7: Gains From Selection Combining: Fades Caused by People Moving

Transmit Antenna	Receive Antenna Alignment	10.0%	1.0%	0.5%	0.1%
Horn	Aligned and 30° Off	0.01	0.75	1.30	2.13
	Each Antenna 15° Off	0.50	1.79	2.31	2.62
Omni	Aligned and 30° Off	0.00	0.01	0.11	0.92
	Each Antenna 15° Off	0.02	0.24	0.86	1.96
Omni (Non LOS)	Each Antenna 15° Off	0.38	1.10	1.37	1.70

Figure 5-17: Effect of Selection Combining on a Typical CDF of Rx Power

5.6 Attenuation of Obstructions

Since both the desired signals and co-channel interferers often must pass through solid materials to reach the receiver, it is useful to examine the amount of attenuation these materials cause. Measurements were taken in several locations with transmit and receive horns aligned, at normal incidence, through obstructions constructed of various substances. One measurement was taken for each case. Results are shown in

Table 5-8 below. These results are used in section 5.10 for link budget calculations.

Table 5-8: Material Attenuation

Type of Obstruction	Attenuation
Narrow glass window (wire-reinforced) set between concrete blocks	0.94 dB
Metal door with large wire-reinforced window.	0.39 dB
Wall made of two sheets of gyprock	approx. 1 dB
Wall made of 6" concrete blocks	13.24 dB

5.7 Circumventing Obstruction Loss

Since the proposed CITR setup would have 12 horns on both the transmitter and receiver, it might be possible to significantly reduce the severity of fades during line of sight obstructions by sending the signal through one of the transmit horns that is pointed well away from the receiver, and receiving the reflected signal in one of the misaligned receive horns. This might bypass the obstruction that is causing the LOS transmission to fade. Several factors would limit the received power of the reflected signal, however: The path length might be significantly longer, some of the transmitted power would be absorbed or pass through the reflecting surface, and a rough or multi-angled surface would cause the transmitted beam to be broken up and scattered. The efficiency of this scheme would also depend strongly on where the obstruction was located: if it was close to the transmitter or receiver, it could easily block the signal through a 120° arc. A full test could not be set up since there was only one transmit horn and two receive horns, but several tests were run where one receive horn was pointed at the transmitter while the other was pointed 30° off to one side. The transmit horn was now swept through a 120° arc, in 10° steps. It was assumed that the obstruction causing the signal on the aligned horn to fade would only obstruct the direct LOS path (i.e. a small obstruction far from the transmitter and receiver). Both transmit and receive stations were located in the room

so that the strongest reflections should enter the receiver when the antennas on both stations were pointed 30° off the direct line. The results of a typical measurement are shown in Table 5-9 below.

Table 5-9: Static Measurement of Relative Received Powers

	Receive horn pointed directly at transmit horn	Receive horn pointed 30° off direct line
Transmit horn pointed directly at receiver	0 dB (reference)	-11.2 dB (mostly due to direct component)
Transmit horn pointed 30° off the direct line	-15 dB	-20.2 dB (mostly due to reflection)

A small portion of the power that was measured in the reflection test was due to a direct line component, since no obstruction was present to block this. However, it was very weak since this component was 11 dB below the reference power when the transmit horn was pointed at the receiver (and the receiver was rotated off the direct line), and during the reflection test it would have dropped by a further 15 dB, to 26 dB below the reference, since the transmit horn was also rotated off the direct line at that time.

As shown in Table 5-9, reflecting a signal off a wall to circumvent an obstruction resulted in a received signal power that was approximately 21 dB below the signal power for an unobstructed case. The direct line component - at 26 dB below the reference level during the test - would not have had much influence on this result, so reflections were the dominant factor. Fades due to obstructions could therefore be limited to approximately 21 dB, again assuming a small obstruction that blocked only line-of-sight transmission, and a fast (<0.5 sec, assuming a 1 second fade) switch between antennas. This reduction would only be effective in a very small percentage of cases, since fades of this depth represented less than 0.1% of all fades.

5.8 Constraints On Accuracy of Results

Since a variety of receive and transmit antenna setups were investigated, the number of measurements for one group was often small, especially for measurements with omni-directional antennas on the transmitter. The large number of samples for each measurement allows a high degree of confidence in the data for a particular location, however measurements in a wider variety of environments might alter the overall results slightly. Also, calibration inaccuracies and the limits of the A/D converter result in a 1 dB margin of error at power levels that are 30 dB or more below the reference point and a 0.5 dB margin of error at power levels that are 20-30 dB below the reference point.

5.9 Constraints On Scope Of Results

The types of results that could be obtained were limited by the amount of available equipment as well as constraints on equipment sensitivity. The noise floor of the receiver was at approximately -20 dBm (with a maximum SNR of 37 dB), limiting non-LOS measurements with an omni transmit antenna to short distances.

In most cases, only one or two people were creating fades at a given time, so the results don't show whether a large number of people in the beam of the transmit antenna will cause greater attenuation. The lack of multiple transmitter stations meant that the effects of co-channel interference on misaligned receive channels during deep fades could not be observed.

5.10 Link Budget Considerations

It is instructive to set up a link budget from the results analyzed in this chapter. This link budget can be used to calculate maximum ranges for typical systems. A single LOS link between a single-horn transmitter and a dual-horn receiver employing angle diversity and selection combining is considered. Effects of possible co-channel interference as well as gains due to a multi-horn transmit setup are not included. The entries in the link budget are described below.

5.10.1 Minimum Required Received Power

This parameter is calculated by using the noise figure results from Chapter 4. In a commercial system, the noise figure might be several dB higher. The noise bandwidth is assumed to uniformly cover the entire 250 MHz bandwidth seen by the receiver detection system. The SNR threshold was taken from calculations in [4] for a system using PSK modulation and a required BER of 10^{-3} .

5.10.2 Antenna Gains

Antenna gains were taken from Appendix E. These gains were measured in the anechoic chamber at the Communications Research Centre in Ottawa.

5.10.3 Transmit Power

Transmit power for EHF systems is limited mainly by the concern over possible health effects. As specified by ANSI-IEEE standards [21], the power density, averaged over a 6 minute period, must not exceed 5 mW/cm^2 at 29 GHz. Since horns radiate along a narrow beam, a maximum output power of 15 dBm was chosen to allow the signal to reach safe levels in a short distance (approx. 10 cm).

5.10.4 Link Attenuation

Path loss for the link was calculated as per the formula for free space loss given in Chapter 2. Two scenarios were examined:

- 1) An average case, with only a window as an obstruction and with one of the two receive antennas lined up with the transmit horn (the other one would then be misaligned by 30°). A mid-range fade margin for the aligned horn was chosen, based on 99.5% availability.
- 2) A worst case, where both antennas were at maximum misalignment (15°) and with three obstructions: a window and two gyprock walls. A worst-case fade margin for 99.5% availability was chosen.

Gains from selection combining were taken from experimental results. These gains show the reduction in required fade margins that 2-antenna angle diversity using selection combining would provide, at a 99.5% availability level.

5.10.5 Signal Strength Decay with Distance

In the absence of many reflections, measurements showed that with horns on both the transmit and receive antennas, the path loss exponent was approximately $n=2$. This figure was used to calculate the maximum distance that could be achieved.

5.10.6 Safety Factor

A 4 dB safety factor was included, to ensure reliable operation. For example, in some cases, noise figures of individual receivers in a commercial system might be worse than the one specified. It also includes an implementation margin to cover minor design flaws in commercial systems which may degrade the BER performance below its theoretical value (e.g. timing jitter).

Table 5-10: Link Budget For An EHF System Using Angle Diversity

Thermal Noise @ 25° C (KT)	-174 dBm/Hz
Noise Bandwidth (250 MHz)	84 dB-Hz
Receiver Noise Figure (our system)	5.2 dB
Min Eb/No For PSK (BER=0.001)	7 dB
Min. Required Received Power	-77.8 dB

Min. Required Received Power	77.8 dB
Transmit Antenna Gain	19 dBi
Receive Antenna Gain	15.2 dBi
Max Transmit Power	15 dBm
Maximum Allowable Link Loss	127 dB

	Average Case	Worst Case
Path loss at 1 metre (29.125 GHz)	61.7 dB	61.7 dB
Fade Margin (Shadow & Multipath)	13 dB	~27 dB
Gain From Selection Combining	1.3 dB	2.3 dB
Loss From Fixed Obstructions	0.4 dB	2.4 dB
Loss From Horn Misalignment	0 dB	~ 3 dB
Link Attenuation at 1 m	76.4 dB	96.4 dB
Implementation Margin	4 dB	4 dB
Margin Left Over For Path loss	46.6 dB	26.6 dB
MAX ACHIEVABLE DISTANCE	214 m	21.4 m

5.10.7 Link Budget Conclusions

As can be seen from Table 5-10, the type of system outlined in this thesis should have adequate range for most indoor environments. Note that even the worst case scenario assumed a relatively open space, with no major fixed obstructions such as concrete walls. As can be seen from the measurements in section 5.6, such walls would drastically reduce the range. This is actually a favorable factor since it allows a high level of frequency reuse.

6. Thesis Summary and Conclusion

6.1 Summary of Thesis

The purpose of this thesis was to design and implement an EHF broadband measurement system capable of recording signals on angle diversity channels, and to analyze measurements made with this system with the objective of examining channel fading characteristics and the effectiveness of angle diversity for mitigating fades. A review of the background theory that would be required for fully understanding the results of this thesis was presented in Chapter Two, and a review of the current state of EHF channel characterization research was presented in Chapter Three. A full analysis of the measurement system and its components was presented in Chapter Four, and the results and analysis of measurements were presented in Chapter Five.

The analysis of the measurements is vital for drawing conclusions about the practical viability of the proposed CITR EHF broadband indoor system, so the results are summarized below. A variety of locations was chosen for the measurements, including computer laboratories, hallways, a factory floor setting and large open areas (the Minto Centre lobby). Several types of measurements were made at each site, both with people moving and with the receiver moving. A single 20° horn or an omni-directional antenna was used on the transmitter, and a 2-channel angle diversity system with 30° horns was used on the receiver.

6.2 Summary of Results

The following is a summary of the results that were obtained from the measurements and the subsequent analysis.

6.2.1 Fading Characteristics

- When people are moving in a room, and tests are performed with horns on both the transmitter and receiver, fades are present for only 40% to 60% of the time. When fades are present, the majority are shallow.
- When people are moving in a room, and occasionally intersecting the line-of-sight transmission, fades greater than 10 dB make up less than 10% of all fade occurrences. They generally last for about 1 second.
- The deepest fades (15-25 dB) occur when a person obstructs the signal between transmit and receive horns that are pointed directly at each other. (15-25 dB at the lowest point of the fade).
- At the 0.1% level of probability, fade margins vary by 8 to 15 dB for most types of measurements. Horn-horn measurements showed the greatest variance.
- Although using an omni-directional antenna on the transmitter increase the occurrence of fades, deep shadow fading tends to be less severe (~5 dB less for cases where the receiver horn is pointed at the transmit antenna)
- As the receive horn is misaligned in a horn-horn setup, fade depths become more shallow at a rate of approximately 3 dB per 15° of rotation (0.1% probability level)
- Deep fades are correlated with an angle diversity setup when the transmit antenna is a horn, but become less so when an omni is used on the transmitter. When the receiver is moved, or when only fast fades are included in the analysis, correlations drop significantly. It should be noted that in a system using angle-diversity, the degree of correlation alone does not define the amount of diversity gain that can be expected, since mean signal powers on the received channels are unequal. This effect lowers diversity gains from uncorrelated channels.
- Multipath fades are lower when they are caused by the movement of people instead of antenna movement. This is especially true for horn-horn measurements, where fades were usually 2-5 dB lower.

6.2.2 Mean Signal Strengths

- As the receive horn is misaligned by 30° in a horn-horn measurement where no people are moving, the average received power drops only about 8 dB (significant power spillover, as demonstrated in the thesis). Samples for this measurement were taken throughout a sphere of 34 cm radius, to eliminate multipath effects.
- Circumventing obstruction loss by relying on signals reflected off a wall would be unlikely to result in significant gains since the average reflected signal strength tended to be below the direct power, even during most deep fades.

6.2.3 Gains from Selection Combining

- Selection combining of two receive channels in an angle diversity setup produced at most a 2.6 dB gain, and only for very deep and rare fades.
- Selection combining of horn-horn measurements with both receive antennas misaligned by 15° was the most effective; measurements with an omni-directional transmit antenna and with one receive horn misaligned by 30° showed the least gain.

6.3 Conclusions

The results of this thesis show that the construction of an indoor EHF system may be a practical possibility, from the SNR point of view. Since the signal is broadband, and the channel is frequency-selective, there is a low probability that fading will affect all frequencies in the channel at one time. Fades therefore tend to be more shallow than would be the case if the channel were narrowband. The use of horns on the receiver increases gain and results in long quiescent periods on the channel, where no fades at all are present even though there is movement in the area. The majority of fades occur when people are moving at the edge of the beam, and these fades are quite shallow. Deep fades due to shadow fading are relatively rare, since a person must block the direct LOS path.

Angle diversity on the receiver may not be an effective way to combat fading, however. Attempting to circumvent a line-of-sight blockage by switching both the

transmit and receive antennas to receive only reflections from nearby objects such as walls does not produce enough signal strength to effectively reduce the depth of shadow fades. It is also unlikely that the receiver and transmitter will be situated in just the right position for a ray to bounce into the centre of one of the receive antennas without being reflected around the room several times and losing much of its strength due to path loss and absorption. This is especially true if either station is situated in a corner, where most of its antennas would be blocked.

When the transmitter is lined up with the receiver, some gain may be achieved by having a receive antenna misaligned and receiving random reflections. However, as the strong correlation between channels for horn-horn measurements shows, the spillover of signal from the direct beam into the misaligned receive antenna produces the dominant effect: when the direct path is obstructed, the power on both the aligned and misaligned horns drops. Although the received power on the misaligned horn drops less during a deep fade, its average received power is also lower. The result is that during deep fades, the signal power on both channels is almost equally low. Another reason for the high correlation is that an obstruction that is relatively close to either station would block both horns at the same time. Also, at short distances, the narrow beam from the transmit antenna would not spread out far enough to reach the walls or other obstructions and cause reflections into the receiver antenna. As a result, in many cases, the only major reflections might be from a wall behind the receiver. These reflections would most likely be blocked at the same time that the line-of-sight was blocked, since the path of these signals from the transmitter would be almost identical to the line-of-sight path.

The unequal mean signal powers on the two channels means that this type of setup will be mostly ineffective against shallow fades on the direct path. The low gains achieved with selection combining at the 90% probability level underscore this point.

Angle diversity is even less effective when an omni directional antenna is used on the transmitter since shadow fades tend to be more shallow. Even when both receive horns are rotated equal distances off the direct line-of-sight path and the mean signal

power on each horn is approximately equal, selection combining produces only small gains since fading is infrequent and shallow to begin with.

In conclusion, then, it can be said that for EHF indoor systems with stationary terminals, angle diversity in conjunction with selection combining is not an efficient way to decrease required fade margins. This setup produces minimal gains against shallow multipath fades, and is only slightly effective against shadow fading, with at most a 2.6 dB gain. The rarity of these deep shadow fades make the extra complexities of this design unattractive. A system which has one horn on the receiver that is aligned with one horn on the transmitter, and is affected by fading from moving obstructions such as people will require, on average, a fade margin of 14.6 dB at the 1% probability level. In a worst-case scenario, this increases to 25 dB.

6.4 Recommendations for Further Research

There is still some analysis that can be done on the data that have been gathered for this experiment. It would be instructive to break down the results to show the differences between the variety of locations that were studied. Also, in most cases, two sets of measurements were done at one location so that both long and short distances between the stations were represented in the results. Preliminary results show that shadow fades were deeper for short distances, and it might be of interest to pursue this. For all of the above cases, an increased number of measurements might be required to produce results that can be analyzed with confidence, since the amount of data in each category will be quite small.

There are several practical variations on the selection combining technique used in this thesis, such as switch-and-stay combining. It might be of interest to determine whether the use of one of these techniques would result in larger gains.

Measurements that were done with a moving receiver could be analyzed to produce a range of K factors for Rician fading, so that a model of a moving indoor EHF system could be developed.

To do a more in-depth examination of indoor EHF systems, it might be useful to gather more information on systems which utilize angle diversity in a non LOS situation. For this purpose, an omni-directional antenna on the transmitter might be used. Since all received signals will consist of reflections, there is some potential for effective gains.

A complete angle diversity setup requires multiple-sector coverage on the transmitter as well as the receiver, so that portables can communicate with any available base station, regardless of orientation. Such a system may also have better diversity gains, since signals can be emitted from the transmitter in various directions. As shown in Chapter 5, however, the extra gains would probably be quite small. To gather conclusive evidence on the gains achievable from such a design, a full system would have to be constructed, complete with an algorithm that would switch between antennas at appropriate times. Replacing the horn on the transmitter with an omni-directional antenna is not an adequate simulation since a transmitter with true angle diversity will only transmit a narrow beam in a specific direction, while an omni-directional antenna will flood the area with its signal. Measurements that were taken for this thesis with an omni-directional transmit antenna would therefore not be adequate for examining diversity gains in such a system, and the setup would require extensive modification for such tests.

Finally, it seems that fading on static indoor EHF systems is heavily dependent on the amount of traffic in the immediate area. In the results outlined above, quiescent portions of the channels were removed before processing, to ensure that results were influenced only by fades on the channels. This was done because the length and frequency of quiescent periods depended very much on the location where the measurements were taken, and time of day when the experiment was run. No statistics regarding the movement of people have been developed yet, so this unpredictable factor had to be removed from the calculations. In an actual environment, the length and frequency of quiescent periods would influence the overall outage probability, and should thus be included in the overall fade margin. A study of the number of people moving through an area, as well as their speed of movement, would therefore go a long way to determining the required fade margin for particular types of locations.

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Appendix A: Photographs of Measurement Setup

Plate A-1: View of Transmitter

Plate A-2: View of Receiver and Close-Up View of Receive Antennas

Appendix B: Characteristics of Transmitter Components

This appendix contains the results of measurements used to characterize and test the components used to construct the transmitter.

Figure B-1: Frequency Response of Attenuator for Ch. A

Figure B-2: Frequency Response of Attenuator for Ch. B

Figure B-3: Frequency Response of Baseband Block, Ch A, 450 MHz Range

Figure B-4: Frequency Response of Baseband Block, Ch A, 230 MHz Range

Figure B-5: Frequency Response of Baseband Block, Ch B, 450 MHz Range

Figure B-6: Frequency Response of Baseband Block, Ch B, 230 MHz Range

Figure B-7: Spectrum of Output Signal-Ch A, 800 MHz Range

Figure B-8: Spectrum of Output Signal-Ch A, 800 MHz Range, No Lowpass Filter

Figure B-9: Spectrum of Output Signal-Ch B, 800 MHz Range

Figure B-10: Spectrum of Output Signal-Ch A, 230 MHz Range

Figure B-11: Spectrum of Output Signal-Ch B, 230 MHz Range

Figure B-12: Frequency Response of the IF Filter

Figure B-13: Gain Sweep of IF amp

Figure B-14: Frequency Response of IF amp

Figure B-15: Frequency Response of IF Attenuator

Figure B-16: Spectrum of IF stage Output, 500 MHz Range

Figure B-17: Gain Sweep of the TX RF Amplifier

Figure B-18: Output Power Spectrum of TX IF LO

Appendix C: Characteristics of Receiver Components

This appendix contains the results of measurements used to characterize and test the components used to construct the receiver.

Table C-1: PIN Switch Insertion Loss for Channel 1

Frequency (Ghz)	Level (dB)	Frequency (GHz)	Level (dB)
28.9189	-1.24	29.1533	-1.38
28.9270	-1.24	29.1614	-1.38
28.9351	-1.25	29.1694	-1.38
28.9432	-1.26	29.1775	-1.37
28.9513	-1.26	29.1856	-1.37
28.9593	-1.27	29.1937	-1.37
28.9674	-1.27	29.2018	-1.36
28.9755	-1.27	29.2098	-1.36
28.9836	-1.27	29.2179	-1.37
28.9917	-1.30	29.2260	-1.36
28.9997	-1.30	29.2341	-1.36
29.0078	-1.31	29.2422	-1.35
29.0159	-1.31	29.2503	-1.35
29.0240	-1.32	29.2583	-1.34
29.0321	-1.32	29.2664	-1.34
29.0402	-1.34	29.2745	-1.32
29.0482	-1.34	29.2826	-1.32
29.0563	-1.34	29.2907	-1.32
29.0644	-1.34	29.2987	-1.32
29.0725	-1.35	29.3068	-1.31
29.0806	-1.35	29.3149	-1.31
29.0886	-1.37	29.3230	-1.31
29.0967	-1.37	29.3311	-1.30
29.1048	-1.38		
29.1129	-1.37		
29.1210	-1.36		
29.1290	-1.36		
29.1371	-1.36		
29.1452	-1.37		

Table C-2: PIN Switch Insertion Loss for Channel 2

Frequency (Ghz)	Level (dB)	Frequency (GHz)	Level (dB)
28.9189	-1.10	29.1533	-1.21
28.9270	-1.10	29.1614	-1.21
28.9351	-1.10	29.1694	-1.21
28.9432	-1.10	29.1775	-1.19
28.9513	-1.10	29.1856	-1.20
28.9593	-1.10	29.1937	-1.21
28.9674	-1.11	29.2018	-1.20
28.9755	-1.12	29.2098	-1.20
28.9836	-1.12	29.2179	-1.19
28.9917	-1.13	29.2260	-1.19
28.9997	-1.13	29.2341	-1.19
29.0078	-1.14	29.2422	-1.18
29.0159	-1.14	29.2503	-1.17
29.0240	-1.17	29.2583	-1.17
29.0321	-1.18	29.2664	-1.17
29.0402	-1.18	29.2745	-1.17
29.0482	-1.18	29.2826	-1.16
29.0563	-1.17	29.2907	-1.15
29.0644	-1.17	29.2987	-1.15
29.0725	-1.17	29.3068	-1.14
29.0806	-1.20	29.3149	-1.14
29.0886	-1.20	29.3230	-1.14
29.0967	-1.21	29.3311	-1.13
29.1048	-1.20		
29.1129	-1.20		
29.1210	-1.19		
29.1290	-1.21		
29.1371	-1.21		
29.1452	-1.21		

Table C-3: PIN Switch Isolation for Channel 1

Frequency (Ghz)	Level (dB)	Frequency (GHz)	Level (dB)
28.9189	-27.38	29.1533	-27.82
28.9270	-27.34	29.1614	-27.86
28.9351	-27.33	29.1694	-27.93
28.9432	-27.35	29.1775	-27.95
28.9513	-27.35	29.1856	-28.02
28.9593	-27.33	29.1937	-28.04
28.9674	-27.34	29.2018	-28.09
28.9755	-27.33	29.2098	-28.15
28.9836	-27.32	29.2179	-28.19
28.9917	-27.33	29.2260	-28.27
28.9997	-27.33	29.2341	-28.30
29.0078	-27.34	29.2422	-28.34
29.0159	-27.35	29.2503	-28.35
29.0240	-27.37	29.2583	-28.39
29.0321	-27.35	29.2664	-28.43
29.0402	-27.40	29.2745	-28.46
29.0482	-27.40	29.2826	-28.47
29.0563	-27.42	29.2907	-28.47
29.0644	-27.44	29.2987	-28.49
29.0725	-27.48	29.3068	-28.52
29.0806	-27.48	29.3149	-28.50
29.0886	-27.52	29.3230	-28.55
29.0967	-27.56	29.3311	-28.54
29.1048	-27.58		
29.1129	-27.63		
29.1210	-27.64		
29.1290	-27.69		
29.1371	-27.71		
29.1452	-27.77		

Table C-4: PIN Switch Isolation for Channel 2

Frequency (Ghz)	Level (dB)	Frequency (GHz)	Level (dB)
28.9189	-25.57	29.1533	-25.97
28.9270	-25.56	29.1614	-25.99
28.9351	-25.56	29.1694	-26.04
28.9432	-25.53	29.1775	-26.10
28.9513	-25.55	29.1856	-26.13
28.9593	-25.53	29.1937	-26.19
28.9674	-25.54	29.2018	-26.23
28.9755	-25.54	29.2098	-26.26
28.9836	-25.53	29.2179	-26.30
28.9917	-25.55	29.2260	-26.35
28.9997	-25.54	29.2341	-26.38
29.0078	-25.54	29.2422	-26.42
29.0159	-25.55	29.2503	-26.46
29.0240	-25.56	29.2583	-26.49
29.0321	-25.58	29.2664	-26.52
29.0402	-25.59	29.2745	-26.56
29.0482	-25.60	29.2826	-26.59
29.0563	-25.63	29.2907	-26.58
29.0644	-25.62	29.2987	-26.58
29.0725	-25.66	29.3068	-26.59
29.0806	-25.66	29.3149	-26.63
29.0886	-25.68	29.3230	-26.64
29.0967	-25.72	29.3311	-26.64
29.1048	-25.75		
29.1129	-25.77		
29.1210	-25.80		
29.1290	-25.85		
29.1371	-25.87		
29.1452	-25.92		

Figure C-1: Gain Sweep of LNA #1 **Figure C-2: Gain Sweep of LNA #2**

Figure C-3: Gain Sweep of RF Stage **Figure C-4: Gain Sweep of Mixer**

Figure C-5: Frequency Response of Bandpass Filter After Mixer

Figure C-6: Frequency Response of Bandpass Filter Before Crystal Detector

Figure C-7: Gain Sweep of the HP 8347A Amplifier

Figure C-8: Frequency Response of the HP 8347A Amplifier

Figure C-9: Gain Sweep of the MCL ZFL-2000 Amplifier

Figure C-10: Gain Sweep of IF Stage

Figure C-11: Output Voltage vs Input Power for Crystal Detector **Figure C-12: Gain Sweep of the Receiver**

Appendix D: Calibration Results

This appendix details the results of one of several calibration runs that was performed to accurately map the output of the data acquisition system to receiver input power values. The process of properly calibrating the equipment is described in section 4.3.2. Table D-1 shows the results for each step in the calibration process in numerical form, and Figures D-1 and D-2 show them in graphical form.

Figure D-1: Channel 1 Calibration - A/D Output vs Receiver Input Signal Power

Figure D-2: Channel 2 Calibration - A/D Output vs Receiver Input Signal Power

Table D-1: A/D Output For All Attenuation Steps Below the Reference Level

Atten	Ch1	Ch2
0.0	1995	1911
-0.1	1990	1907
-0.2	1984	1896
-0.3	1972	1884
-0.4	1963	1872
-0.5	1954	1861
-0.6	1946	1851
-0.7	1937	1839
-0.8	1925	1827
-0.9	1910	1815
-1.0	1894	1802
-1.1	1880	1788
-1.2	1868	1776
-1.3	1855	1765
-1.4	1845	1753
-1.5	1831	1737
-1.6	1825	1726
-1.7	1812	1712
-1.8	1800	1700
-1.9	1785	1687
-2.0	1767	1673
-2.1	1757	1661
-2.2	1743	1647
-2.3	1726	1635
-2.4	1710	1621
-2.5	1695	1608
-2.6	1681	1595
-2.7	1667	1582
-2.8	1655	1565
-2.9	1641	1555
-3.0	1627	1541
-3.1	1616	1528
-3.2	1600	1514
-3.3	1587	1501
-3.4	1573	1489
-3.5	1555	1476
-3.6	1543	1458
-3.7	1530	1446
-3.8	1517	1432
-3.9	1504	1417
-4.0	1487	1404
-4.1	1475	1389
-4.2	1463	1377
-4.3	1450	1361
-4.4	1439	1348
-4.5	1427	1332
-4.6	1415	1319

Atten	Ch1	Ch2
-4.7	1403	1305
-4.8	1390	1292
-4.9	1378	1280
-5.0	1366	1268
-5.1	1354	1256
-5.2	1342	1244
-5.3	1330	1232
-5.4	1318	1220
-5.5	1306	1208
-5.6	1294	1196
-5.7	1282	1184
-5.8	1270	1173
-5.9	1257	1161
-6.0	1244	1149
-6.1	1232	1137
-6.2	1221	1125
-6.3	1209	1113
-6.4	1196	1101
-6.5	1186	1089
-6.6	1173	1077
-6.7	1161	1065
-6.8	1151	1053
-6.9	1139	1045
-7.0	1126	1032
-7.1	1114	1025
-7.2	1102	1015
-7.3	1091	1007
-7.4	1079	996
-7.5	1068	986
-7.6	1052	973
-7.7	1043	965
-7.8	1032	953
-7.9	1021	944
-8.0	1009	936
-8.1	998	925
-8.2	990	915
-8.3	980	904
-8.4	968	895
-8.5	958	886
-8.6	950	876
-8.7	939	877
-8.8	926	868
-8.9	917	856
-9.0	908	849
-9.1	898	841
-9.2	887	831
-9.3	878	819

Atten	Ch1	Ch2
0.0	1995	1911
-0.1	1990	1907
-0.2	1984	1896
-0.3	1972	1884
-0.4	1963	1872
-0.5	1954	1861
-0.6	1946	1851
-0.7	1937	1839
-0.8	1925	1827
-0.9	1910	1815
-1.0	1894	1802
-1.1	1880	1788
-1.2	1868	1776
-1.3	1855	1765
-1.4	1845	1753
-1.5	1831	1737
-1.6	1825	1726
-1.7	1812	1712
-1.8	1800	1700
-1.9	1785	1687
-2.0	1767	1673
-2.1	1757	1661
-2.2	1743	1647
-2.3	1726	1635
-2.4	1710	1621
-2.5	1695	1608
-2.6	1681	1595
-2.7	1667	1582
-2.8	1655	1565
-2.9	1641	1555
-3.0	1627	1541
-3.1	1616	1528
-3.2	1600	1514
-3.3	1587	1501
-3.4	1573	1489
-3.5	1555	1476
-3.6	1543	1458
-3.7	1530	1446
-3.8	1517	1432
-3.9	1504	1417
-4.0	1487	1404
-4.1	1475	1389
-4.2	1463	1377
-4.3	1450	1361
-4.4	1439	1348
-4.5	1427	1332
-4.6	1415	1319

Atten	Ch1	Ch2
-4.7	1403	1305
-4.8	1390	1292
-4.9	1378	1280
-5.0	1366	1268
-5.1	1354	1256
-5.2	1342	1244
-5.3	1330	1232
-5.4	1318	1220
-5.5	1306	1208
-5.6	1294	1196
-5.7	1282	1184
-5.8	1270	1173
-5.9	1257	1161
-6.0	1244	1149
-6.1	1232	1137
-6.2	1221	1125
-6.3	1209	1113
-6.4	1196	1101
-6.5	1186	1089
-6.6	1173	1077
-6.7	1161	1065
-6.8	1151	1053
-6.9	1139	1045
-7.0	1126	1032
-7.1	1114	1025
-7.2	1102	1015
-7.3	1091	1007
-7.4	1079	996
-7.5	1068	986
-7.6	1052	973
-7.7	1043	965
-7.8	1032	953
-7.9	1021	944
-8.0	1009	936
-8.1	998	925
-8.2	990	915
-8.3	980	904
-8.4	968	895
-8.5	958	886
-8.6	950	876
-8.7	939	877
-8.8	926	868
-8.9	917	856
-9.0	908	849
-9.1	898	841
-9.2	887	831
-9.3	878	819

Appendix E: Receiver Noise Figure Calculations

Notes on Noise Figure Calculations

Noise figures for the receiver were calculated according to the method developed in section 2.9.1. The formula is:

$$F = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \frac{F_4 - 1}{G_1 G_2 G_3} \dots + \frac{F_x - 1}{G_1 G_2 \dots G_{(x-1)}}$$

where F_x is the noise figure for component number 'x' in the chain, starting at the front end, and G_x is the gain for component number 'x' in the chain. The results of these calculations are displayed on the next page.

Explanation of Column Headings For Noise Figure Calculations

(1) Gain	Gain of each component
(2) Noise Figs	Noise figure for each component
(3) Gain Sums	The denominator for each expression in the above equation
(4) $F_x - 1$	The numerator for each expression in the above equation
(5) NF Addition	The addition to the overall noise figure from each component (ie. the evaluation of each of the above expressions)
(6) Total NF	The noise figure of the system after the inclusion of all previous components. The final figure in this column is the receiver noise figure.

All values are in dB.

Component	(1) Gain		(2) Noise Figs		(3) Gain Sums		(4) Fx-1		(5) NF Addition		(6) Total NF	
	Ch 1	Ch 2	Ch 1	Ch 2	Ch 1	Ch 2	Ch 1	Ch 2	ch 1	ch 2	ch 1	ch 2
Flexible Waveguide		-0.3		0.30				-11.46		0.30		0.30
PIN Switch	-1.3	-1.2	1.20	1.20		-0.3	-4.97	-4.97	1.20	-4.67	1.20	1.50
WR-28 Termination	-0.3	-0.3	0.30	0.30	-1.3	-1.3	-11.46	-11.46	-10.16	-9.96	1.51	1.80
LNA#1	24.5	24.5	3.01	3.01	-1.6	-1.6	0.00	0.00	1.60	1.80	4.56	4.81
Suhner cable	-1.0	-1.0	1.00	1.00	22.9	22.9	-5.87	-5.87	-28.77	-28.57	4.57	4.81
LNA#2	24.4	24.4	3.43	3.43	21.9	21.9	0.80	0.80	-21.10	-20.90	4.58	4.82
WR-28 Termination	-0.1	-0.1	0.10	0.10	46.3	46.1	-16.33	-16.33	-62.63	-62.43	4.58	4.82
Overall RF Gain	46.2	46.0						RF Stage Noise Figure			4.58	4.82
Harmonic Mixer	-25.6	-25.6	25.60	25.60	46.2	46.0	25.59	25.59	-20.61	-20.41	4.59	4.84
Filter and Cable	-0.5	-0.5	0.50	0.50	20.6	20.4	-9.14	-9.14	-29.74	-29.54	4.59	4.84
ZHL 1042J Amp	25.0	25.0	4.50	4.50	20.1	19.9	2.60	2.60	-17.50	-17.30	4.62	4.86
Step Attenuator	-34.0	-34.0	34.00	34.00	45.1	44.9	34.00	34.00	-11.10	-10.90	4.73	4.98
Cables to Atten	-0.2	-0.2	0.20	0.20	11.1	10.9	-13.27	-13.27	-24.37	-24.17	4.74	4.98
ZFL-2000 Amp	20.5	20.5	7.00	7.00	10.9	10.7	6.03	6.03	-4.87	-4.67	5.19	5.43
HP 8347A Amp	29.2	29.2	15.00	15.00	31.4	31.2	14.86	14.86	-16.54	-16.34	5.22	5.46
Filter	-0.4	-0.4	0.40	0.40	60.6	60.4	-10.16	-10.16	-70.76	-70.56	5.22	5.46
Overall IF Gain	14	14			Receiver Noise Figure (dB)						5.22	5.46
Overall Rx Gain	60.2	60			Equivalent Noise Temperature (deg. K)						674.575	729.359
					Additional Noise Power (dBW/Hz)						-140.11	-139.97
					Output Noise Power (dBW/Hz)						-138.56	-138.52
					Noise Floor (dBW/250 MHz BW)						-54.578	-54.539
					SNR (dB) Note: 17.2 dBm max output						41.7784	41.7385

Appendix F: Antenna Calibration

Calibration of the antennas for the receiver and transmitter was done in an anechoic chamber at the Communication Research Centre in Ottawa, Ontario.

Figure F-1: Radiation Pattern for Tx Horn Antenna at 28 GHz (H-Plane)

Figure F-2: Radiation Pattern for Tx Horn Antenna at 29 GHz (H-Plane)

Figure F-3: Radiation Pattern for Tx Horn Antenna at 30 GHz (H-Plane)

Figure F-4: Radiation Pattern For Omni Antenna at 29 GHz (H-Plane)

Figure F-5: Radiation Pattern For Omni Antenna at 29 GHz (E-Plane)

Figure F-6: Radiation Pattern for Rx Horn at 29 GHz, Ch 1 (H-Plane)

Figure F-7: Radiation Pattern for Rx Horn at 29 GHz, Ch 1 (E-Plane)

Figure F-8: Radiation Pattern for Rx Horn at 29 GHz, Ch 2 (H-Plane)

Figure F-9: Radiation Pattern for Rx Horizontal at 29 GHz, Ch 2 (E-Plane)

Appendix G: Measurement Locations

Figure G-1: Layout of 4th floor Minto Centre

Note: Measurement areas are indicated with diagonal stripes

Figure G-2: Layout of Minto Centre Lobby and Civil Engineering Lab

Figure G-3: Layout of 4th floor, Mackenzie Engineering Building

BLOCK A MACKENZIE BUILDING

LEVEL 4

SCALE : 0 1 2 3 4 5 10 METERS

Appendix H: Listings of Major Programs Used For Data Analysis

This section describes, in general terms, several of the programs that were used for analyzing the data. Only a few programs were selected since they were used to perform the bulk of the analysis, and a complete listing and explanation of all programs would be too cumbersome. In most cases, programs and subroutines were shortened, and have unimportant sections (e.g. file management, automation procedures, and integral graphing/printing routines) removed for clarity.

Calibration of Data

The results of equipment calibration are shown in Appendix G. To attain these results, a set of measurements was created for each channel, where the input to the receiver was manually attenuated by a certain amount and the A/D output was sampled at 100 Hz for 10 seconds at each step. A program then found the average value of this A/D output, which was then rounded to the nearest integer (e.g. the average A/D output value over several hundred samples for 1.6 dB of attenuation below the reference level might be 1963.85, which would then be rounded to 1964). This program is not detailed here. The results of the equipment calibration were used to calibrate the data, as detailed below:

Three procedures were used in the data calibration process:

- 1) A program to use the calibration data reproduced in Appendix G to produce a calibration vector of length 2048 for each channel, so that each A/D output value (range: 1 to 2048) could be mapped directly to an equivalent received input power.
- 2) A program to load in the data from a measurement (with results from channels 1 and 2 still interleaved, since sampling from each channel was done in sequence), separate the data into results for the two channels, and calibrate it using the procedure outlined in (3).

3) A procedure to calibrate a segment of data by comparing the measured A/D values for a specific channel to the matrix generated by program #1 for that channel, and substituting the correct power level (in dB).

The programs are outlined in order below.

Program To Create Calibration File

```
%access this function with: adcal=caladlev(calval,channelcals)
%This file creates a 2048 point vector that corresponds
%to the 2048 levels that the A/D board can put out. Each of the points in the
%matrix is a calibrated (dB) value corresponding to the particular
%level that the A/D board has put out. For example, level 208 coming from
%the A/D board corresponds to a relative power level of -32 dB. Thus, row 208
%of the matrix contains the value -32.To calibrate a data file, use the
%data value as an index. For example, if a particular data value is 208,
%an index of 208 into the matrix is used, and the value is extracted.
%'calval' should contain the dB values corresponding to the A/d
%levels found in vector 'Channelcals'(the actual calibration file that matches
%A/D output values to dB values).
%NOTE: channelcals is assumed to be monotonically decreasing
%(calibration starts at highest value and works down to lowest)
```

```
function adcal=caladlev(calval,channelcals)
adcal=zeros(2048,1);
len=length(channelcals); %ASSUME calval is the same length
halfabove=2049;
for loop=1:(len-1) %process all points in calibration file by matching
                    %A/D levels to them
    %process this point by matching all the corresponding A/D levels to it
    thispoint=channelcals(loop);
    nextpoint=channelcals(loop+1);
    halfbelow=round((thispoint+nextpoint)/2);
    for loop2=halfbelow:(halfabove-1) %A/D levels from halfway below cal pt to
                                    %halfway above
        adcal(loop2)=calval(loop); %match dB value
    end; %for loop over A/D vals for one cal pt
    halfabove=halfbelow; %bottom boundary of A/D vals for this cal pt is top
                        %boundary for the next cal pt
end; %for loop over all cal pts except the last
```

```

%now do last point
for loop2=1:halfabove %from lowest possible A/D point to upper boundary
    %for lowest cal pt
    adcal(loop2)=calval(len); %match dB value
end; %for loop over A/D vals for last cal pt

```

Program To Calibrate all Data From A Set Of Measurements

```
%procedure allcal
%Note: the raw data about the two channels won't be saved.
%major program variables: data1, data2: raw data for channels 1 and 2
%caldat1,caldat2: calibrated data for the two channels
%adcal1,adcal2: Vectors containing power levels that A/D values corresponding to the
%index into the matrix should be converted to.
```

%LOAD THE DATA FILE AND SEGREGATE DATA FOR THE TWO CHANNELS
(procedure to load raw data from .measurement)

```
[data1,data2]=demult(Ch); %the datafile holds points from both channels:  
                        %1 point from each channel, in order. Therefore, the  
                        %data must be demultiplexed and stored seperately.  
disp(['pulled data apart into 2 channels: data1 and data2' 10]);  
clear Ch; %clear out the raw data-it's no longer needed
```

%CALIBRATE THE DATA FOR CHANNEL 1

```
save chantemp data2 %get rid of info on channel 2 for now
clear data2
```

```
disp(['loading calibration file ' calfilename]);
```

```
eval(['load ' calfilename]);
```

```
% This loads calibration data for 2 channels, and the dB values to convert the data to  
disp('calibrating the data from channel 1...result in caldat1');
```

```
numdata=length(data1);
```

```
if numdata>500000
```

```
disp ('TOO MUCH DATA');
```

exit

end

```
numtempfiles=ceil(numdata/80000)
```

%max number of points to process at a time
%is about 80000.

```
pointsperfile=round(numdata/numtempfiles); %Each temp file has this many
```

%points. The last one may have
%some less due to rounding error.

%SET UP THE DATA FILES FOR EACH DATASET

```

for temploop=1:1:numtempfiles
startpoint=(temploop-1)*pointsperfile+1; %first point to store for this
                                     %data set
if temploop==numtempfiles %last set- there may be less data points
                                     %from rounding
    endpoint=numdata; %set 'endpoint' to last point in file
else %an earlier data set
    endpoint=temploop*pointsperfile;
end %ifelse
pointstostore=data1(startpoint:endpoint);
name=['',(temploop+48)]; %temploop ranges from 1 to a max of 9 (data broken down
                                     %into a maximum of 9 sections)

fil=name(2);
eval(['save caltemp'fil ' pointstostore']);
end

```

clear data1 %it's now all packed into several temp files

%NOW PROCESS EACH FILE IN TURN, AND SAVE RESULTS

```

for temploop=1:1:numtempfiles
name=['',(temploop+48)]; %temploop ranges from 1 to a max of 9.
fil=name(2);
eval(['load caltemp'fil]);
cali=newcal(adcal1,pointstostore); %calibrate the data points to known
                                     %dB values from the calibration curve
                                     %for channel 1

clear pointstostore; %don't need the raw data anymore
eval(['save called'fil ' cali']);
clear cali;
end

```

%NOW PULL ALL THE CALIBRATED FILES TOGETHER

```

startpoint=1;
for temploop=1:1:numtempfiles
name=['',(temploop+48)]; %temploop ranges from 1 to a max of 9.
fil=name(2);
eval(['load called'fil]);
endpoint=startpoint+length(cali);
caldat1(startpoint:(endpoint-1),1)=cali(1:length(cali),1); %add the data from
                                                             %this file to caldat1

clear cali;
startpoint=endpoint; %index into caldat1 where next data set should be put
end %for

```

(procedure to save data 'caldat1' is saved, as well as
'channel1cal' for future reference)

```
clear caldat1;
```

```
%NOW CALIBRATE CHANNEL 2
```

(same process, using calibration curve for Channel 2)

```
load chantemp %get data2 back
```

```
disp(['calibrating the data from channel 2...result in caldat2']);
```

```
.
```

```
.
```

```
.
```

```
%delete temporary files and save calibrated data in zipped-up format
```

(procedure to save calibrated data for channel 2 and remove temporary files)

```
end
```

Procedure to Calibrate a Block Of Data

```
% call this function with: caldat=newcal(adcal, data)
```

```
% 'data' is a 1-d column of results, to be converted to dB values
```

```
% the converted data is a vector in 'caldat'
```

```
% 'adcal' has dB values for every possible point that the A/D board
```

```
% can give out (range 1-2048)
```

```
function caldat=datacal(adcal, data)
```

```
numdata=length(data);
```

```
caldat=zeros(numdata,1);
```

```
loop=ones(numdata,1);
```

```
loop=1:1:numdata;
```

```
caldat=adcal(data(loop));
```

Fade Margin Analysis: CDF Calculations For Both Receive Channels

To calculate the CDF of a set of measurements, the following was needed:

- 1) A program to examine the data from each channel, and remove quiescent periods if necessary before calling the procedure outlined in (2)
- 2) A procedure to calculate the pdf of a set of data by using the 'hist' function from Matlab.

The majority of measurements had long quiescent periods with no fading, and it was essential to remove them before CDF's were examined since the measurements were not designed to examine typical traffic patterns, but rather to study the characteristics of fades when they occurred. Most measurements where the receiver was stationary fell into this category. Hallway measurements were an exception, but measurements with omnidirectional antennas on the transmitter were not. The procedure for removing fades was done independently on the two channels. The data was divided into bins ranging from -40 to 7 dB, and the bin with the largest number of data points, as well as the two adjacent bins, was discarded. The number of bins was chosen such that a resolution of 0.1 dB could be obtained, if calibration accuracy in the region of the discarded data allowed this.

Program to Calculate the PDF For Both Channels of One Measurement

```
%function allcdf.m
%if 'shorten' (a manual input to the program) is 1, the largest three bins (the largest
%and the closest non-zero one to either side) are removed from the cdf and
%average value calculations.

%PROCESS CHANNEL 1
disp(['processing channel 1, file ' datafile])
eval(['load ch1cal' datafile]);

if shorten==1 %remove quiescent periods before processing
len=length(caldata1);
binvalues=[-40:2:0]; %get a feel for where the largest value is by dividing the data into
                    %coarse bins
[pdf,binvalues]=hist(caldata1(1:ceil(len/4)),binvalues);
mostval=binvalues(find(pdf==max(pdf))); %get most numerous value (within 2 dB)
```



```

lowval=mostval-4;
highval=mostval+4; %check +- 4dB of the largest value accurately
binvalues=[lowval:0.1:highval];
fprintf ('found that most values in %g range\n', mostval);
[pdf,binvalues]=hist(caldat1,binvalues);
pdf(1)=0; %disregard lowest and highest pts in pdf because if there's
pdf(81)=0; %a lot of points in the file that are much lower or higher
           %than the most common value (ie. an extended fade) they
           %would show up as one value in the extreme range of the pdf
           %and possibly overshadow the most common single value
nummostval=max(pdf); %the number of points in the largest bin
binnum=find(pdf==nummostval); %the bin number of this largest bin
fprintf ('After accurate sampling, max value is at %g\n',binvalues(binnum));

%find the next non-zero bin up
%(i.e. if calibration is in 1 dB steps in this area, the resolution into 0.1 dB bins
%means that 9 bins with no data points in them will exist between the bin with
%the largest number of data points and the adjacent one)
upnum=binnum+1;
while pdf(upnum)==0
upnum=upnum+1;
end %while
upval=binvalues(upnum); %the dB value of this bin

%find the next non-zero bin below
downnum=binnum-1;
while pdf(downnum)==0 %this bin has no points in it
downnum=downnum-1; %so go to the next one down
end %while
downval=binvalues(downnum); %the dB value of this bin

%find out what percentage these points make up of the total file
numupval=pdf(upnum); %number of points in the upper bin
numdownval=pdf(downnum);
totalofthree=numupval+nummostval+numdownval; %total points in the three bins
percentage=totalofthree/len*100;
fprintf ('removing %g points, %g percent from file\n',totalofthree,percentage);
fprintf ('bins being removed are at %g %g %g\n', binvalues(downnum),
binvalues(binnum), binvalues(upnum));

%remove all points from the data file that fall between the upper and
%lower bins above.
shortdat1=caldat1(find(caldat1 >= (upval+0.1) | caldat1 <= (downval-0.1)));
clear caldat1; %no longer needed

```

```

disp ('now getting the average value of the shortened file (linear vals used)')
avgval=10*log10(mean(dbtolin(shortdata1)));
disp ('normalizing the data...')
shortdata1=shortdata1-avgval;

%finally, get the pdf
disp ('Getting the PDF...');
pdf=getpdf(shortdata1);
clear shortdata1;

%IF NO SHORTENING NEEDED:
else
disp ('now getting the average value of the file (linear vals used)')
avgval=10*log10(mean(dbtolin(caldata1)));
disp ('normalizing the data...')
caldata1=caldata1-avgval;

%get the pdf of the normalized file
disp ('Getting the PDF...');
pdf=getpdf(caldata1);
clear caldata1;
end %else if

%save the info for channel 1
disp (['saving the PDF and average value in ch1cdf datafile']);
eval (['save ch1cdf datafile ' avgval shorten pdf]);

%DO CHANNEL 2 (same process as for channel 1)
disp (['processing channel 2, file ' datafile])
.
.
.

```

Procedure to Obtain a PDF from Data

```

%format for use:pdf=getpdf(data)
%where 'data' is a vector of calibrated data points
%PDF is a vector returning the number of points found in each bin
function pdf=getpdf(data);
binvalues=[-40:0.1:7]; %To get max resolution, go in 0.1 dB steps from
    %-40dB to +7dB, since calibration is this exact at
    %high power levels. However, error bar is 1 dB

```

```
%at least (coarsest cal when non-normalized power  
%is less than -30dB)  
[pdf,binvalues]=hist(data,binvalues); %'pdf' now holds # of points in each bin
```

Fade Margin Analysis: CDF Calculations For Selection Combining

The following program calculates the pdf for the combination of the two channels, as well as individual channel pdf's for comparison of gains. Removal of quiescent periods was done beforehand with the procedure used for cross-correlations, since any data that is removed from one channel must be removed from the other for valid comparison of signal strengths at a given point in time.

Program To Calculate PDF's For Selection Combining

```
%Program to calculate the pdf's for both channels as well as the pdf
%for the best combination of the two channels, using selection
%combining
%Note: The pdf's for individual channels are not normalized to allow comparison of
%gains %with the pdf from selection combining.
%'data1' and 'data2' are needed.
%The pdf is sorted into 471 bins, starting at -40 dB and stopping at +7 dB,
%going up in steps of 0.1 dB.

eval(['load corsht' datafile]);
binlevels=[-40:0.1:7];
le=length(data1);
bestpdf=zeros(1,length(binlevels));
ch1pdf=zeros(1,length(binlevels));
ch2pdf=zeros(1,length(binlevels));
bestavg=0;

%get means of both channels
ch1avg=10*log10(mean(dbtolin(data1)));
ch2avg=10*log10(mean(dbtolin(data2)));

%process data vectors into 25000 pt sections
numplits=ceil(le/25000); %includes last part, that's not 25000 pts long
for loop=1:1:numplits
    startpt=(loop-1)*25000+1;
    if loop==numplits %last, incomplete section
        endpt=le;
    else
        endpt=loop*25000;
    end %if
```

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NBS 1010a ANSI/ISO #2 EQUIVALENT**

```

diffvect=data1(startpt:endpt)-data2(startpt:endpt);
bestvect=((abs(diffvect)+diffvect)/2)+data2(startpt:endpt); %abs value gets
                    %rid of negative points, and adding it to the original
                    %vector makes sure that only positive points count
                    %ie. [10 5 -5] +[10 5 5]=[20 10 0] and [20 10 0]/2=
                    %[10 5 0]. Add this to data2
bestavg=bestavg+sum(dbtolin(bestvect));
%get pdf of best combination
[partpdf,binlevels]=hist(bestvect,binlevels);
bestpdf=bestpdf+partpdf;
%get pdf of ch1
[part1,binlevels]=hist(data1(startpt:endpt),binlevels);
ch1pdf=ch1pdf+part1;
%get pdf of ch2
normvect=data2(startpt:endpt)-ch2avg;
[part2,binlevels]=hist(data2(startpt:endpt),binlevels);
ch2pdf=ch2pdf+part2;
end %for
bestavg=10*log10(bestavg/le);

%now save results
eval(['save cdf' datafile ' bestpdf ch1pdf ch2pdf ch1avg ch2avg bestavg']);

```

Cross-Correlation Calculations: All Fades

The code in this section was used to determine the amount of cross-correlation between receive channels according to the equations presented in Chapter 2. As with the analysis for obtaining PDF's, quiescent periods in the channels had to be removed. However, the channels were no longer independent since a valid correlation analysis could only be done if the data on both channels came from the same time period. This meant that quiescent periods were kept on both channels if fading occurred on Channel 1 or channel 2.

Program to Calculate Cross-Correlations For A Measurement

```
%function allcor.m
%if 'shorten' (an input to the program) is 1, the largest three bins
%(the largest and the closest non-zero one to either side) are removed
%from the data before correlation calculations are done
%'data_rate' is the number of samples per second that the raw data for
%ONE channel was taken at (ie. if overall rate =4000, rate per channel=2000)
%it's only used to display the window and section lengths in terms of time
%on the correlation graph
%'slide' is the number of points to slide the window at one time
%'slide' can be a vector. A cross-correlation will be done for each value.
%'window' is the number of points in the sliding window.

if shorten==1 %remove 'flat' data before processing

%PROCESS CHANNEL 1
%load the calibrated data
disp(['processing channel 1, file ' datafile])
eval(['load ch1cal' datafile]);

len=length(caldata1);
binvalues=[-40:2:0]; %get a feel for where the largest value is
[pdf,binvalues]=hist(caldata1(1:ceil(len/4)),binvalues);
mostval=binvalues(find(pdf==max(pdf))); %get most numerous value (within 2 dB)
lowval=mostval-4;
highval=mostval+4; %check +- 4dB of the largest value accurately
binvalues=[lowval:0.1:highval];
fprintf('found that most values in %g range\n', mostval);
[pdf,binvalues]=hist(caldata1,binvalues);
pdf(1)=0; %disregard lowest and highest pts in pdf because if there's
```

```

pdf(81)=0;    %a lot of points in the file that are much lower or higher
              %than the most common value (ie. an extended fade) they
              %would show up as one value in the extreme range of the pdf
              %and possibly overshadow the most common single value
nummostval=max(pdf); %the number of points in the largest bin
binnum=find(pdf==nummostval); %the bin number of this largest bin

%find the next non-zero bin up
upnum=binnum+1;
while pdf(upnum)==0
upnum=upnum+1;
end %while
upval=binvalues(upnum); %the dB value of this bin

%find the next non-zero bin below
downnum=binnum-1;
while pdf(downnum)==0 %this bin has no points in it
downnum=downnum-1;    %so go to the next one down
end %while
downval=binvalues(downnum); %the dB value of this bin

%Find the index into caldat1 of which points to keep
keeppoints1=find(caldat1 >= (upval+0.1) | caldat1 <= (downval-0.1));
clear caldat1; %no longer needed
save cortemp keeppoints1;
clear keeppoints1;

%PROCESS CHANNEL 2 (same process as for channel 1)
disp(['processing channel 2, file ' datafile])
.
.
.
%Create a mask for each channel, so that a logical 'or' function can be performed
mask2=zeros(len,1);
mask2(keeppoints2)=keeppoints2; %set all points that should be kept to
                                %a non-zero value in the mask ie. all points
                                %whose indices appear in 'keeppoints2'
clear keeppoints2;

load cortemp; %get keeppoints1 back
mask1=zeros(len,1);
mask1(keeppoints1)=keeppoints1; %set all points that should be kept to
                                %a non-zero value in the mask ie. all points
                                %whose indices appear in 'keeppoints1'

```



```

clear keeppoints1;
%PERFORM LOGICAL 'OR' BETWEEN THE TWO CHANNELS
endmask=mask1 | mask2;    %if points in either mask are non-zero, they
                        %are marked as '1'. (they should be kept)

clear mask1;
clear mask2;
pointslst=find(endmask==1); %condense the info: pointslst now has overall
                        %indices of points that should be kept, just
                        %as 'keeppoints1' did for ch 1.

clear endmask;
fprintf ('keeping %g points\n',length(pointslst));

%NOW PULL THE CORRECT DATA FROM BOTH CHANNELS
disp(['Shortening channel 1, file ' datafile])
eval(['load ch1cal' datafile]);
data1=caldat1(pointslst); %get values of points that should be kept
clear caldat1;

disp(['Shortening channel 2, file ' datafile])
eval(['load ch2cal' datafile]);
data2=caldat2(pointslst); %get values of points that should be kept
clear caldat2;
clear pointslst;
end %if data needs to be shortened

%IF SHORTEN NOT NEEDED
else
disp(['Loading channel 1, file ' datafile])
eval(['load ch1cal' datafile]);
data1=caldat1; %get values of points that should be kept
clear caldat1;

disp(['Loading channel 2, file ' datafile])
eval(['load ch2cal' datafile]);
data2=caldat2; %get values of points that should be kept
clear caldat2;

end %else if data doesn't have to be shortened (e.g. receiver was moving,
    %and fades were continuous)

%NOW HAVE ALL DATA IN FORM NEEDED FOR GETTING CORRELATIONS
slen=length(slide);
for sloop=1:1:slen %process all cross-correlations for the file

```

```

currentslide=slide(sloop);
disp(['Getting correlations for file ' datafile])
disp(['Getting the correlation for a ' int2str(currentslide) '-point slide']);
dlen=length(data1);

if dlen>80000 %too long-split file into 3
sect=floor(dlen/3);
v1=getcor(data1(1:sect),data2(1:sect),currentslide>window);
startv2=sect-window+currentslide; %overlap some points to get all corrs
v2=getcor(data1(startv2:sect*2),data2(startv2:sect*2),currentslide>window);
startv3=(sect*2)-window+currentslide; %overlap some points to get all corrs
v3=getcor(data1(startv3:dlen),data2(startv3:dlen),currentslide>window);
corvect(1:length(v1))=v1;
corvect((length(v1)+1):(length(v2)+length(v1)))=v2;
corvect((length(v1)+1+length(v2)):(length(v1)+length(v2)+length(v3)))=v3;

else
corvect=getcor(data1,data2,currentslide>window);
end %if-else dlen>80000

%GET THE HISTOGRAM FOR THE CORRELATION
binvalues=[-1:0.05:1];
[corpdf,binvalues]=hist(corvect,binvalues);

%Get the mean value of the correlation
avgval=mean(corvect);

(procedure to save the mean value of the correlation - avgval -, the vector of correlation
values - corvect - and other relevant information to file)

clear corvect;
clear corpdf;

end %for loop over all correlations

```

Procedure to Calculate the Cross-Correlation For a Set of Data From 2 Channels

```

%function corvect=getcor(data1,data2,slide>window)
%A vector of all correlation values is returned
%the vector is of size: floor of (datalength-window)/slide
%approx max data size is 180000 points
%execution time: approx 12000 correlations/minute
%correlations are done using the data converted to linear values

```

%NOTE: The correlation of two vectors where one vector has a variance of
%less than 1e-11 is defined to be zero, since a vector with a variance of
%zero gives an undefined result. Also, the correlation of two vectors
%where each vector's variance is less than 1e-11 is defined to be 1, for
%the same reason.
%The cutoff of 1e-11 was chosen since rounding errors occur with smaller
%numbers that may set a variance of zero to a negative result, causing
%complex and erroneous correlation values to be returned.

```
function corvect=getcor(data1,data2,slide>window)
len=length(data1); %and data2

datmw1=dbtolin(data1); %do correlation using linear values
datmw2=dbtolin(data2);

bigsum1=cumsum(datmw1); %overall, cumulative sum of points in data1
bigsum2=cumsum(datmw2);
bigsum1sum2=cumsum(datmw1.*datmw2);
bigsumsq1=cumsum(datmw1.*datmw1);
bigsumsq2=cumsum(datmw2.*datmw2);

clear datmw1;
clear datmw2;

numcor=floor(len>window)/slide;
corvect=zeros(numcor,1);

%do first slide (since startpt is 0 here)
endpt>window;
sum1=bigsum1(endpt);
sum2=bigsum2(endpt);
sum1sum2=bigsum1sum2(endpt);
sumsq1=bigsumsq1(endpt);
sumsq2=bigsumsq2(endpt);
mean1=sum1>window; %get mean in dB's
mean2=sum2>window;
nmeansq1=sum1*sum1>window;
nmeansq2=sum2*sum2>window;
var1=sumsq1-2*mean1*sum1+nmeansq1;
var2=sumsq2-2*mean2*sum2+nmeansq2;
covar=sum1sum2-mean2*sum1-mean1*sum2>window*mean1*mean2;
if (var1<1e-11 | var2<1e-11) %at least vector's variance is zero
corvect(1)=0;
    if ((var1<1e-11) & (var2<1e-11)) %both variances are zero (quiescent channels)
```

```

        corvect(1)=1;
    end %if
else
    corvect(1)=covar/sqrt(var1*var2); %see Chapter 2 for the equation
end %else if

%do the other correlations
loop=2;
for startpt=slide:slide:(numcor*slide) %starts at second cor value
    endpt=startpt+window;
    sum1=bigsum1(endpt)-bigsum1(startpt);
    sum2=bigsum2(endpt)-bigsum2(startpt);
    sum1sum2=bigsum1sum2(endpt)-bigsum1sum2(startpt);
    sumsq1=bigsumsq1(endpt)-bigsumsq1(startpt);
    sumsq2=bigsumsq2(endpt)-bigsumsq2(startpt);

    mean1=sum1/window; %get mean in dB's
    mean2=sum2/window;

    nmeansq1=sum1*sum1/window;
    nmeansq2=sum2*sum2/window;

    var1=sumsq1-2*mean1*sum1+nmeansq1;
    var2=sumsq2-2*mean2*sum2+nmeansq2;

    covar=sum1sum2-mean2*sum1-mean1*sum2+window*mean1*mean2;
    if (var1<1e-11 | var2<1e-11) %at least vector's variance is zero
        corvect(loop)=0;
        if ((var1<1e-11) & (var2<1e-11)) %both variances are zero (quiescent channels)
            corvect(loop)=1;
        end %if
    else
        corvect(loop)=covar/sqrt(var1*var2); %see Chapter 2 for the equation
    end %else if

    loop=loop+1;
end

```

Cross-Correlation Calculations: Fast Fades Only

To determine the cross-correlations for channels when only fast fades are considered, an extra section of code was added to the cross correlation program that referenced the function presented below to filter slow fades before correlations were calculated. This function examines each data point and normalizes it using the mean power of data points in a 1 sec window around it.

Procedure to Filter Slow Fades

```
%A vector of all normalized values is returned
%the vector is of size: floor of (datalength-window)
%means are done using the data converted to linear values
%window size must be odd
%points are normalized by finding the mean of a window around them and
%dividing the points by that mean value(linear case). ie. the mean in db's
%is subtracted from the point value in dB's.
%NOTE: window must be odd so that it can be distributed evenly around a point
%ie. a window of size 11 means an average over the point and 5 points to
%either side of it
```

```
function nor=normaliz(data>window)
```

```
len=length(data);
```

```
bigsum=cumsum(dbtolin(data)); %do means using linear values. Bigsum has
                                %cumulative sum of points in 'data'
```

```
numnorm=len-window+1;
```

```
nor=ones(numnorm,1)*1000;
```

```
firstdatpt=ceil(window/2); %first point in data file that can be normalized
```

```
                                %(points less than half a window from the start
```

```
                                %and end of the file can't be)
```

```
lastdatpt=len-floor(window/2); %last data point that can be processed
```

```
%do first point (since startpt is 0 here)
```

```
sum=bigsum(window);
```

```
dbmean=10*log10(sum>window); %get mean in dB's
```

```
nor(1)=data(firstdatpt)-dbmean;
```

```
%do other points
lowsum=bigsum(window+1:1:length(bigsum)); %same vector, but displaced so
                                         %that subtracting elements of
                                         %bigsum from lowsum will give the sum
                                         %of points over a certain window size

allmeans=(lowsum-bigsum(1:length(bigsum)-window))/window; %all means except
                                                         %for first point

clear lowsum;
clear bigsum;
allmeans=10*log10(allmeans);
nor(2:numnorm)=(data((firstdatpt+1):lastdatpt)-allmeans);
```

END

2 1 1 1 - 9 6

FIN