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Diversity measurements for indoor radio communications

by
Jean-François Lemieux, B. Eng, RMC

**A thesis submitted to the
Faculty of Graduate Studies and Research
in partial fulfilment 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**

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Abstract

Signal propagation for the indoor radio channel has been shown to suffer from multipath fading and strong attenuation. In this work, a transmitter and diversity receiver were designed and implemented to characterize the indoor radio channel using frequency, space and polarization diversity. Selection combining was used. Statistical analysis of the data gathered was performed and envelope level distributions were calculated. Frequency diversity results showed that maximum improvement occurs with any frequency spacing above 10 MHz. Partial improvement occurs with a spacing of 5 MHz. Space diversity results showed that maximum improvement occurs with any antenna spacing above $3/4$ wavelength. Polarization diversity results showed that partial improvement is achieved, although smaller than that achievable with two uncorrelated channels. Attenuation results were presented in the form of the exponent of the distance-power law: it was between 3.56 and 3.49, in agreement with published results for similar types of building construction.

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Chapter 1

Introduction

1.1 General

Mobile communication is increasingly becoming a part of our society. Paging is a well known and accepted method of reaching somebody on the move. Over the past few years, cellular telephone systems have experienced phenomenal growth and are becoming an indispensable tool for more and more people. Beginning as a vehicle based system, the latest developments in the cellular telephone industry are bringing on the market portable cellular telephones. Research and development specifically on indoor radio propagation and communication systems came from the development of these portable telephones.

Indoor radio communication, sometimes called indoor wireless communication, does not have any major practical applications yet but that time is fast approaching. The cordless telephones presently available are more a gadget than a new communication system. Their restrictions and crude features will prevent them from having any serious future [1], [2]. They are only an early application of a technology still in its infancy. The possible implementations of indoor wireless include narrowband systems for voice or low speed data as with a wireless PEX [3] or for communication and control in factories [4], [5], [6]. They also include wideband systems for high rate data transmission, possibly in a wireless LAN [7].

It is tempting to use the cellular telephone system and its popularity as a possible example for the future of indoor radio communication. Just as cellular radio fulfilled a need for quality, high capacity mobile telephone communication, indoor radio could mature along similar lines to provide portable and flexible communication within a large building or factory. Integrated systems where one portable phone will allow its user to make calls from any office building or public place are not that far fetched [1], [2].

Market penetration of the indoor radio technology will be linked more or less directly to its cost. As any other high technology application, cost reduction will come with mass production of simple and robust instruments. The key to this simplicity is a better knowledge of the channel and its characteristics. All types of radio channel have some common elements but the frequencies and the in-building environment of the indoor radio channel bring a number of unknowns that must be explored.

1.2 Fading channel

The indoor radio channel is not a friendly environment for signal propagation. All kinds of obstacles (people, walls, furniture etc) attenuate and reflect a signal to various degrees. At any given location, the received signal is the sum of many replicas of the transmitted signal that propagated along different paths. These obstacles, the paths and the propagation conditions in general are moving and changing with time. The changes in the channel transfer function have a great influence on communication systems performance. They alter the amplitude and the phase of the received signal. It is convenient although somewhat imprecise to separate these changes in two groups based on their time scale: short-term and long-term fading. Communication links experience variations in performance mostly due to the effects of short term fading. These effects are usually described in terms of signal-to-noise ratio and probability of error. Longer term variations of the channel structure are also taking place. Looking at it from the point of view of statistical communication theory, the short term fading is concerned with the instantaneous values of parameters described by longer term statistics [8]. We will review short term fading and will show which statistical model can be used to explain it.

Figure- 1.1: Vectorial sum of received signals

Short term fading can also be called multipath fading because the energy is arriving at the receiver along several distinct paths. If either the receiver position or the elements affecting those paths undergo changes, it will produce a moving pattern of signal maxima and minima that causes short term fading. The arrival of the signal along these multiple paths is caused by changes in the physical medium that cause scattering and ray splitting and by reflections on various objects. Each path provides a component of the resulting signal with its own amplitude and phase. Even if all components are from the same source, the small variations in path length caused by reflections, scattering and medium changes may amount to large changes in the electrical phase of the received signal. Figure 1.1 shows this concept in vector format. It is easy to see that the resulting vector can have sudden changes in amplitude with small variations in the phase of some components. This is called destructive interference: opposing phase vectors of roughly the same intensity added together can result in a signal of very small magnitude. The phase and attenuation of arriving rays is changed randomly over the different paths.

The effect of multipath fading can be studied mathematically. Using the complex low pass equivalent notation as in [9], the transmitted signal is

$$s(t) = \text{Re} [u(t) e^{j\omega_c t}] \quad (1.1)$$

where $u(t)$ is the complex low pass envelope and ω_c is the angular carrier frequency. Each of the multiple propagation paths has a certain attenuation and a certain phase. The received signal is :

$$x(t) = \sum_n a_n(t) s(t - \tau_n(t)) \quad (1.2)$$

for a discrete numbers of paths and

$$x(t) = \int_{-\infty}^{\infty} a(t; \tau) s(t - \tau) d\tau \quad (1.3)$$

for a continuous number of paths (e.g. HF propagation); where $a_n(t)$ is the attenuation and $\tau_n(t)$ is the delay associated with the n th path. $a(t; \tau)$ is the attenuation of signal components at delay τ and time t . To simplify the expression, assume there is a finite number of paths with a fixed attenuation and delay. This is an idealized static situation where neither the receiver nor any scatterer/reflector are moving and the propagation medium is changing slowly enough to be considered static. With this assumption, 1.2 becomes

$$x(t) = \sum_n a_n s(t - \tau_n) \quad (1.4)$$

or substituting 1.1 in 1.4

$$x(t) = \text{Re} \left[\sum_n a_n u(t - \tau_n) e^{j\omega_c(t - \tau_n)} \right] \quad (1.5)$$

This equation can be rearranged to show the equivalent low pass received signal envelope:

$$x(t) = \text{Re} [r(t) e^{j\omega_c t}] \quad (1.6)$$

where $r(t)$ is:

$$r(t) = \sum_n a_n u(t - \tau_n) e^{-j\omega_c \tau_n} \quad (1.7)$$

or for the continuous case:

$$r(t) = \int_{-\infty}^{\infty} a(\tau) u(t-\tau) e^{-j\omega_c \tau} d\tau \quad (1.8)$$

In the case of an unmodulated carrier at $f = f_c$, $u(t)=1$ for all t and $r(t)$, the signal envelope, is independent of time.

$$r(t) = \sum_n a_n e^{-j\omega_c \tau_n} \quad (1.9)$$

This is the case shown in Figure 1.1 where each path contributes to the resultant signal. If the scatterers are moving and/or if the receiver is moving, the attenuation and delay will be constantly changing. It is reasonable to assume in this case that the attenuation and delay would be changing in a random fashion. Each path component is a random variable. When there are large numbers of paths, the central limit theorem can be applied and $r(t)$ can be modeled as a zero mean complex gaussian random process. The received signal envelope can be expressed as:

$$r(t) = R + jS \quad (1.10)$$

where:

$$\begin{aligned} R &= \sum_n a_n \cos \omega_c \tau_n \quad \text{and} \\ S &= \sum_n a_n \sin \omega_c \tau_n \end{aligned} \quad (1.11)$$

$r(t)$ can be also expressed in the polar form as:

$$r(t) = A(t) e^{(-j\psi(t))} \quad (1.12)$$

with

$$\begin{aligned} A(t) &= \sqrt{R^2 + S^2} \quad \text{the envelope and} \\ \psi(t) &= \tan^{-1} \frac{S}{R} \quad \text{the phase} \end{aligned} \quad (1.13)$$

For a zero mean complex Gaussian process, the joint density function in R and S is a bivariate Gaussian distribution:

$$\begin{aligned} p(R, S) &= \frac{1}{2\pi\sigma^2} \exp\left(-\frac{R^2 + S^2}{2\sigma^2}\right) \\ \sigma^2 &= \frac{1}{2} E\{|r^2|\} \quad \text{mean power in waveform} \end{aligned} \quad (1.14)$$

The joint probability density function (pdf) of the random variables $A(t)$ and $\psi(t)$ (from 1.13) can be obtained by a transformation of 1.14. By integrating the transformed joint pdf over each variable, a pdf for the envelope and a pdf for the phase is obtained:

$$\begin{aligned} p(\psi) &= \frac{1}{2\pi} \quad 0 \leq \psi \leq 2\pi \\ p(A) &= \frac{A}{\sigma^2} \exp\left(\frac{-A^2}{2\sigma^2}\right) \quad 0 \leq A \leq \infty \end{aligned} \quad (1.15)$$

The distribution obtained from the envelope density function is a Rayleigh distribution with 0 mean. The term Rayleigh fading is often attached to those channels or signals displaying this type of envelope behavior. The distribution obtained from the phase density function is a uniform distribution. These two distributions are widely known and discussed in many manuals and published documents. This is the most usual model of envelope and phase distribution in mobile radio. It is a valid model when all path attenuations and delays are zero mean random variables, for small values of envelope and when there are enough scatterers to spread the phase uniformly. When there is a single non-fading component in addition to a number of fading ones, the envelope has a Rice distribution. Its pdf is:

$$\begin{aligned} p(A) &= \frac{A}{\sigma_f^2} \exp\left(-\frac{2\sigma_0^2 + A^2}{2\sigma_f^2}\right) I_0\left(\frac{A\sqrt{2}\sigma_0}{\sigma_f^2}\right) \quad 0 \leq A \leq \infty \\ \sigma_f^2 &= \text{mean power of fading component} \\ \sigma_0^2 &= \text{mean power of non-fading component} \end{aligned} \quad (1.16)$$

The topic of frequency selective fading will now be addressed. Equation 1.8 can be seen as the convolution of $u(t)$ with the equivalent low pass channel impulse response $c(\tau, t)$:

$$c(\tau, t) = a(\tau, t) e^{-j\omega_c \tau} \quad (1.17)$$

From this impulse response, the channel transfer function can be obtained by doing a Fourier transform on $c(\tau, t)$ in the τ variable:

$$C(f, t) = \int_{-\infty}^{\infty} c(\tau, t) e^{-j2\pi f \tau} d\tau \quad (1.18)$$

The autocorrelation function obtained from $C(f,t)$ is called ϕ_C :

$$\phi_C(f_1, f_2; \Delta t) = \frac{1}{2} E \{ C^*(f_1; t) C(f_2; t + \Delta t) \} \quad (1.19)$$

Under the assumption of wide sense stationarity of the channel and that scattering at τ_1 and τ_2 is uncorrelated, ϕ_C is a function only of $\Delta f = f_1 - f_2$ and not of a specific frequency. $\phi_C(\Delta f; \Delta t)$ is the spaced-frequency spaced-time correlation function. From this function, many other functions can be derived.

For $\Delta t=0$, $\phi_C(\Delta f; 0) \equiv \phi_C(\Delta f)$. This autocorrelation in the frequency variable gives a measure of the *coherence bandwidth* of the channel, Δf_c . Practically speaking, it means that two tones spaced by more than Δf_c will be affected differently by the channel. This is the basis of frequency diversity. If Δf_c is smaller than the transmitted signal bandwidth, the received signal will be very distorted: this is frequency selective fading. If it is the other way around, this is frequency non-selective fading or often called flat fading. For $\Delta f=0$, $\phi_C(0; \Delta t) \equiv \phi_C(\Delta t)$. This function of Δt gives a measure of the *coherence time* Δt_c of the channel. The concept is similar to the coherence bandwidth just discussed. This coherence time is an indication that two signals at the same frequency will be affected differently by the channel if transmitted at a time interval $> \Delta t_c$. This is the basis of time diversity.

Since $C(f, t)$ is the Fourier transform of $c(\tau, t)$, the inverse Fourier transform of $\phi_C(f; \Delta t)$ is $\phi_c(\tau; \Delta t)$ which is by definition the autocorrelation of $c(\tau, t)$:

$$\phi_c(\tau; \Delta t) = \frac{1}{2} E \{ c^*(\tau_1; t) c(\tau_2; t + \Delta t) \} \quad (1.20)$$

The inverse Fourier transform of $\phi_C(f; 0)$ is then $\phi_c(\tau; 0)$. This is the average power output of the channel as a function of the delay τ . It is also called the multipath intensity profile of the channel. The multipath spread of the channel T_m is the range of τ for which $\phi_c(\tau)$ is non zero. Figure 1.2 shows a typical profile. This means that for delay $\tau > T_m$, the delayed versions of the transmitted signal are too weak to be of any significance. By way of its inverse Fourier transform relation to $\phi_C(f)$, $\frac{1}{T_m} \approx \Delta f_c$, the coherence bandwidth. T_m is easier to measure than the coherence bandwidth. Saleh

and Valenzuela [10] and Devasirvatham [11] explain the procedure and give results of such measurements.

Figure- 1.2: *multipath intensity profile $\phi_c(\tau)$*

1.3 Diversity

1.3.1 General

Diversity combining has been in use for many years in the world of HF radio, with the first experiments reported in the late twenties [12], [8]. It is based on the fact that if two or more independently fading transmission paths carry the same message, the probability that all paths will experience a deep fade at the same time is lower than for a single path. There are a few methods widely used either to choose the path with the strongest signal or to combine the signals from each path to make a stronger one. The original method to realize independent fading paths consisted in using more than one receiving antenna separated by some distance. The distance between each pair of antennas had to be large enough to insure independent fading. Other methods used to achieve independent fading consisted of transmitting on different frequencies, at different time or with different polarizations. There is no limit on the number of diversity branches

that can be used but the amount of improvement decreases with every additional branch. Performance improvement must then be balanced with receiver complexity.

Space diversity is a simple technique which does not require additional frequency spectrum. A spacing between each pair of antennas of at least $\lambda/2$ [12] should be sufficient. Practically, the required spacing depends of the position of the scatterers causing the multipaths.

Frequency diversity requires more spectrum but uses only one antenna at the receiver. The frequency separation will depend on the coherence bandwidth of the channel.

Polarization diversity takes advantage of the de-polarization of the transmitted signal as it propagates through the channel. A vertically polarized signal leaving the transmitter will have both vertical and horizontal components when it reaches the receiver. The physical process of de-polarization or cross-polarization is very complex and is not within the scope of this work. Polarization diversity brings in focus problems associated with radiation patterns and type of antenna. A dipole antenna has a null along any plane containing the antenna. A vertical dipole has uniform azimuthal directivity. By turning it 90° , there are two directions from which no significant energy will be collected [13].

1.3.2 Combining methods

Some techniques must be used to turn the independent fading in the branches of the receiver into an improvement in performance. Combining is the mechanism used to do this. There are three principal combining methods: selection, maximal ratio and equal gain combining. When using linear detection for demodulation, there is no difference between doing the combining before or after detection. When using non-linear detection (e.g. square law), predetection combining yields a 3 dB gain over post-detection combining [14].

Selection combining consists of selecting the branch with the best signal among all branches. In predetection combining, this branch is used for demodulation. In post-detection combining, the signal from this branch is sent to the receiver output. The algorithm consists of monitoring all branches and picking the strongest. When

the signal of the selected branch falls lower than any other one, a new best signal is selected. At first this method seems very simple but is difficult to implement because the threshold where a decision to switch branch is made is always changing. Reference [12] shows a 10 dB improvement between 1 branch and 2 branches for $P_{12} = 1\%$ (P_{12} is the probability that either signal in branch 1 or 2 is lower than a given level L). For $P_{12} = .1\%$ the improvement is almost 15 dB. A 10 dB improvement means that for the same probability, L is 10 dB higher with 2 branch diversity than with no diversity.

Switched combining is an easier method to implement. The signal from one branch is selected until it becomes lower than a fixed threshold A . When this happens, the next branch is selected. Hopefully its signal is higher than A . The performance improvement from this method is at best equal to selection combining. This happens only close to the threshold level. At all other points, it is worse than selection combining but it is still better than no diversity.

Maximal ratio combining consists of weighting the signal from each branch according to its level and then summing them all. Before summing them, all signals are co-phased. Ideally, the strongest signal, having the greatest weight, contributes the most to the final signal and the weakest signal contributes the least. For $P_{12} = 1\%$ the improvement from 1 to 2 branch is 11.5 dB as shown in reference [12]. For $P_{12} = .1\%$ the improvement is 17 dB. This method requires complex circuitry at the receiver to obtain the correct weighting factors and to phase align all signals.

Equal gain combining is a variation of maximal ratio combining. The signals are co-phased before summing but each one is given the same weight. The performance improvement is generally only a fraction of a dB lower than maximal ratio combining. Taking the limit for an infinite number of branches, the difference is 1.05 dB [12].

The results mentioned earlier were for a 2 branch receiver where the branches were uncorrelated. If they are partially correlated, the performance improvement will not be as high. Ultimately, when correlation is total, there is no improvement. In selection combining, for a 2 branch receiver at $P_{12} = 1\%$, the improvement is 3 dB poorer for an envelope correlation of .8 compared to an envelope correlation of 0. For correlation of

.6 and .3, the improvement is 2 dB and 1 dB poorer respectively. With maximal ratio combining, the improvement is about 1 dB poorer for a correlation of .3, 2 dB for a correlation of .6 and about 4 dB poorer for a correlation of .8.

1.4 Thesis objectives

The objectives of this thesis are threefold:

1. **Design and implementation of an experimental installation** for the characterization of the indoor portable channel using C.W. transmission. The transmitter section should be capable of transmitting one or two tones with variable frequency spacings. Its output power should also be variable. The receiver section is to be designed such that it allows for the individual reception of two channels simultaneously. The receiver antenna system is to be either one monopole, two monopoles or two cross-polarized monopoles. The receiver is to be interfaced to a high speed data acquisition system to allow for the gathering and storage of data from the received channels with minimal intervention.
2. **Performance of a comprehensive set of experiments**, using the installations outlined above, to gather sufficient data to characterize the portable radio channel using diversity reception. The experiments are to be conducted at a set of locations that is representative of the different conditions in such places as an engineering building.
3. **Analysis on a statistical basis** of the large volume of data gathered from such experiments. The analysis will extract important parameters such as signal loss, signal distribution, outage probability and how this is influenced by different kind of diversity and by the level of activity in the locations where the experiments took place.

1.5 Organization

The remainder of this thesis is organized in four chapters.

Chapter 2 will review previous experiments done in the area of indoor radio communication and the use of diversity. The experimental installations, procedures and results obtained in these experiments will be presented. From this review, the initial design parameters of our experiment will be defined. An outline of the experiment will be presented

Chapter 3 will describe the experimental installations built in the course of this thesis, the procedures used to calibrate the receivers and to record the data. Transmitters, receivers and antennas design will be described. The data acquisition system will be covered in detail.

Chapter 4 will describe the statistical analysis carried out on the recorded data. The data processing steps will be explained and the algorithms used will be described. The main type of data analysis done was the computation of average cumulative distribution functions for the signal of each branch and then for a signal obtained by doing selection combining on the recorded data. Limited results (distance-power law) on attenuation versus distance will be presented.

Chapter 5 will conclude by reviewing the material presented, specially the data and the results. The principal findings of this thesis will then be compared to theoretical models and to previously published results. Recommendations for further related research work will be given at the end of the chapter.

Chapter 2

Review of current research

2.1 Introduction

Measurements of indoor radio propagation characteristics can be divided into narrowband measurements and wideband measurements. Narrowband experiments are usually done using continuous wave (C.W.) transmitter to get information on attenuation, coverage and signal envelope distribution. These cumulative distributions, usually done for small-scale areas, are an important tool to predict the error performance of narrowband transmission systems. The attenuation measurements must be done over a larger area to find how the signal is attenuated when propagating within a building. It is often referred to in the literature as the distance–power law. The free space received signal power has an inverse relationship to the square of the distance. Indoor propagation usually has a different exponent: most of the time it is greater e.g. 3, 4, 5 [15], [16] and sometime smaller e.g. 1.5, 1.8 [10], [17].

Wideband measurements are usually done to find out the time delay spread of the building tested. This is a measure of how many replicas of the transmitted signal will be received along the multiple paths of the building. A very short pulse (large bandwidth) is sent at a certain pulse repetition frequency. The receiver output is the summation of all received signal at that time point. A plot of this output vs time might show many distinct pulses or a large smear for every pulse transmitted. This will depend on whether there are some well defined paths or more diffused paths. The time delay

spread is the delay between the arrival of the first and the last pulse. The shortest delay does not always mean the largest amplitude. A few strong signals can add destructively to produce a small output. The delay depends on the path traveled. In simple situations, the received signal can be traced back to a given physical reflector by using the time delay to calculate the distance traveled by the signal. The information extracted from these type of measurements is usually the power delay profile and the root mean square (rms) value of the time delay spread. These parameters have a direct impact on the maximum data rate that could be achieved for a given probability of error. Cumulative distributions of the rms time delay spread are produced. The measurement resolution will depend on the length of the pulse used as a probe: the shorter the pulse the better the resolution. Data on signal attenuation must be recorded in the course of these experiment and is often presented.

Some authors do not directly report experimental measurements but present results from various applications of indoor communication systems [18], [19], [20], [21]. They base their work on experimental data previously reported and usually support it with computer simulations. These systems usually include many of the following aspects: modulation/demodulation, media access scheme, equalization, channel assignment, use of diversity, system availability to the user, error correction code, elimination or diminution of co-channel interference and use of antennas or slotted coaxial cable.

The next section will review currently published results of indoor radio experiments. In section 2.3, we will review the experiments and use the information to design our experiment. The experiment's outline will be described in section 2.4.

2.2 Current results

In the early eighties, Hoffman and Cox [22], Cox, Murray and Norris [23] did some work on radio wave propagation and signal level distribution, both with CW signals. The transmitter for the first experiment was outside, illuminating a small metallic building with the receiver scanning some locations inside. In the second experiment, Cox *et al* were in a suburban area and had the transmitter moved around inside one house and three

small buildings with the receiver located outside some distance away. They provided part of the data to support the concept later put forward by Cox [2] of universal portable radio communication. Both experiments reported that signal envelope variations in a small area closely followed a Rayleigh distribution. Large scale variations of the signal level followed a log-normal distribution once distance attenuation effects were removed.

Alexander's experiments [16], [24] are truly about indoor radio propagation with both transmitter and receiver inside the same building. A CW signal was transmitted from a portable source moved around a number of rooms on the same floor while the receiver was fixed. Signal envelope distributions were also close to a Rayleigh distribution. He also characterized distance-power laws for many types of building materials [15]. Steel partitions gave a distance-power law of 5.7, brick walls gave about 3.9 and plasterboard and open plan gave 2.5 to 2.8.

In continuation of these experiments, Alexander and Motley [25] used a 2-branch space diversity receiver with selection combining to see the possible improvements in two systems under consideration. The first system used two frequencies spaced by 45 MHz to transmit and receive. The second one used burst-mode operation on one frequency where each time period was divided in a transmit and a receive period. The antenna spacing varied from $\lambda/4$ to λ in $\lambda/4$ increments. At the 1% probability point (as explained in section 1.3.2: the probability that the signal in either branch is lower than a given level) an advantage of 10 dB was obtained, well in agreement with the theoretical improvement of selection combining. The variation in antenna spacing made no significant difference in the signal strength distribution.

Patsiokas, Johnson and Dailing [26] did in-building attenuation measurements at 150, 450 and 850 MHz with CW signal. They gave an histogram of signal envelope level for each frequency but no cumulative distribution. They reported attenuation versus distance for the three frequencies in the form of slopes (dB/foot) of their attenuation versus distance graph. At 850 MHz, the slope was .42 dB/foot up to 70 feet then .27 dB/foot. They attribute the change of slope to the slower attenuation of the indirect propagation waves composing the received signal for distances over 70 feet. They

recorded 40 dB fades at 850 MHz and noted that fades were much sharper at 850 than at 450 and 150 MHz.

Devasirvatham conducted time delay spread measurements using a wideband (80 MHz) signal. He reported measurements in a small and in a large building [27]. For all locations in the large building the median rms time delay spread was 120 ns and the worst case was 250 ns. For all locations in the small building the median rms time delay spread was 60 ns and the worst case was 218 ns, a surprising result considering the large difference in building size and geometry. A large hill behind the small building might have increased its effective size as far as those measurements were concerned. As might have been expected, the effect of a good direct path reduced rms time delay spread to a smaller value. The maximum delay was 88 ns in the small building and 70 ns in the large one. This effect was also noted in [11] where he made measurements on an inside to outside path from two houses in a suburban area (same location as in [23]). When there was no direct path from the van (receiver) to the house (transmitter), the maximum time delay spread was 422 ns and the minimum was 220 ns. With a good direct path, the maximum delay was under 100 ns. Attenuation measurements were also made by calculating the total power of each profile. With an 80 MHz wide signal, the measurement gave an average over a large bandwidth. In reference [11] they were in agreement with measurements reported by Cox *et al* [23] considering the large uncertainties of such measurements. It was noted that for longer time delays the signal strength was lower in the small building than in the large one.

Saleh and Valenzuela [10] conducted wideband measurements in a medium size two-story office building. They used 10 ns pulses at 1.5 GHz with a pulse repetition period of 600 ns. Based on these time delay spread and attenuation measurements they proposed a statistical model of the indoor radio channel. By sweeping the RF frequency over 200 MHz instead of keeping it fixed, they were able to get a 5 ns resolution. The distance-power law obtained varied from 1.5 in hallways to 3 in offices. There was one occurrence of a value over 4 in one large open machine shop. The median value for the rms time delay spread was 25 ns with a maximum of about 50 ns. For these calculations

they limited themselves to pulse responses with less than 100 dB of total path loss. The lower pulse responses were discarded. No hallways measurements were used: these situations had line-of-sight rays plus strong rays at long delays. Line-of-sight situations do not really need a model to predict performance: the reception should be very good. Saleh and Valenzuela's rms time delay spread values were much smaller than those obtained by Devasirvatham [27] in the same building. His data included signals of low power level (100–140 dB path loss). Taking a subset of this data where path loss was less than 100 dB, his results were similar.

In the proposed model, the rays of the received signals arrive in clusters. The rays' amplitudes are independent Rayleigh random variables. Their variances decay exponentially with cluster delay as well as with ray delay within a cluster. The phase angles are uniformly distributed. The clusters and the rays within a cluster form Poisson arrival processes with different but fixed rates. The cluster formation depends on building structure. The rays in a cluster are formed by multiple reflections from objects close to the transmitter and receiver.

Bultitude [17] made some CW envelope, phase and attenuation measurements in two types of building. The first one was partitioned in small offices and the second was an open concept building. Temporal fading measurement where the transmitter and the receiver are static, were taken for long periods of time. In the first building the fading intervals were 20 to 40 seconds long with periods of almost constant envelope level in between them. In the second type, the fading intervals were less well defined in time: there was not really any period of constant envelope level between fading intervals. The fading dynamic range was 30 dB in the first building and 17 dB in the second. Both buildings showed wide phase variations.

Cumulative distribution functions of the received signal envelope were computed for the fading intervals only. They followed fairly well a Rician distribution with $K = -6.8$ dB ($K = \text{random power component} / \text{specular power component}$) in the first building. For the second building, a Rician distribution with $K = -11$ dB was a good fit to the experimental curves. Experimental and simulation work directed by Bultitude, reported

in internal reports, showed that the spatial distribution of the received signal with mean variation removed was also Rician with $K = -2\text{dB}$ when the physical environment in the building was static. He also calculated time autocorrelation functions for some fading intervals. They showed that the channel statistics stayed approximately the same for time periods of up to 4 seconds.

Rappaport and McGillem reported narrowband and wideband measurements in 5 factories [4], [5]. Attenuation measurements showed a distance-power law of 1.5 to 2.8 depending on factory topography. For narrowband measurements the cumulative distribution function of the large scale area signal level were well described by a log-normal distribution. Small scale fading range was typically 30–35 dB. The distributions of small scale envelope fading were of three types. For 17 out of 26 *line-of-sight with light clutter* (one of four categories of building used by the authors) measurement locations, the distribution was Rayleigh or Rice with $K < 2\text{ dB}$ (K is defined here as the ratio specular power/random power, the opposite of Bultitude in [17]). The other 9 were distinctly Rician. At 3 out of 8 *obstructed with light clutter* locations and at 9 of 15 *obstructed with heavy clutter* locations, the curve fitted a log-normal distribution. The curve produced using all the data combined (20 000 signal measurements) followed a log-normal distribution below the median value and a Rayleigh distribution above the median value. In temporal fading measurements, the curves were in good agreement with a Rician distribution with $K = 10\text{ dB}$. Fading dynamic range was less than 20 dB.

Wideband measurements taken in the factories were reported by Rappaport in [6]. Overall attenuation measurements gave a power-distance law (2.16) very close to the narrowband measurements. Rms time delay spreads were found to be uncorrelated to the transmitter-receiver distance or the factories topography. They seemed to depend more on the building age, inventory, wall location and ceiling height. The median time delay spread was 96 ns over line-of-sight (LOS) paths and 105 ns over obstructed paths. These time delay spreads are on the order of values reported by Devasirvatham in [11].

Lee and Yeh [28] in the seventies and more recently Cox, Murray, Arnold, Norris and Wazowicz [13] took measurements to collect data on vertically and horizontally polarized

waves and cross-polarization coupling in the 800 MHz range. Cross-polarization coupling is defined in [13] as

$$20 \log \frac{E_x}{E_t} \quad (2.1)$$

where E_x is the average field amplitude of the orthogonal polarization and E_t is the average field amplitude of the transmitted polarization. Although for the mobile radio channel, the results of [28] can be applied to the indoor channel. Both types of polarization were observed to have Rayleigh distributed envelope levels and low correlation ($< .2$) for any antenna spacing. Their local means were within 3 dB of each other. Cross-polarization coupling measurements done on the mobile to base path showed that the orthogonal polarized signal was about 6 dB lower than the original polarization.

Cox, Murray *et al* took cross-polarization coupling measurements inside two laboratories buildings (same building as in [27]) and inside and outside of eight houses with vertical and horizontal dipoles. As mentioned earlier in section 1.3, a horizontal dipole has two nulls in its radiation pattern. Under the assumption of a uniform azimuthal distribution of horizontally polarized components, the power reading from a horizontal dipole would be 3 dB less than the actual value. Under this assumption however, the effect of the two nulls takes less importance: the energy density is the same for all direction in the horizontal plane, so the orientation of the dipole matters less. They were able to crudely verify the validity of the assumption by comparing end-on (antenna pointing towards the transmitter) and broadside (antenna being 90° away from pointing towards the transmitter) signal medians. For a uniform azimuthal distribution, median values should be close together.

Using the dipole for measurements, the definition of cross-polarization coupling now becomes

$$\frac{(L_{h(broadside)} + L_{h(end-on)})}{2} + 3 - L_v \quad (2.2)$$

L_h and L_v are the horizontally and the vertically polarized components. The mid-point between the broadside and end-on value was used as the horizontal component. A

cross-polarization coupling value of 0 dB indicates that the power is equally distributed between the horizontally and the vertically polarized components. A negative value means that the vertically polarized component is stronger. These results will be reliable only if horizontal plane wave propagation is assumed. With their dipole measurements, they cannot calculate the correction for vertical angle of arrival. Random angle of arrival and random polarization would give a cross-polarization coupling value of +3 dB. This led them to believe that positive values could mean that wave propagation is not restricted to the horizontal plane.

In the two buildings the uniform azimuthal distribution assumption was valid. The cross-polarization coupling median was .9 dB and 1.8 dB for the small and the large building respectively which means that significant propagation occurred outside the horizontal plane. The measurements made inside and outside the houses gave median cross-polarization coupling of -2.5 dB and -3.5 dB respectively. The lower value for outside was due to these locations having a more direct path to the transmitter: there were less scatterers to de-polarize the vertically polarized transmitted signal. Cross-polarization coupling measurements were also taken in [23] and were all greater than -10 dB with most above -6 dB.

The radiating elements discussed so far were only single antenna. They are widely used because they are easy to deal with: quick installation and removal, low cost to buy or build and their radiating characteristics are well known. Distributed antennas are another method that could be used [29]. Slotted coaxial cables, also known as leaky feeders, are another option to use in an indoor radio communication system. Motley and Palmer [4] and Reed and Stevenson [30] compared the propagation characteristics of antennas, slotted coax and distributed antennas. They concluded that slotted coaxial cables give better signal coverage and lower floor to floor penetration than a single antenna. Motley and Palmer established the performance cross-over point of the slotted coax versus single antenna at 20 m. In a building with large open areas (> 20 m), single antennas would give a better performance but for a building with many hallways and closed rooms the slotted coax would be the better choice.

2.3 Review of previous experiments

This section will review some installation details of experiments mentioned in the previous section. This was done to help design our experimental installations (detailed in chapter 3). The main points we were looking for in these experiments were: the data sampling rate, the type of data acquisition system, the type of antenna used, the type of detection/demodulation, the parameters measured, the general design of the experiment.

Hoffman and Cox [22] used a corner reflector at the transmitter and a vertical $\lambda/2$ skirted coaxial antenna at the receiver. The receiver was a spectrum analyzer connected to an x-y plotter. The receiver antenna was scanned horizontally in 4 feet square areas at three different heights. Paper charts produced by the plotter were digitized every $.1 \lambda$ and stored on computer for processing. All recordings were scaled in dB relative to a point just outside of the building.

Cox, Murray and Norris [23] were using a modified walkie-talkie transmitting at 815 MHz with a power output of .8 Watt. It had a $\lambda/2$ coaxial sleeve dipole on top of the unit. The unit was hand-held at arm's length 4.5 feet above the floor and scanned along a certain pattern for about 20 seconds. The receiving antenna was a 815 MHz collinear array consisting of 4 dipole elements mounted vertically on top of a 27 feet mast. The receiver was a modified commercial FM communication unit: an 11.7 MHz intermediate frequency (IF) was extracted before the limiter, down-converted to 13 kHz, bandpass filtered (3dB bw=8 kHz) and linearly detected. It had a 45 dB measurement range. Since the transmitter power was fixed, a step attenuator was used to control the RF input signal level. The data acquisition system consisted of a 12 bit resolution digital storage oscilloscope set to record 2048 samples in the 20 seconds period (≈ 100 samples/second). After collection, the data was scaled to convert recorded voltage to dB relative to signal level at a reference location (transmitter 14.2 feet away from the receiver). The scaling took into account the reference location voltage level, the DC offset and the attenuator setting. Scaled data was stored on a desk-top computer. For the cross-polarization coupling measurements, the transmitter was held horizontally and scanned once with the antenna pointing toward the receiver and a second time broadside

to it. These two values were then averaged to get a horizontal power component. Low values of cross-polarization coupling could not be measured accurately due to the uncertainty of the measurement technique: the horizontal antenna was held within 10° of the horizontal plane. This misalignment would, by itself, produce a residual cross-polarization of -15 dB [13].

The method used by Alexander [24] was designed to provide data on signal strength distribution on a per-room basis, without the need for a precise location reporting system. A small hand-held transmitter was used as a mobile source. It was moved around each room, 1 m above the floor, at a slow walking pace to cover as much area as practically possible. The transmitter had a power output of 30 mW and was using a vertically polarized $\lambda/4$ monopole with ground plane. The receiver was stationary and consisted of a CW receiver (dynamic range 60 dB, $\text{bw}=120$ kHz) linked to a computer via an 8 bit analog-to-digital converter. The receive antenna was a vertically polarized $\lambda/2$ dipole, 1.2 m above the floor. Each room was surveyed for 1 minute, with about 10^6 samples taken (sampling rate $\approx 16\,500$ samples/second).

Devasirvatham used a 40 Mbits/s pseudo-noise sequence to modulate (BPSK) a 850 MHz carrier in the transmitter (as used by Cox in the mobile radio environment [31]). The receiver used the same sequence but running 4 kbit/s slower. As this sequence swept past the time smeared received signal, it would produce correlation peaks whenever it became in phase with the transmitted signal (4000 times per second), thus producing the power-delay profile of the received signal. The receiver output was sampled by a computer-based data acquisition system every $10\,\mu\text{s}$, an effective sampling interval of 1 ns [31]. The first 2048 points per recording were stored. The transmitter and receiver antennas were both vertically polarized sleeve dipoles. The transmitter antenna was 1.3 m above the floor. The receiver antenna was on top of a 27 feet mast [11] for indoor to outdoor measurements and 1.8 m above the floor for indoor to indoor measurements [27]. For each transmitter and receiver location, 8 profiles were recorded with the receiver or transmitter placed on 8 positions on the perimeter of a 1.2 m square. All recordings were referenced to the power level obtained when the receiver was 3 m away. The

8 profiles were averaged point by point to form one profile for which the root mean square (rms) time delay spread was calculated.

Bultitude [17] used a synthesizer at 910 MHz and an amplifier connected to a $\lambda/4$ monopole with ground plane as his transmitter. The receiver, a commercial telemetry receiver, down-converted the signal to an IF of 21.4 MHz. This signal was then down-converted to baseband in a quadrature double-balanced mixer circuit. The in-phase and quadrature components were sampled by a 12 bit analog-to-digital converter then stored in a micro-computer. The receiver automatic gain control (AGC) output was also recorded. This output gave a reading on the signal strength to do attenuation measurements. Each synthesizer in the system (receiver and transmitter) was slaved to a single rubidium frequency standard to maintain phase accuracy. The data sampling rate was 10 samples/second.

The transmitter used by Rappaport and McGillem [5] consisted of a pulse generator modulating a CW signal at 1282 MHz. The mixer output was amplified and fed to a discone antenna through a decade attenuator for power control. The receiver was configured to do attenuation and pulse measurements without reconfiguration. The received signal was amplified and passed through a 15 dB directional coupler. One output was sent to an envelope detector while the other was fed to a modified commercial receiver whose AGC was inactivated. The DC level of the receiver's last IF stage was proportional to the CW signal level. A digitizing oscilloscope recorded the detector's output. These recordings were sent to a computer via an HP-IB connection. Path loss, small-scale fading, temporal fading and shadowing measurements were made. The small-scale fading measurements were done by moving the receiver cart along a 1.2 m track, recording 128 IF voltage values in 5 seconds. For each receiver location, 3 runs were made on two paths 1 to 3 m apart. On the first track, two runs were made with antenna height of 2 m and 1.5 m. The third run was made on the second track with an antenna height of 2 m. Temporal fading measurements (receiver and transmitter stationary) were made in areas of significant channel motion. They were made in 100 seconds run and data was recorded at 10.24 samples/second. The dynamic range of the

envelope measurements was limited to the linear scale of the oscilloscope, about 25 dB. The measurement range of the system was 90 dB.

2.4 Experiment outline

Our experimental program is aimed at characterizing the indoor channel using frequency, space and polarization diversity for narrowband communications system. Certain parameters are needed to effectively characterize the channel for diversity. These parameters can be obtained by doing wideband or narrowband measurements. Wideband measurements are more delicate to perform and involve a very large amount of data. The interpretation of this data and the extraction of the parameter needed is always more difficult than with narrowband measurements. Narrowband communications system do not require any phase measurements because for bandwidth < 200 kHz, the information on the phase is not critical to the performance of the system. We will be using tones (very narrow bandwidth) in our experiment and will not make any provision to record any phase information.

The frequency spacing (in frequency diversity) will be within a 25 MHz band (903 to 928 MHz) which is allocated to indoor radio communications. In the space diversity experiment, the antenna separations will remain within a reasonable distance. To keep the conduct of the experiment within a reasonable time frame and to keep the quantity of data to a manageable level, five frequency separations and five antenna separations will be studied.

To cover the frequency band evenly, the following frequency separations will be selected: 1, 5, 10, 15 and 20 MHz. The maximum distance between antennas is limited to 2λ . To avoid the risk of having the two receive antennas coincide with two signal nulls simultaneously, the separations will be: $.75\lambda$, 1λ , 1.5λ , 1.66λ and 2λ .

Signal combining will not be implemented in hardware to keep the receiver hardware simple and to allow flexibility in the choice of combining method. A software algorithm will implement the combining using the recorded data as input. By simply changing the algorithm, many combining methods can be tried. Selection combining was chosen

for analysis purposes because of its good performance and its relative simplicity and practicality.

The experiment area should be as large as possible to allow a good range of LOS and non-LOS locations to be studied. Power limitation of the transmitter will probably be a limiting factor in the actual area covered. Recordings at a large number of locations will improve the statistical validity of our measurements. This will have to be balanced against available storage for the data collected and against processing time. As reported in many of the experiments described in the previous section, floor to floor penetration is poor; therefore our area will be limited to locations on the same floor.

Power levels, attenuation readings and calibration data will be recorded during the measurements to present some results on the attenuation characteristics of the building.

Chapter 3

Description of experimental installations and procedures

3.1 Introduction

The design and building of the experimental installations was a critical part of the experiment. The calibration of the installations and the actual measurement procedure were also an important aspect. This chapter will describe the details of each.

Before finalizing the design of the transmitter and receiver, we felt that some preliminary measurements were needed to get additional knowledge on the channel and the attenuation of the building and to get some practical experience with the equipment. A transmitter and receiver were built to conduct initial measurements. The information obtained from these initial measurements was used to refine our final design.

Although studying three different types of diversity, there were only two basic configurations for the transmitters and receivers: one for frequency diversity and the other for space and polarization diversity. For these two types, the transmitter remains the same and the receivers differed only in the position and orientation of the second antenna. In all cases the experimental installation can be broken down into four building blocks: transmitter, receiver, antenna and data acquisition. They will be studied in this sequence. The data acquisition block was further divided in two sections for clarity of

presentation: a section on the analog-to-digital converter and a section on the Vectra personal computer.

To extract meaningful information from the data produced by the analog-to-digital converter, some scaling had to be done. To achieve this, the system had to be calibrated. The calibration procedure and its results will be presented. Finally, the detailed measurement procedure used to collect the data will be outlined.

The preliminary measurements will be discussed in the next section. Section 3.3, 3.4, 3.5 and 3.6 will cover the transmitters and the receivers. The antenna system will be discussed in section 3.7. The analog-to-digital converter and the Vectra PC will be discussed in section 3.8 and 3.9. The chapter will conclude with a description of the calibration steps and the measurement procedure.

3.2 Preliminary measurements installation

3.2.1 Transmitter and Receiver.

The preliminary measurements were done only for frequency diversity. The transmitter design used here was very similar to the one used for the final measurements. Figure 3.1 shows the block diagram for the transmitter. A 915 MHz local oscillator signal was fed to a double balanced mixer. The IF (intermediary frequency) signal was provided to the mixer at five different frequencies : .5, 2.5, 5.0, 7.5 and 10 MHz. The mixer output was two equal amplitude tones equally spaced (by the IF frequency value) above and below the 915 MHz carrier. This output was amplified and fed to a simple antenna consisting of a 10 cm monopole connected to a female SMA adapter. The transmitter remained static in one room.

Figure- 3.1: Block diagram of preliminary transmitter

The receiver was based on a Hewlett-Packard 8590A spectrum analyzer. Figure 3.2 shows the block diagram for this receiver. The signal from the receive antenna was amplified and then fed to the spectrum analyzer. Through the Hewlett Packard Interface Bus (HP-IB) a HP Vectra PC controlled the functions of the analyzer and recorded the power level of each tone. The receiver was installed on a wheeled cart so it could be moved to various locations.

Figure- 3.2: Block diagram of preliminary receiver

3.2.2 Lessons learned

These measurements were aimed at getting additional knowledge on the channel to help with the final design of the transmitter and receiver and to become familiar with the equipment and its limitations. The spectrum analyzer, with its variable frequency span and large dynamic range, was a good instrument to use in preliminary measurements but it had drawbacks which ruled out its use for more precise measurements.

With the fastest program loop possible, the PC could take one pair of readings per 420 ms (≈ 2.3 samples/second). This was not fast enough to acquire fast fading events. The analyzer's fastest sweep time was 20 ms: this meant there was about 15 to 18 ms between the sampling of the first and the second tone. This does not seem like much but ideally each tone should be sampled simultaneously to see how much they differ at a given instant.

The need for some kind of spatial averaging quickly became evident. With the transmitter and the receiver antenna staying static during recordings, the resulting signal amplitudes would not show much variation. One signal was consistently much higher than the other. Some measurements were made with the receiving antenna being moved by hand in a circular fashion. The majority of the fading events then recorded spanned a dynamic range of about 40 dB. These measurements gave a good indication of the output power needed to cover the area selected for the final measurements.

Some other practical facts became evident during these measurements: SMA rigid or semi-rigid coaxial cables are better suited for use in the 900 MHz frequency range than BNC coaxial cables. They have much better signal attenuation characteristics.

3.3 Frequency diversity transmitter

In its final version, the frequency diversity transmitter was not very different from the one used in the preliminary measurements. Figure 3.3 is a block diagram and plate 1 is a photograph of the transmitter. The two frequencies are produced by the same method: a lower frequency F (.5, 2, 5.0, 7.5 and 10 MHz) is mixed with a carrier to

Figure- 3.3: Block diagram Frequency diversity transmitter

produce two tones spaced by F (1, 5, 10, 15 and 20 MHz), centered around 915 MHz. The actual output frequencies were 914.5 and 915.5 MHz, 912.5 and 917.5 MHz, 910 and 920 MHz, 907.5 and 922.5, 905 and 925 MHz. The output of the mixer was fed to an amplifier through a two stage step attenuator. It provided attenuation in 1 dB steps from 0 to 121 dB. By changing the attenuation at the transmitter, the average power at the receiver could be adjusted to be in the dynamic range of the receiver, no matter where the transmitter was located.

The transmitter power amplifier was in two stages and had a total gain of 41 dB. Two heat sinks were installed on the power amplifier to avoid significant changes in the amplifier gain. From the amplifier, the signal was sent to the transmit antenna 1.6

m above the floor. The maximum output power of the transmitter (attenuator at 0 dB) is listed in table 3.1.

OUTPUT POWER					
Spacing	1 MHz	5 MHz	10 MHz	15 MHz	20 MHz
Power (dBm)	30.4	31.0	30.3	30.5	31.0

Table 3.1 Transmitter maximum output power

The upper part of figure 3.3 shows how the 915 MHz carrier was produced with a low frequency synthesizer. A 76.25 MHz tone (from synth 1) was fed to a 26 dB gain amplifier (amp 1) then to a frequency doubler. Many harmonics of the 76.25 MHz tone were produced at the output of the amplifier. This output was fed to the frequency doubler. At the output of the doubler, all the harmonics, except the one at 915 MHz, were filtered out with a K&L tunable band-pass filter. At the amplifier's output, the lower harmonics were much stronger than the higher ones. Better results were achieved by passing the 457.5 MHz tone (6th harmonics) through a doubler and amplify it rather than trying to directly use the tone at 915 MHz (12th harmonics). This carrier was then amplified to +6 dBm and fed to the mixer. The carrier produced this way was as accurate in frequency as the initial synthesizer (± 10 Hz).

The frequency spectrum (for each frequency spacing) at the transmitter output is shown in figure 3.4 to 3.8. It would not have been practical to try removing the sidebands at the transmitter output. In order to pass the desired tones, the required filter bandwidth would have been so large that many sidebands would get through and the filter would be of minimal use. Since the intermodulation products are much weaker (< -20 dB) than the desired tones, no attempt was made to remove them. The carrier was also low enough that it did not interfere in the measurements.

Figure- 3.4: *Power spectrum of frequency diversity transmitter output, 1 MHz spacing*

Figure- 3.5: *Power spectrum of frequency diversity transmitter output, 5 MHz spacing*

Figure- 3.6: *Power spectrum of frequency diversity transmitter output, 10 MHz spacing*

Figure- 3.7: *Power spectrum of frequency diversity transmitter output, 15 MHz spacing*

Figure- 3.8: *Power spectrum of frequency diversity transmitter output, 20 MHz spacing*

3.4 Space and polarization diversity transmitter

This transmitter was very similar to that used for frequency diversity. Figure 3.9 shows a block diagram of the transmitter. Only one frequency was required so the mixer was removed and the carrier was fed directly to the step attenuator. To compensate for the absence of the mixer's insertion loss, a 10 dB coaxial attenuator was added after the step attenuator. The maximum output power of this transmitter was 33.8 dBm.

Figure- 3.9: Block diagram of the spacel polarization diversity transmitter

3.5 Frequency diversity receiver

The receiver can be broken down in three main parts: the front end, the IF part and the demodulation part. Figure 3.10 shows a block diagram of the receiver and plate 2 shows a photograph of it.

The receiver was set on a wooden cart. The antenna was on a cardboard frame resting on top of the two synthesizers, about 1.75 m above the floor. The front end part consisted of the antenna, a preamplifier, an amplifier and a power divider. The antenna was of the same type as the transmit antenna and will be covered in more detail in section 3.7. The preamplifier and amplifier are combined in one module, the HP

Figure- 3.10: Block diagram of frequency diversity receiver

8447F, for a total gain of 47 dB. The power divider was balanced to ± 1 dB with an insertion loss of 3.1 dB.

The IF portion of the receiver had two identical branches in parallel. Each branch consisted of a mixer, fed from the power divider, an HP 8656B signal synthesizer to drive the mixer, a bandpass crystal filter and an amplifier to boost the down-converted signal. The nominal center frequency of the two filters was set as 70 MHz. The

exact frequency was confirmed by slowly sweeping a signal at the input and noting the frequency where the power output was maximum. The center frequencies of the two filters were off from each other by about 3 kHz and their insertion loss by .6 dB (2.1 and 2.7 dB). The transfer function of the filters, as shown in figure 3.11 and 3.12 is good in the region around the center frequency. It has spurious responses a few hundreds of kHz away which is characteristic of crystal filters.

The synthesizer frequency for each branch was set to down-convert the lower and higher tone to the center frequency of its respective filter. The power spectrum at the output of each filter is shown in figure 3.13 and 3.14 for a spacing of 1 MHz. We can see that the undesired tone and the carrier were not completely blocked out but they were weak enough (< -40 dB and < -55 dB) that their effect was negligible. The situation for the other spacings is similar but the undesired tone and the carrier were even weaker.

The demodulation portion was straightforward. We were interested only in the envelope and not in the phase of the signal. The demodulator used a crystal detector specifically matched to a 50Ω load resistor. In addition, the two detectors were a matched pair to ensure that their response would be as close as possible to each other. The crystal detectors and the 50Ω resistors produced a voltage proportional to the envelope of the input signal. The data acquisition system was connected to these resistors.

Figure- 3.11: *Transfer function of receive crystal filter channel 1*

Figure- 3.12: *Transfer function of receive crystal filter channel 2*

Figure- 3.13: Output of crystal filter, frequency diversity receiver, channel 1

Figure- 3.14: Output of crystal filter, frequency diversity receiver, channel 2

3.6 Space and polarization diversity receiver

The receiver configuration was the same for space and polarization diversity with only the antenna positions differing. The IF and the demodulation sections of this receiver are the same as the frequency diversity receiver but the front end part is different. Figure 3.15 and plate 3 show a block diagram and a photograph of the receiver.

Figure- 3.15: Block diagram of space and polarization diversity receiver

The front end portion differs from the frequency diversity case in that there are two receiving antennas. Each antenna requires a preamplifier and amplifier before going to the IF section. Although nominally the same, it was discovered that the gain of the two amplifiers were different. They were fine tuned by using some coaxial attenuators at the output of the first and second stage of each amplifiers. This was important in order to situate the output of each branch within the range of the analog-to-digital converter and to insure that both branches were at the same level.

The center frequencies of the two filters were about 3 kHz apart. Contrary to the frequency diversity case, the input frequency of each mixer was the same. Only one synthesizer tuned to the half-way point between the center frequencies of each filter and a power divider could have been used to drive both mixers. This would have given a mixer output frequency slightly off from the center frequencies of their respective filters. The use of two synthesizers allowed for more precision.

3.7 Antennas

The three antennas used for the experiments were $\lambda/4$ monopole with ground plane. Two of them were borrowed from the Department of Communication's Communication Research Council (CRC) laboratory in Ottawa. The third one was made locally using the CRC's design. Plate 4 shows a photograph of the antenna. They are made of sheet copper and 1/8" copper tubing. The radiating (vertical) element is soldered to the center pin of a female BNC flange connector. It is electrically isolated from the radial (angled) elements. The four angled radial arms are electrically equivalent to a ground plane. They are soldered to the star shaped base which is grounded to the flange connector. Instead of being at a 90° angle with the radiating element, to be closer to an actual plane, they are angled such as to provide a better horizontal radiation pattern. The styrofoam and cardboard bases provide support and facilitate handling. These bases are fastened to the antennas with a silicone compound.

For consistency, always the same antenna was used at the transmitter. The other two were used with the receiver (only one in the case of frequency diversity). The

space and polarization diversity measurements involved the placing of the two receiver antennas in a specific position in reference to each other. This was accomplished by building two frames holding the antennas in the desired position. The frames could be moved without changing the relative position of the antennas. The frame for space diversity consisted of a piece of wood with a holding handle. Some holes were drilled in the wood at the proper spacing to allow plastic tie-down to hold the antenna bases in place. Plate 5 shows a photograph of the frame with two antennas. The frame for polarization diversity was made using the corner of a cardboard box. This frame was then fastened with plastic tie-down to the previous wooden frame for easier handling. Plate 6 shows a photograph of the cardboard frame.

The radiating element's physical length was calculated based on a frequency of 915 MHz. The electrical length was not adjusted for that frequency. To improve the antenna's performance, the length of the radiating element was adjusted by trial and error with a network analyzer to give a low return loss between 903 and 928 MHz. The return loss is an indication of how much energy is reflected back to the transmitter due to an impedance mismatch between the transmission line and the load (in this case the antenna). Return loss is given as a ratio of the power reflected over the power supplied to the antenna. This was done by looking at the network analyzer, cutting a little piece of the tubing, looking again at the analyzer, etc. When the return loss notch was close to 915 MHz, the return loss at each end of the band was between -15 to -20 dB.

3.8 Analog-to-digital converter

3.8.1 General

The output provided by the envelope detectors had to be recorded and stored for later retrieval and processing. This was the role of the data acquisition system. The output was first digitized with an analog-to-digital converter and, after some transformations, stored in the computer. The DASH-16F, from Metrabyte Corporation is a multifunction high speed analog/digital I/O expansion board designed for the IBM PC and compatible machines. Its analog-to-digital converter was used to digitize the envelope detectors

output. It was installed in an expansion slot in the HP Vectra PC, which was the computer used for data storage and processing. The DASH-16F has many features and options that can be easily tailored to a specific application. We will limit ourselves to cover only the modes used in our experiment. More details on this board are available in Appendix C.

The board's installation into the PC consisted of a few steps. After selecting a suitable I/O base address (hex 300 in our case), the input channel configuration switch was set to 16 single ended channels, the input voltage switch set to unipolar, 1 V. The DMA priority level was set to 1. The interrupt priority level was set at 2. These were proposed as a standard setting.

Utility software was provided with the board and included:

- an assembler language driver subroutine to control the board functions;
- program example of each of the driver modes in BASIC;
- a set-up, a calibration and a test program; and
- various demonstration programs.

3.8.2 Modes of operation of the analog-to-digital converter

The control program uses the driver subroutine with up to 7 parameters: mode, data (array of 5 values) and error flag. The mode parameter indicates which function the subroutine is to perform. The error flag provides some information about the status of the operation just performed. There are 21 possible modes, 0 to 20, that can be used to operate the DASH-16F. We used 6 of these modes. A complete list of modes and their functions is given in Appendix C. Table 3.2 lists, in order of appearance in the data acquisition control program ADSAMP.BAS (listing of this program is in Appendix A), the modes used in the experiment.

DASH-16F MODES	
Mode	Function
Mode 0	initialize board
Mode 1	set multiplexer scan limits
Mode 17	set programmable timer rate
Mode 20	data acquisition
Mode 8	monitor status of data acquisition
Mode 9	data transfer from memory

Table 3.2 List of DASH-16F modes used in the experiment

The modes will now be explained in more detail and the parameters required will be listed. The parameters are MD%, DIO%(0:4) and FLG%. % indicates an integer variable (2 bytes).

Mode 0

It initializes the board. The entry parameters are:

MD%=0, mode number;
 DIO%(0)= I/O address;
 DIO%(1)= interrupt level;
 DIO%(2)=DMA level;
 DIO%(3) and DIO%(4) : not used.

The return parameter is FLAG%, providing a value indicating if the operation was successful or if something has gone wrong. The entry parameters do not change.

Mode 1

This mode sets the scan limits of the input channels' multiplexer to the desired values. There is one analog-to-digital converter module and a maximum of 16 input channels. The multiplexer is used to activate in sequence the input channels on which the next

conversion will be performed. Input channel 0 and 1 were used in our experiment. The entry parameters are:

MD%=1, mode number;
 DIO%(0)= channel 0, lower scan limit;
 DIO%(1)= channel 1, upper scan limit;
 DIO%(2), DIO%(3) and DIO%(4): not used.

On return FLAG% provides a diagnostic of the operation and the entry parameters are not changed.

Mode 17

This mode sets the rate of the internal programmable timer. If the option is selected, the timer will trigger the start of a conversion at the selected interval. The rate is selected by specifying two numbers N1 and N2 that will divide the timer's input clock frequency.

$$Rate = \frac{clock\ freq}{N1 \times N2}$$

The clock is a precision crystal oscillator operating at 1 MHz. N1 and N2 can be between 2 and 65535 and must be integer. This allows rates from less than 1 pulse/hour to the maximum possible by the analog-to-digital converter, about 100,000 samples/s.

On entry the parameters are:

MD%=17, mode number;
 DIO%(0)= N2;
 DIO%(1)= N1;
 DIO%(2),DIO%(3) and DIO%(4) : not used.

On return the parameters do not change.

Mode 20

There were many modes offered for the actual data acquisition. Different modes allowed much flexibility. We selected mode 20 because it fitted best our needs. An explanation on how it was selected can be found in Appendix C. Mode 20 performs C conversions (= 1 scan) per trigger, up to N total conversions, on successive channels. The value of

C depends on the channels specified earlier by mode 1. The trigger is provided by the timer or externally. Data is transferred to memory by interrupt. The number of scan is N/C. The time between a conversion on channel 1 and on channel 2 was is 60 μ s. The calculations to obtain this value are explained in Appendix C. The entry parameters are:

MD%=20, mode number;
 DIO%(0)= total no of conversions, N, 1 to 32 767;
 DIO%(1)= address of memory segment to receive data;
 DIO%(2)= trigger: 1=external or 0= internal timer;
 DIO%(3) = 1; recycle: data continuously written to the same memory segment.
 =0; no recycle: conversions stop when N is reached;
 DIO%(4)=not used.

On return the parameters are not changed and FLAG returns any error codes.

Mode 8

This mode is used to monitor the status of conversions initialized by mode 20. On entry the parameters are:

MD%=8, mode number;

The other parameters are not used. On return the parameters are:

DIO%(0)= operation type, i.e. mode 5, 6,18 or 20;
 DIO%(1)= status of operation: active=1, done=0;
 DIO%(2)= current value count;
 DIO%(3) and DIO%(4) are not used.

Mode 9

This mode is used to transfer data from memory to an array defined in the control program. It transfers the specified number of values from the specified memory segment into a dimensioned array. It is also possible to retrieve data at a given offset from the starting memory address. Each value is an integer i.e. 2 bytes. The analog-to-digital converter produces a 12 bit number. The other 4 bits represent the number of the

originating channel. There is an option to have this number also put into an array. The entry parameters are:

MD%=9, mode number;
 DIO%(0)= number of values to transfer, 1 to 32767;
 DIO%(1)= starting address of memory segment;
 DIO%(2)= offset from beginning of segment, 0 to 32766;
 DIO%(3) = VARPTR(data array), VARPTR is a function pointing to the beginning of the data array;
 DIO%(4)=VARPTR(channel number array).

On return, FLAG% gives the error code and the other parameters are not changed.

3.9 HP Vectra PC

3.9.1 General

The Vectra PC is an IBM AT compatible machine running at a clock frequency of 8 MHz with 640 Kbytes of RAM, a 20 Mbytes hard drive and a 1.2 Mbytes floppy drive. It is also equipped with the Language Processor card, a completely separate processor emulating the HP 200/300 BASIC workstation. The processor, a MC68000, has a separate 3 Mbytes of RAM and run HP BASIC 5.0. It accesses the same disk drives and other peripherals. The card has a built-in HP Interface Bus (HP-IB) connector. This is the HP version of the IEEE-488 bus. Having used the Language Processor in the preliminary measurements and having gained experience with it, it was decided to use it for data storage and processing. Unfortunately there were not many communication links between the Language Processor (referred from now on as the HP BASIC environment) and the AT compatible processor (the DOS environment). Once in the HP BASIC environment, there was no direct access possible to any expansion boards in the DOS environment. This precluded using HP BASIC to control the analog-to-digital converter board. GW-BASIC was used to do this function. This version of BASIC was not

nearly as powerful as HP BASIC. Section 3.9.2 will explain the data acquisition control program.

3.9.2 Data acquisition and conversion programs

The data acquisition control program was run in GW-BASIC. Using the machine language driver subroutine the measurements were taken and the data stored on disk in ASCII files. This format was selected because it was the simplest way to store the data files. These files were not very practical in term of disk storage and access time but they were the easiest type of files to import into HP BASIC. Some utility programs were provided with the Vectra and one of them converted ASCII files produced in the DOS environment to HP BASIC ASCII files. The data acquisition program is called ADSAMP.BAS and is listed in appendix A. Before starting the measurements some decisions had to be made on sampling rate and the length of the sampling interval. The product of these two values was the number of samplings, referred earlier as N in the description of mode 20.

In the review of previous experiments done in chapter 2, sampling rates for narrow-band system experiments varied from about 10 samples/s [17], [4] to 16,000 samples/s [24]. In our case the measurements would be taken during the day when people are moving in the building. A more significant factor would be the fact that the receiver antenna would be moving to get some spatial averaging. The actual measurement procedure will be discussed in section 3.11. This dictated a rate that should be faster than 10 samples/s but how fast did it need to be? A test run was made at 2000 samples/s for 5 to 10 seconds. The output of the envelope detector was plotted against time, using all the samples. The plot was then redone using every second point, every fourth point and every eight point. This was equivalent to using a rate of 1000, 500 and 250 samples/s. There were no noticeable difference using 1000 samples/s. The 500 samples/s plot was very close to the original one. The last plot showed a large difference with the first one. A 500 samples/s rate was fast enough to satisfy our requirement.

The duration of the sampling interval had to be long enough to reflect the status of the channel at that time. An interval too short might reflect only a transition of the channel and would diminish the statistical value of the measurements. 10 000 readings would give a minimum probability of 10^{-4} . Twice that amount would give a minimum probability of 5×10^{-5} . The file size was also a factor to keep in mind so that file management and storage would not be too complex. In double precision arithmetic, $2 \times 30\,000$ values was about the maximum practical number of values that could be processed at the same time.

With a total number of samples of 60 000 (30 000 per channel), a sampling rate of 1000 samples/s would give a 30 seconds sampling interval. At 500 samples/s, the interval would be 60 seconds. It was estimated that there would be enough fading events in 1 minute to make the measurements worthwhile.

Before explaining the data acquisition program, we will deal with some of the limitations of the driver subroutine and GW-BASIC. The maximum number of conversions (referred earlier as N) allowed by the driver subroutine in one run was 32 767. The 60 000 conversions were taken in 2 blocks of 30 000 samples by using a loop repeated twice. There was a gap of 10 to 20 ms between the each iteration of the loop.

GW-BASIC has a total work space of 64 Kb. There had to be a trade-off between program space and space allocated for arrays. A 25 000 integer array was about the maximum array size that could be included in the workspace with the data acquisition program. The 60 000 readings were therefore handled in blocks of 15 000 values.

The data acquisition program ADSAMP.BAS using the driver subroutine took 60 000 readings in 60 seconds and wrote the data on disk. The listing is included in appendix A. The program structure is listed in the following statements.

1. the machine language driver is loaded in memory at address hex 8000;
2. program parameters are initialized;
3. board is initialized using *mode 0* ;
4. channel scan limits are set using *mode 1*;
5. programmable timer rate is set using *mode 17*;

6. start first 30 000 conversions using *mode 20*. data to be written in memory starting at address hex 3000;
7. start second 30 000 conversions using *mode 20*. data to be written in memory starting at address hex 4000;
8. transfer first block of 15 000 values from memory to array in program using *mode 9*. write data from array to disk;
9. repeat with second block;
10. repeat with third block;
11. repeat with fourth block.

3.10 Calibration

The raw data produced by the analog-to-digital converter did not have much significance by itself. It needed some sort of scaling to represent something meaningful. It was decided to relate the values produced by the analog-to-digital converter to the power received at the antenna. The power input at the receiver's front end amplifier is a known parameter in many systems and it means something tangible: how much energy was picked up by the antenna.

To do this calibration effectively the input power into the receiver's amplifier had to be measured and controlled. It was not practical to try using the transmitter as the source due to the variations that would be introduced, even by transmitting over a short distance. A signal synthesizer directly connected to the receive amplifier was used instead. Its output power was easily controlled by changing the front panel settings. For each power level of the synthesizer, 200 readings from the analog-to-digital converter were averaged. After each set of reading, the synthesizer power level was incremented by 1 dBm. That process was repeated on both channels and for each frequency. This produced six sets of measurements, five for frequency diversity and one for space and polarization diversity. These were the backbone of the scaling files. Figure 3.16 shows an example of these values for frequency diversity and figure 3.17 shows the values for space and polarization diversity (combined because they use the same transmitter).

Figure- 3.16: Calibration curve, 1 MHz spacing frequency diversity

Figure- 3.17: Calibration curve, space and polarization diversity

The analog-to-digital converter readings went from 0 to 4095. The measurements explained earlier covered this range with power levels from about -83 dBm to -33 dBm. The aim of the calibration was to generate a dBm value for each analog-to-digital converter reading. Taking the values already measured as reference, we used linear interpolation to generate a dBm value for all remaining analog-to-digital converter readings. That was done for all six sets of measurements and the results were stored in six files. Each file consisted of 4096 dBm values for channel 1 and 4096 more for channel 2. Once this was done for dBm values, they were converted in milliwatt and stored these results in six other files.

During data processing, the first operation performed was to read the proper scaling file and to replace each raw analog-to-digital converter reading by the corresponding power (dBm or mW) value. This could be done quickly and allowed the data files to be stored in the raw form, as integer numbers using two bytes per value instead of in the scaled form as real numbers using 8 bytes per value.

3.11 Measurements procedure

3.11.1 General

Before taking measurements, 20 locations were selected on the same floor around the laboratory. We were on the top floor of the School of Architecture building at Carleton University, Ottawa, Ontario. The other rooms on this floor consisted of two drafting rooms, one other lab, a computer terminals room, five study rooms for students, a number of offices and two mechanical rooms. Since this floor was an addition on top of an existing building, the two mechanical rooms contained ventilation equipment that was originally on the roof. The drafting rooms were furnished with about 40 wooden drafting tables. The two laboratories had wooden cabinets and tables but only a small quantity of equipment. There were six computer terminals in the terminal room. The offices and the study rooms had wood and metal furniture. The hallways were laid out in a H form. In the center of that H was a large skylight going right through to the floors below. The windows in the lab, the skylight, in some study rooms and on some outside windows

had wire mesh in their center. The walls were mainly made of plasterboard with wood studs and some metal pillars. The floors were made of cement covered with tiles. The ceilings were all cement with soundproofing panel held by a metal frame about .5 meter from the ceiling. In this inter-space, many cables were routed. Appendix E shows a floor plan of the area and a table of locations and distances. The conditions (rooms full or empty, hall busy...) at the time of measurements are also listed in appendix E. The receiver was located in one of the labs. The transmitter was moved to various locations on the floor. At each location, the exact antenna location was recorded by measuring the distance from the antenna to an easily recognizable building feature. With these measurements, the transmitter could be brought again to the exact same location for the other diversity measurements. This location was reported on a scaled floor plan and the exact distance from the transmitter to the receiver was calculated.

3.11.2 Procedures

After bringing the transmitter to the desired location, the antenna position was recorded. This was usually done by aligning it with the floor tiles (12 " by 12 ") and counting the number of tiles to the next wall corner or angle. After a warm-up period the transmitter attenuation would be set to a given level. A short test run was done to confirm the attenuator setting. If the signal level was not fully within the receiver's dynamic range, the attenuator would be adjusted and another test run would be done. The conditions around the transmitter were recorded (full, half full or empty room etc). Using the frame shown in plate 7 as a guide, the receive antenna was moved by hand along the circumference of a circle cut inside the frame. The circle had a diameter of 1λ for the lowest used frequency (in our case 905 MHz, $\lambda=33$ cm). The antenna was moved along that circle for two reasons. First it was to get a average received power that would be a local average and second, to create fading events. This occurred for 1 minute then the program wrote the data on disk while the antenna spacing or the frequency settings were adjusted. The five sets of measurements for a given type of diversity (one set in the case of polarization diversity) were done within 30 minutes of each other at one location without moving the transmitter.

The frequency diversity measurements were made first. The transmitter and receiver were then modified to conduct the space diversity measurements. The same sequence was repeated at all the locations. The antennas at the receiver were then adjusted for 90° polarization separation and again the same sequence of events was repeated.

Chapter 4

Analysis and results

4.1 Introduction

The analysis of the data collected was done keeping in mind the purpose of the experiments. We wanted to establish the effect of increased spacing on diversity improvement in indoor radio communication. Data was recorded for 20 transmitter locations for each type of diversity. For frequency and space diversity, 5 frequency and antenna spacings were tested. There was only one configuration for polarization diversity: 90° separation. Each data file was processed to produce a cumulative distribution of the envelope level for channel 1 and 2 separately and for a set of points which were the result of selection combining done on both channels. Average path loss was calculated for each location using the average received power values, the transmitter power output and the attenuator setting at the transmitter. With this information, scatter diagrams were produced and the distance–power law for each set of measurements (total of 11) was calculated.

The next section will explain the steps required to produce the distributions and the attenuation scatter diagrams. Section 4.3 will present tables of improvements obtained with diversity. It will also show samples of the received signal envelope and the various cumulative distributions. Section 4.4 will show the attenuation scatter diagrams and present the table of distance–power laws. Section 4.5 will summarize the results obtained in the previous sections.

4.2 Analysis

4.2.1 General

The initial data recorded was ASCII files produced by the sampling program ADSAMP.BAS (listing in Appendix A) in GW-BASIC. The files contained some miscellaneous information and 30 000 pairs of readings from the analog-to-digital converter. Since they were read in sequence channel 1 then channel 2, they were stored in the same way. Before processing the files in the HP BASIC environment, a 2 step conversion was performed. The files were initially stored as ASCII files to make storage simpler and to facilitate their conversion to HP BASIC format. A utility program provided with HP BASIC was used to convert the files to HP BASIC ASCII format. This was the first step of the conversion. The HP BASIC ASCII files were also very cumbersome: not only did they use 1 byte per digit (up to 4 digits per value) but each value had a 4 byte header. This was a waste of disk space and very slow to access. The second step of the conversion was to change the files from ASCII to binary format. A program read the HP BASIC ASCII file, re-ordered the data to bring each channel's data in sequence (instead of having them alternating) and re-wrote the data on floppy disk as HP BASIC binary files. Each value was now stored as an integer at 2 bytes per value. Appendix B describes the structure of the binary data files.

4.2.2 Processing for cumulative distribution

The first calculation task was to obtain the average received power of each channel for all files. The steps to do that were to read in the raw data, read in the proper scaling file, scale the data and calculate the average. At this point, a results file was created for each data file and the average received power was stored in it. Appendix B describes the structure of the results files. In order to have an arithmetic average and not a geometric average, the scaling files used were the ones in milliwatts, as discussed previously in section 3.10

The next step was the calculation of cumulative distribution for each channel individually and for a set of points obtained by doing selection combining with both

channels. Calculating the distributions involved three operations which will be listed here and described later.

1. establishing sub-range boundaries for the histogram;
2. producing the histogram by finding in which sub-range each data point will be counted in; and
3. normalizing the histogram and summing its values to obtain the cumulative distribution.

Diversity will bring the greatest improvement when the two channels have the same average received power.

Histogram boundaries.

The histogram boundaries (or thresholds) were based on the receiver dynamic range. It was a little above 40 dB so it was decided to partition this range into sub-ranges of 1 dBm each from 30 dBm below average power to 10 dBm above average power. These 41 thresholds were calculated for each channel based on its average power value. 41 thresholds gave us 40 sub-ranges of 1 dBm plus one that included all events below the first threshold (average power - 30 dBm) and another one that included all events above the last threshold (average power +10 dBm), for a total of 42 sub-ranges. The thresholds values were placed in two 41 value vectors, X1 and X2. X1 for example, was:

$$X1(1) = \text{average power of channel 1} - 30 \text{ dBm}$$

$$X1(41) = \text{average power of channel 1} + 10 \text{ dBm}$$

Associated with the 42 sub-ranges was a counting vector of 42 values. There was one vector per channel plus one for the selection combining. They were called Pd1, Pd2 and Pdjt. Pd1 for example, was:

$$Pd1(1) = \text{counter for first sub-range (all values} \leq X1(1))$$

$$Pd1(41) = \text{counter for second last sub-range (all values} > X1(40) \text{ and} \leq X1(41))$$

$$Pd1(42) = \text{counter for last sub-range (all values} > X1(41))$$

Producing the histogram.

This step took the most time. Two similar subroutines were used to do the histograms. One subroutine carried out the computations for channels 1 and 2 while the other was intended for the computations related to the selection combining. The algorithm was designed to make maximum use of the built-in matrix and vector manipulation commands of HP BASIC. P1(1:30 000) and P2(1:30 000) contained the power values (scaled) of channel 1 and 2. The algorithm consisted of taking each value of P1 and searching for which sub-range it belonged by comparing it in sequence to each threshold value in X1. The sub-range was found at the first instance where the threshold value was greater than the P1 value. Once found, the cell (in the counting vector Pd1) associated with that threshold value (i.e. Pd1(3) and X1(3)) is incremented by one. If the value of P1 is not smaller than any threshold values, it belongs to the last sub-range. So as not to waste time during each search, the search was started from St1, the first threshold $\geq$ the minimum value of P1.

M1 ← Minimum of P1

Search X1 for first value $\geq$ M1; St1 ← index X1

Search X1, starting at St1, for value $\geq$ P1(I); D1 ← index of X1

increment Pd1(D1)

if X1(41) not $\geq$ P1(I) then increment Pd1(42)

start over with next I

The same process was repeated for channel 2.

Selection of the best of both channels was done at each time point. The aim of the search was to find the threshold which was greater or equal to the value in both channels at this time point. The sub-range associated with that threshold had its cell incremented in the counting vector. The absolute power values were not used to do the selection. Instead the difference between a value and the average power for that channel was used. The selected channel was the one whose value for that time point was the highest compared to its average value. This meant the indexes of vector X1, X2 (not the values in the vectors) were used as a selection factor. The minimum values of P1 and P2 and St1, St2 were found earlier. The selected channel's value will always be

greater than St , the largest of $St1$ and $St2$. The first part of the search is the same as for the individual channels. We started by searching $X1$ from St to find the first value $\geq P1$. The index $D1$ for that sub-range was used as the starting index of a loop comparing $P2$ with the value of $X2$. When the threshold in $X2$ became $\geq P2$, the appropriate counter was incremented. This counter was $Pdjt()$, for the joint distribution.

```

search  $X1$ , starting at  $St$ , for value  $\geq P1(I)$ ;  $D1 \leftarrow$  index of  $X1$ 
for  $J = D1$  to 41
  find if  $P2 \leq X2(J)$ 
  if yes,  $Pdjt(J)$  is incremented; start over with next  $I$ 
  if no,  $J$  incremented, comparison redone
  if  $X2(41)$  not  $\geq P2(J)$  then  $Pdjt(42)$  is incremented
  start over with next  $I$ 

```

Summing the histogram

With the histogram values in $Pd1(1:42)$, $Pd2(1:42)$ and $Pdjt(1:42)$, these numbers were normalized by dividing by 30 000, the total number of points. In these individual distributions, the minimum probability that could be measured will $1/30\ 000$ or 3.33×10^{-5} . The cumulative distribution was done by adding the next Pd value to the previous cumulative sum.

$$Cd1(1) = Pd1(1)$$

$$Cd1(I) \leftarrow Cd1(I-1) + Pd1(I)$$

The three $Cd(\dots)$ vectors were stored in the results files created earlier.

Once these three steps were completed for all files, the distributions for each type of diversity were read into three $20 \times 5 \times 42$ matrices (location, spacing, number of points), for channel 1, channel 2 and the joint distribution. The matrix was $20 \times 1 \times 42$ for polarization diversity. For each type of diversity, a new file containing the three matrices was created. The distributions were now stored together by diversity type. Each spacing was then averaged over all locations: the 20 sets of 42 points were averaged together to give one distribution (42 points) representing all locations for that spacing. This

operation was done on each matrix. In addition to the average, the maximum value and the minimum value of each point in the distribution were recorded. All these values were stored in a second set of matrix $5 \times 42 \times 3$ (spacing, number of points, average/max/min value). The minimum probability of the averaged distribution could be lower than 3.33×10^{-5} because each point is added up with the other 19 and then the sum is divided by 20.

4.2.3 Channel 1 correction

After calculating the cumulative distribution function for all files and averaging them for all locations, it became evident that something was wrong with all the channel 1's distributions. They were supposed to be more or less the same as channel 2's distributions since they were taken at the same time and in the same conditions but they were diverging from the channel 2's distributions, especially at low power values. It was as if the analog-to-digital converter would never give a low power reading even if the signal was very weak. A DC offset present during the measurements would have this effect: the constant voltage would cause the analog-to-digital converter to always give a reading even when the signal level we being measured was very low.

In order to correct this problem, a numerical solution was tested and proved to be satisfactory in correcting the problem. A correction factor was subtracted from all channel 1's readings. This factor was such that it brought the lowest value in channel 1 in line with the lowest in channel 2. The histogram for channel 1 and for the selection combining points were re-calculated. The corrected distributions for channel 1 became much closer to channel 2's distributions. Figure 4.6 to 4.16 show all the distributions.

4.2.4 Processing for attenuation results

The individual results files mentioned earlier contained the average power of channel 1, channel 2 and the corresponding transmitter attenuator value. As for the cumulative distributions, these values were collected in one matrix per diversity type. The matrices ($20 \times 5 \times 3$) contained the above mentioned values for each location and spacing. The transmitter output power was then recorded for both transmitter configurations at different

attenuator settings. For the space/polarization diversity configuration, there was only one tone at the transmitter's output: the measurements were done directly with a power meter. For the frequency diversity configuration, at least three tones were present at the transmitter output: to get an accurate figure of each tone's power the measurements were taken with a spectrum analyzer. Under computer control, the spectrum analyzer took 80 power readings of each tone. The readings were averaged to produce the power output. These power values were stored in a vector using the attenuator value as the index.

With all the above values available, the average path attenuation was calculated for both channels for all 20 locations and 5 spacings. Using the attenuator value as an index, the average path loss was computed simply by doing the subtraction

$$\text{Path loss} = \text{Tx power (attenuator)} - \text{Average received power}$$

To get the transmitter-receiver separation for all locations, a scaled floor plan was used (1/8"=1'). During measurements, the transmitter location had been recorded precisely by measuring distances from the antenna to an obvious building feature. These distances were reported on the floor plan, the locations marked and the transmitter-receiver separation measured. To get the distance-power law, the first step was to plot relative attenuation against distance (on a log scale). A plot (or scatter diagram) was produced for each spacing (total of 11). Three of these scatter diagrams are shown in section 4.4. The absolute path attenuation calculated above was scaled to a relative attenuation using as the reference the lowest attenuation value (from the location closest to the receiver, 1.72 m). The scatter diagrams obtained were fitted using a log fit. Channel 1 and channel 2 were fitted separately. The gradient of these fitted curves was the value representing the exponent of the distance-power law. The calculation to obtain the gradient were as follows:

$$\text{gradient} = \frac{\Delta \text{Att (dB)}}{\Delta \text{Dist (dB)}}$$

where ΔDist in dB is $10 \log(\text{Dist2}) - 10 \log(\text{Dist1})$. Dist2 and Dist1 are two arbitrary distances used to calculate the x axis variation. ΔAtt is the corresponding attenuation variation.

4.3 Diversity systems measurements

4.3.1 General

To calculate the diversity improvement, a plot of the averaged distribution for channel 1, channel 2 and the selection combining distribution was produced for each separation. Figure 4.6 to 4.16 show the plots for frequency diversity separations of 1 MHz to 20 MHz, space diversity separations of $.75\lambda$ to 2λ and polarization diversity. Two other curves were plotted, for comparison purposes only. The first curve is a Rayleigh cumulative distribution (identified as *1 Rayleigh* in figure 4.6 to 4.16). The second curve (identified as *2 independ. Rayleigh* in figure 4.6 to 4.16) represents the distribution resulting from selection combining with 2 independent Rayleigh channels. The Rayleigh distribution was obtained by calculating the cumulative distribution function

$$P(x < A) = 1 - e^{-10^{\frac{A}{10}}} \quad (4.2)$$

for A values ranging from -31 to +11 dB. A is the value of a given threshold relative to the average power. The *2 independ. Rayleigh* curve was obtained by squaring the Rayleigh distribution. With two independent, uncorrelated channels the probability becomes $P(A, B) = P(A) \times P(B)$. By presenting these two curves, we are not implying that channel 1 and channel 2 should be Rayleigh or that the selection combining curve should be close to the other curve. We want to use them to show the maximum diversity improvement that could be obtained when two channels had a correlation of exactly 0.

The improvement calculations were done at the .1% (10^{-3}), .5% (5×10^{-3}), 1% (10^{-2}) and 5% (5×10^{-2}) probability level, that is to say 10^{-3} is the probability that the received signal is lower than a given value below average power. The improvement values were obtained graphically by measuring the relative improvement between the two sets of curves. This was done by reporting on the x axis the position of each curve (the individual channels and of the joint channel) at the four probability levels. We used a feature of the graphic software allowing the tracking and reporting of the x-y coordinates

of a mouse controlled marker. The mid-point between channel 1 and 2 was used to compute the improvement value.

The following five figures, 4.1 to 4.5, show a sample of the envelope detector's output for each type of diversity and for various spacings. We can deduce intuitively by looking at the fades in figure 4.1 and in figure 4.2 that the diversity advantage could be greater at 20 MHz than at 1 MHz.

Figure- 4.1: *Envelope detector output, frequency diversity, 1 MHz*

Figure- 4.2: Envelope detector output, frequency diversity, 20 MHz spacing

Figure- 4.3: Envelope detector output, space diversity, .75 λ spacing

Figure- 4.4: *Envelope detector output, space diversity, 2 λ spacing*

Figure- 4.5: *Envelope detector output, polarization diversity*

4.3.2 Frequency diversity

The graphs of average cumulative distribution functions for frequency diversity cases are shown in figures 4.6 to 4.10. From table 4.1 we can see that the improvement is monotonically increasing with frequency separation. A three step progression can be seen in the improvement values: 1 MHz, 5 MHz and 10 to 20 MHz. This progression is well defined at the 5%, 1% and .5% level. At the .1% level, the progression is not as well defined: the 5 and 20 MHz improvements are almost the same. The range of improvement values for the 10 to 20 MHz spacings is very tight at the 5%, 1% and .5% level: within .7, 1.1 and 1.1 dB of each other and within .6, .9 and 1.7 dB of the Rayleigh value respectively. At the .1% level this range is more than twice these numbers. The larger spread of improvement values at the .1% level may be due to the fact that computations near the lower end of the curves are based on fewer samples and therefore do not present accurate statistics of the received signal.

FREQUENCY IMPROVEMENT TABLE				
	5 % Level	1 % Level	.5% Level	.1% Level
1 MHz	2.7	3.9	4.8	7.4
5 MHz	5.5	7.4	8.7	11.8
10 MHz	6.3	9.3	10.2	13.4
15 MHz	7.0	10.4	11.2	13.6
20 MHz	6.3	10.0	10.1	11.3
Rayleigh	6.9	10.2	11.8	15
	all values in dB			

Table 4.1 Frequency diversity improvement values

The improvements calculated between the Rayleigh and the theoretical combining distribution were 6.9 dB at 5%, 10.2 dB at 1%, 11.8 dB at .5% and 15 dB at .1%.

Reference [12] shows a channel availability curve (this curve is complementary to ours: i.e. our 1% level= their 99% level) for selection combining using a different X axis variable: the $(S+N)/N$ ratio is used instead of absolute power. The diversity

improvements obtained from their curve is about 10 dB at the 1% level and 14 dB at the .1% level. These numbers are close to ours for the same levels.

Figure- 4.6: Cumulative distribution function for 1 MHz spacing

Figure- 4.7: Cumulative distribution function for 5 MHz spacing

Figure- 4.8: Cumulative distribution function for 10 MHz spacing

Figure- 4.9: Cumulative distribution function for 15 MHz spacing

Figure- 4.10: Cumulative distribution function for 20 MHz spacing

4.3.3 Space diversity

The graphs of cumulative distribution functions for the space diversity cases are shown in figures 4.11 to 4.15. The diversity improvement values for the five spacings are displayed in table 4.2. By examining the table, we can see that, contrary to the frequency diversity case, the improvement is not increasing with separation. At the 5%, 1% and .5% probability level, all the improvement values are basically the same. Their range is again very tight: within 1.2, 1.2 and .8 dB respectively from each other and within 1.1, .9 and .8 from the Rayleigh value. The range of improvement values for .1% is again much larger. The explanation to this is also the same: the small number of samples reducing their value as accurate representation of the the received signal's statistics. These results indicate that there is no advantage to separate the antennas by more than $.75 \lambda$. This is in agreement with theory [12] and previous experiments [25].

SPACE IMPROVEMENT TABLE				
wave length	5% Level	1% Level	.5% Level	.1% Level
.75	7.0	10.5	11.8	12.5
1	7.1	10.1	11.6	16.3
1.5	6.4	9.9	11.3	15.7
1.66	6.0	9.5	11.0	14.8
2	5.8	9.3	11.0	16.6
Rayleigh	6.9	10.2	11.8	15
	all values in dB			

Table 4.2 Space diversity improvement values

Figure- 4.11: Cumulative distribution function for $.75 \lambda$ spacing

Figure- 4.12: Cumulative distribution function for 1λ spacing

Figure- 4.13: Cumulative distribution function for 1.5λ spacing

Figure- 4.14: Cumulative distribution function for 1.66λ spacing

Figure- 4.15: Cumulative distribution function for 2λ spacing

4.3.4 Polarization diversity

The graph of cumulative distribution functions for polarization diversity is shown in figure 4.16. The diversity improvement values for level 5% to .1% are displayed in table 4.3. It is interesting to see that polarization diversity yields an improvement over no diversity. The improvement values are lower than for frequency and space diversity but they are still significant. The lower value would indicate a higher correlation between each channel. These values were obtained from cumulative distributions calculated assuming the same average received power in both channels. The horizontal antenna has a 3 dB disadvantage due to the nulls in its radiation pattern (see section 2.2). With an antenna design without this inconvenience, full advantage could be taken of these improvements.

Although great care was taken to have the two antennas very close, it is possible that part of the improvement could be due to a space diversity effect. It can be seen in plate 6 that the antenna were not exactly at the same location. The two radiating elements were separated by about 1 cm ($\approx \lambda/30$) at their center point.

Figure- 4.16: Cumulative distribution function for 90 ° spacing

POLARIZATION IMPROVEMENT TABLE				
	5% Level	1% Level	.5% Level	.1% Level
90 degrees	5.8	7.8	9.0	13.5
Rayleigh	6.9	10.2	11.8	15
	all values in dB			

Table 4.3 Polarization diversity improvement

4.4 Attenuation measurements

The values of the exponents in the distance-power law calculated from the scatter diagrams, as explained in section 4.2.4, are listed in table 4.4. Figures 4.17, 4.18 and 4.19 show three diagrams as examples for each diversity type. The remainder of the diagrams (8) are in appendix D. Except for one value in channel 1 (4.13, for spacing of 1.5λ), they all are between 3.17 and 3.71. The averages values are 3.62 and 3.49 for channel 1 and channel 2 respectively. If we remove the high value in channel 1, its average becomes 3.56, even closer to channel 2's.

SCATTER DIAGRAM GRADIENT				
	channel 1	channel 2	correlation coeff channel 1	correlation coeff channel 2
1 MHz	3.69	3.60	.945	.942
5 MHz	3.60	3.47	.959	.928
10 MHz	3.65	3.43	.940	.940
15 MHz	3.71	3.54	.929	.925
20 MHz	3.67	3.59	.940	.908
.75 wavel.	3.49	3.53	.971	.967
1 wavel.	3.43	3.51	.973	.965
1.5 wavel.	4.13	3.47	.982	.966
1.66 wavel.	3.34	3.17	.947	.941
2 wavel.	3.65	3.48	.972	.958
polarization	3.40	3.56	.942	.958
average	3.62	3.49		

Table 4.4 Exponents of distance-power law for all diversity cases

These two values are in agreement with other reported distance power law measurements [15]. In one office building with plasterboard walls, Alexander measured a gradient of 3.7. In all scatter diagrams, the data for the 20 locations was used to do

the fitting. The floor plan and the list of locations presented in Appendix E must be used to relate the distances to the attenuation values. There was no attempt to separate the line-of-sight (LOS) locations (usually short distances) from the non-LOS locations (usually longer distances). As mentioned in [26], the attenuation rate is not the same for the two cases. The LOS locations usually give exponent values closer to the free-space value of 2.

For frequency and space diversity, all curves (except figure D.6, space diversity, 1.5λ) are almost one over the other. This is an indication that the relative attenuation is consistently the same for both channels and that the overall rate of attenuation (distance-power law exponent) is also very close for all spacings and all frequencies.

The polarization diversity scatter diagram shows the two curves almost parallel but well separated. Channel 2's relative attenuation is consistently higher than channel 1's. Channel 2 had the horizontal antenna. From this curve, the horizontal dipole has a relative attenuation about 3.4 dB worse than the vertical antenna. The difference is due to the non-uniform radiation pattern of an horizontal dipole, as discussed in section 2.2. The difference mentioned by Cox *et al* in [13] is 3 dB.

The correlation coefficients calculated when the fitting was done are also presented in the table. These coefficients give an indication of the degree of fit of the curve.

Figure- 4.17: Scatter diagram for frequency diversity 1 MHz, all locations

Figure- 4.18: Scatter diagram for space diversity .75 λ , all locations

Figure- 4.19: Scatter diagram for polarization diversity, all locations

4.5 Summary

This section will summarize and regroup the results obtained in section 4.3 and 4.4.

For frequency diversity, at all but the .1% level, the largest diversity improvement were achieved for spacings of 10 to 20 MHz. Lower improvement was achieved with a 5 MHz spacing and minimal improvement was achieved with a 1 MHz spacing. The results at the .1% level were less definite as they were based on fewer numbers of samples.

For space diversity, the results for the 5%, 1% and .5% levels indicated that all five spacings are good to get the maximum improvement. As for frequency diversity, the .1% level results were less definite for the same reason.

The polarization diversity results showed that the two channels were partially correlated but significant improvement was measured using polarization diversity. A space diversity effect could be the cause of some of the improvement because there was still a small physical separation between the two antennas ($\approx \lambda/30$).

The overall exponent of the distance-power law was 3.56 for channel 1 and 3.49 for channel 2. These values are well within previously reported values for similar types of building construction. The attenuation results for frequency and space diversity were very consistent except for one curve. The attenuation results for polarization diversity showed that the channel with the horizontal antenna had a relative attenuation 3.4 dB higher than that of the vertical antenna: a normal characteristics of horizontally polarized dipole antenna.

Chapter 5

Conclusion and Recommendations

5.1 Conclusion

We started by reviewing theory on the fading channel and on diversity. Currently published experiments were then reviewed in order to study proven experimental installations, procedures and results. The design and implementation of our transmitters, receivers, antennas and data acquisition system were presented and explained in detail. The analysis and data processing steps conducted to obtain our results were also explained.

Three types of diversity were used: frequency, space and polarization. Frequency and antenna spacings were used as a parameter to study diversity improvement. Polarization diversity used one horizontally and one vertically polarized dipole.

The overall results are in agreement with theory and previously published results. Frequency diversity results showed that maximum improvement is achieved for spacings of 10 MHz or more. Lower improvement is achieved with a spacing of 5 MHz. As a comparison, the maximum improvement figures were within .6, .9 and 1.7 dB from the improvement obtained with 2 uncorrelated Rayleigh distributions.

Space diversity results showed that any antenna spacing above $.75 \lambda$ will give the maximum improvement. This confirms theory and previously reported results which

indicated that a spacing of $.5 \lambda$ would get the maximum improvement. The improvement figures obtained were within 1.1, .9 and .8 dB from the improvement obtained with 2 uncorrelated Rayleigh distributions..

Polarization diversity results showed that significant improvement is achieved. The improvement obtained ranged from 5.8 dB at the 5% level to 13.5 dB at the .1% level. The lower value of the improvement (compared to frequency and space diversity) would mean that our two channels were somewhat correlated. The physical separation of the two antennas might have had a small space diversity effect.

Attenuation results in the form of the distance-power law's exponents were calculated using data from all 20 locations. The average value of the distance-power law's exponent was 3.49 and 3.56. This matches previously reported results for similar types of building construction. The scatter diagram for polarization diversity showed that overall attenuation for the horizontal dipole was 3.4 dB higher than for the vertical one, as expected for that type of antenna.

5.2 Contribution

The indoor radio channel is still a new channel. Some results are reported in the literature about attenuation, the type of envelope level distribution and difference between static and dynamic measurements. Except for one case of space diversity, there were no direct results on the use and effectiveness of diversity in the indoor channel. This thesis provides such results on diversity and the relative merit of frequency, space and polarization diversity. Results on frequency diversity showed what minimum spacing is good to get the most improvement for that type of building construction. It was shown that polarization diversity could be used effectively as method to obtain a second partially uncorrelated channel.

5.3 Recommendations for future research

The raw data collected in the course of this work was gathered in such a way as to allow a great flexibility in studying other aspects of the indoor portable channel.

Some analysis on the distribution of fade durations and time between fades could be performed. Studies on the difference between the attenuation and envelope distribution characteristics of LOS and non-LOS locations could be done. As done in many previous work, the cumulative distribution or the probability density function could be studied to find out more precisely what type they match (e.g. Rayleigh, Rician or log-normal).

Since the combining was done by a software algorithm, other combining method such as equal gain, maximal ratio or switched combining could be easily implemented. The data gathered could be used to refine or produce a channel model. Channel error performance could be calculated for different types of modulation to see their relative performance in this environment.

Plate 1 *Frequency diversity transmitter*

Plate 2 *Frequency diversity receiver*

2 OF/DE 2

MICROCOPY RESOLUTION TEST CHART
NATIONAL BUREAU OF STANDARDS
STANDARD REFERENCE MATERIAL 1010a
(ANSI and ISO TEST CHART No. 2)

Plate 3 Space and polarization diversity receiver

Plate 4 *Example of antenna used*

Plate 5 *Holding frame for space diversity antenna*

Plate 6 *Holding frame for polarization diversity antenna*

Plate 7 *Guiding frame used to take measurement*

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Appendix A

Program Listings

The program listings shown here are the program used for data acquisition and the main calculation programs. The short programs are listed in their entirety but the longer ones are not: only the appropriate segments are listed. The algorithms for these programs were explained in chapter 3 and 4.

The programs listed in this appendix are the following:

1. ADSAMP.BAS;
2. DATASCALE;
3. DATAVG;
4. DATACDF; and
5. LOSS.

A.1 ADSAMP.BAS

This program is the data acquisition control program explained in section 3.9.2. It is written in GW-BASIC.

```

10 ' ADSAMP.BAS
20 ' DATA ACQUISITION CONTROL PROGRAM
30 ' using mode 0,1,17,20,8 and 9
40 ' Mode 0,17,1,20,8 and mode 9 are used. 2-03-89
50 ' _____
60 COLOR 11,0 'screen setting
70 DEF SEG = &H8000 ' memory address to load machine language driver
80 DIM DIO%(4),DT%(20000)
90 ' ——— PROGRAM INITIALIZATION ———
100 BLOAD "DASH16.BIN",0 ' load driver in memory at offset 0 from
DEF SEG

```

```

110 RATE1%=20 :RATE2%=100 ' timer rate settings
120 DMA%=&H2000: DAT%=&H2000 ' DMA and DATA address
130 INPUT "NAME OF FILE TO STORE DATA:",FLNM$
140 INPUT "ANTENNA SPACING:1 ,2 ,3,4 ,5 OR 0 IF SPACING NOT
IMPORTANT: ",SP%
150 INPUT "TOTAL NO OF CONVERSIONS (default=60 000):",INCON-
VER% 'no entry=60000
160 INPUT "SAMPLING RATE? (default=500 Hz) ",INRATE%
170 IF INRATE% <>0 THEN RATE1%=10000/INRATE% ' for different
rates
180 IF INCONVER% <=20000 THEN FLG=1 ' an entry < 20 000 is a test
run
190 IF INCONVER% = 0 THEN FLG=2
200 IF INCONVER% >20000 THEN GOTO 150
210 ON FLG GOTO 220,270
205 ' case for test run
220 CONVER%=INCONVER%
230 TOTAL=CONVER%
240 TRNS%=INCONVER%
250 CNT=0 :TRNSCNT=0
260 GOTO 320
265 ' case for data run :
270 CONVER% = 30000 'default samples: 2 × 30 000
280 CNT=2 :TRNSCNT= 4 'CNT:sampling run counter,
290 TOTAL=CNT*CONVER% 'TRNSCNT:data transfer counter
300 TRNS%=15000 ' default transfer:4 × 15 000
310 '——MODE 0 Board Initialization——
320 DIO%(0)=768 ' I/O address of board
330 DIO%(1) = 2 :DIO%(2)=1 'interrupt level

```

```

340 DASH16 = 0 'call offset from DEF SEG
350 FLAG% = 0 'error variable
360 MD% = 0 'mode 0 - initialize
370 CALL DASH16 (MD%, DIO%(0), FLAG%)
380 IF FLAG% <> 0 THEN PRINT "INSTALLATION ERROR":STOP 'Halt
on error
390 '——MODE 17- Sampling rate——
400 '
410 DIO%(0)=RATE1% ' timer rate = 1MHz/ (Rate1× Rate2)
420 DIO%(1) = RATE2% 'product = 2 000 =500 HZ
430 MD% = 17 'timer set mode
440 CALL DASH16 (MD%, DIO%(0),FLAG%)
450 IF FLAG% <> 0 THEN PRINT "Error in setting timer. Error #
";FLAG%.STOP
460 '——mode 1 block scan limits——
470 LL%=0 :UL%=1
480 DIO%(0) = LL% 'lower limit
490 DIO%(1) = UL% 'upper limit
500 MD% = 1
510 CALL DASH16 (MD%, DIO%(0),FLAG%)
520 IF FLAG% <> 0 THEN PRINT "Error in setting scan limits # ";FLAG%
: STOP
530 '—— MODE 20 block conversions——
540 DMA%=DMA%+&H1000 : PRINT DMA% ' increment address of
memory location to store conversions
550 DIO%(0)=CONVER% 'No of conversions= 30 K
560 DIO%(1) = DMA% 'Memory segment to dump data
570 DIO%(2) = 1
580 DIO%(3) = 0

```

```

590 MD%=20 'mode 20 - block scan on interrupt
600 BEEP
610 CALL DASH16 (MD%, DIO%(0), FLAG%) 'set it going
620 IF FLAG%<>0 THEN PRINT"Error in block scan setup =
";FLAG%:STOP
630 '—— MODE 8 waiting for the end of conversions
640 MD%=8
650 CALL DASH16 (MD%, DIO%(0), FLAG%) 'monitoring block scan
660 IF DIO%(1)=1 THEN PRINT DIO%(1)
670 IF DIO%(1)=0 THEN GOTO 690
680 GOTO 640
690 CNT=CNT-1 'block counter decreased
700 IF CNT<=0 GOTO 730
710 GOTO 400
720 '——Opening of transfer file——
730 BEEP
740 PIP$="\HPBAS"+FLNM$ ' name of storage file
750 OPEN "O",#1,PIP$ ' establishing path to file
760 PRINT #1,FLNM$ ' printing file info
770 WRITE#1,SP%,TOTAL
780 '——MODE 9 data from memory to array
790 MD%=9
800 DAT%=DAT%+&H1000 ' address of mem location to get data
810 DIO%(0)=TRNS% ' no of data to transfer: 15 K default
820 DIO%(1) = DAT% ' address to get data
830 DIO%(2) = 0 ' offset from address
840 PRINT DIO%(1)
850 DIO%(3) = VARPTR(DT%(0)) 'indicates beginning of array in program
860 DIO%(4)=0

```



```

870 CALL DASH16 (MD%, DIO%(0), FLAG%)
880 IF FLAG%<>0 THEN PRINT"Error data transfer = ";FLAG%:STOP
'any errors?
885 ' writing array to file on disk, 20 values per loop
890 FOR I =0 TO TRNS%-1 STEP 20
900 WRITE#1,DT%(I),DT%(I+1),DT%(I+2),DT%(I+3),DT%(I+4),
DT%(I+5),DT%(I+6), DT%(I+7),DT%(I+8),DT%(I+9)
910 WRITE#1,DT%(I+10),DT%(I+11),DT%(I+12),DT%(I+13),DT%(I+14),
DT%(I+15),DT%(I+16),DT%(I+17),DT%(I+18),DT%(I+19)
920 NEXT I
930 IF TRNS%=CONVER% GOTO 1120 'test to see if any more data
940 TRNSCNT=TRNSCNT-1 ' decrease transfer counter
950 IF TRNSCNT<=0 GOTO 1120
960 MD%=9 'starting 2nd half of 30 K block
970 DIO%(0)=TRNS%
980 DIO%(1) = DAT%
990 DIO%(2) = TRNS% ' offset from memory address: 2nd half
1000 PRINT DIO%(1)
1010 DIO%(3) = VARPTR(DT%(0))
1020 DIO%(4)=0
1030 CALL DASH16 (MD%, DIO%(0), FLAG%)
1040 IF FLAG%<>0 THEN PRINT"Error data transfer = ";FLAG%:STOP
'any errors?
1045 ' writing array to file
1050 FOR I =0 TO TRNS%-1 STEP 20
1060 WRITE#1,DT%(I),DT%(I+1),DT%(I+2),DT%(I+3),DT%(I+4),
DT%(I+5),DT%(I+6),DT%(I+7),DT%(I+8),DT'      )
1070 WRITE#1,DT%(I+10),DT%(I+11),DT%(I      )DT%(I+13),DT%(I+14),
DT%(I+15),DT%(I+16),DT%(I+17),DT%(I+18),DT%(I+19)

```

```
1080 NEXT I
1090 TRNSCNT=TRNSCNT-1 ' decrease transfer counter
1100 IF TRNSCNT<=0 GOTO 1120
1110 GOTO 790
1120 CLOSE#1
1130 STOP
1140 END
```

A.2 DATASCALE

This program does the scaling of data file: it transforms the raw analog-to-digital converter values to power values in dBm.

```
10 ! SCALING PROGRAM
20 !
30 ! DATASCALE1
40 !
50 ! THIS PROGRAM TAKES THE DATA FROM A FILE, RUNS IT
    THROUGH
60 ! A SCALING ROUTINE TO CONVERT THE A/D READINGS TO
    DBM.
70 ! IT THEN LOADS "DATAACDF"
80 ! THE VALUES IN P1(*),P2(*) ARE PASSED ALONG THROUGH A
    COM BLOCK.
90 ! —LOAD BIN "IO" BEFORE RUNNING THE PROG
100 WAIT 1
110 OPTION BASE 1
120 INTEGER Min1,Min2,Delmin
```

The COM block is used to pass information between the various programs being loaded

```

130 COM REAL P1(30000), P2(30000), Total, Name$(12), Dir$(115)[80],
Vol$(3)[12], INTEGER Sp, Prog, Filno, Rtype, Maxfil, V
140 ALLOCATE INTEGER D1(30000)
150 DIM Filnm$(21)
160 GOSUB Clears
170 Prog=1
180 Ms$=""
190 IF Rtype<>0 THEN GOTO S2
    Allows for batch mode if program is to run unattended
200 S1:INPUT "do you want to run in batch mode: 1=yes, 2=no",Ba
210 INPUT "enter initial value and max value of FILNO", Startfil, Maxfil
    This loop allows to read data from more than one directory
220 FOR K=1 TO 3
230 LINPUT "ENTER VOLUME WHERE THE DATA FILES ARE:",
Vol$(K)
240 PRINT Vol$(K)
250 NEXT K
260 V=1 !
270 Filno=Startfil
280 ON Ba GOTO S2,S3
290 S2:IF Rtype=2 THEN GOTO S3 ! INTERAC MODE
300 S21:IF Filno>=Maxfil+1 THEN
310 V=V+1
320 Filno=1
330 ON V GOTO S21,S21,Fin
340 Fin:BEEP
350 GOSUB Clears
360 PRINT USING "5/"
370 PRINT " .....THE END....."

```

```
380 PRINT "all files are done "  
390 Filno=1  
400 STOP  
410 ELSE  
420 Rtype=1! batch mode  
430 IF Filno=1 THEN GOSUB Catdump ! Catdump write the directory to  
the vector Dir$  
440 IF LEN(TRIM$(Dir$(Filno)[1,12]))<7 THEN ! non data files have  
smaller file name: they are not selected  
450 Filno=Filno+1  
460 GOTO S21  
470 ELSE  
480 Name$=TRIM$(Dir$(Filno)[1,12]) ! data file names is read into Name$  
490 Filnm$=Name$&Vol$(V)  
500 PRINT Name$  
510 GOTO S5  
520 END IF  
530 END IF
```

The program section from line 540 to 770 deals with non batch mode and no scaling operations

```
780 S5:In$=Name$[1,1] ! first letter of file name indicates the diversity type  
790 ASSIGN @F TO Filnm$ ! path to file  
800 ENTER @F;N$,Sp,Total  
810 REDIM D1(Total),P1(Total),P2(Total)  
820 ENTER @F;Tim,D1(*) ! entering channel 1 data  
830 IF In$="F" THEN  
840 Scale$=In$&"SCALE"&VAL$(Sp) ! loading the proper scaling file  
850 ELSE  
860 Scale$="SSCALE" ! loading the proper scaling file
```

```
870 END IF
880 ALLOCATE REAL Ad(0:4095)
890 ASSIGN @S TO Scale$ ! path to scaling file
900 ENTER @S;Ad(*) ! entering scaling vector for chan 1
910 ALLOCATE INTEGER D2(30000)
920 REDIM D2(Total)
930 ENTER @F;D2(*) ! entering channel 2 data
    Doing manipulations needed to do the correction on channel 1's data
940 Min1=MIN(D1(*))
950 Min2=MIN(D2(*))
960 OUTPUT 26;N$,"MIN1";Min1,"MIN2"; Min2
970 IF Min1<Min2 THEN GOTO S6
980 Delmin=Min1-Min2
990 MAT D1= D1-(Delmin)
1000 !
1010 S6:DISP "S6: scaling D1"
1020 FOR I=1 TO Total STEP 15
1030 P1(I)=Ad(D1(I)) ! Scaling channel 1
1040 P1(I+1)=Ad(D1(I+1))
1050 P1(I+2)=Ad(D1(I+2))
1060 P1(I+3)=Ad(D1(I+3))
1070 P1(I+4)=Ad(D1(I+4))
1080 P1(I+5)=Ad(D1(I+5))
1090 P1(I+6)=Ad(D1(I+6))
1100 P1(I+7)=Ad(D1(I+7))
1110 P1(I+8)=Ad(D1(I+8))
1120 P1(I+9)=Ad(D1(I+9))
1130 P1(I+10)=Ad(D1(I+10))
1140 P1(I+11)=Ad(D1(I+11))
```

```
1150 P1(I+12)=Ad(D1(I+12))
1160 P1(I+13)=Ad(D1(I+13))
1170 P1(I+14)=Ad(D1(I+14))
1180 NEXT I
1190 !
1200! ASSIGN @F TO *
1210 !
1220 ENTER @S;Ad(*) ! entering scaling vector for chan 2
1230! ASSIGN @S TO *
1240 !
1250 DISP "S7:scaling D2"
1260 S7:FOR I=1 TO Total STEP 15
1270 P2(I)=Ad(D2(I)) ! SCALING P2
1280 P2(I+1)=Ad(D2(I+1))
1290 P2(I+2)=Ad(D2(I+2))
1300 P2(I+3)=Ad(D2(I+3))
1310 P2(I+4)=Ad(D2(I+4))
1320 P2(I+5)=Ad(D2(I+5))
1330 P2(I+6)=Ad(D2(I+6))
1340 P2(I+7)=Ad(D2(I+7))
1350 P2(I+8)=Ad(D2(I+8))
1360 P2(I+9)=Ad(D2(I+9))
1370 P2(I+10)=Ad(D2(I+10))
1380 P2(I+11)=Ad(D2(I+11))
1390 P2(I+12)=Ad(D2(I+12))
1400 P2(I+13)=Ad(D2(I+13))
1410 P2(I+14)=Ad(D2(I+14))
1420 NEXT I
1430 DEALLOCATE Ad(*),D2(*)
```

```
1440 !
1450 ON Rtype GOTO Cd,S8 ! for batch mode Rtype =1
1460 S8:INPUT "1—work on CDF, 2—plot on the screen",En
1470 ON En GOTO Cd,Scrn
1480 Cd:PRINT " LOADING DATACDF"
1490 LOAD "DATACDF"
1500 Scrn:PRINT
1510 PRINT " LOADING DATASCREEN"
1520 PRINT
1530 LOAD "DATASCREEN"
1540 !
1550 !
1560 Catdump:CAT Vol$(V) TO Dir$(*);COUNT T,NO HEADER ! ,8
1570 REDIM Dir$(T)
1580 RETURN
1590 Clears:OUTPUT KBD USING "#,B";255,75
1600 RETURN
1610 END
```

A.3 DATAVG

This program calculates the average power of channel 1 and channel 2 and write the two values in a results file. To calculate the arithmetic average the scaling is done using the milliwatt scaling files.

```
10 ! AVERAGE CALCULATION PROGRAM
20 !
30 ! DATAVG
40 !
50 ! THIS PROGRAM TAKES THE DATA FROM A FILE , RUNS IT
THROUGH
```

```

60 ! A SCALING ROUTINE TO CONVERT THE A/D READINGS TO
mW.
70 ! IT THEN CALCULATES THE ARITHMETIC AVERAGE OF BOTH
SIGNAL.
80 ! THESE AVERAGE ARE STORED IN A RESULT FILE CREATED
IN \HPBAS\RES1
90 OPTION BASE 1
100 COM REAL P1(30000),P2(30000),Total,Name$(12), Dir$(20)(80),
INTEGER Sp, Prog, Filno, Rtype, Maxfil
110 INTEGER Disk
120 REAL Mw(0:4095),Atten(10)
130 DIM Rname$(21),Filnm$(21)
140 Prog=5
150 Vol$=":CS80,1500,9"
160 GOSUB Catdump
170 PRINT Dir$(*)

```

The program section not shown was dealing with entering the attenuation for each channel and writing to the table of content file.

```

320 ALLOCATE INTEGER D1(30000)
330 ASSIGN @F TO Filnm$&Vol$
340 ENTER @F;N$,Sp,Total
350 REDIM D1(Total),P1(Total),P2(Total)
360 ENTER @F;Tim,D1(*)
370 PRINT " ",K,N$
380 IF In$="F" THEN
390 Scale$=In$&"SCALE"&VAL$(Sp)&"W" ! the W indicates it is a mW
file
400 ELSE
410 Scale$="SSCALEW"!MEM,0,1"

```



```
420 END IF
430 ASSIGN @S TO Scale$
440 ENTER @S;Mw(*) ! TABLE FOR P1
450 !
460 S3: DISP " scaling D1"
470 FOR I=1 TO Total STEP 10
480 P1(I)=Mw(D1(I)) ! SCALING P1
490 P1(I+1)=Mw(D1(I+1))
500 P1(I+2)=Mw(D1(I+2))
510 P1(I+3)=Mw(D1(I+3))
520 P1(I+4)=Mw(D1(I+4))
530 P1(I+5)=Mw(D1(I+5))
540 P1(I+6)=Mw(D1(I+6))
550 P1(I+7)=Mw(D1(I+7))
560 P1(I+8)=Mw(D1(I+8))
570 P1(I+9)=Mw(D1(I+9))
580 NEXT I
590 !
600 DEALLOCATE D1(*)
610 ALLOCATE INTEGER D2(30000)
620 REDIM D2(Total)
630 ENTER @F;D2(*)
640 ASSIGN @F TO *
650 !
660 ENTER @S;Mw(*) ! TABLE FOR P2
670 ASSIGN @S TO *
680 !
690 DISP " scaling D2"
700 S4: FOR I=1 TO Total STEP 10
```

```

710 P2(I)=Mw(D2(I)) ! SCALING P2
720 P2(I+1)=Mw(D2(I+1))
730 P2(I+2)=Mw(D2(I+2))
740 P2(I+3)=Mw(D2(I+3))
750 P2(I+4)=Mw(D2(I+4))
760 P2(I+5)=Mw(D2(I+5))
770 P2(I+6)=Mw(D2(I+6))
780 P2(I+7)=Mw(D2(I+7))
790 P2(I+8)=Mw(D2(I+8))
800 P2(I+9)=Mw(D2(I+9))
810 NEXT I
820 DEALLOCATE D2(*)
830 ! calculating the average in mW
840 DISP "CALCULATING AVERAGE OF ",N$
850 Ave1=SUM(P1)
860 Ave2=SUM(P2)
870 Ave1=Ave1/Total
880 Ave2=Ave2/Total

```

Converting the mW values into dBm values

```

890 Ave1=10*LGT(Ave1)
900 Ave2=10*LGT(Ave2)
910 OUTPUT 26;N$
920 OUTPUT 26;Ave1," ",Ave2
930 OUTPUT 26;" "

```

Creating the result files: the file name is based on the data file name

```

940 IF POS(Dir$(K)[1,9],"_")=4 THEN
950 Rname$=Dir$(K)[1,3]&"_R"&"":CS80,1500,9"
960 ELSE

```

```

970 Rname$=Dir$(K)[1,4]&"_R"&":CS80,1500,9"
980 END IF
990 CREATE BDAT Rname$,10
1000 ASSIGN @R TO Rname$
1010 OUTPUT @R;N$,Sp,Ave1,Ave2,Atten(K)
1020 NEXT K
1030 BEEP
1040 ASSIGN @R TO *
1050 PRINT USING "5/"
    Printing hardcopy of floppy disk directory
1060 OUTPUT 19;"WAIT_OFF"
1070 OUTPUT 19;"DIR A:\ >PRN"
1080 CAT Vol$
1090 DISP "....THE END....."
1100 STOP
1110 Catdump:CAT Vol$ TO Dir$(*);COUNT T, NO HEADER
1120 REDIM Dir$(T)
1130 RETURN
1140 Clears:OUTPUT KBD USING "#,B";255,75
1150 RETURN
1160 END

```

A.4 DATACDF

This program is the main calculation program. It took the scaled data from the DATASCALE program and the average power value stored in the result file and calculated the histogram boundaries, the histogram itself and the cumulative distribution function. The distribution was stored in the result file for future access.

The program was designed to allow interactive or batch operations. The interactive mode allowed many functions to be accessed including plotting on the monitor and on an hard copy plotter connected to the PC. These parts of the program will not be listed.

10 ! DATACDF: THE STATISTICS PROGRAM

20 !

30 ! THIS PROGRAM TAKES THE DATA FROM A COM BLOCK AND
CALCULATES PDF, CDF

40 ! FOR P1,P2 AND MAX (P1,P2) .OPTION TO PLOT ON SCREEN AND
PLOTTER.

50 ! OPTION TO CREATE RESULTS FILE AND STORE CDF VALUES.

60 !

70 OPTION BASE 1

COM block used to pass data from the DATASCALE program.

80 COM P1(30000), P2(30000), Total, Name\$(12), Dir\$(115)[80], Vol\$(3)[12],
INTEGER Sp, Prog, Filno, Rtype, Maxfil,V

Vectors used for boundary, histogram and cdf counters

90 REAL X1(41), X2(41), Pd1(42),Pd2(42),Cd1(42),Cd2(42), Pd3(42), Cd3(42)

100 REAL Dra(42),Cd12(42),Cdjt(42),Pdjt(42),Cd(42), Pd(42)

110 DIM Ylab\$(10)[3],Ylab1\$(25),Xlab\$(21)[3], Spac\$(11)[6], Title1\$(50]

120 DIM Div\$(0:11)[8]

130 REDIM P1(Total),P2(Total)

140 GOSUB Clears

These are the function allowing interactive use of the program.

150 ON KEY 5 GOSUB Cdf12 ! F1 FOR CDF OF P1,P2

160 ON KEY 6 GOSUB Getresf ! F2 TO GET A RESFILE

170 ON KEY 7 GOSUB Jnt12 ! F3 FOR JOINT CDF OF P1,P2

180 ON KEY 8 GOTO More ! F4 END PROGR

190 ON KEY 0 GOSUB Resfile ! Sh-F1 TO STORE INFO IN RESULT FILE

200 ON KEY 2 GOSUB Med ! Sh-F3 MEDIAN OF P1,P2

```

210 ON KEY 1 GOSUB Hpplo ! Sh-F2 TO PLOT ANY CDF
220 ON KEY 3 GOSUB Screen ! Sh-F3 TO PLOT ON SCREEN
230 !
240 Bdy=31 ! FROM AVE-30 TO AVE+10
250 GOSUB Pdfbdry ! TO GET X LABEL VALUES FOR PLOT
260 C2:IF Rtype=1 THEN GOTO Batchrun ! batch mode, no interaction
270 Ch:GOSUB Menu ! menu for interactive mode
280 Nex=1
290 Wait:IF Nex=2 THEN GOTO Ch
300 GOTO Wait
310 More:GRAPHICS OFF
320 INPUT "TYPE 1-STOP, 2-SCALE NEW FILE, 3-CARRY ON", En
330 ON En GOTO Fin, Sc, Ch
340 Fin:PRINT "...THE END ....."
350 STOP
360 Sc:DISP "LOADING DATASCALE"
370 Prog=3
380 LOAD "DATASCALE"
390 !
400 !-----
410 Batchrun:GOSUB Cdf!2 ! doing individual channels cdf
420 GOSUB Jnt12 ! doing joint cdf
440 Filno=Filno+1
450 LOAD "DATASCALE1" ! when cdf are done for this file, return to scaling
program to do next file
460 !-----
470 !
480 Cdf12: GOSUB Getresf ! to get data from result file
490 GOSUB Pdfbdry ! calculating the histogram boundaries

```

```

500 GOSUB Pdf ! doing histogram (or pdf)
510 MAT Pd= Pd1
520 GOSUB Cdf ! doing cdf
530 MAT Cd1= Cd
540 MAT Pd= Pd2
550 GOSUB Cdf
560 MAT Cd2= Cd
570 GOSUB Resfile ! storing cdf data in result file
580 Nex=2
590 Cdf121:RETURN
1040 !-----
1050 ! doing cdf for joint event ( selection combining)
1060 Jnt12:GOSUB Getresf
1070 GOSUB Pdfbdry
1080 DISP " doing joint PDF of P1 and P2 ",Name$
1090 M1=MIN(P1(*))
1100 MAT SEARCH X1,LOC(>=M1);St1
1110 M2=MIN(P2(*))
1120 MAT SEARCH X2,LOC(>=M2);St2
1121 ! The joint event will always be at least greater than the largest minimum
of each P1 or P2.
1130 St=MAX(St1,St2)
1140 MAT Pdjt= (0)
1150 FOR I=1 TO Total STEP 2
1160 MAT SEARCH X1,LOC(>=P1(I)) ; D1,St
1170 FOR J=D1 TO 41
1180 IF P2(I)<=X2(J) THEN
1190 Pdjt(J)=Pdjt(J)+1
1200 GOTO Jnt2

```

```

1210 END IF
1220 NEXT J
1230 Pdjt(42)=Pdjt(42)+1
1240 Jnt2:MAT SEARCH X1,LOC(>=P1(I+1)) ; D1,St
1250 FOR J=D1 TO 41
1260 IF P2(I+1)<=X2(J) THEN
1270 Pdjt(J)=Pdjt(J)+1
1280 GOTO Jnt3
1290 END IF
1300 NEXT J
1310 Pdjt(42)=Pdjt(42)+1
1320 Jnt3:NEXT I
1330 DISP "doing CDF",Name$
1340 MAT Pd= Pdjt
1350 GOSUB Cdf
1360 MAT Cdjt= Cd
1370 Nex=2
1380 GOSUB Resfile
1390 Jnt12l:RETURN
2880 !-----
2890 !
2900 Cdf:! general routine to do CDF
2910 DISP "DOING CDF"
2920 MAT Pd= Pd/(Total)
2930 MAT Cd= (0)
2940 Cd(1)=Pd(1)
2950 FOR I=2 TO 42
2960 Cd(I)=Cd(I-1)+Pd(I)*Inc
2970 NEXT I

```

```
2980 DISP "DONE"
2990 Cdf1:RETURN
3000 !-----
3010 !
3020 Pdf:MAT Pd1= (0)
3030 MAT Pd2= (0)
3040 GOSUB Pdfbdry
3050 DISP "doing PDF for P1 and P2",Name$
3060 IF St1<>0 AND St2<>0 THEN GOTO Pdf1
3070 M1=MIN(P1(*))
3080 MAT SEARCH X1,LOC(>=M1);St1
3090 M2=MIN(P2(*))
3100 MAT SEARCH X2,LOC(>=M2);St2
3110 Pdf1:FOR I=1 TO Total STEP 2
3120 MAT SEARCH X1,LOC(>=P1(I));D1
3130 Pd1(D1)=Pd1(D1)+1
3140! MAT SEARCH X2,LOC(>=P2(I));D2
3150! Pd2(D2)=Pd2(D2)+1
3160 MAT SEARCH X1,LOC(>=P1(I+1));D1
3170 Pd1(D1)=Pd1(D1)+1
3180! MAT SEARCH X2,LOC(>=P2(I+1));D2
3190! Pd2(D2)=Pd2(D2)+1
3200 NEXT I
3210 DISP "DONE"
3220 Pdf1:RETURN
3230 !-----
3240 !
3870 Pdfbdry:Inc=1
3880 DISP "Calculating,boundary values for the PDF ave-30 to ave + 10"
```



```

3890 FOR I=1 TO 41
3900 X1(I)=Ave1+(I-Bdy)*Inc
3910 X2(I)=Ave2+(I-Bdy)*Inc
3920 NEXT I
3930 FOR I=1 TO 21
3940 Xlab$(I)=VAL$(2*I-Bdy-1)
3950 NEXT I
3960 DISP
3970 Pdfbdryl:RETURN
5420 END

```

A.5 LOSS

This program does the calculation of the path losses. These losses were used to produce the scatter diagrams in section 4.4.

```

100 ! LOSS: PROGRAM TO CALCULATE TRANSMISSION LOSS
200 ! FROM TX TO RX.
300 OPTION BASE 1

```

The three vectors Fatten, Satten, Patten have the data required to do the calculation. For all locations, the transmitter attenuator and the average received power in channel 1 and 2 is recorded in these vectors.

```

400 REAL Dist(25),Fatten(20,5,3),Satten(20,5,3),Patten(20,1,3)

```

The vectors Sptxpwr and Ftxpwr contain the power values corresponding to the attenuator values for the two types of transmitters: frequency diversity and space/polarization diversity.

```

500 REAL Sptxpwr(-10:50), Ftxpwr(5,0:50,2),Spathloss(20,5,2),Ppathloss(20,1,2),

```

```
Fpathloss(20,5,2)
600 Vol$=":CS80,1500,7"
```

The file Atten contains all the vectors needed.

```
700 ASSIGN @F1 TO "ATTEN:CS80,1500,7"
800 ENTER @F1;Fatten(*),Satten(*),Patten(*),Dist(*), Sptxpwr(*), Ftxpwr(*)
900 RESET @F1
1000 ! Path loss for space diversity, chan 1 and chan2
1100 FOR K=1 TO 20
1200 FOR I=1 TO 5
```

Satten(K,I,1) is the attenuator value, Satten(K,I,2) is the average received power for channel 1 and Satten(K,I,3) is the average received power for channel 2.

```
1300 Spathloss(K,I,1)= Sptxpwr(Satten(K,I,1))-Satten(K,I,2)
1400 Spathloss(K,I,2)= Sptxpwr(Satten(K,I,1))-Satten(K,I,3)
1500 NEXT I
1600 NEXT K
1700 ! Path loss for polarization diversity
1800 FOR K=1 TO 20
1900 I=1
2000 Ppathloss(K,I,1)= Sptxpwr(Patten(K,I,1))-Patten(K,I,2)
2100 Ppathloss(K,I,2)= Sptxpwr(Patten(K,I,1))-Patten(K,I,3)
2200 NEXT K
2300 ! FOR FREQ DIV
2400 FOR K=1 TO 20
2500 FOR I=1 TO 5
```

```
2500 Fpathloss(K,I,1)= Ftxpwr(I,Fatten(K,I,1),1)-Fatten(K,I,2)
2700 Fpathloss(K,I,2)= Ftxpwr(I,Fatten(K,I,1),2)-Fatten(K,I,3)
2800 NEXT I
2900 NEXT K
```

The file Pathloss contains the resulting vectors and the table of distance and location number.

```
3000 ASSIGN @P TO "PATHLOSS"&Vol$
3100 OUTPUT @P;Dist(*), Fpathloss(*), Spathioss(*), Ppathloss(*)
3200 END
```

Appendix B File Format

B.1 Data files

The data files containing the collected data are configured in the following format:

N\$, Sp, Total, Tim, D1(30000),D2(30000)

note: in addition of 1 byte per character, all alphanumeric (\$) variables have an extra 4 byte header plus a pad byte at the end if they have an odd number of characters.

N\$: name of data file; 8 or 9 characters string; 14 bytes

Sp: spacing used in this file; from 1 to 5, 0 used for polarization diversity;
INTEGER, 2 bytes

Total: number of samples per channel ;REAL , 8 bytes, 30 000 for all files

Tim: time of conversion to HP BASIC binary data file, REAL, 8 bytes

D1(30000): vector containing raw data for channel 1, INTEGER, 2 bytes times
30 000=60 000 bytes

D2(30000): vector containing raw data for channel 2, INTEGER, 2 bytes times
30 000 = 60 000 bytes

B.2 Result files

The result files that were created when processing the data files had the following format:

N\$, Sp, Ave1, Ave2, Att, Cd1(42), Cd2(42), Cd3(42), Cdjt(42)

N\$: name of data file, same as for data file

Sp: spacing, same as for data file

Ave1: average power of channel 1, in dBm; real REAL, 8 bytes

Ave2: average power of channel 2, in dBm; REAL, 8 bytes

Att: attenuator setting at the transmitter, in dB; REAL, 8 bytes

Cd1(42): cumulative distribution function for channel 1; REAL, 42×8 bytes

Cd2(42): cumulative distribution function for channel 2; REAL, 42×8 bytes

Cd3(42): not used; REAL, 42×8 bytes

Cdjt(42): cumulative distribution function for joint distribution; REAL, 42×8 bytes

B.3 Result matrix files

The result matrix file contains all the information from the respective result files collected in one large matrix. There are three files: **FREQ_RES**, **SPAC_RES** and **POL_RES** corresponding to the three diversity type. **FREQ_RES** and **SPAC_RES** have exactly the same format. Where information about 5 spacings is used for the previous two files, only one is used in **POL_RES**. The files have the following format:

Typ\$, Validloc(20), Jnt(20, 5, 42), Cum1(20, 5, 42), Cum2(20, 5, 42), Locavejnt(5, 42, 3), Locave1(5, 42, 3), Locave2(5, 42, 3)

Typ\$: type of diversity, 4 character string, 8 bytes;

Validloc(20): flag to indicate if given location is valid: 1 if valid, 0 if not; integer 40 bytes;

Jnt(20, 5, 42): matrix containing all joint cumulative distribution functions for 20 locations and 5 spacings; real, 4200×8 bytes;

Cum1(20, 5, 42) and Cum2(20, 5, 42): same as Jnt but for channel 1 and channel 2;

Locavejnt(5, 42, 3): point per point average of all cumulative distribution functions in Jnt. Locavejnt(*,*,1) is average value, Locavejnt(*,*,2) is the maximum value and Locavejnt(*,*,3) is the minimum value of all points

Locave1(5, 42, 3) and Locave2(5, 42, 3): same as Locavejnt but for channel 1 and 2.

Appendix C

DASH-16F

C.1 Features and options

The main feature of this board is the Harris HI-774 12 bit successive approximation analog-to-digital converter (conversion time of 8 μ s). Two input channels configurations are available: 16 single ended channels or 8 differential channels. Many input voltage ranges are available: +1, +2, +5, +10 Volt unipolar or ± 5 , ± 1 , ± 2.5 , ± 5 , ± 10 Volt bipolar. Conversions can be initiated in three ways: by software command, by the internal programmable timer, or by direct external trigger. After the conversion is done, there are also three ways of transferring the data to the computer: using program transfer, using interrupts or by direct memory access (DMA). In addition to this analog-to-digital converter function, the board has two channels of multiplying 12 bit digital-to-analog output, a 4 bits TTL/DTL compatible output and 4 bits of digital input. This is a list of the 21 modes available on DASH-16F. Some are used for the analog-to-digital converter functions and some are used for the digital-to analog functions.

Mode 0: initialize the board

Mode 1: set multiplexer scan limits

Mode 2: read multiplexer scan limits

Mode 3: do 1 A/D conversion and increment multiplexer

Mode 4: do N A/D conversions direct to array

Mode 5: do N A/D conversions and transfer to memory on interrupt

Mode 6: do N A/D conversions and transfer to memory via DMA

Mode 7: disable DMA/interrupt operation

Mode 8: monitor status of operation

Mode 9: data transfer from memory to program

Mode 10: set configuration of counter 0

- Mode 11: load counter 0
- Mode 12: read counter 0
- Mode 13: write digital output line OP0-3
- Mode 14: read digital output line IP0-3
- Mode 15: output data to 1 D/A
- Mode 16: output data to both D/A
- Mode 17: set timer rate
- Mode 18: analog output/input on interrupt
- Mode 19: analog trigger function
- Mode 20: do A/D on block of channel

The DASH-16F was only used as an analog-to-digital converter and we will now focus on the data acquisition modes

C.2 Data acquisition modes

There were many modes offered for the actual data acquisition. Different modes allowed much flexibility. We selected mode 20 because it fitted best our needs. In our case the ideal system should sample the channels simultaneously at the desired rate. A secondary requirement would be to keep the system as simple as possible. The following modes perform analog-to-digital conversion:

- mode 3: performs a single conversion, returns data and increments the multiplexer address. This is done under program control.
- mode 4: performs 1 conversion per trigger, up to total N, on successive channels. The trigger is provided internally by the timer or externally. It transfers data directly to an array. The channels are scanned according to the limits set by mode 1.
- mode 5: performs 1 conversion per trigger, up to total N, on successive channels. The trigger is provided internally by the timer or externally. It transfers data to memory on interrupt. The channels are scanned according to limits set by mode 1.

- mode 6: same as mode 5 except that data transfer is done by DMA (direct memory access) to the specified memory location.
- mode 20: performs C conversions (= 1 scan) per trigger, up to total N, on successive channels. C is the number of channels specified by mode 1. The trigger is provided by the timer or externally. Data is transferred to memory by interrupt. The number of scan is N/C.

We will study in more detail the performance of each mode and explain why mode 20 was selected.

Mode 3 seems adequate but since each conversion is done under program control, it is too slow and will not be considered any further. Some limitations of the driver subroutine and the hardware must be highlighted before going any further. Without going into too much detail, the DMA page register (1 page = 64 Kbytes of memory) cannot be incremented by the DMA controller. Any DMA data transfer is limited to a maximum of 64 Kbytes worth of data. Each value being 2 bytes, 32 767 values is the maximum number of values that can be transferred in one attempt. In addition, the parameter N mentioned in the previous paragraph being a two bytes 2's complement integer, any number greater than 32 767 is treated as a negative number and returns an error in the FLAG parameter.

Mode 4, 5, and 6 are very similar in that they perform $N (\leq 32\,767)$ conversions on C channels. The conversions are equally spaced in time by $1/\text{timer rate}$. The fastest conversion speed available on the DASH-16F is 100,000 samples/s (10 μ s between samples). As far as the signal envelope is concerned, nothing of significant importance happens faster than 10 μ s, so this is as good as simultaneous. At this rate, 1 second worth of data would occupy close to 200 Kb of memory space, 30 seconds would use close to 6 000 Kb. This quickly becomes unmanageable.

This brings us to the topic of interval between sampling on each channel versus interval between each pair of samples. Obviously the time between each samples should be short enough so that simultaneous readings can be done on both channels. The time between each pair of sample might not need to be so short. Once this is taken into

account, the solution offered by mode 20 looks promising. The DASH-16F manual states that with a 4.77 MHz clock, a PC using mode 20 has a time interval between sampling on two successive channels of about 100 μ s. The Vectra with a clock of 8 MHz should be faster.

To find out more precisely what was the time interval between the sampling of input channel 0 and channel 1, the following test was done. A sawtooth signal was fed to an oscilloscope. The slope of the rising part of the wave was calculated, in mV/ms. The sawtooth signal was then connected to the two channels' input. With a timer rate much faster than the signal frequency, a few hundred samples were taken. The difference in readings between each channel for readings taken on the rising edge of the waveform was averaged. With the slope calculated earlier and the voltage range of the converter, the calculated time difference was 60 μ s $\pm$ 10 %. This time was considered short enough to be satisfactory and mode 20 was selected for the data acquisition.

The analog-to-digital converter had many options available but once each one was understood, the choice between them was fairly obvious. Although with hardware and software limitations, the DASH-16F was a useful tool. The implementation of each option was demonstrated in the example programs provided with the board. The data acquisition control program was written using as building blocks the various example programs provided and by adding the customized functions required by our specific application.

Appendix D

Scatter Diagrams

The following 8 scatter diagrams are the remainder of the diagrams not shown in chapter 4. For frequency diversity, spacing 5 MHz, 10 MHz, 15 MHz and 20 MHz are in figure D.1 to D.4. For space diversity, spacing 1λ , 1.5λ , 1.66λ and 2λ are shown in figure D.5 to D.8.

Figure- D.1: Scatter diagram for frequency diversity 5 MHz, all locations

Figure- D.2: Scatter diagram for frequency diversity 10 MHz, all locations

Figure- D.3: Scatter diagram for frequency diversity 15 MHz, all locations

Figure- D.4: Scatter diagram for frequency diversity 20 MHz, all locations

Figure- D.5: Scatter diagram for space diversity 1λ , all locations

Figure- D.6: Scatter diagram for space diversity 1.5λ , all locations

Figure- D.7: Scatter diagram for space diversity 1.66λ , all locations

Figure- D.8: Scatter diagram for space diversity 2λ , all locations

Appendix E

Measurement area and Conditions

E.1 Measurement locations

Figure E.1 shows a floor plan of the area where the measurements were taken. The receiver (marked as R on the floor plan) was located in one of the laboratory and the transmitter was moved to all the locations marked by a number. Table E.1 shows the location number and the corresponding separation from the receiver.

Locations and Distances			
Locations	distances (meters)	Locations	distances (meters)
1	1.68	11	16.89
2	3.78	12	22.68
3	5.33	13	22.62
4	7.19	5	20.85
5	12.19	15	27.19
6	12.07	16	34.14
7	14.94	17	32.55
8	20.97	18	11.34
9	14.51	19	34.63
10	13.84	20	17.50

Table E.1 *Locations and distances*

Figure- E.1: Floor plan of the measurement area and locations position

E.2 Measurement conditions

The frequency diversity measurements were taken from 20 March to 30 March 89 from about 1200 hrs to 1630 hrs during the day to get the maximum of activity in the halls and the classrooms. Most of the time, the two classrooms were half to three quarter full. About half the offices were occupied at one moment or another. The study rooms were also occupied. In the lab where the receiver was located, the person taking the measurements was always present plus one or two more. The doors of that lab were regularly open and closed as people would go back and forth to the adjacent computer terminal room.

The space diversity measurements were taken from 5 to 14 April 89 during the same period of the day. The activity level was about the same. The polarization diversity measurements were taken from 21 to 24 April also during the same period of the day. The activity level at that time was lower then for the two previous sets of experiments. The study rooms were still used but the classrooms were close to empty most of the time. Less people were walking in the halls. The activity level in the receiver lab and the adjacent terminal room stayed the same.

E.3 Hardware stability

There is two aspects that this section will cover: the frequency stability of the synthesizers and the gain stability of the transmitter power amplifier. The synthesizers used had a stability of 1 in 10^{-7} : this meant that for a transmit frequency of 915 MHz, there could be a drift of 90 Hz. Since the filter at the receiver had a bandwidth of 25 kHz and that we were using single tones only, the drift had no effect on the validity of the results.

After an initial warm up period (50 minutes), the gain of the transmitter power amplifier was stable within .2 dB. This warm up was done before every measurement. In addition, the diversity and the attenuation results were normalized so that variations in the absolute power level would not affect the results. The cumulative distributions

were calculated relative to the average received power of each set of measurements. A variation in the absolute power would not affect the results. The attenuation results were calculated relative to the attenuation of the closest location (1.72 m). This also removed any effect of absolute power variations on the results.

END

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