PROPAGATION MEASUREMENTS AT 55 GHz IN AN URBAN ENVIRONMENT

THESIS PRESENTED FOR EXAMINATION FOR THE DEGREE OF PhD AT THE UNIVERSITY OF LONDON

BY

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To Fiona

and

my Mother and Father

and

my Mother and Father-in-law
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The work described in this Thesis is concerned with the study of the propagation characteristics of 55 GHz millimetre waves between a fixed transmitter site and a mobile receiver terminal in an urban environment. Transmitter-receiver ranges up to 400 m are used and measurements are obtained in the presence of high traffic densities. The results of both analogue and digital (15 Mb/s data rate) measurements conducted on the link are presented.

The work has been motivated by the suggestion that a millimetre wave micro-cellular mobile radio system could form the basis of future urban mobile radio systems (Ref: Ruthruff C L, 1974; Steele R S, 1983; Huish P W et al, 1985; Norbury J, 1987).

The propagation of radio waves in random time variant media is discussed, and a product type model is developed which describes the non-stationary statistics of the received signal envelope over three different distance scale sizes. The variations on the different scale sizes describe: the general trend of the received signal to fall with increased receiver transmitter separation; the variation of the local average signal; and the "fast fading" of signal envelope.

Results indicate that transmitted power is contained within the micro-cell by the buildings which surround it. A power law ($r^{-16}$) describes the trend of the signal to fall with distance along the cell. Outside the micro-cell, the power law describing the reduction in signal power with distance is far steeper (approximately $r^{-10}$).

Comparison of experimental results with theoretical models of channel behaviour suggests that the received signal is dominated by line of sight propagation and specular reflections from principal scatterers. It is suggested that the prime cause of the variation of the statistical estimators of channel behaviour throughout the micro-cell is determined by the strength of the line of sight and specular component at each location.

The variation of the first order statistics is comparable with results reported from measurements at 900 MHz in suburban environments (Ref: IEEE VT Special Issue, 1988). This suggests that the propagation mechanism is similar in both cases. The results show that the distribution of the "fast fading" envelope of the signal is seldom described by a Rayleigh distribution.

The experimental results show an extreme range of coherence bandwidths from 17 MHz to 150 MHz throughout the micro-cell. The result is comparable with urban propagation measurements at 57 GHz (Ref: Violette et al, 1988).
The potential for applying space, frequency, and polarization diversity reception to the channel is investigated.

i) Auto-correlation measurements are used to estimate the antenna separation required to generate two un-correlated inputs for a space diversity system. The estimate varies from $2.5\lambda$ (1.4 cm) up to $22\lambda$ (12 cm).

ii) Two frequency measurements are used to estimate the frequency separation required between two input channels to provide de-correlated inputs for a diversity receiver. The estimated frequency separation to produce a correlation of 0.5 varies from 17 MHz to 150 MHz. These values are considerably larger than values of 25 KHz and 840 KHz reported from measurements in urban and suburban environments at 900 MHz (Ref: Clarke RH, 1968). This reflects the greater distances between major reflectors in the different environments.

iii) The effect of the micro-cellular environment on transmission of orthogonal polarizations is measured. The two co-polar coefficients $\Gamma_{vv}$ and $\Gamma_{vh}$ are found to be de-correlated as are the two cross polar coefficients $\Gamma_{vh}$ and $\Gamma_{hv}$. In addition, the value of $\Gamma_{hv}$ is shown to be greater than $\Gamma_{vh}$. These results agree with those reported at 900 MHz (Ref: Lee W C Y, Yeh Y S, 1975).

A simulation of two branch diversity reception, using experimental data, is undertaken. The results show a median signal level improvement of 1 dB and at the 99% reliability level, 9 dB. This compares with theoretical values assuming Rayleigh fading of 1.7 dB and 10.5 dB. This suggests that first order predictions of the performance of a diversity receiver operating in a micro-cell can be made by assuming Rayleigh fading input channels to a diversity receiver.

Experimental results obtained from a digital experiment operating at 15 Mb/s across the mobile link are compared with theoretical predictions. The results indicate that errors are primarily due to fading and that inter-symbol interference is not important. This is in agreement with the coherence bandwidth measurements. A prediction is made of the performance which would be expected from an ideal digital receiver using bi-phase-shift-keying and employing two branch maximal ratio combination. The prediction suggests that an error rate better than $10^{-3}$ could be obtained at 15 Mb/s data rate operating over a 400 m range.

The experimental apparatus used for both sets of experiments is described. The computer programs used for data analysis written by the author are included in the Appendix.
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Chapter 1 INTRODUCTION

This work is primarily concerned with the propagation of millimetre waves in the urban mobile radio environment. The motivation to undertake this work is based on the suggestion that a future mobile communication system, using a modified cellular radio principle, could be implemented in the millimetre wave frequency band (Ref: Steele R, 1983; Huish et al, 1985), thus easing spectrum congestion and providing the potential for wide-bandwidth communication. To put this work in context, a brief introduction to the development of cellular mobile radio is given.

1.1 THE MOBILE NETWORK

It is only recently that the possibility of a mass market for mobile communications has existed. Before the advent of cellular mobile radio, mobile communication was a scarce and expensive commodity. Figures 1.1 and 1.2 show how the structure of a cellular mobile radio system differs from that of a conventional mobile radio.

The concept of high capacity cellular mobile radio was originated at Bell Laboratories in the USA in about 1947. However, the idea was put forward in the public domain, in its most comprehensive form, in January 1979 (Ref: BSTJ Special Issue, 1979). Cellular mobile radio systems now in use, or being commissioned, are based on the advanced mobile ‘phone system (AMPS) developed by Bell Laboratories, or its Nordic equivalent the Nordic Mobile Telephone System (Ref: Makitalo O, 1988). The UK employs a system which is a variant of AMPS called Total Access Communication System (TACS) which has modified channel spacing and frequency allocations. The key concepts which are essential to the development of cellular radio are: trunking, frequency re-use, cell splitting and sectoring.

Trunking: Trunking is the ability of a system to combine several channels into a single group so that a mobile unit can be connected to any unused channel in the group for either an in-coming or an out-going call. This arrangement reduces blocking probability, thus considerably increasing traffic handling efficiency over the situation in which the mobile unit can only use one fixed channel. Dynamic channel allocation allows both transmit and receive channels to be re-allocated during a ‘phone call based on signal strength or interference parameters. The later generations of conventional mobile radio network use trunking. However, to obtain a significant increase in efficiency, a relatively large number of channels are
Chapter 1

needed. This is not generally possible with private mobile radio (PMR) and so spectral efficiency is not appreciably different from the single carrier per user situation (Ref: Dunlop J, Smith D, 1984; Young W R, 1979).

Figure 1.1 Structure of Conventional Mobile Radio

![Figure 1.2 Structure of Cellular Mobile Radio](image)

Figure 1.2 Structure of Cellular Mobile Radio
Chapter 1

**Frequency Re-Use:** Frequency re-use is not a new idea. It is used in the conventional mobile radio networks, public broadcasting and most other radio services. The new concept, however, is to employ frequency re-use on a small geographical scale. Instead of covering a large area with a single high power base station, as in PMR, the service is provided by distributed transmitters of moderate power throughout the coverage area.

**Cell Splitting:** The idea of cell splitting is that as communication traffic reaches capacity within one particular cell, that cell can be split in two with each new cell having a different frequency allocation. This process halves the distance between base stations and reduces the area to be covered by each cell by one quarter. Thus, a cellular system will have large cells to serve areas of low traffic density and small cells to serve areas of high traffic density. In this way, all users can be accommodated within a limited bandwidth allocation. At its inception, the system will have a relatively low number of users served by a few large cells each with omni-directional antennas. Cell splitting allows the possibility of a cellular radio system to grow organically according to demand. Thus, start up costs for a cellular network are low since only a few base stations need be constructed.

**Sectoring:** Sectoring is the process of splitting up individual cells by the use of directional antennas. This process allows capacity to grow without investing in new base station sites. Figure 1.3 shows a network which uses three 120° beamwidth antennas at each base station to split every cell into three sectors. The resulting system can be regarded as a corner fed network. The frequency allocations for each cell are divided into three sub-sets and each sector of the cell is fed by one of these sub-sets. With this arrangement, it is possible to employ a smaller re-use distance because the front to back ratio of the antenna reduces the co-channel interference. Thus, because the re-use distance is smaller, fewer cells are required per cluster, and more frequencies can be allocated per cell.
1.1.1 ANALOGUE CELLULAR RADIO

It is possible to employ any type of modulation in a mobile radio system. However, today's cellular mobile radio systems based on the AMPS system use narrow band FM with discriminator detection for voice transmission and FSK for data transmission. The exact channel spacing used varies between systems, but it is between 25 and 30 KHz. The maximum frequency deviation is about 12 KHz which does not leave a large guard band between adjacent channels. To avoid interference, frequencies are allocated so that adjacent channels are not used in the same cell. With current cellular radio systems, the maximum achievable traffic density is around 15 Erlangs per square kilometre (based on a cell radius of 3 Km). An Erlang is a measure of telephone traffic defined as the product of the number of calls per unit time multiplied by the average call length. If a circuit carries one call continuously, it is said to carry one Erlang of traffic (Ref: Dunlop J, Smith D, 1984). Estimates of the channel capacity required by an individual mobile vary, depending on assumptions made about call rate, duration, chance of lost calls, etc. However, a typical estimate is 0.025 Erlang per mobile user which gives a density of mobile users of 600 per square Km (Ref: Wiest G, 1987).
1.1.2 DIGITAL CELLULAR RADIO

The use of digital modulation and channel coding has several advantages over analogue techniques in the context of mobile radio. The effects of co-channel interference can be reduced by the use of dynamic channel assignment. Dynamic channel assignment allows the same frequency to be used more often in the cell plan and, as a result, more frequencies can be allocated per cell. Typically this will allow a further doubling of the traffic density (Ref: Swerup J, 1987). In addition, narrow band TDMA allows the use of cell radii down to 100 m. This is possible because of the very fast and reliable "hand-off" procedures as the mobile moves from one cell to another (Ref: Swerup J, 1987). Using a TDMA system with 100 m cell spacing, the maximum traffic handling capacity can be increased from 15 Erlangs per square Km which is a typical value for analogue cellular radio, to 200 Erlangs per square Km. This increases user density from 600 to 8000 per square Km.

1.2 FUTURE MOBILE RADIO SYSTEMS

If the growth in the demand for mobile telecommunication continues, the enhanced capacity of 900 MHz digital radio will be insufficient. However, there seems to be no immediate ceiling to the demand for mobile services. In some Scandinavian countries the number of subscribers has reached several percent of the population and the demand is still growing. Some reports suggest that a mass market reaching 50% of the population might be the ultimate result (Ref: Makitalo O, 1988). Also, the growth in data communication over the public network can be expected to lead to a demand for high rate data communications to mobile terminals.

It is apparent that a mobile radio system serving a mass market, approximately 10% of the UK population, will not be able to function using the current frequency allocation. In addition, if broad-band capabilities are offered on the mobile network, then the pressure on the current frequency allocations will become even more intense.
1.3 COMPETITION FOR AVAILABLE BANDWIDTH

At present, all mobile radio systems are located in the 100 MHz to 1 GHz band. However, mobile communication radio must compete for spectrum with a variety of other users. This competition for spectrum is particularly fierce in the region between 100 MHz and 1 GHz, because the radio technology is very well established at these frequencies below 1 GHz and it is relatively cheap and simple to establish a communication system. In addition, above 10 GHz for mobile communication systems must take atmospheric and climatic conditions into account (Ref: Lee W C Y, 1982).

As a result of this competition, there is considerable interest in identifying new frequency allocations for mobile radio which lie outside the 100 MHz to 1 GHz region. However, although there are some frequency bands below 1 GHz which may provide extra bandwidth for mobile communications (Ref: Macario R, Grimm F, 1985), the demand is such that new frequency allocations above 1 GHz must be identified.

1.3.1 MILLIMETRE WAVE OPTION

In the search for new frequency allocations, attention has recently turned to the use of frequencies in the millimetre waveband for land mobile radio use. The millimetre waveband covers the range of frequencies from ~10 GHz through to ~300 GHz. However, the part of the band which has been suggested for mobile use lies between 40 GHz and 60 GHz. Millimetre wave frequencies potentially offer a vastly greater bandwidth than is currently available and they are relatively free from other users (Ref: Huish et al, 1985; Steele R, 1983).

Propagation measurements in the millimetre waveband have primarily been confined to measurements over static links. This is because the concept of using the millimetre waveband for mobile applications is relatively new. However, several initial propagation studies have been undertaken which indicate that, over short distances, with predominantly line of sight propagation, millimetre wave frequencies are able to support mobile communication links (Ref: Ruthruff C L, 1974; Kibler L U, 1974; Huish et al, 1985; McGeehan J P, 1985; Thomas H J et al, 1986).
The use of millimetre wave frequencies has been suggested as a solution to spectral overcrowding of mobile radio channels in dense urban centres. Steele has proposed a system architecture based on a network of micro-cells (Ref: Steele R, 1983). A system such as this would use quasi line of sight propagation and, as such, would be in accord with the observed propagation characteristics of millimetre waves.

1.3.2 THE MICRO-CELLULAR CONCEPT

Micro-cellular radio is a development of sectoring. However, instead of using antenna directivity to reduce co-channel interference, the local geometry of the buildings is used to define cell boundaries. In addition, because the cells are shaped to the geometry of the streets, typically being 200 - 300 m along particular sections of road, a micro-cellular network can provide a more uniform coverage of the service area. This may reduce the possibility low signal strength "holes" occurring in the network, where a mobile could not get access to the system due to insufficient signal strength.

The micro-cellular concept proposed by Steele is different to a cellular system which simply uses very small cells, because, rather than using constant frequency switching as the mobile moves between micro-cells, each mobile unit is allocated a single frequency slot for the whole duration of its stay within the group of micro-cells constituting a cell. The aim of this procedure is to reduce the number of "hand-offs" which the mobile unit has to deal with. The problem of "hand-offs" is passed onto the local base station which becomes responsible for the co-ordination of frequency allocations within the cell.

However, a system using digital modulation would most likely not need to use this frequency handling procedure, because digital modulation facilitates the use of very fast "hand-offs" between cells. Indeed, the planned 900 MHz Pan European digital mobile radio system, based on time domain multiple access (TDMA), will, in theory, allow the use of cells of 100 m radius without resorting to the use of special frequency allocations for mobile units (Ref: Swerup J, 1987). Thus, a millimetre wave micro-cellular mobile radio system might conceivably resemble a scaled down, wide-bandwidth, digital version of today's radio systems.
1.3.3 PERSONAL COMMUNICATION

One idea for a future millimetre wave mobile radio system which has been suggested is that it could be used to provide personal hand-held communication. The very small user-base station distances found in a micro-cellular network allow the use of very low power hand-held transceiver units and the small wavelength used allows the use of small antennae. A system of this type could be extended beyond car mounted ‘phones to include the possibility of personal communication, whereby each individual citizen would have their own hand-held, "go anywhere, speak to anyone in the world", pocket telephone (Ref: Steele R, 1983; Makitalo O, 1988).

1.4 THE CASE FOR THE MILLIMETRE WAVE SOLUTION

There are several factors which have to be considered to assess the potential for millimetre wave mobile radio. First, there are a number of advantages which a millimetre wave system would possess:

i) Very large available bandwidth which could be utilized to provide a combination of a very large number of channels and high data rate communication.

ii) Possibility of very small size implementation for personal transceiver unit.

iii) Small re-use distance due to quasi line of sight propagation and atmospheric absorption, thus allowing a high user density.

However, there are three possible disadvantages which are of prime importance in assessing the feasibility of commercial urban millimetre wave mobile radio (Ref: Thomas H J et al, 1986):

i) The cost of millimetre wave components is prohibitively high.

ii) Multipath fading potentially provides a very difficult propagation medium for mobile radio at millimetre wave frequencies.

iii) The complexity of network co-ordination. A mobile communication system based on a network of micro-cells will require a very complex and sophisticated co-ordination methodology for dealing with communication between micro-cells, connection to the public network and "hand-offs".
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The first of these problems, component expense, is expected to become less severe as millimetre wave component costs drop with the introduction volume production and advances in MMIC technology (Ref: Booton R C, Fetterman H R, 1988). The second problem, the difficult propagation channel at millimetre wave frequencies, can only be addressed when a full understanding of the propagation characteristics of millimetre waves in the urban environment is available. The third problem, the network architecture of a micro-cellular radio system has not been considered in detail. However, it is expected that if the propagation problems can be overcome, and given the expected drop in component prices, then a solution will be found to this problem (Ref: Steele R, 1983).

1.5 MILLIMETRIC MOBILE RADIO STUDIES

Several propagation studies have been carried out by various workers in the field to measure the characteristics of millimetre waves in the mobile environment. The studies have been carried out in a variety of different mobile radio environments. Brief overviews are given of these studies, their various shortcomings are discussed and, against this background, the relevance of the work described in this Thesis is presented.

Kibler (Ref: Kibler L U, 1974) conducted a basic propagation study at 60 GHz which probed the received signal strength distribution along a few streets in Denver, Colorado. The study was carried out using directive antennas and a low sensitivity receiver and so it did not address the possibility of using scattered radiation to provide a communication path in the absence of a line of sight signal.

Violette (Ref: Violette et al, 1983) conducted a propagation study in downtown Denver, Colorado. This is a densely built up area. The study was carried out at night in the absence of other cars, and hence no moving scatterers were present. In addition to these measurements, a large amount of data was gathered about the reflective properties of common building materials. A second study by the same group (Ref: Violette et al, 1988) has recently been carried out using a wideband digital technique to probe the impulse response of the mobile path. This study was carried out at 30 GHz. However, neither of these studies was focused solely on the possibility of millimetre wave cellular radio. In particular, the antennae used were narrow beamwidth, 1.2° at 57.6 GHz and 30° at 30 GHz, and the study was carried out in the absence of other vehicles. These factors distort the received signal characteristics compared to those expected in a mobile radio system (Ref: Lee W C Y, 1982).
Chapter 1

Pugliese and Alexander (Ref: Pugliese G, Alexander S E, 1983) were responsible for conducting one of the first studies of millimetre wave propagation inside a building to see whether a millimetre wave communication system were a possibility.

McGeehan (Ref: McGeehan J P, 1985) has conducted a preliminary propagation study on a university campus. This study was carried out without any moving scatterers or large buildings present. The study examined the effects of blocking by trees and multipath propagation.

Huish et al (Ref: Huish et al, 1985) conducted a propagation study in a rural environment. Some moving scatterers were present, but the density of scatterers, both moving and stationary, was very low. The study also examined the effects of propagation on digital error rates.

1.5.1 CRITIQUE OF THE STUDIES

The studies summarized above are very valuable for indicating that a millimetre wave cellular radio system may be a feasible proposition in the future. However, they have been made in environments with low traffic density, and in some cases in low rise suburban areas. Thus, these studies do not fully address the problems which would be encountered in an urban millimetre wave mobile radio scenario.

In particular, the studies do not address the problem of propagation in the presence of other moving vehicles. Carrying out measurements in the presence of other vehicles is very important for the case of micro-cellular radio, because it is other vehicles which provide some of the main obstructions to the LOS signal in a micro-cell.

The argument which is put forward to justify carrying out propagation measurements in zero traffic conditions for conventional cellular mobile radio at 900 MHz is that with cell radii of 2 - 3 Km, it is buildings and other major structures which provide the main sources of scatterers, with cars only adding a small perturbation (Ref: Cox D C, 1973). However, this argument is clearly not valid in a micro-cell which does not have any buildings obstructing the LOS path. Thus, to be valid, propagation measurements must address the influence of traffic levels on propagation characteristics.
1.6 DESCRIPTION OF PRESENT WORK

The object of this Thesis is to investigate propagation effects in the urban environment at millimetre wave frequencies.

The study is divided into two parts. The first deals with analogue measurements. The second subsidiary study deals with an investigation of digital propagation. The propagation measurements are carried out at a carrier frequency of 55 GHz. The choice of this frequency and the validity of generalizing the results to the 40 - 60 GHz band which has been suggested for mobile use, is justified in Section 1.6.1.

1.6.1 THE CHOICE OF 55 GHz

The suggested frequencies for millimetre wave mobile radio lie in the range 40 - 60 GHz (see Section 1.3.1). However, it would be very difficult and expensive to determine the propagation characteristics over the whole of this band. Thus, it was decided to choose a single carrier frequency, 55 GHz, with which to characterize the general properties of the whole band of frequencies. 55 GHz was chosen for a number of reasons:

i) Availability of experimental equipment. A number of 55 GHz oscillators and mixers are available in the Department for carrying out propagation measurements.

ii) 55 GHz lies in the 40 - 60 GHz band where interest in micro-cellular use has been shown.

The generalization of results obtained in the vicinity of 55 GHz for determining the approximate characteristics over the range of frequencies from 40 - 60 GHz is justified, because the line of sight characteristics of millimetre waves across the frequency band are known (Ref: Gibbins C J, 1988) and the relationship between the wavelength, \( \lambda \), and the average size of the scatterers in the mobile environment, \( L \), does not change significantly. For all frequencies in the band the inequality \( \lambda \ll L \) is true. In addition, because the wavelengths are similar, the scattering characteristics (eg road, building surfaces) will be much the same across the band.
1.6.2 ANALOGUE MEASUREMENTS

In Chapter 2, the statistical character of the received signal in the mobile radio environment is considered theoretically. The idea of describing the channel as the product of two processes is introduced and the mathematical characterization of each process presented. The predictions of the theoretical models available in the literature are then summarized. This Section establishes the mathematical framework on which the propagation study is based. The experimentally measurable parameters which are of interest for predicting the performance of a millimetre wave system over a mobile radio channel are then examined.

The experimental equipment used to carry out the analogue propagation measurements is described in Chapter 3. The various measurements require several different experimental arrangements and these are described in Sections 3.2 through 3.9. The analogue results are presented in Chapter 4.

1.6.3 DIGITAL MEASUREMENTS

In the second part of the study, the effect of multipath propagation on digital error rates is addressed. The effect of a multipath channel, with Rayleigh fading, on the bit error rate (BER) of a digital link is examined theoretically in Chapter 5. The digital receiver used in this Thesis uses a fixed threshold detection scheme and the predicted error rate for this type of detection is derived.

The transmitter and receiver assemblies, built for the analogue measurements, were modified to allow the mobile link to be used with digital modulation. The experimental equipment used for the digital measurements is described in Chapter 6. The propagation parameters which this study measured were BER, received signal level and car position. The results of the digital measurements are presented in Chapter 7.

1.6.4 CONCLUSION

Chapter 8 summarizes the results of the analogue and digital propagation studies illustrating how the performance of a future digital mobile link may be inferred from the results of the analogue study.
Chapter 2  THE MOBILE RADIO CHANNEL

2.1  INTRODUCTION

The received signal in the mobile radio environment is observed to suffer rapid and extreme variations in amplitude and phase as the mobile moves distances of the order of the free space wavelength $\lambda$. In addition, the mean value of the signal intensity is observed to fluctuate as the mobile moves over larger distances.

The response of the mobile radio channel is often considered as the product of two processes with quite different characteristics. The first process, commonly referred to as "fast fading", is responsible for the rapid and extreme variations in the amplitude and the phase of the received signal as the mobile unit moves over short distances. This is attributed to multipath fading. The received signal is modelled as the super-position of a number of delayed replicas of the transmitted signal. Displacement by a short distance randomly re-arranges the phase of these component waves producing constructive or destructive interference. The second process, commonly referred to as "slow fading", is responsible for the gradual change in the average signal power from location to location throughout the micro-cell. This is attributed to variations in the local topology of the scattering environment which result in shadow loss or signal enhancement. Signal enhancement can be caused by very large buildings which act as narrow beamwidth reflectors (Ref: Cox D C, 1973).

In addition to these two random processes, there are deterministic effects to be considered. First, there is free space propagation loss, or spreading loss, which results in an $r^{-2}$ dependence on signal strength. In most cases, this power law is modified by the mobile radio environment. For example: channelling effects may reduce the attenuation with distance (Ref: Pugliese G, Alexander S E, 1983), or diffraction and obstruction effects may increase the attenuation with distance (Ref: Bullington K, 1957). Secondly, at millimetre wave frequencies there are atmospheric effects to be considered; oxygen causes significant attenuation, and precipitation causes a combination of scattering and absorption. However, it has been shown (Ref: Huish et al, 1985) that, over the short ranges involved in micro-cellular radio, the effects of atmospheric absorption are not major contributors to propagation loss (Ref: Huish et al, 1985).
However, before we consider the task of decomposing the channel response into precisely defined "fast fading" and "slow fading" processes with distinct statistical characteristics, we need to define the expressions for the transmitted and received signals, introduce the concept of the equivalent low-pass response of the essentially band-pass mobile radio channel and present the description of the mobile radio channel in terms of a time variant impulse response and a time variant transfer function. It must be emphasized that the envelope of the received signal is equal to the magnitude of the complex signal envelope. When "envelope" appears alone, without the preceding "complex", it refers to the magnitude of the complex envelope.

### 2.2 DEFINITION OF TRANSMITTED SIGNAL

The transmitted signal can be described as the real part of a complex exponential

\[ S_0(t) = u(t)e^{j(2\pi f_c t + \phi_0)} \tag{2.1} \]

where: \( S_0(t) \) = transmitted signal at time \( t \), \( u(t) \) = envelope of transmitted signal, \( f_c \) = carrier frequency, \( \phi_0 \) = phase.

The complex exponential form is used because it is easier to manipulate mathematically than the equivalent trigonometric expression. However, it is important to note that the power in the signal is related only to the real part of the complex waveform. The envelope of the transmitted signal is constrained such that \( f_u \ll f_c \) where \( f_u \) is the maximum frequency component of \( u(t) \).

### 2.3 DEFINITION OF RECEIVED SIGNAL

The received signal is the sum of all the reflections from the scatterers in the mobile radio environment. Each reflection will be a delayed, weighted and Doppler shifted replica of the transmitted signal. In general, the weighting factors will be complex. For the case where the mobile unit moves with velocity \( V \) and the scatterers are fixed, the received signal is given by the sum over the set of all reflections:

\[ S(t) = \sum_{i=1}^{N} a_i S_0(t - \tau_i) e^{-j2\pi(Vt/\lambda \cos \theta_i)} \tag{2.2} \]
where: $\tau_i$ = delay time of the $[i]$th path, $a_i$ = amplitude weight of the $[i]$th path, $N$ = total number of paths, $V$ = vehicle speed, $\theta_i$ = angle between $[i]$th wave and vehicle motion.

**Complex Envelope of the Received Signal:** An expression for the complex envelope of the received signal, $x(t)$, can be obtained by substituting 2.1 into 2.2 and dropping the carrier term and steady state phase $\phi_o$, thus:

$$x(t) = \sum_{i=1}^{N} a_i e^{-j2\pi(V/\lambda \cos \theta_i)(t - \tau_i)}$$

2.3

The expression $x(t)$ is the output of an ideal linear receiver consisting of a distortionless down converter followed by a low-pass filter. This gives rise to the description of the mobile radio channel which is an essentially band-pass phenomenon, in terms of its equivalent, low-pass filtered, baseband response. This concept is used frequently to describe communication channels. An exposition of the mathematical consequences of this description is given by Mortensen (Ref: Mortensen RE, 1987, pp 147 - 148). In the general case both $a_i$ and $\theta_i$ are dependent on time. In effect, a different set of $a_i$ and $\theta_i$ are used to evaluate $x(t)$ at each time, $t$. To proceed further, some assumptions must be made about the character of multipath propagation. In particular, we need to estimate the distribution of the arrival angles and delays.

**Time Variant Impulse Response:** In order to characterize the equivalent low-pass response of the mobile radio channel, we need to develop the input-output relationship between the complex input envelope $u(t)$ and the complex output envelope $x(t)$. One way in which this relationship may be represented is via the equivalent low-pass impulse response of the channel. However, the properties of the mobile radio channel vary with time. Thus, the impulse response must also vary with time. For the case where the mobile vehicle moves at velocity, $V$, and the scatterers are fixed, the input-output relationship may be expressed [1]:

$$x(t) = \int_{-\infty}^{\infty} h(t, \tau) u(t - \tau) d\tau$$

2.4

---

1 The integration limits are set to $\pm$ infinity. However, in reality, the analytic function $h(t, \tau)$ will be zero for $t < 0$. Thus, the lower limit on the integral could be replaced by 0 without loss of generality.
Chapter 2

Time Variant Transfer Function: The input-output relationship of the channel may be represented in the frequency domain by the time variant transfer function $T(t,f)$. $T(t,f)$ may be obtained by taking the Fourier transform of the impulse response function $h(t,x)$. Thus:

$$T(t,f) = \int_{-\infty}^{\infty} h(t,\tau)e^{-j2\pi f\tau}d\tau$$  \hspace{1cm} 2.5

Alternatively, $T(t,f)$ may be viewed as the complex envelope of the channel resulting from an excitation $\cos [2\pi (f_c + f)]$, where $f_c$ is the carrier frequency at which the channel is excited. The transfer function is denoted as $T(t)$, rather than as the usual $H(t)$, to avoid confusion with the generalized two frequency transfer function of a time variant process which is the double Fourier transform of $h(t,\tau)$ (Ref: Bello P A, 1963; Parsons J D, Bajwa A S, 1982; Bendat J S, Piersol A G, 1986). For an arbitrary input envelope $u(t)$ which has a frequency spectrum $U(f)$, the time varying complex envelope of the output may be expressed:

$$x(t) = \int_{-\infty}^{\infty} U(f)T(t,f)e^{j2\pi ft}df$$  \hspace{1cm} 2.6

2.4 ANALYSIS OF NON-STATIONARY PROCESSES

The analysis of non-stationary processes is simplified by making assumptions about the character of the non-stationarity involved. In some circumstances, this can lead to the use of a transformation to decompose the non-stationary process into a stationary process and an associated non-stationary, or deterministic, trend. Bendat and Piersol present an exposition of some approaches which can be made to decompose non-stationary processes in this way (Ref: Bendat J S, Piersol A G, 1986). Several simplifying assumptions are commonly applied when dealing with the analysis of mobile radio channels:

---

1 The convention of using $2\pi f$ rather than $\omega$ for Fourier transforms is adopted to avoid confusion with factors of $1/2\pi f$ when performing inverse transforms.
Chapter 2

i) The process is regarded as quasi-stationary. That is, for input bandwidths, $W$, and output time durations, $T$, of interest, the statistical properties of the channel can be regarded as constant (Ref: Bello P A, 1963). The process is also regarded as quasi-ergodic. That is, the statistical properties of the channel can be estimated from a single sample record of the process of duration $T' < T$.

ii) The channel is regarded as a product type non-stationary process which can be decomposed into a stationary "fast fading" process and a non-stationary, but slowly varying function representing the local mean value of the envelope at time, $t$. The channel response is decomposed into a "fast fading" and a "slow fading" process which may be separated on a single sample record of the received signal using a "moving average" process. The channel response is then normalized with respect to the slowly varying mean value to yield a stationary process.

iii) In addition, it is common to assume Gaussian statistics for the channel which means that the statistical properties of the channel are completely defined from a knowledge of its mean value and auto-correlation function. This assumption is based on application of the central limit theorem. It is valid so long as the channel possesses a sufficient number of scatterers to meet the criteria of the theorem and provided that exact modelling of the extremes of the statistical distribution is not required. The various correlation properties of the Gaussian distribution are well known which considerably simplifies the prediction of the spectrum and auto-correlation of the channel.

The adoption of these assumptions leads to the description of the channel response by a two stage model, one stage describing the statistics of the locally stationary "fast fading", and the other, the statistics of the slower variation in the local mean value.

2.4.1 THE QUASI-STATIONARY MOBILE RADIO CHANNEL

The description of the mobile radio channel as a quasi-stationary channel follows intuitively from the observed character of the response of the mobile radio channel which exhibits very rapid amplitude and phase fluctuations for small changes in the mobile location and more gradual changes in the received signal intensity over longer distances. The quasi-stationary model describes the channel as a locally stationary process with a slowly varying mean square value.
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From a statistical point of view, the variation of the channel response to translations in time and frequency is described by the auto-correlation function $T(t,f)$. When $t$ and $f$ are fixed at $t'$ and $t''$, the auto-correlation function $R_{t',t''}(t,f)$ describes the way in which $T(t,f)$ becomes de-correlated for a frequency interval $\Omega$ and a time interval $\tau$ centred on the "local" time frequency co-ordinates $t', t''$. For the general case this may be expressed:

$$R_{t',t''}(t,f) = E \left[ T^* \left( t' - \frac{\tau}{2}, f' - \frac{\Omega}{2} \right) T \left( t'' + \frac{\tau}{2}, f' + \frac{\Omega}{2} \right) \right]$$

where "E" is the expectation operator.

The non-stationary variation of the channel response with time and frequency translations is described by the variation of $R_{t',t''}(t,f)$ with changes in the "local" co-ordinates $t'$ and $f'$. However, it is sufficient to describe the slow non-stationary changes simply using $R_{t',t''}(0,0)$ (Ref: Bello P A, 1963). In the case where the statistics of the channel are stationary, $R_{t',t''}(t,f)$ would become independent of $t'$ and $f'$, ie:

$$R_{t',t''}(t,f) = R(t,f)$$

The time selective behaviour of the channel, caused by the variation of $T(t,f)$ with time, is commonly referred to as fading. However, it is useful to generalize this concept of fading to describe the frequency selective behaviour of the channel caused by the variation of $T(t,f)$ with frequency, as frequency fading.

To define the class of quasi-stationary channels, assume that the input waveform has a constraint on bandwidth and that the output waveform has a constraint on time duration (Ref: Bello P A, 1963). Let the bandwidth and time constraints be denoted by $W$ and $T$, centred on $f'$ and $t'$ respectively ($f = 0$ represents the carrier frequency). The quasi-stationary channel is defined as the set of channels for which $R_{t',t''}(t,f)$ changes negligibly with translations by intervals of $w < W$ and $t < T$ in frequency and time respectively. This situation can be represented by the following inequalities:

$$\frac{\partial R_{t',t''}(t,f)}{\partial f} \big|_{\text{max}} \leq \frac{1}{W}, \quad \frac{\partial R_{t',t''}(t,f)}{\partial t} \big|_{\text{max}} \leq \frac{1}{T}$$
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Bello then demonstrates that if the inequalities of Equation 2.9 hold, the actual channel may be replaced by a hypothetical wide-sense-stationary uncorrelated scattering (WSSUS) channel, subject to the constraints on time duration and input bandwidth, at least in so far as the determination of the correlation function of the channel output is concerned (Ref: Bello P A, 1963). If Gaussian statistics are assumed, then the channel may be replaced by a WSSUS channel for the estimation of any statistical quantity. The WSSUS channel is frequently used in the literature to describe mobile radio channels because it is the simplest non-degenerate channel which exhibits both frequency and time fading.

2.4.2 DEFINITION OF "SLOW FADING" AND "FAST FADING"

The term "fast fading" is used frequently throughout the literature. It refers to the fluctuation of the transfer function with small scale translations in time or frequency which are very much smaller than T or W respectively. "Slow fading", on the other hand, refers to the change of the "local mean value" of the envelope of the transfer function with translations in time and frequency which are of similar order to the constraints T and W.

It should be noted that in a large number of experimental applications the time variant transfer function is regarded as a distance variant transfer function, where the distance constraint, in the case where the car is moving with constant velocity, is given by \( D = V \times T \) where V is the velocity of the mobile vehicle. In this case, "fast fading" and "slow fading" are defined in a similar manner to that above, but in terms of translations by intervals of distance, d, which are either much smaller than D or on a similar scale to D.

2.4.3 PRODUCT MODEL OF THE QUASI-STATIONARY CHANNEL

This model describes the situation where the channel has statistical quantities which are locally constant, but which are subject to a modulation by a slowly varying deterministic function \( a(t) \). The description of the quasi-stationary channel as a product type non-stationary process, implies that:

\[
T(f,t) = a(f,t)T_0(f,t)
\]  \hspace{1cm} 2.10
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where \( a(f, t) \) is a slowly varying deterministic function of frequency and time and \( T_0(f, t) \) is a unity variance stationary process. In all practical cases, however, we assume that \( a(f, t) \) does not vary with frequency. This is equivalent to assuming that the reflective and refractive properties of the terrain and buildings remain unchanged for frequency bandwidths of practical interest which is a reasonable physical assumption (Ref: Gans M J, 1972). Thus:

\[
a(f, t) = a(0, t) = a(t)
\]

The auto-correlation function of \( T(t, f) \) is given by:

\[
R_{f, f}(\tau, \Omega) = a^2(t')E\left[T_0\left(f' - \frac{\Omega}{2}, t' - \frac{\tau}{2}\right)T_0\left(f' + \frac{\Omega}{2}, t' + \frac{\tau}{2}\right)\right]
\]

\[
= a^2(t')R_0(\tau, \Omega)
\]

If we assume that \( a(t) \) is a non-negative function of \( t \) which is a reasonable assumption for physical processes (Ref: Bendat J S, Piersol A G, 1986), then we can write:

\[
|T(t, f)| = a(t)|T_0(t, f)|
\]

If we re-define the unity variance process \( \{T_0(f, t)\} \) such that the expectation value of its modulus is unity, and using Equation 2.11, we can write:

\[
E[|T(t, f)|] = a(t)E[|T_0(t, f)|] = a(t)E[|T_0(t, 0)|] = a(t)
\]

where \( a(t) \) is equal to the local mean value of the envelope of the channel response.

The mobile radio channel response is usually considered to be a spatial phenomenon, where the time dependence of the envelope only arises from the movement of the vehicle. If we express Equation 2.14 in terms of \( Vt = s \), then the spatial aspect of the phenomenon is shown explicitly:

\[
E[|T(s, 0)|] = a(s)E[|T_0(s, 0)|] = a(s)
\]

where \( a(s) \) is equal to the local local mean value of the envelope of the channel response. The process is stationary over a distance interval \( S = VT \), where \( T \) is the time constraint for which the process can be assumed to be stationary.
2.4.4 NORMALIZING THE CHANNEL RESPONSE

If we assume that the channel response is approximately an ergodic function of distance, ie if the variations of a(t) are very slow compared to the lowest frequency of T_0(f,t), then we can estimate a(s) from a single sample record of the envelope of the channel response by low-pass filtering. In this Section we assume a transmitted carrier frequency with unity amplitude. Thus, the envelope of the received signal is equivalent to the envelope of T(s,0), ie \{r(s)\} = \{T(s,0)\}. The estimate of a(s), denoted \(\hat{a}(s)\), can be used to transform the locally stationary received signal envelope, r(s), to an estimate of the unity mean stationary process \(r_0(s)\). Hence:

\[
\hat{r}_0(s') = \frac{r(s')}{\hat{a}(s')}
\]

2.16

However, since we cannot form the ensemble average of the process \{r(s)\}, we must estimate a(s) by the distance integrated average of the received signal envelope for a transmitted carrier frequency. If we assume a(s) is constant over the integration interval \(S'\), we can write:

\[
\hat{a}(s') = \frac{1}{S'} \int_{s'-S'/2}^{s'+S'/2} \left| T(s,0) \right| ds
\]

\[
= \frac{1}{S'} \int_{s'-S'/2}^{s'+S'/2} r(s)ds
\]

\[
= a(s') \frac{1}{S'} \int_{s'-S'/2}^{s'+S'/2} r_0(s)ds
\]

2.17

where: s = distance, a(s') = true local mean at s', \(\hat{a}(s')\) = estimate of local mean at s', x(s) = complex envelope of received signal, r(s) = absolute magnitude of x(s), \(r_0(s)\) = normalized magnitude with respect to the local mean value a(t), \(f_c\) = carrier frequency, T(s,0) = complex envelope of transfer function t(s,f) at \(f_c\).

However, to apply Equation 2.16, we must know the value \(S'\) over which Equation 2.17 is to be integrated. The distance S over which the statistics of the process are approximately stationary is determined by the distance over which the effects of terrain and buildings remain approximately unchanged. Gans suggests
that for urban propagation this is approximately equal to the distance between the
centre of a block and an intersection (Ref: Gans M J, 1972). The interval over
which the integral is evaluated must satisfy $S' \ll S$. However, this is of no practical
use unless we know the details of the local topography.

Lee took a different route to estimate the integration length. Using a simplified
mathematical model of the channel, he calculated the variance of the estimate of
the local mean value as a function of the integration length. To produce his result
he used Clarke’s model of the mobile radio channel (Ref: Clarke R H, 1968) to give
estimates for the auto-correlation for the received signal envelope and for the
auto-correlation of the envelope squared. The expectation value of the variance of
$\hat{a}(s)$, using Papoulis to express the variance in terms of the auto-correlation
function (Ref: Papoulis A, 1965), is given by:

\[
\sigma_{d(s')} = E[\hat{a}^2(s')] - a(s')^2
\]

\[
= a(s') \frac{1}{S'} \int_{0}^{S'} \left( 1 - \frac{s}{S'} \right) (R_{\sigma}(s) - 1) ds
\]

2.18

Lee then tabulates the numerical evaluation of Equation 2.18, for various values of
integration length, giving the standard deviation of the estimate as a fraction of the
true mean value, $a$. This Table is reproduced as Table 2.1.

<table>
<thead>
<tr>
<th>Integration Length</th>
<th>Standard Deviation of Mean $\sigma$, in terms of true mean $a$</th>
<th>Ratio of Error to True Mean Value for 1$\sigma$ and 2$\sigma$ spreads</th>
</tr>
</thead>
<tbody>
<tr>
<td>$5\lambda$</td>
<td>0.165$a$</td>
<td>3 dB</td>
</tr>
<tr>
<td>$10\lambda$</td>
<td>0.122$a$</td>
<td>2.1 dB</td>
</tr>
<tr>
<td>$20\lambda$</td>
<td>0.09$a$</td>
<td>1.6 dB</td>
</tr>
<tr>
<td>$40\lambda$</td>
<td>0.06$a$</td>
<td>1 dB</td>
</tr>
</tbody>
</table>

Table 2.1 Variance of Local Mean Estimate Versus Integration Length

The Table indicates the 1$\sigma$ and 2$\sigma$ dB spread of the mean value estimate of
envelope amplitude for various values of integration length. Thus, for $40\lambda$, the
mean value estimate is within 1 dB of the true mean value for 86% of the time.

If square law detection is used, then we are concerned with the estimate of the
variance of the amplitude squared, i.e. the mean power. Define the true mean
power as, $p$, and its time integrated estimated value as $\hat{p}$. For a Rayleigh
distribution the mean amplitude and the mean power are related by $\hat{p} = 4\hat{a}^2 / \pi$. To
evaluate the variance of the mean power estimate for different integration lengths,
Lee uses a similar argument to that above, but using the Clarke expression for the auto-correlation function of the envelope squared. The numerically evaluated Table 2.2 is reproduced below.

<table>
<thead>
<tr>
<th>Integration Length $S'$</th>
<th>Standard Deviation of Mean Power $\sigma_p$ in terms of true mean $p$</th>
<th>Ratio of Error to True Mean Value for $1\sigma$ and $2\sigma$ spreads</th>
</tr>
</thead>
<tbody>
<tr>
<td>$5\lambda$</td>
<td>0.33$p$</td>
<td>2.98 dB</td>
</tr>
<tr>
<td>$10\lambda$</td>
<td>0.244$p$</td>
<td>2.14 dB</td>
</tr>
<tr>
<td>$20\lambda$</td>
<td>0.18$p$</td>
<td>1.55 dB</td>
</tr>
<tr>
<td>$40\lambda$</td>
<td>0.12$p$</td>
<td>1.0 dB</td>
</tr>
</tbody>
</table>

Table 2.2 Variance of Local Mean Power Estimate Versus Integration Length

The accuracy of the distance integrated estimates of $a(s')$ are similar for calculations based on the envelope $r(s)$ or the power $p(s)$. The above Tables represent the situation where there are an infinite number of scatterers in the mobile radio environment. However, in reality, there will only be a limited number of scatterers. To investigate the dependency, the variance of the estimate on $N$, the number of scatterers, Lee produced a computer simulation for several different values of $N$. His results show that the above spreads in the mean values are conservative, the true spreads being somewhat less than indicated in Table 2.1 (Ref: Lee W C Y, Yeh Y S, 1974). Thus, the above values are valid for use in filtering routines for separating the "fast fading" and "slow fading" processes of the received signal envelope.

Thus, we have a mechanism for separating the "fast fading" and "slow fading" processes which constitute the received signal envelope. We shall now consider the detailed analysis of each process separately.

2.5 ANALYSIS OF "FAST FAADING"

2.5.1 INTRODUCTION TO THE THEORETICAL MODELS

The techniques of statistical communication theory have been applied to the problem of analysing the "fast fading" of the received signal in the mobile radio environment by a number of authors; for example (Ref: Ossanna J Jr, 1964), (Ref: Clarke R H, 1968), (Ref: Gilbert E N, 1965), (Ref: Gans M J, 1972), (Ref: Jakes W C Jr, 1974), (Ref: Aulin T, 1979), etc. These analyses have predicted many of the observable properties of the received signal.
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The theoretical work has concentrated on the propagation of 400 MHz - 900 MHz radio waves in urban and suburban environments where today's mobile cellular systems operate. However, the theories are based on the "high frequency" approximation that the free space wavelength $\lambda$ is very small compared to the average size of scatterers in the mobile radio environment. Thus, it seems reasonable to use existing theory to model the millimetre wave scenario.

There are two aspects of the mobile radio propagation phenomenon which the models must address. First, there is the distribution of the received signal components with angle. From this distribution, the spatial auto-correlation and power spectra may be predicted. Secondly, there is the distribution of the received signal with delay which is used to predict the coherence bandwidth of the channel.

We first consider geometrical models of the channel which deal only with the angular distribution of the received signal. These models are implicitly based on the assumption that the standard deviation in time delay of the signal components is very much greater than the period of the carrier. This results in the angle of arrival and phase of the received signal components being described by independent random processes.

Then, secondly, we consider a simple model which deals explicitly with the distribution of delays of the received signal components. The model assumes that the angle of arrival and delay distributions are independent. Some evidence is quoted which supports this assumption at 900 MHz for short delays. However, in the general case it is very unlikely that this assumption will be accurate.

The theoretical models develop the case for a very simple transmitted signal consisting of either an un-modulated carrier wave or an impulse. The effects of the mobile radio channel on more complex signals can be derived by using these results to calculate the transfer function of the mobile radio channel.

One of the first models was developed by Ossanna (Ref: Ossanna J Jr, 1964). He assumed that the time variations of the signal observed at the moving vehicle could be represented as a spatially varying function. Based on this assumption, he proposed a statistical model of the received signal in terms of a set of vertically polarized horizontally travelling plane waves. However, to generate the fading of the received signal, he only considered a single scatterer at a time. His model provided good agreement with experiment for suburban areas, but soon ran into difficulties for more densely built up areas, as would be expected for his use of a single scatterer at a time to generate the received signal.
A more realistic model was presented by Clarke (Ref: Clarke R H, 1968), based on a model suggested by Gilbert (Ref: Gilbert E N, 1965) which included Ossanna's as a special case. N vertically polarized horizontal plane waves are superimposed, every wave having statistically independent phase shift $\phi_i$ and angle of arrival $\alpha_i$ with uniform distribution between 0 and $2\pi$.

However, the predictions of the model are still at variance with experiment and the conceptual basis of the model contradicts a physical understanding of the propagation process, for in an urban environment the component waves at the receiver cannot be travelling horizontally as this would result in virtually no propagation taking place due to obstruction by buildings. Aulin (Ref: Aulin T, 1979) proposed a generalization to the model of Clarke to allow the component waves to have a distribution of arrival angles in the vertical plane as well as in the horizontal plane. Aulin demonstrated that the power spectra of the received signal is substantially changed even for small deviations from horizontal propagation. However, the correlation properties are relatively insensitive to the distribution of vertical arrival angles. Figure 2.1 shows the geometry of Aulin's model for a single component plane wave. The wave has angle of arrival $\alpha_i$ to the x-z plane and $\beta_i$ to the x-y plane. It also has an associated complex amplitude $a_i$ which we shall represent in terms of amplitude $c_i$ and phase $\phi_i$. The parameters $\alpha_i, \beta_i, c_i, \phi_i$ are all assumed to be random and statistically independent. In addition, the amplitudes of the component waves are arranged such that the ensemble average of the power in each wave is given by:

$$E[c_i^2] = \frac{\xi_0}{N}$$

2.19

where $\xi_0$ is the incident E-field.
The resultant field E-field at a point \((x_0, y_0, z_0)\) is given by the sum of the in-phase components of the incident waves:

\[
\xi(t) = \sum_{i=1}^{N} \xi_i(t)
\]

where:

\[
\xi_i(t) = c_i \cos\{2\pi ft - \phi_d + \phi_0\},
\]

\[
\phi_d = \frac{2\pi}{\lambda} \{\cos \beta_i (x_0 \cos \alpha_i + y_0 \sin \alpha_i) + \sin \beta_i z_0\}
\]
Aulin then gives the point \((x_0, y_0, z_0)\) a velocity, \(V\), at angle, \(\gamma\), to the x-z plane and expresses the resultant field in terms of the in-phase and quadrature parts of the received signal \(T_c(t)\) and \(T_s(t)\). Thus:

\[
\xi(t) = T_c(t) \cos 2\pi ft - T_s \sin 2\pi ft \tag{2.22}
\]

where \(T_c(t)\) and \(T_s(t)\) are defined as:

\[
T_c(t) = \sum_{i=1}^{N} c_i \cos(2\pi f_i t + \theta_i), \quad T_s(t) = \sum_{i=1}^{N} c_i \sin(2\pi f_i t + \theta_i) \tag{2.23}
\]

and

\[
f_i = \frac{V}{\lambda} \cos(\gamma - \alpha_i) \cos \beta_i, \quad \theta_i = \frac{2\pi z_0}{\lambda} \sin \beta_i + \phi_0 \tag{2.24}
\]

The maximum Doppler frequency is defined as \(\langle f_i \rangle_{\text{max}} = \frac{V}{\lambda} = f_m\). The \(T_c(t)\) and \(T_s(t)\) defined here are equal to the quadrature components of the normalized transfer function \(T_0(t,f_c)\) defined in Equation 2.10.

From the central limit theorem, \(T_c(t)\) and \(T_s(t)\) will be approximately Gaussian when the number of component waves, \(N\), is sufficiently large. Bennett has shown that this approximation is very good if \(N \geq 6\) (Ref: Bennett W R, 1948). However, the extreme values of the distribution will depart from the Gaussian assumption even for relatively large \(N\) (Ref: Melsa J L, Sage A P, 1973). Thus, it is necessary to take some care in using the Gaussian approximation, particularly when considering the extreme values of the distribution. We shall return to this point when considering the loss deviation of the signal envelope which provides an important measure of the distribution of the signal for low signal levels. Aulin assumes that \(T_c(t)\) and \(T_s(t)\) can be regarded as absolutely Gaussian. Due to this assumption, the process \(\xi(t)\) is completely characterized by its mean and auto-correlation function.

### 2.5.2 PROBABILITY DISTRIBUTION FUNCTION (PDF)

Aulin's model is concerned with predicting the properties of the received field strength, for the transmission of a vertically polarized signal, with the assumption of no appreciable depolarization effects due to the environment. The received
signal is thus represented by the \( E_r \) component of the signal strength at the receiver. This is denoted \( \xi \) in the text to avoid confusion with the expectation operator \( E[\cdot] \). Aulin shows that the mean of the received E-field is zero.

\[
E[T_e(t)] = E[T_s(t)] = 0 \Rightarrow E[\xi(t)] = 0
\]

The phase and envelope of the received signal are described by two random statistically independent processes \( r(t) \) and \( \phi(t) \) which are given by:

\[
r(t) = \left[ T_e^2(t) + T_s^2(t) \right]^{\frac{1}{2}} \quad \phi(t) = \tan^{-1}\left[ \frac{T_s(t)}{T_e(t)} \right]
\]

The phase of the received signal envelope \( \phi(t) \) is assumed to be uniformly distributed over \( 0 - 2\pi \). This assumption leads to the probability distribution of the envelope being Rayleigh distributed.

\[
p_r(r) = \frac{r}{\sigma^2} e^{-r^2/2\sigma^2}
\]

with cumulative distribution:

\[
P(r \leq R) = 1 - e^{-r^2/2\sigma^2}
\]

### 2.5.2.1 PDF Line of Sight (LOS) Case

With the very small transmitter-receiver separations used for micro-cellular radio, there is a very high probability that the received signal will contain a line of sight component. The quadrature components of the received signal \( T_e \) and \( T_s \) will remain Gaussian distributed, but their mean values will no longer be zero. This causes \( r(t) \) and \( \phi(t) \) not to become statistically dependent. The resultant distribution of the envelope has been analysed by Rice (Ref: Rice S.O, 1944 and 1948). It is known as the Rician distribution and has the form:

\[
p(r) = \frac{r}{\sigma^2} e^{-r^2 + A^2/2\sigma^2} I_0\left(\frac{rA}{\sigma^2}\right)
\]
where $I_0$ is a zero order modified Bessel function and $A$ is the amplitude of the LOS component. The cumulative distribution has the form:

$$ P(r \leq R) = 1 - Q \left( \frac{A}{2\sigma^2}, \frac{R}{2\sigma^2} \right) $$  

where $Q(\alpha, \beta)$ is Marcum's Q-function.

The Rician distribution is a cross between a Rayleigh and a Gaussian distribution. When $A = 0$, $p(r)$ reduces to a Rayleigh distribution, whilst, in the limit of large $A$, the distribution tends to a Gaussian distribution:

$$ p(r) \lim_{A \to \infty} = \frac{1}{\sigma\sqrt{2\pi}} e^{-\frac{(r-A)^2}{2\sigma^2}} $$

with corresponding cumulative distribution function:

$$ P(r \leq R) = \text{erf} \left( \frac{r-A}{\sigma} \right) $$

where erf() is the error function.

### 2.5.3 AUTO-CORRELATION FUNCTION

The auto-correlation and power spectrum of a stationary random process are related by the Fourier transform relation:

$$ R(\tau) = \int_{-\infty}^{\infty} S(f) e^{j2\pi f \tau} df $$

Some correlation properties of the received signal in the mobile radio environment are more easily approached via the power spectrum of the received signal. This approach has been adopted, to varying degrees, by Gans, Ossanna, Clarke, Jakes and Aulin, and is also used here. However, we first consider the analysis of the auto-correlation function in a direct manner. Aulin considers the auto-correlation function:

$$ R_\xi(\tau) = E[\xi(t)\xi(t + \tau)] $$
in terms of the quadrature components of the received signal. Thus:

\[ R_x(\tau) = E[T_c(t)T_c(t + \tau)] \cos 2\pi f_c \tau - E[T_c(t)T_s(t + \tau)] \sin 2\pi f_c \tau \]

\[ = g(\tau) \cos 2\pi f_c \tau - h(\tau) \sin 2\pi f_c \tau \hspace{1cm} 2.35 \]

where:

\[ g(\tau) = \frac{\xi_0}{2} E[\cos 2\pi f \tau], \quad h(\tau) = \frac{\xi_0}{2} E[\sin 2\pi f \tau] \hspace{1cm} 2.36 \]

where the \( f \) in Equation 2.36 is the Doppler frequency shift given by Equation 2.24.

The derivation carried out by Aulin, and summarized above, uses the results of Rice for the various quadrature correlations (Ref: Rice S O, 1944). In particular:

\[ E[T_c T_c] = E[T_c T_s] = g(\tau), \quad E[T_c T_s] = -E[T_s T_c] = h(\tau) \hspace{1cm} 2.37 \]

Alternatively, the quadrature correlation functions \( g(\tau) \) and \( h(\tau) \) may be represented in terms of the power spectral density of the received signal (Ref: Jakes W C Jr, 1974; Rice S O, 1944). Thus:

\[ g(\tau) = \int_{f_c-f_m}^{f_c+f_m} S(f) \cos[2\pi(f-f_c)\tau]df, \]

\[ h(\tau) = \int_{f_c-f_m}^{f_c+f_m} S(f) \sin[2\pi(f-f_c)\tau]df \hspace{1cm} 2.38 \]

For the case where the \( \alpha_i \) are uniformly distributed throughout \( 0 - 2\pi \), Aulin shows that:

\[ g(\tau) = \frac{\xi_0}{2} \int_0^{2\pi} \cos \beta J_0 \left( \frac{2\pi V \tau}{\lambda} \cos \beta \right) p_\theta(\beta) d\beta, \quad h(\tau) = 0 \hspace{1cm} 2.39 \]

where:

\[ p_\theta(\beta) = \text{distribution of elevation angles} \]

\[ J_0(0) = \text{zero order Bessel function} \]
The distribution of the vertical arrival angles, \( p(\beta) \), does not affect the probability distribution of the envelope or phase of the received signal. However, it does affect the correlation properties of the received signal and its envelope. In the case where all the component waves at the receiver travel horizontally, and \( p(\beta) \) is represented by a delta function \( \delta(\beta) \), Aulin's model reduces to that of Clarke. The correlation for this case is denoted \( g_0(\tau) \). Thus:

\[
g_0(\tau) = \frac{\xi_0}{2} J_0 \left( \frac{2\pi V \tau}{\lambda} \right)
\]

Aulin arbitrarily imposes a distribution of vertical arrival angles on his model of the form:

\[
p_\beta(\beta) = \begin{cases} 
\frac{\cos \beta}{2 \sin \beta_m}, & |\beta| \leq |\beta_m| \leq \frac{\pi}{2} \\
0, & \text{elsewhere}
\end{cases}
\]

where \( \beta_m = \) maximum vertical arrival angle.

Aulin states that this distribution is a realistic density function for small \( \beta_m \). Also, the distribution has the property of tending to a delta function as \( \beta_m \) tends to zero. However, the main reason for this choice is that the Fourier transform \( g(\tau) \) can be found analytically when \( p_\beta(\beta) \) takes this form, thus facilitating the calculation of power spectra. Figure 2.2, reproduced from Aulin's paper, shows the difference between \( g_0(\tau) \) and \( g(\tau) \) when the distribution of vertical arrival angles takes the form of Equation 2.41 with \( \beta_m = 45^\circ \). Although Aulin argues that this choice of \( \beta_m \) is non-physically large and would not occur in a real situation, there is in fact some evidence that, at least at 900 MHz, vertical arrival angles as large as 40\(^\circ\) occur (Ref: Jakes W C Jr, 1974, pp 134). The difference between the two curves is relatively small, especially over the region between the first maximum and the first zero. However, when we consider the power spectrum, the effect of the vertical distribution of arrival angles is considerably more profound.
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2.5.4 POWER SPECTRUM

As stated at the beginning of Section 2.5.3 above, the power spectrum follows directly from the Fourier transform of the auto-correlation function which, for a uniform arrival angle distribution and omni-directional receiver antenna, reduces to the Fourier transform of \( g(\tau) \). For Clarke’s model we denote the power spectrum, \( S(f) \), as \( S_0(f) \). Substituting the auto-correlation function of Equation 2.40 into 2.33 and carrying out the inverse Fourier transform with respect to delay, \( \tau \), gives Clarke’s spectrum:

\[
S_0(f) = \begin{cases} 
\frac{\xi_0}{4\pi f_m \sqrt{1 - (f/f_m)^2}}, & |f| \leq f_m \\
0, & |f| > f_m
\end{cases}
\]  

2.43
For Aulin's model the Fourier transform of Equation 2.39 cannot in general be carried out exactly. However, for the distribution of vertical arrival angles in Equation 2.41, substituting 2.39 into Equation 2.33 and carrying out the inverse Fourier transform with respect to delay, $\tau$, yields:

$$s(f) = \begin{cases} 
\frac{\xi_0}{4\pi \sin \beta_m f_m} \left[ \frac{\pi}{2} - \sin^{-1} \left( \frac{2\cos^2 \beta_m - 1 - (f/f_m)^2}{1 - (f/f_m)^2} \right) \right], & |f| < f_m \cos \beta_m \\
\frac{\xi_0}{4\pi \sin \beta_m f_m}, & f_m \cos \beta_m \leq |f| \leq f_m \\
0, & |f| > f_m
\end{cases}$$

$$2.44$$

Relative Power dB

Clarke's Model

Aulin's Model

Doppler Frequency

Theoretical Mobile Radio Doppler Spectrum
Illustrating Clarke's Model and Aulin's Model
($f_m$ = maximum Doppler frequency Omni Antenna)

Figure 2.3 Comparison of Doppler Spectra of Clarke and Aulin Models
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Figure 2.3 shows the two spectra from Equations 2.43 and 2.44, where $\beta = 30^\circ$ for Aulin's model. The major differences are that $S(f)$ is finite at $f = f_m$ whereas $S_0(f)$ tends to infinity. Also, there is much more energy at low frequencies in $S(f)$ than in $S_0(f)$. The infinity arises from the geometrical nature of Clarke's model. It does not imply that an infinite amount of power arrives from the small band of frequencies around $f_m$.

2.5.4.1 Power Spectrum by Geometrical Considerations

The power spectrum for Clarke's model can be derived from a geometrical analysis of the channel. This approach is particularly valuable for assessing the effect of the use of directional antennae on the power spectrum. An expression for the Doppler spectra of the received signal is obtained using the relationship between Doppler shift and angle of arrival in Equation 2.24. For $\beta = 0$ and $\gamma = 0$, this reduces to:

$$f = \frac{V}{\lambda} \cos(\alpha)$$  \hspace{1cm} (2.45)

The power contributed to the received signal by waves arriving within a small element of angle $\delta \alpha$ from direction $\alpha$ by a receiver antenna of gain $G(\alpha - \Gamma)$, where the bearing of the antenna is $\Gamma$ to the direction of travel, is given by:

$$S(\alpha)\delta \alpha = p(\alpha)G^2(\alpha - \Gamma)\delta \alpha$$  \hspace{1cm} (2.46)

where $p(\alpha)\delta \alpha$ is the power within $\delta \alpha$ from direction $\alpha$ that would be received by an isotropic antenna of the correct polarization. The power spectral density $S(f)$ can be expressed in terms of the angular power density $S(\alpha)$ by:

$$S(f) = S(\alpha) \left| \frac{d\alpha}{df} \right|$$  \hspace{1cm} (2.47)

Using the expression for the Doppler frequency, the derivative of $\alpha$ with respect to $f$ is given by:

$$\frac{d\alpha}{df} = -\frac{1}{f_m\sqrt{1 - (f/f_m)^2}}$$  \hspace{1cm} (2.48)
The model assumes that the incident waves are uniformly distributed with respect to angle \( \alpha \). The final expression for the Doppler spectra is calculated by adding the components from \( \pm \alpha \) which are equal due to symmetry:

\[
S(f) = G(\alpha - \Gamma)^2 \frac{1}{\sqrt{1 - (f/f_m)^2}} \tag{2.49}
\]

where: \( \alpha = \text{modulus}[\cos^{-1}(f/f_m)] \)

### 2.5.4.2 Power Spectrum for Directional Antenna

The effect of using a directional receiver antenna assuming a horizontal distribution of incoming waves, i.e., Clarke's model, is predicted by examining Equation 2.49. For an omni-directional antenna, Equation 2.49, above, reduces to Equation 2.43. However, when an antenna of beamwidth, \( B \), is used at the receiver, with gain function approximated by a step function in the angle domain given by:

\[
G(\alpha - \Gamma) = \begin{cases} 
G_0, & |\alpha - \Gamma| < \frac{B}{2} \\
0, & \text{elsewhere} 
\end{cases} \tag{2.50}
\]

The effect is to remove part of the Doppler spectrum depending on the direction of the antenna bearing, relative to the direction of vehicle travel [1].

Figure 2.4 shows the received Doppler spectra for a directional antenna pointing along and at right angles to the direction of travel respectively.

---

1 This can be seen by substitution of Equation 2.50 into 2.49.
Figure 2.4 Doppler Spectra for 120° Beamwidth Antenna (Illustrated using Clarke Spectrum)
2.5.4.3 Power Spectrum for Directional Antenna (Aulin's Model)

The situation for Aulin's model is not so simple. As can be seen from Equation 2.24 for the Doppler frequency, although the antenna beam may cut off sharply at ± B/2 the Doppler spectrum will not. Qualitatively, this can be seen to round off the sharp edges of the spectrum. For a cut-off angle of B/2, the nominal Doppler cut-off frequency is defined as $f_\text{B} = f_m \cos B/2$. For the case where the antenna points along the direction of motion, the spectrum will be rounded off over a region extending from:

$$\frac{f_B}{\cos \beta_m} \rightarrow f_B \cos \beta_m,$$  \hspace{1cm}  2.51

Whilst for the case where the antenna points at right angles to the direction of motion, the spectrum is rounded off over a region extending from:

$$f_B \rightarrow f_B \cos \beta_m$$  \hspace{1cm}  2.52

However, for the case where the distribution of vertical arrival angles is relatively narrow, say $\beta_m \leq 10^\circ$, the correction to the simple cut-off effect of the antenna on the spectrum is small and is ignored.

2.5.4.4 Power Spectrum with a LOS Component

As mentioned in Section 2.5.2.1, discussing the PDF of the received signal, there is a high probability that the received signal will contain a direct component from the transmitter. This component will be represented by a delta function in the power spectrum at frequency:

$$f_{\text{LOS}} = f_m \cos \alpha_{\text{LOS}} \cos \beta_{\text{LOS}}$$  \hspace{1cm}  2.53

For the geometry in the micro-cellular environment, the LOS component will occur very close to the maximum or minimum Doppler frequency. Figure 2.5 illustrates the modified spectrum for a directional antenna with a LOS component.
2.5.5 SQUARED ENVELOPE AUTO-CORRELATION FUNCTION

In this Thesis, the experimental apparatus measures the square law detected envelope of the received signal. Thus, we need to develop expressions for the auto-correlation of the squared envelope, $r^2$, of the received signal. The auto-correlation function of the output of a square-law device in response to a Gaussian input, expressed in terms of the quadrature correlations $g(\tau)$ and $h(\tau)$, is given by (Ref: Bendat J S, Piersol A G, 1986):

$$R_{r^2}(\tau) = 4\sigma^4 + 4 \{ g^2(\tau) + h^2(\tau) \}$$

where the proportionality constant of the detector is assumed to be 1. Thus, for the case where the received signal follows Clarke’s model and the receiver antenna is omni-directional, the auto-correlation of the envelope squared is given by:
Thus, the correlation coefficient as a function of time delay, \( \rho(\tau) \), and that as a function of antenna separation, \( \rho(d) \), are given by:

\[
\rho_{r2}(\tau) = J_0^2(2\pi f_m \tau), \quad \rho_{r2}(d) = J_0^2\left(\frac{2\pi \frac{d}{\lambda}}{\lambda}\right)
\]

\[2.56\]

### 2.5.5.1 Envelope Squared Auto-Correlation (General Case)

In the general case, where the received signal follows the Aulin model, the receiver has a directional antenna and the received signal contains a line of sight component, the auto-correlation function is more easily obtained using the power spectrum relations in Equation 2.38 to derive \( g(\tau) \) and \( h(\tau) \). The correlation coefficient of the squared envelope is given in terms of \( g(\tau) \) and \( h(\tau) \), using Equation 2.38, by:

\[
\rho_{r2}(\tau) = \frac{g(0)^2}{g(0)^2 + h(\tau)^2}
\]

\[2.57\]

Figure 2.6 shows the numerically evaluated, squared envelope, auto-correlation function for a 120° beamwidth receiver antenna, where the received signal has an "Aulin type" vertical angle distribution with \( \beta_m = 30^\circ \). Superimposed on the same Figure is the auto-correlation function predicted for an omni-directional antenna, subject to a horizontal, uniform, angle of arrival distribution; ie the model of Clarke, Equation 2.40. The Figure is normalized and illustrates the shape of the auto-correlation for the case where the antenna points along (or opposite to) the direction of vehicle motion.

Also shown on the same Figure is the normalized squared envelope auto-correlation obtained when the received signal contains a line of sight signal (LOS). The LOS component is incident at \( \alpha_{\text{LOS}} = 0 \) and \( \beta_{\text{LOS}} = 10^\circ \) and has a power which is 10 times greater than the combined power in the scattered components. These conditions are similar to the experimental conditions described later in this Thesis. On Figure 2.6, curve 1 is the squared envelope auto-correlation for the directional
antenna, curve 2 is directional antenna auto-correlation demonstrating the effect of the LOS component, and curve 3 is the squared envelope auto-correlation for Clarke's model in Equation 2.56.

Figure 2.6 Squared Envelope Auto-Correlation function for a Directional Antenna. Also shown on the Figure is the auto-correlation for an omni-directional antenna and the auto-correlation function for a directional antenna containing a LOS component.

The effect of the directional antenna in broadening the squared envelope auto-correlation function is immediately apparent in Figure 2.6. The effect of the LOS component in broadening the squared envelope auto-correlation function is also apparent. Section 2.5.7 presents a simple model which relates the width of the auto-correlation function to the strength of the line of sight component in relation to the power in the scattered component. This follows from the Fourier transform relationship between auto-correlation and power spectra. Spatial filtering, caused by the directional antenna, narrows the spectral width in the Doppler frequency domain which, due to the Fourier transform relation between the two domains, causes a broadening of the auto-correlation function in the time delay (or distance) domain. In a similar way, as the power in the LOS component increases with respect to the scattered power, the auto-correlation function of squared envelope broadens, tending, in the limit, to a constant value for all times.
2.5.6 SQUARED ENVELOPE POWER SPECTRUM

The power spectrum of the output of a square law device is given by the Fourier transform of the auto-correlation $R_{x x}$. Hence:

\[ S_{r^2} = \sigma_x^4 \delta(f) + 2 \int_{-\infty}^{\infty} R_x^2(\tau) e^{-2\pi j f \tau} d\tau \]

\[ = \sigma_x^4 + 2 \int_{-\infty}^{\infty} S(l)S(f-l)dl \]  

2.58

Thus, the output spectral density is composed of two parts: an impulse at zero frequency equal to the output mean value, and a part corresponding to the random variations of the output which is given by the convolution of the Doppler spectra $S(f)$. The output of the square law device is then low pass filtered to remove the parts of the spectrum centred around $\pm f_c$ and the DC component is removed. After the low pass and DC blocking operation, the varying part of the spectrum may be represented for positive frequencies by:

\[ S_{r^2} = 2 \int_{f_c - f_m}^{f_c + f_m - f} S(l)S(f-l)dl, \quad 0 \leq f \leq 2f_m \]  

2.59

For Clarke's model, Equation 2.59 may be solved analytically giving (Ref: Clarke RH, 1968):

\[ S_{r^2}(f) = \frac{\xi_0^2}{2} K \left[ \sqrt{1 - \left( \frac{f}{2f_m} \right)^2} \right] \]  

2.60

where $K()$ is the complete elliptic integral of the first kind. For Aulin's model, the spectrum must be evaluated numerically.
Figure 2.7 shows the difference between the square law spectra predicted by the models of Clarke and Aulin respectively for the case where the antenna is a $120^\circ$ beamwidth horn pointing along the direction of travel. Aulin's spectrum has a greater low frequency content than Clarke's. The difference is due to the flat portion of the Doppler input spectrum predicted by Aulin's model. The degree of accord which each model has with experiment will be dependent on the local geometry surrounding the mobile vehicle at the time when a measurement is recorded. Experimental results are given in Chapter 4.
2.5.6.1 Spectrum in Presence of LOS Component

The line of sight component, amplitude A, modifies the Doppler spectra to include a delta function with power A^2/4 Doppler shifted by f_{LOS} away from the positive and negative carrier frequencies. Thus:

\[ S'(f) = S(f) + \frac{A^2}{4} [\delta(f + f_c - f_{LOS}) + \delta(f - f_c + f_{LOS})] \]  \hspace{1cm} 2.61

The low pass square law detected output spectrum is thus (Ref: Davenport W B Jr, Root W L, 1958; Jakes W C Jr, 1974):

\[ S_r^2 = (A^2 + \sigma_n^2)^2 \delta(f) + 2 \int_{f_c-f_m}^{f_c+f_m} S(l)S(f-l)dl \]

\[ + A^2 [S(f + f_c - f_{LOS}) + S(f - f_c + f_{LOS})], \quad 0 \leq f \leq 2f_m \]  \hspace{1cm} 2.62

The AC part of the spectrum is simply the original square law spectrum S_r(f), plus a shifted folded section of the input Doppler spectrum S(f). Figure 2.8 shows the effect caused by the presence of a LOS component on the square law spectrum presented in Figure 2.7. The obvious effect is to increase the spectral content at low frequencies in comparison to the non-LOS square law spectrum. The enhancement to the low frequency content of the Clarke spectrum is greater than the corresponding enhancement for the Aulin spectrum. This is because of the interaction between the pole at the maximum Doppler frequency in the Clarke model with the delta function representing the line of sight signal when the square law spectrum is formed from the convolution of the Doppler spectrum of Figure 2.5.
Figure 2.8 Squared Envelope Power Spectrum for the Models of Aulin and Clarke Using a Directional Antenna and Including a Line of Sight Component with 10 Times the Power of the Scattered Component

2.5.7 AUTO-CORRELATION WIDTH VERSUS LINE OF SIGHT SIGNAL STRENGTH

It is possible to derive a simple approximate relationship between the relative strength of the line of sight component and the width of the auto-correlation function. By definition, there is a Fourier transform relationship between the auto-correlation function and the power spectrum of the received signal Doppler spectrum. Therefore, since the Fourier transform is a linear operation, reducing the width of the Doppler spectrum by a factor \(1/K\) will increase the width of the auto-correlation by a factor \(K\). The standard deviation of the Doppler spectrum is used as a measure of the width of the spectrum. This is only an approximation. It relies on the assumption that the Doppler spectrum in the absence of the line of sight component is uniform across the beamwidth of the receiver antenna and that the strength of the line of sight component is the only influence on the apparent width of the spectrum. This may not be the case experimentally because the limited number of scatterers in the urban micro-cell may lead to a limited angular spread.
of received components even in the absence of a line of sight component. However, even given these shortcomings, it is still useful to have a measure of the relative strength of the line of sight component in relation to the strength of the scattered component.

The standard deviation of the Doppler spectrum may be expressed in terms of the normalized Doppler frequency \( \frac{1}{f} = \frac{f}{f_m} \) and the Doppler spectral density \( S(l) \) as (Ref Bendat J S, Piersol A G 1986):

\[
\sigma^2 = \frac{\int l^2 S(l) dl - \int l S(l) dl}{\int S(l) dl} \tag{2.63}
\]

where: \( \sigma^2 \) = standard deviation squared of the Doppler spectrum, \( l \) = normalized Doppler spectrum, \( S(l) \) = theoretical Aulin or Clarke Doppler spectrum in the absence of a line of sight component.

When a line of sight component at the maximum Doppler frequency is included in the Doppler spectrum, the expression for the standard deviation is modified:

\[
\sigma'^2 = \frac{\int l^2 \{S(l) + \alpha \delta(1 - l)\} dl - \int l\{S(l) - \alpha \delta(1 - l)\} dl}{\int S(l) + \alpha \delta(1 - l) dl - \int S(l) \alpha \delta(1 - l) dl} \tag{2.64}
\]

where: \( \delta(1 - l) \) = Kronecker delta function, \( \alpha \) = power in the line of sight component, \( \sigma^2 \) = modified standard deviation squared.

The total power in the randomly scattered component is \( P \) and that in the scattered component is \( \alpha \), where:

\[
P = \int S(l) dl, \quad \alpha = \alpha \int \delta(1 - l) dl \tag{2.65}
\]
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The modified standard deviation is then expressed:

\[
\sigma^2 = \frac{\int l^2 S(l)dl - \int l S(l)dl}{P + \alpha} + \frac{\alpha [l^2 - l]}{P + \alpha}
\]

\[
= \frac{P}{P + \alpha} \sigma^2 + 0
\]

2.66

The standard deviation of the Doppler spectrum which contains a line of sight component is therefore related to that without one by:

\[
\sigma' = \sqrt{\frac{P}{P + \alpha}} \sigma
\]

2.67

Thus, the width of the auto-correlation function which contains a line of sight component, is related to the width without one by:

\[
L' = \sqrt{1 + \alpha P} L
\]

2.68

where: \( L = \) width of auto-correlation function with no LOS component to the first zero, \( L' = \) width of auto-correlation function with a LOS component to the first zero, \( \alpha / P = \) ratio of power between direct and scattered component.

Thus, the approximate relationship between the direct and scattered power is given by the relationship:

\[
\alpha / P = (L' / L)^2 - 1
\]

2.69

This equation provides a simple relationship between the direct and scattered power in terms of the width of the auto-correlation function.

2.6 COHERENCE BANDWIDTH

In a real communication system, information must be transmitted between the base station and the mobile unit. Multipath propagation introduces both amplitude and phase distortion effects which interfere with the transmission of information. Thus, a means of characterizing multipath distortion is needed.
One effect of multipath propagation is to cause two carrier waves, transmitted simultaneously, with frequency separation $\Omega$, to become de-correlated as $\Omega$ increases. This leads to the concept of the coherence bandwidth of the channel. This is defined as the point where the correlation coefficient of the squared envelope of the received signals, at two separate frequencies, drops to some predetermined value. However, since there is a degree of choice as to what this value may be, there are several different definitions of correlation bandwidth. Following Jakes (Ref: Jakes W C Jr, 1974), a value of 0.5 for the correlation coefficient of the squared envelope is chosen to define the correlation bandwidth in this Thesis.

The correlation bandwidth may be interpreted in various ways. It can be considered as a bandwidth limitation on the channel, whereby transmissions which exceed the bandwidth become badly distorted and, on digital channels, inter-symbol interference results. The distortion is caused by the signal interfering with delayed replicas of itself and, on a digital link, it becomes severe when different components of the signal are delayed by longer than one bit period. Alternatively, it can be considered as the minimum separation required for a frequency diversity system. A diversity system needs input channels with un-correlated fading. With a frequency diversity system this can only occur if the frequency separation of the channels is greater than the coherence bandwidth.

The models presented so far have only considered the angular distribution of the received power. However, to examine the effect of the channel on signals transmitted at different frequencies, we must take account of the distribution of the delay times of the components, because it is the phase differences caused by the delay distribution which give rise to the de-correlation of two signals.

It is important to note that the correlation coefficient of the envelope is approximately the same as the correlation coefficient of the squared envelope. This follows from the description of the envelope auto-correlation in terms of a confluent hypergeometric function. This can be approximated by a constant term plus a quadratic term in the envelope of the RF auto-correlation (Ref: Jakes W C Jr, 1974), whilst the auto-correlation of the squared envelope is given exactly by the same quadratic term. Thus, the correlation bandwidths defined in terms of the envelope or its squared value are equivalent.

To develop the model for the correlation as a function of carrier frequency separation $\Omega$, where $\Omega = f_2 - f_1$, we consider Equation 2.3 for the case of transmission of an un-modulated carrier wave. The equation represents the received signal in terms of a summation over index $[i]$, where $[i]$ identifies each individual
reflection with delay \( \tau_i \) and angle of arrival \( \alpha_i \). If we consider that each wave with angle of arrival \( \alpha_i \) is composed of the summation of a number of waves with a distribution of delays \( T_{ij} \), then we can represent the received signal in terms of a summation over arrival angles \( \alpha_i \) and delay times \( T_{ij} \). This representation can then be used to derive the received signal statistics, so long as the joint distribution of arrival angles and time delays, \( p(\alpha, T) \), is known. The \([ij]\)th component wave in this representation, suppressing the explicit carrier frequency dependence, is given by:

\[
\begin{align*}
  c_{ij} e^{-j2\pi f_{\tau}T_{ij}}
\end{align*}
\]

where: \( f_{\tau} = \) carrier frequency shift, \( f_i = \) Doppler frequency of \([ij]\)th wave, \( G(\alpha) = \) gain pattern of antenna, \( T_{ij} = \) delay of \([ij]\)th wave, \( c_i = \) amplitude of \([i]\)th wave, \( p(\alpha, T) = \) angle-delay distribution, \( f_i = f_m \cos(\alpha_i) \), \( c_i = G(\alpha_i)p(\alpha_i T_{ij})d\alpha dT \).

The model is only concerned with the horizontal distribution of the waves and \( G(\alpha) \) is the horizontal directivity pattern of the receiver antenna. Following the description of the quasi-stationary model of the channel, \( \alpha_i \) and \( T_{ij} \) are assumed constant for motion over distances of the order of several tens of wavelengths. In addition, the phase shift between two component waves, given by \( 2\pi(f_{\tau}T_{ij} - f_{\tau}T_{in}) \), is assumed much greater than \( 2\pi \) for \( i \neq m, j \neq n \). In fact, at 55 GHz, even for delay times as short as \( 10^{-9} \) seconds which represent only 30 cm path difference, the phase difference between component waves is still greater than \( 100\pi \). This leads to the assumption that scattering at two different delay times is uncorrelated, i.e., we have the quasi-wide-sense-stationary uncorrelated scattering channel which is so popular in mobile communications. The cross-correlation of the received signal at two different times and frequencies can be represented in terms of the respective time and frequency differences:

\[
R_{\xi}(\Omega, \tau) = E[\xi^*(f, t)\xi(f + \Omega, t + \tau)]
\]

The quadrature components of the received signal at times \((f_1, t_1)\) and \((f_2, t_2)\) are:

\[
T_{C1,2}(f, t) = \xi_0 \sum_{ij} c_{ij} \cos 2\pi(f, t - f_{1,2}T_{ij}),
\]

\[
T_{S1,2}(f, t) = \xi_0 \sum_{ij} c_{ij} \sin 2\pi(f, t - f_{1,2}T_{ij})
\]
In the limit $N \to \infty$, $M \to \infty$, the quadrature components are jointly Gaussian and the cross-correlation function in Equation 2.71 may be represented in terms of the quadrature components in Equation 2.72. Using Equation 2.37 with $\Omega = f_2 - f_1$ and $\tau = t_2 - t_1$, we obtain:

$$R_q(\Omega, \tau) = 2E[T_{c1}T_{c2}] \cos 2\pi f_c t - 2E[T_{c1}T_{s2}] \sin 2\pi f_c t$$ \hspace{1cm} 2.73

The correlations $E[T_{a1}T_{a2}]$ and $E[T_{a1}T_{a1}]$ may be expressed in terms of time and frequency separations $\tau = t_2 - t_1$, $\Omega = f_2 - f_1$, as $g(\Omega, \tau)$ and $h(\Omega, \tau)$ respectively. Following the work of Jakes, these may be expressed as:

$$g(\Omega, \tau) = \frac{\xi_0}{2} \int_0^{2\pi} \int_0^\infty G(\alpha)p(\alpha, T) \cos(2\pi f_c \tau \cos \alpha - \Omega T) dT d\alpha$$

$$h(\Omega, \tau) = \frac{\xi_0}{2} \int_0^{2\pi} \int_0^\infty G(\alpha)p(\alpha, T) \sin(2\pi f_c \tau \cos \alpha - \Omega T) dT d\alpha$$ \hspace{1cm} 2.74

In general, the equations of 2.74 are very difficult to solve. However, if the distribution of arrival angles and delays are independent, i.e. $p(\alpha, T)$ may be expressed as $p(\alpha)p(T)$ and a simple exponential delay distribution is adopted, the integration over $T$ may be carried out separately from that over $\alpha$. There is some evidence from measurements at 900 MHz that this assumption is approximately correct for small delays where most of the received power is concentrated (Ref: Suzuki H, 1977). However, for the longer delay times the angles of arrivals are predominantly from the opposite direction to that of the transmitter, as could be expected intuitively (Ref: Cox D C, 1973).

Jakes assumes a distribution $p(\alpha, T)$ which is uniform in angle and exponential in delay, given by:

$$p(\alpha, T) = p(\alpha)p(T) = \frac{1}{2\pi\sigma_T} e^{-T/\sigma_T}$$ \hspace{1cm} 2.75

where $\sigma_T$ is the standard deviation of the delays. The expression for $g(\Omega, \tau)$ then becomes:
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\[
g(\Omega, \tau) = \frac{\xi_0}{2} \int_0^{2\pi} \int_0^{\infty} G(\alpha) e^{-T/\sigma_T} \cos(2\pi f_m \tau \cos \alpha - \Omega T) dT d\alpha \tag{2.77}
\]

Since \( \alpha \) and \( T \) are assumed independent, we can calculate the integral of Equation 2.76 with respect to time separately from that with respect to frequency. For convenience, we define three quantities: \( \Phi(\alpha, \Omega, \tau) \) which is equal to the integrand of Equation 2.76 after the time integral, and \( A \) and \( B \) which contain the angle dependent terms in Equation 2.76. Thus, at a fixed delay \( \tau \) we can write:

\[
A = \frac{\xi_0}{4\pi} G(\alpha), \quad B = 2\pi f_m \tau \cos \alpha \tag{2.77}
\]

\[
\Phi(\alpha, \Omega, \tau) = A \int_0^\infty \frac{1}{\sigma_T} e^{-T/\sigma_T} \cos(B - \Omega T) dT
\]

\[
= -\frac{A}{\Omega^2 \sigma_T^2 + 1} \left[ e^{-T/\sigma_T} \{ \cos(B - \Omega T) + \Omega \sigma \sin(B - \Omega T) \} \right]_0^\infty
\]

\[
= \frac{A}{\Omega^2 \sigma_T^2 + 1} \{ \cos B + \Omega \sigma \sin B \} \tag{2.78}
\]

It should be noted, only a single assumption that \( \alpha \) and \( T \) are independent is made about \( B \) in order to carry out the above integration. Thus, the integral could also be carried out by replacing \( B \) with Aulin's expression for the Doppler shift.

The quantity \( g(\Omega, \tau) \) is then equal to the integral of \( \Phi(\alpha, \Omega, \tau) \) with respect to angle. Substituting \( A \) and \( B \) back into Equation 2.78, we obtain the expression for \( g(\Omega, \tau) \). In a similar manner, we can also obtain the expression for the quadrature correlation \( h(\Omega, \tau) \). Using Clarke's model which assumes a horizontal distribution of components, the two correlations are given by:
\[ g(\Omega, \tau) = \frac{\xi_0}{4\pi} \left[ \frac{1}{\Omega^2 \sigma_T^2 + 1} \right] \times \int_0^{2\pi} G(\alpha) \{ \cos(2\pi f_m \tau \cos \alpha) + \Omega \sigma_T \sin(2\pi f_m \cos \alpha) \} d\Omega, \]

\[ h(\Omega, \tau) = -\frac{\xi_0}{4\pi} \left[ \frac{1}{\Omega^2 \sigma_T^2 + 1} \right] \times \int_0^{2\pi} G(\alpha) \{ \Omega \sigma_T \cos(2\pi f_m \tau \cos \alpha) - \sin(2\pi f_m \lambda \tau \cos \alpha) \} d\Omega \quad 2.79 \]

For an omni-directional antenna, using the symmetry property that \( G(\alpha) = G(-\alpha) \) and using the result of Equation 2.40, Equation 2.79 can be expressed as:

\[ g(\Omega, \tau) = \frac{\xi_0}{2} \int_0^{2\pi} J_0(2\pi f_m \tau), \quad h(\Omega, \tau) = -[\Omega \sigma_T] g(\Omega, \tau) \quad 2.80 \]

In this Thesis we are interested in the correlation coefficient of the squared envelope which, for a zero mean complex envelope input to the square law detector (ie there is no line of sight component), is expressed in terms of the quadrature correlations of the received field by (Ref: Brown W B, Root W L, 1958, Bendat J S, Perisollo A G, 1986).

\[ \rho_{r^2}(\Omega, \tau) = \frac{g^2(\Omega, \tau) + h^2(\Omega, \tau)}{g(0,0)^2} \quad 2.81 \]

Thus, for an omni-directional antenna, the correlation coefficient of the squared envelope is:

\[ \rho_{r^2} = \frac{J_0^2(2\pi f_m \tau)}{\Omega^2 \sigma^2 + 1} \quad 2.82 \]

In the general case for an antenna of beamwidth \( \theta \), the auto-correlation coefficient of the envelope squared, using Equation 2.79, can be shown to be:
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\[ \rho_{r,2}(\Omega, \tau) = \frac{\psi^2(\tau) + \xi^2(\tau)}{\sigma_f^2\Omega^2 + 1} \quad 2.83 \]

where:

\[ \psi(\tau) = \frac{1}{2\pi} \int_{r-\theta/2}^{r+\theta/2} \cos(2\pi f_m \tau \cos \alpha) d\alpha, \]

\[ \xi(\tau) = \frac{1}{2\pi} \int_{r-\theta/2}^{r+\theta/2} \sin(2\pi f_m \tau \cos \alpha) d\alpha \quad 2.84 \]

The correlation bandwidth, \( B_c \), is defined as the point where \( \rho_{r,2}(\Omega) \) falls to 0.5. This will occur when \( \Omega \sigma = 1 \). Thus it is given by:

\[ B_c = \frac{1}{2\pi \sigma_f} \quad 2.85 \]

From inspection of Equation 2.83, it can be seen that the coherence bandwidth is independent of the distribution of arrival angles. Thus, the same value of coherence bandwidth would be expected regardless of whether the receiver antenna experiences a vertical distribution of arrival angles, or when a narrow beamwidth receiver antenna is used. This behaviour arises from the assumed independence of the angle and delay distributions. However, this result is not backed up by experiment where it has been found that the shorter time delays predominate at angles near to the bearing of the transmitter (Ref: Jakes W C Jr, 1974). Thus, reducing antenna beamwidth tends to increase coherence bandwidth by reducing the standard deviation of time delays (Ref: Violette et al, 1988).

Figure 2.9 shows the predicted variation of the auto-correlation coefficient of the squared envelope for several \( \sigma_f \). The various \( \sigma_f \) are chosen from examination of the likely geometry of the urban micro-cell where path differences between the reflected rays will be of the order of 1 - 10 m. The chosen average path delays which correspond to \( \sigma_f \) in Figure 2.9 are 1.4 m, 0.96 m, and 0.48 m. The resultant coherence bandwidths, from the Figure, are respectively 50, 100, and 200 MHz. These predictions are supported to some extent by the experimental work of Violette (Ref: Violette et al, 1988) on propagation at millimetre wave frequencies using narrow beamwidth antennas in downtown Denver, Colorado. Violette reported delays of the order of 10 - 100 ns with a coherence bandwidth of the order
of 200 MHz. However, if an omni-directional or wide beamwidth antenna is used the observed, delay spreads will be larger because more of the scattered radiation will be received.

\[ \sigma_r = \text{time delay spread} \]

\[ a_c = 4.8 \text{ ns} \]

\[ a = 3.2 \text{ ns} \]

\[ \sigma_r = 1.6 \text{ ns} \]

![Figure 2.9 Squared Envelope Correlation Coefficient as a Function of Carrier Frequency Separation for Three Values of Delay Spread \( \sigma_r \)]

2.6.1 VARIATION OF THE COHERENCE BANDWIDTH WITH THE DELAY DISTRIBUTION

The exponential model for the delay distribution, adopted to derive the results in Section 2.6, has good support from the measurements which are available at 900 MHz (Ref: Gans M J, 1972). However, Gans showed that the coherence bandwidth, \( B_c \), is relatively insensitive to the actual shape of the delay distribution. To show this, Gans calculated \( B_c \) for a delay distribution which consisted of two equal power delta functions located at zero and \( 2\sigma_r \), respectively.

\[
W(T) = \frac{1}{2} [\delta(T) + \delta(T - 2\sigma_r)]
\]  

2.86
The coherence bandwidth, $B_c$, for the delay distribution of Equation 2.86 is only ~20% different to that for the exponential distribution. This has important implications for the prediction of $B_c$ in the urban micro-cell where there is a high probability that propagation will be dominated by a few strong specular or line of sight components. This is because it suggests that a good estimate of $B_c$ can be made from simple geometrical considerations about the likely path delays of the specular reflections in the micro-cell.

To illustrate the relationship between delay distribution, $W(T)$, and coherence bandwidth, $B_c$, Gans used the correlation function of the complex envelope of the received signal $R(\Omega, \tau)$. The complex correlation function involves both the quadrature correlations $g(\tau)$ and $h(\tau)$ which were defined in Equation 2.75. Representing the quadrature phase relationship between the correlations by using $<\phi>$, rather than by $\cos[2\pi f_0 t]$ and $\sin[2\pi f_0 t]$, and using Equation 2.75 and Equation 2.74 we obtain:

$$R_\xi(\Omega, \tau) = \frac{\xi_0}{2} \int_0^\infty \int_0^\infty G(\alpha)p(\alpha, T)e^{i2\pi(\alpha\tau - \Omega T)}d\tau d\alpha \quad 2.87$$

It is apparent that there is a symmetry between the two products $f(\alpha)\tau$ and $\Omega T$ in the exponent of Equation 2.87. Gans identifies this symmetry as a duality between the delay distribution and the angle of arrival distribution. Taking the Fourier transform of Equation 2.87 with respect to time delay $\tau$, at zero frequency separation, yields the power spectrum of the data $S(f)$. The dual relation, taking the Fourier transform of Equation 2.87 with respect to frequency separation at zero time delay, yields the power delay spectrum of the data. In other words, it gives the delay distribution $W(T)$. Thus:

$$W(T) = \int_{-\infty}^{\infty} R_\xi(\Omega, 0)e^{i2\pi\tau\Omega}d\Omega \quad 2.88$$

Gans uses the inverse Fourier transform (Equation 2.87) to obtain the expression for the normalized correlation coefficient as a function of carrier frequency separation from the suggested delay distribution functions for $W(T)$. Calculating the inverse Fourier transform of Equation 2.87 for the two suggested $W(T)$ yields the correlation function as a function of frequency. This can be used to obtain the
squared envelope correlation coefficient following Equation 2.81. The correlation coefficient for the delta function distribution of Equation 2.86, at zero time delay, is:

\[ \rho_{r^2}(\Omega) = \cos^2(\Omega \sigma_T) \]  \hspace{1cm} 2.89

The corresponding correlation coefficient for the exponential distribution is:

\[ \rho_{r^2}(\Omega) = \frac{1}{\sigma_T^2 \Omega^2 + 1} \]  \hspace{1cm} 2.90

The resulting coherence bandwidth, \( B_c \), is obtained by setting the correlation coefficient of Equation 2.90 to 0.5, ie:

\[ \rho_{r^2}(\Omega) = 0.5 \Rightarrow B_c = 2\pi \Omega = \frac{1}{8\sigma_T} \]  \hspace{1cm} 2.91

Figure 2.10 shows a plot of the predicted squared envelope correlation coefficient for the exponential and delta function delay distributions. The respective coherence bandwidths for the delta function and exponential delay distributions are \( 1/2\pi\sigma_T \) and \( 1/8\sigma_T \). This indicates that the coherence bandwidth, defined as the 0.5 value of the coefficient, is relatively insensitive to the delay distribution.

It should be noted that although the correlation coefficient of the delta function delay distribution is periodic, in reality there would be more than two discrete components each with independent phase and the resultant correlation coefficient would, in general, be non-periodic. The simple case, however, can be used to illustrate the behaviour of the more general one. The relationship between \( \sigma_T \) and \( B_c \) for the delta function delay distribution is used in the deterministic model presented in Section 2.7 to predict the coherence bandwidth \( B_c \), for an urban micro-cell, based on geometrical considerations.
2.6.2 COHERENCE BANDWIDTH AND DELAY SPREAD ESTIMATION

Equation 2.85 suggests that the coherence bandwidth and delay spread could be measured by undertaking a series of measurements of correlation at different frequencies, or by directly probing the channel in the time domain in order to measure the distribution of delays. However, it is also possible to estimate the two quantities using the cross-correlation from a single two frequency measurement. Methods for estimating the coherence bandwidth from two frequency measurements are presented in Section 2.6.3. The direct approach is approximated by the swept frequency measurements and is discussed in Section 2.6.4.

2.6.3 ESTIMATE VIA TWO FREQUENCY MEASUREMENTS

The simplest way in which to use the results of the two frequency measurements to estimate the coherence bandwidth and delay spread of the channel, is to calculate the correlation coefficient between the two received envelopes at each carrier frequency separation and plot the resulting correlation coefficients against frequency. The coherence bandwidth, $B_c$, is given by the frequency separation
point at which the normalized, squared envelope, correlation coefficient falls to 0.5. The delay spread is estimated from this value of coherence bandwidth using Equation 2.85.

In addition, it is possible to estimate the coherence bandwidth from a single measurement of the channel characteristics at two frequencies, in two different ways: directly via estimation of the cross-correlation function; or indirectly, in the frequency domain, via the coherency squared function.

2.6.3.1 Estimate via Cross-Correlation Function

A single, simultaneous, measurement of the squared envelope at two separate frequencies, with separation $\Delta$, can be used to estimate $\rho(\Delta, \tau)$. This estimate can be used to estimate the standard deviation of delays $\sigma_T$ and hence the coherence bandwidth of the channel $B_c$. Thus:

$$
\hat{B}_c = \frac{1}{2\pi\sigma_T} = \Delta \left[ \frac{1}{\hat{\rho}} - 1 \right]^{\frac{1}{2}} \tag{2.92}
$$

where: $\Delta =$ frequency separation between carriers in Hz, $\hat{\rho} =$ estimate of correlation coefficient, $\hat{B}_c =$ estimate of coherence bandwidth.

Estimates carried out at different frequency separations may then be used to calculate an average value for $\sigma_T$ and $B_c$.

2.6.3.2 Estimate via Cross Spectrum

To estimate the coherence bandwidth from the cross spectrum, we define the coherency squared function in terms of the co-spectra and the cross-spectra of the complex envelope at two separate carrier frequencies $f_1$ and $f_2$ (Ref: Bendat J S, Piersol A G, 1986).

$$
\gamma_{f_1f_2}^2(f) = \frac{|S_{f_1f_2}(f)|^2}{S_{f_1}(f)S_{f_2}(f)} \tag{2.93}
$$
Note that we are not concerned with the DC parts of the spectra since it is impossible to measure them. However, the relationship for the coherency squared, as a ratio of delta functions, would still hold. The various co-spectral and cross-spectral densities can be calculated from the Fourier transform of Equation 2.83. Define \( F(f) \) as:

\[
F(f) = \int \{ \psi^2(\tau) + \zeta^2(\tau) \} e^{-j2\pi f \tau} d\tau
\]

The respective co spectra and cross spectra can then be expressed in terms of \( F(f) \) using the linearity properties of the Fourier transform operation. Thus:

\[
S_{f_1}(f) = S_{f_2}(f) = F(f),
\]

\[
S_{f_1 f_2}(f) = \left( \frac{1}{\sigma_f^2 \Omega^2 + 1} \right) F(f)
\]

where \( \Omega = 2\pi \Delta \).

The equation for \( \gamma(f) \) then reduces to:

\[
\gamma_{f_1 f_2}(f) = \left( \frac{1}{\sigma_f^2 \Omega^2 + 1} \right)^2 \frac{F^2(f)}{F(f) F(f)} = \left( \frac{1}{\sigma_f^2 \Omega^2 + 1} \right)^2
\]

Thus, \( \gamma(f) \) is a function of the carrier frequency separation, \( \Delta \), alone and is independent of frequency in the Doppler frequency domain. The standard deviation of delays, \( \sigma_T \), and the coherence bandwidth, \( B_c \), can be estimated from \( \gamma(f) \) in an analogous way to that used for the estimation from the auto-correlation function. The standard deviation of delays, \( \sigma_T \), is derived from Equation 2.90 and this result is used to derive the coherence bandwidth, \( B_c \), using Equation 2.85. Thus:

\[
B_c = \frac{1}{2\pi \sigma_T} = \Delta \left[ \frac{1}{\sqrt{\gamma}} - 1 \right]^{-\frac{1}{2}}
\]
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2.6.4 ESTIMATE VIA SWEPT FREQUENCY TECHNIQUES

The swept frequency technique is used to synthesize a wideband input channel to enable the impulse response of the channel to be probed in a direct way. The output of the transmitter consists of a set of repeated sweeps across a 66 MHz bandwidth. The mobile vehicle is stationary during the measurements and the sweep duration of 0.1 second is rapid compared to the rate of change of the received signal due to the movement of scatterers. Thus, the properties of the channel can be assumed to be constant for the duration of the sweep. Each sweep forms a snapshot of the squared envelope of the channel transfer function, \( T(f, t') \), at time \( t' \), Section 2.3. The frequency cross-correlation and time delay distribution properties of the channel may be estimated from \( T^2(f, t') \). The details of how the measurement is undertaken are discussed in Section 3.9.

2.6.4.1 Estimate of Frequency Cross-Correlation

The normalized squared envelope correlation function as a function of frequency separation can be calculated directly from the swept frequency data, for all frequency separations from zero up to the maximum deviation of the sweep. Thus, where received signal envelope is \( |T(f, t)|^2 \) we can write:

\[
\hat{R}_{T^2}(\Delta, t) = \frac{1}{F} \int_{f' - F/2}^{f' + F/2} |T(f, t)|^2 |T(f - \Delta, t)|^2 df
\]

\[
\hat{\rho}_{T^2}(\Delta) = \frac{\hat{R}_{T^2}(\Delta)}{\hat{R}_{T^2}(0)} \tag{2.98}
\]

where: \( \hat{R}_{T^2}(\Delta, t) = \) estimate of transfer function auto-correlation function, \( F = \) bandwidth of sweep, \( \Delta = \) frequency separation, \( f' = \) centre frequency of sweep.

The coherence bandwidth, \( B_c \), is estimated from the 0.5 point on the \( \rho_{T^2}(\Delta) \) curve.
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2.6.4.2 Estimate of Delay Spread

The delay distribution of the received signal $W(T)$ is equal, using the argument expressed in Section 2.6.1, to the inverse Fourier transform of the frequency separation auto-correlation. This is obtained by applying Equation 2.94 to Equation 2.98 derived above in Section 2.6.4.1.

2.6.5 CORRELATION OF ORTHOGONAL POLARIZATIONS

This Section summarizes the work presented in the literature on the analysis of the effect of the mobile radio environment on the correlation between two co-frequency orthogonally polarized signals. The work is essentially qualitative in nature and concentrates on examining the scattering and cross-coupling mechanisms to show that they are incoherent. Thus, it is not necessary to give a detailed analysis of the different scattering coefficients for the two polarizations and of the polarization cross-coupling effects.

It has been suggested that orthogonally polarized co-frequency channels could be used to provide diversity reception for mobile radio (Ref: Jakes W C Jr, 1974; Lee W C Y, 1982; Lee W C Y, Yeh Y S, 1975). Potentially, orthogonally polarized channels may provide a means of diversity without increasing the bandwidth requirement of the system and without necessitating the use of spatially separated antennae. The use of co-located antennae for diversity would be particularly advantageous for base station diversity where large antenna separations are required (Ref: Lee W C Y, 1982).

Figure 2.11 illustrates the propagation scenario for transmission of two orthogonal polarizations. The transmitted signal consists of a vertically polarized component and a horizontally polarized component which are transmitted from co-located antennae with identical radiation patterns. The received signal is similar, consisting of a vertical and a horizontal component received by co-located antennae with identical radiation patterns.
The respective horizontal and vertically polarized received signals can each be considered in terms of two coupling constants; one representing the power transmitted and received on the same polarization, the other representing the power transmitted on one polarization and received on the other. Considering only the E-field, the received signal may be represented in terms of vertical and horizontal components as:

\[
E_{rv} = \Gamma_{vv} + \Gamma_{hv}, \quad E_{rh} = \Gamma_{hh} + \Gamma_{vh}
\]

where \( \Gamma_{xy} \) represent respectively the \( V \rightarrow V, H \rightarrow H, V \rightarrow H, \) and \( H \rightarrow V \) coupling coefficients, and \( E_{rv} \) and \( E_{rh} \) represent the vertical and horizontal components of the received signal. From examination of Figure 2.11, it can be seen that, for transmission of two orthogonal components from the mobile unit, the received signal at the base station, \( E_{rv} \) and \( E_{rh} \), can be expressed as:

\[
E_{rv} = \Gamma_{vv} + \Gamma_{vh}, \quad E_{rh} = \Gamma_{hh} + \Gamma_{hv}
\]

The model described in Section 2.6 can be modified to examine the effects of the mobile radio channel on \( \Gamma_{vv} \) and \( \Gamma_{hh} \). Consider \( \Gamma_{vv} \) as a summation of scattered waves, of the type represented in Equation 2.70, the total field is given by:

\[
\Gamma_{vv} = \frac{\zeta_0}{2} \sum_{i=1}^{N} \sum_{j=1}^{M} c_{ij} e^{j(2\pi/\lambda - 2\pi/\lambda_i + \phi_{ij})}
\]
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where \( c_{ij} \) and \( \phi_{ij} \) represent the amplitude and phase respectively of the \([ij]th\) component wave with Doppler shift \( f \) and delay \( T_{ij} \). The horizontal and vertically polarized antennae are co-located at both the transmitter and the receiver terminals. Thus, it follows that the waves will follow the same propagation paths between the transmitter and receiver with the same associated delay and Doppler shift. However, the reflection coefficients for the vertical and horizontal waves are different. Following Lee (Ref: Lee W C Y, 1982), the reflection coefficients for the vertically [1] and horizontally polarized [2] waves are respectively:

\[
-1 \leq \rho_V \leq -1, \quad \text{ground reflected wave} \\
\rho_V \equiv -1, \quad \text{building reflected wave} \quad 2.102 \\
\rho_H \equiv -1, \quad \text{ground reflected wave} \\
-1 \leq \rho_H \leq 1, \quad \text{building reflected wave} \quad 2.103
\]

Thus the horizontal component of reflection, defined with respect to the orientation of the scatterer itself, unconditionally has \( \pi \) phase change; whilst the vertical component has a phase continuously variable between \( \pi \) and 0. For a smooth surface, Lee gives the following inequalities for the vertical reflection coefficient defined with respect to the scatterer itself:

\[
\rho_H \equiv 1, \quad \text{incident angles } \ll 10^\circ \\
\rho_H \equiv -1, \quad \text{incident angles } \sim 90^\circ \quad 2.104
\]

The horizontally polarized wave at the receiver, resulting from the horizontally polarized transmitted signal, may then be represented as:

\[
\Gamma_{hh} = \frac{\xi_0}{2} \sum_{i=1}^{N} \sum_{j=1}^{M} (c_{ij} + c'_{ij})e^{j(2mf_{ij}T_{ij} + \phi_{ij} + \phi'_{ij})} \quad 2.105
\]

where \( c'_{ij} \) and \( \phi'_{ij} \) represent the differences in amplitude and phase respectively between the \([ij]th\) horizontal and vertical waves with Doppler shift \( f \) and delay \( T_{ij} \) respectively. In the limit of large \( N \), the amplitude and phase differences will be

---

1 Polarisation is horizontal with respect to the buildings.

2 Polarisation is vertical with respect to the ground.
respectively Gaussian and uniformly distributed. The expectation $E[\Gamma_{vv}\Gamma_{\text{hv}}]$ will then be zero (Ref: Lee W CY, Yeh Y S, 1975). However, we have not accounted for the cross coupling terms $\Gamma_{\text{hv}}$ and $\Gamma_{\text{vh}}$.

The Radar Cross Section Handbook (Ref: Ruck G T et al, 1970, pp 708) states that the polarization cross-coupling terms arise from multiple scattering phenomena. Thus, models which do not explicitly take account of multiple scattering cannot predict polarization cross-coupling. The necessity of multiple scattering can be seen from a physical argument. The polarization cross-coupling process can be considered as a rotation of one polarization into the orthogonal polarization, and from basic geometry, a rotation can be considered as the result of two consecutive reflections through different axes of symmetry. Thus, a minimum of two reflections are required to provide the rotation of the polarization axis of the incoming wave. The uncorrelated nature of the $H \rightarrow V$ and $V \rightarrow H$ coupling coefficients follows from two arguments. First, the two coupling coefficients will be different in both phase and in amplitude for a given pair of scatterers. Secondly, the two processes will, in general, result from scattering by different scattering pairs and, from the WSSUS assumption, Section 2.4.1, scattering from different scatterers is uncorrelated. These two arguments also suggest that the two scattering coefficients will have different amplitudes. Evidence that the coupling coefficients are unequal is presented by Lee, who observes that at 900 MHz $\Gamma_{\text{hv}}$ is much larger than $\Gamma_{\text{vh}}$. Lee accounts for this in terms of the predominantly vertical orientation of scatterers in the urban environment, where the scatterers largely consist of the vertical walls of buildings (Ref: Lee W CY, 1982). This leads to a far smaller probability for vertical to horizontal cross coupling than for the reverse process.

### 2.7 DETERMINISTIC MODEL OF CHANNEL RESPONSE

At first sight it may seem inappropriate to model the micro-cellular mobile radio channel with statistical models which assume a uniform angular distribution of reflected power and a uniform distribution of delays at the mobile unit. This is because, in reality, there are so few obstructions and potential scatterers in a micro-cell that propagation is dominated by specular and line of sight signals. To examine the foundation of these models, a simple deterministic model of propagation in a micro-cell was constructed.

The first part of the model calculates the path differences which occur between the line of sight path and the various specular reflection paths. These are used to provide estimates of the time delays for these paths and hence an estimate of the coherence bandwidth of the channel when the specular paths are dominant. The
relationship developed in Section 2.6 between the standard deviation of the time delays and coherence bandwidth is used for these calculations. The second part of the model generates an expression for the amplitude and phase for the received signal at the mobile unit as a function of distance between the base station and the mobile. This is used to create an "equivalent time series" of the received signal by assuming that the mobile moves at a constant speed through the micro-cell. The power spectrum of the square law detected received signal is derived from the Fourier transform of this equivalent "time series".

The model is a useful addition to the more complex models discussed in Section 2.6 because it considers the effects of specular reflections and line of sight components. These play a far larger role in the micro-cellular environment where the transmitter is always close at hand, than in a conventional cellular environment where the received signal is made up of components which have an approximately uniform angle of arrival distribution.

### 2.7.1 SPECULAR REFLECTION APPROXIMATION

The model considers only the line of sight signal and reflections from principal scatterers in the micro-cell for constructing the received signal; the principal scatterers being the road surface and the two rows of buildings lining the sides of the street. Figure 2.12 illustrates the geometry of the model. Specular reflections occur in the mobile radio environment when radiation is incident on a scatterer near to grazing angles such that the Rayleigh smoothness criterion is satisfied:

\[
C = \frac{4\pi \sigma}{\lambda \sin \theta}
\]

where: \( \sigma \) = standard deviation of surface roughness, \( \theta \) = incidence of angle to surface.
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Figure 2.12 Micro-Cellular Geometry for the Deterministic Model

Experimental evidence shows that for $C < 0.1$ specular reflection results and the surface may be considered smooth. Surfaces are considered rough for values of $C > 10$. Under these conditions the reflected wave is very small in amplitude (Ref: Jakes W C Jr, 1974). In this Thesis, we define surfaces with values of surface roughness from 0.1 through to 1 as strong or quasi-specular reflectors. For a wavelength of 5.5 mm and a locally defined standard deviation of surface heights of 3 mm which corresponds to a typical tarmac road surface or a pored concrete
girder, the maximum incidence angle for strong reflection is 13°. In a typical micro-cell of 300 m length and 30 m width, with the mobile vehicle located in the middle of the cell, all three reflections, road surface and two side walls, have angles of incidence less than 6° which gives a value for C of ~0.5. Thus, all three surfaces can generate strong, if not completely specular, reflections. Thus, to a first approximation, the received signal at the mobile unit can be described by the summation of a direct component and simple specular reflections with reflection coefficients of approximately -1.

2.7.2 COHERENCE BANDWIDTH

Using the relationship in Equation 2.91 between the delay spread, $\sigma_T$, and the coherence bandwidth, $B_c$, an estimate of the coherence bandwidth of a communication channel in a micro-cell can be made from simple geometrical considerations. Figure 2.12 shows the simplified geometry of a typical micro-cell. The micro-cell is 300 m in extent along a section of street which is 30 m wide. For simplicity, we assume that the mobile is a typical location in the middle of the street and halfway along the micro-cell. It is now possible to calculate the path differences between the direct path and the reflected paths shown in Figure 2.12. The reflected path from the road surface has a path difference of 3 m with respect to the direct path. This gives rise to a time delay of 10 ns. The resulting coherence bandwidth, if this path is dominant and the power is equally distributed between the direct and reflected rays such that $\sigma_T = 5$ ns, is 33 MHz. The reflected paths from the side walls have path differences of 1.5 m with respect to the direct path when the mobile is located on the mid-line of the street; these give rise to a time delay of 5 ns. The resulting coherence bandwidth, if these paths are dominant and power is equally distributed between reflected and direct rays, is 67 MHz.

The resultant coherence bandwidth on the channel will depend on the disposition of power between the various reflections and the line of sight signal. The greater the amount of power contained in the line of sight signal the larger the coherence bandwidth will be.

2.7.3 POWER VERSUS DISTANCE

The above model can be used to produce an approximate expression for the amplitude and phase of the received signal at the mobile unit as a function of distance. The antenna pattern was modelled with a Gaussian function. This
provides a close fit to the actual radiation pattern for a standard gain horn. The received power is approximated by a summation of the reflections from the side walls and the road surface and a direct line of sight component.

\[
A(x) = \sum_k \left( \theta \sqrt{2\pi} \right)^{-\frac{1}{2}} \frac{1}{r_k^3} e^{-\frac{4\pi B^2}{r_k^2}} e^{j2\pi r_{LOS}^2 - r_k^2} \lambda
\]

where: \( \phi \) = angular deviation from Tx bore-sight, \( \theta \) = beamwidth of Tx antenna, \( r_{LOS} \) = LOS path length, \( r_k \) = k\textsuperscript{th} path length, \( \lambda \) = free space wavelength.

The received intensity is calculated by forming the square of this quantity and is given in dB by:

\[
P(x) = 20 \log \left| A(x) \right|
\]

The length of the LOS path is:

\[
r_{LOS} = \sqrt{x^2 + (H - h)^2}
\]

where: \( x \) = Tx-Rx separation (0 - 300 m), \( H \) = Tx height (15 m), \( h \) = Rx height (1.5 m).

Using equal triangles, the length of the reflected path from the road is:

\[
r_{REF} = \sqrt{H^2 + \left( \frac{hx}{H + h} \right)^2 + \left( h^2 + \left( \frac{Hx}{H + h} \right)^2 \right)}
\]

The length of the path for reflections from the side wall, neglecting the effect of antenna height, is:

\[
r_{REF} = \sqrt{x^2 + (W \pm \delta)^2 + (H - h)^2}
\]

where: \( W \) = width of street (30 m), \( \delta \) = displacement of car from mid-line of street.

Figure 2.13 shows the predicted variation of power with distance along the micro-cell calculated using the geometrical parameters which occurred in the experimental measurements described in Chapter 3. The fall off in power with distance along the street is a simple \( r^{-2} \) power law.
2.7.4 POWER SPECTRUM

The theoretical prediction of the power spectrum of the square law detected received signal was calculated from a short section of the predicted power versus distance variation using a Fourier transform routine. The process was carried out numerically. A section of the power-distance curve at the mid-point of the micro-cell was selected. Then a moving average routine was used to filter the data to prevent aliasing. A moving average routine is equivalent to a convolution, i.e., it represents a low pass filter in the frequency domain. The routine used a Gaussian window to reduce the possibility of sidelobes being imposed on the data in the frequency domain. Finally, a triangular window function was applied to the data and its FFT was calculated. Figure 2.14 shows the resulting power spectrum. The square law power spectrum shows a very large low frequency content followed by a steep fall off in power at higher Doppler frequencies. The fall off at the higher Doppler frequencies is greater than that predicted by the models of Clarke and Aulin, Figure 2.8. This is expected because the model only considers a very narrow distribution of arrival angles which leads to a correspondingly narrow range in Doppler frequency components.
2.8 ANALYSIS OF "SLOW FADING"

2.8.1 INTRODUCTION

In Section 2.4.3 it is shown that the distribution of the received signal in the mobile radio environment may be described as the product of a "fast fading" and "slow fading" process. In this Section we are concerned with the characterization of the "slow fading" process. The approach to "slow fading", presented here, is a largely
empirical scheme for classifying the variations which occur in the mean value of the received signal, over different distance scale sizes, in terms of the signal strength versus distance power law and the location variability parameters. Section 2.8.2 defines the signal strength versus distance power law and location variability parameters which are used to give a statistical description of the variation of the mean power in an urban micro-cell. In addition, the Section presents the qualitative explanation, put forward by Suzuki (Ref: Suzuki H, 1977), for the origin of the log-normal distribution of the local mean variation.

Although a variety of experimentally or theoretically based models have been developed to predict radio propagation in land mobile radio systems, reference IEEE VT Special Edition, 1988 gives a summary of the standard ones. None of the models is applicable to propagation within a micro-cell because they are all concerned with propagation at ranges of greater than 1 Km.

No quantitative relationship is given between the characteristics of the micro-cellular radio environment such as average street width, average building height, etc because the relationship has not been determined. Indeed, the purpose of this Thesis is to carry out measurements to provide an experimental basis to allow the scattering mechanism to be examined.

In general, transmission paths from mobiles to base stations are extremely varied, ranging from direct line of sight paths, to paths severely shadowed by large terrain obstructions. Accurate prediction of transmission loss is only possible if:

i) a detailed topographical map of the local propagation area is available;

ii) the disposition of moving scatterers is known;

iii) all reflection properties are well defined.

This information is not available for the millimetre wave case and, even if it were, the analysis is cumbersome and of interest only to that one location. However, under certain idealized conditions, when loss mechanisms are determined solely by free space propagation, propagation over a plane earth or diffraction over a knife edge, the value of transmission loss may be calculated exactly. A summary of these propagation modes is presented in Section 2.8.2. In addition, there are three sources of atmospheric attenuation: rain, water vapour, and oxygen. Their effects are discussed in Section 2.8.3. Using a piece-wise application of these deterministic effects, it is possible to undertake a qualitative analysis of individual experimental time series data, and thereby explain the gross changes in mean signal level and correlate them with local geographical factors.
In a practical mobile radio situation, the resultant propagation mode is a mixture of all three modes of propagation described in Section 2.8.2 and quantitative analysis of "slow fading" is not feasible. However, it is usually observed that the balance between the different propagation mechanisms alters in different mobile radio environments. This gives rise to a degree of correlation between "environment type" and the signal versus distance power law and location variability parameters. For example, in an open country area, propagation over a plane earth predominates, resulting in a lower variance in the location variability, whilst in an urban area multiple scattering predominates resulting in a higher variance in the location variability. Classification of environment type is imprecise because of the wide variations in building heights, street widths, and street orientations which occur over relatively small distances within urban areas.

Measurements presented in this Thesis were made in two different types of environment: a high density inner city area with a large number of buildings four to five storeys high with a few buildings approximately 20 storeys high; and an open rural area with no significant scatterers apart from the road surface.

2.8.2 TRANSMISSION LOSS ON IDEAL PATHS

This Section presents a summary of the three principal propagation modes in a micro-cell: free space propagation, propagation over a plane earth, and knife edge diffraction. In very simple situations, where propagation is determined by a single mode, exact calculation of transmission loss may be made. The calculations are described respectively in Sections 2.8.2, 2.8.2.2 and 2.8.2.3.

2.8.2.1 Free Space Transmission Loss

If two antennae are separated by an unbounded region of free space, provided that there are no objects within the region to absorb or reflect energy, the propagation loss between them is governed by a simple $r^{-2}$ loss given by:

$$P_{Rx} = P_{Tx} \left( \frac{\lambda}{4\pi r} \right)^2 G_{Tx} G_{Rx}$$

where: $P_{Tx}$ = transmitted power, $P_{Rx}$ = received power, $G_{Tx}$ = transmitter antenna gain, $G_{Rx}$ = receiver antenna gain, $r$ = transmitter-receiver separation, $\lambda$ = free space wavelength.
The power law arises from the application of the principle of conservation of energy. In a mobile radio situation the constraints of the free space transmission model can never be met, but the model yields a good estimate of propagation loss provided that there are no obstructions within the first Fresnel zone of the direct path between base station and mobile unit.

### 2.8.2.2 Propagation Over a Plane Earth

The next model to consider, one step up in complexity from the free space model, is that for propagation over a smooth, conducting, flat plane, i.e., we have introduced a single well defined scatterer into the free space model. Bullington (Ref: Bullington K, 1977) has produced an analytic solution for this model:

\[
P_{\text{Rx}} = P_{\text{Tx}} \left( \frac{\lambda}{4\pi r} \right)^2 G_{\text{Rx}} G_{\text{Tx}} \left[ 1 + R e^{i\Delta} + (1 - R) A e^{i\Delta} + \ldots \right]^{2.113}
\]

where: \( P_{\text{Rx}} \) = received power, \( P_{\text{Tx}} \) = transmitted power, \( r = \text{Tx} \rightarrow \text{Rx} \) separation, \( G_{\text{Rx}} \) = Rx antenna gain, \( G_{\text{Tx}} \) = Tx gain, \( R \) = reflection coefficient, \( A \) = surface wave attenuation factor, \( \Delta \) = phase difference between reflected and direct ray.

Within the absolute value symbols, the terms represent respectively, the direct wave, the reflected wave and the surface wave. Subsequent terms represent the induction field and secondary effects of the ground. The surface wave term has significance only in a region extending at most a few wavelengths above the ground and can thus be neglected.

The reflection coefficient, \( R \), of the ground depends on the angle of incidence, \( \theta \), the polarization of the wave, and the ground characteristics; conductivity, permittivity and surface roughness. However, following the argument in Section 2.7.1, except for very small transmitter-receiver separations, the angle of incidence of the transmitted wave with the ground will be less than 5°, and it is reasonable to assume near specular reflection with a reflection coefficient approaching -1.

The phase difference between the reflected and direct paths, \( \Delta \), is expressed in terms of the receiver height, \( H \), the transmitter height, \( h \), and their separation, \( r \), as:

\[
\Delta = \frac{2\pi r}{\lambda} \left( \sqrt{\left( \frac{H + h}{r} \right)^2 + 1} - \sqrt{\left( \frac{H - h}{r} \right)^2 + 1} \right) \quad 2.114
\]
Equation 2.114 may be expanded as a binomial series. If $H^2 + h^2 \ll r^2$ the expression may be approximated by the first order term of the expansion Equation 2.115. In this case, the error will be of the order of $H^2 + h^2/2r^2$. For $H = 15$m, $h = 1.5$ m and $d > 50$ m the error will therefore be less than 5%.

$$\Delta \approx \frac{4\pi H h}{\lambda r}$$

2.115

Under conditions where $A$ can be neglected and $R = -1$, Equation 2.113 reduces to:

$$P_{R\Delta} = P_{T\Delta} \sin^2 (A/2)$$

2.116

In conventional mobile radio situations the term $\Delta$ will be small. Thus, the approximation $\sin\Delta/2 \approx \Delta/2$ can be made. This gives rise, when the approximation represented by Equation 2.115 can be made, to an $r^{-4}$ fall off in received power with distance. However, in the millimetric micro-cellular situation, the very short wavelength and comparatively short Tx-Rx separation result in $\Delta$ typically having a value of about 500. Thus, the $\sin^2\Delta/2$ factor results in oscillatory behaviour with an average value of 1. This gives rise to an $r^{-2}$ fall off in received power with distance.

2.8.2.3 Diffraction Around Simple Objects

Often in a micro-cell the line of sight path from base station to mobile is obscured by obstructions such as trees, high sided vehicles, and buildings. When the obstruction is caused by a single object, it is possible to estimate the diffraction loss using the knife edge diffraction approximation. The electric field, $E$, within the shadow region of a knife edge is then given by:

$$E = E_0 \alpha e^{i\psi}$$

2.117

where $\alpha$ and $\psi$ are expressed as Functions of the Fresnel integrals and $E_0$ is the electric field incident at the knife edge (Ref: Jakes W C Jr, 1974). However, for most millimetre wave mobile applications, several assumptions can be made to simplify the calculation of knife edge diffraction (Ref: Jakes W C Jr, 1974).
Figure 2.15 shows a typical geometry which results in knife edge diffraction around a junction into a street at right angles to the micro-cell. The building on the corner is treated as an infinite, absorbing, half-plane. Provided that the distances, $d_1$ and $d_2$, from the half-plane to the transmitter and receiver antennae are large compared to effective height of the obstruction, $h$, and provided, $h$, itself is large compared to the wavelength, $\lambda$, then the diffracted power can be approximated by:

$$P_{rx} = P_0 \frac{1}{2\pi^2 D^2}$$

where:

$$D = h \sqrt{\frac{2}{\lambda} \left( \frac{1}{d_1} + \frac{1}{d_2} \right)}$$

and where $P_0$ is the power incident at the knife edge. The result is independent of polarization if the condition $d_1, d_2 \gg h \gg \lambda$ is met.
2.8.3 ATMOSPHERIC ABSORPTION

One major obstacle with using frequencies in the millimetre waveband over fixed and satellite radio links is the influence of atmospheric conditions on propagation. In particular, atmospheric attenuation at millimetre wave frequencies is far greater than that at microwave frequencies, the main cause for this attenuation being due to atmospheric gaseous components and precipitation (Ref: Medeiros Filho F C, 1981). However, signals transmitted in the micro-cellular radio environment are subject to path losses which are dominated by shadowing and multipath propagation. Compared to these losses, excess attenuation due to atmospheric absorption, over the short base-mobile separations used, is very small (Ref: Huish et al, 1985). Section 2.8.3.1 considers the effect of absorption by atmospheric gaseous components and Section 2.8.3.2 considers the effect due to precipitation, rain, snow, fog, etc, on millimetric transmission.

2.8.3.1 Gaseous Attenuation

Figure 2.16 shows the curve of atmospheric attenuation versus frequency for the millimetre waveband at sea level for a pressure of 1 atmosphere, temperature 20 °C and water vapour density of 7.5 g m⁻³. The major features on the curve are due to an oxygen absorption peak at 60 GHz and a water vapour absorption peak at 24 GHz. The most important of these two absorption peaks is that of oxygen which results in a maximum attenuation of \(-15\ \text{dB Km}^{-1}\) at 60 GHz. The attenuation factor is approximately \(5\ \text{dB Km}^{-1}\) at 55 GHz. Thus, an excess attenuation of approximately 1.5 dB will result at a transmitter-receiver range of 300 m (Ref: Gibbins C J, 1988). However, absorption due to gaseous water vapour is less than 0.2 dB and can be neglected.
Figure 2.16 Atmospheric Absorption Versus Frequency (Ref: Meeks M L, 1976)
2.8.3.2 Precipitation Attenuation

Attenuation due to precipitation has been investigated at millimetre wave frequencies by a number of workers, a recent summary is presented by Gibbins (Ref: Gibbins C J, 1988). Measurements under the different conditions caused by fog, rain, sleet, and snow, indicate that the attenuation is primarily due to the amount of liquid water present. The attenuation factor for very heavy rain is of the order of 12 dB Km\(^{-1}\) (Ref: Gibbins C J, 1988). This results in an excess attenuation of \(-3.6\) dB for a transmitter-receiver range of 300 m. However, this condition occurs only for approximately 1 hour per year (Ref: Gibbins C J, 1988) and the effects of rain over micro-cellular ranges can effectively be ignored.

2.8.4 PROPAGATION IN AN URBAN MICRO-CELL

In the complex geometry of an urban micro-cell, the simple propagation mechanisms described in Section 2.8.4.2 combine in a random manner to produce the modulation of the mean signal level, referred to as "slow fading". An additional contribution is also made by the atmospheric absorption factors described in Section 2.8.4.3. This Section describes how "slow fading" is characterized approximately in terms of two parameters: signal strength versus distance power law, and location variability. A qualitative argument, reproduced from Suzuki (Ref: Suzuki H, 1977), is presented to explain the origin of the log-normal distribution of the location variability.

2.8.4.1 Description of Mean Power Variation

Figure 2.17 illustrates the general variation of signal strength which is observed throughout a micro-cell on a simulated time series. The variations over the different scale sizes are respectively described by a signal versus distance power law Section 2.8.4.2, location variability Section 2.8.4.3, and on the smallest scale by the transmission loss distribution. The variation over the smallest scale size is the multipath or "fast fading" which was theoretically analysed in Section 2.5.
2.8.4.2 Signal Versus Distance Power Law

On the largest scale size there is a tendency for the average signal power to decrease with distance. For the free space case, this tendency is described by an $r^{-2}$ power law. However, in the mobile radio environment, the complex net of scatterers, atmospheric absorption losses and obstruction by obstacles such as buildings and vegetation, markedly changes this power law. The smooth curve in Figure 2.17 illustrates the signal versus distance power law for a typical micro-cell.
Examination of the plane earth propagation model in Section 2.8.2.2 shows that a power law ranging from $\sim r^{-2}$ through to $\sim r^{-4}$ can be predicted, depending upon the base station height, base-mobile separation, and the wavelength. In the absence of other scatterers, the model would predict an $\sim r^{-2}$ power law for the millimetre wave micro-cell.

In the mobile radio environment (here we include propagation measurements within buildings) power laws from $r^{-1}$ to $r^{-5}$ have been observed in millimetre wave experiments (Ref: Pugliese G, Alexander S E, 1983). The power laws in each case have respectively been enhanced by channelling effects of corridors within buildings and attenuated by diffraction over objects outside buildings.

The power law is usually obtained from experimental data by low pass filtering, plotting on log-log graph paper, and visually plotting a straight line through the data. With the non-stationary nature of the data there is not much merit in using a linear regression technique to obtain a more "accurate" result.

### 2.8.4.3 Location Variability

On a smaller scale size, the mean signal power varies from location to location throughout the micro-cell in a random manner. The mean power is respectively enhanced and attenuated as the disposition of obstructions between the mobile and base station changes with the movement of the mobile through the micro-cell. Signal enhancement can result from high amplitude reflections from large scatterers, ie glass sided buildings, which act as narrow beamwidth antennas with high sidelobe levels (Ref: Cox D C, 1973). Location variability is represented by the fluctuations of the solid line in Figure 2.17.

Location variability is defined as the standard deviation of the distribution of the logarithm of the local mean power, obtained by averaging the received signal over a number of small local areas at different locations in the micro-cell, but all at similar distances from the base station.

Measurements at 900 MHz have shown that location variability follows a log-normal distribution with a standard deviation characteristic of the "type" of mobile radio environment, ie urban, suburban and rural. However, no comparable data exists for correlating environment type with location variability in a micro-cell. The correlation is thus left to empirical analysis of experimental data which is presented in Chapter 4.
Chapter 2

The log-normal distribution and its corresponding cumulative distribution are given respectively by:

\[
p(r) = \frac{1}{r \sigma \sqrt{2\pi}} e^{-\frac{(\ln r - \mu)^2}{2\sigma^2}}, \quad r > 0
\]

\[
P(r \leq R) = \text{erf} \left( \frac{\ln r - \mu}{\sigma} \right), \quad R > 0
\]

where \( \sigma \) is the standard deviation of the variable \( \ln(r) \), \( \mu \) is the mean of variable \( \ln(r) \), and \( r \) and \( R \) are distances.

The log-normal character of the location variability can be checked by plotting the cumulative distribution of the local average of the received signal on a probability plot. This type of plot has its axes arranged so that a log-normal distribution is a straight line. It is a simple matter to estimate the log variance and mean from the probability plot.

A qualitative argument was put forward by Suzuki (Ref: Suzuki H, 1977), based on the work of Turin (Ref: Turin G L, 1972), to explain the origin of the log-normal character of the location variation distribution. A log-normal distribution arises when a large number of randomly distributed errors combine in a multiplicative process (Ref: Aitchison J, Brown J A C, 1957). Suzuki suggests that a multiplicative mechanism may arise in the mobile radio scenario if two distinct scattering processes give rise to the received signal. He suggests that because the base station antenna is at a greater elevation than the mobile antenna which is the case even for micro-cellular radio, it will illuminate a widely dispersed array of scatterers each of which will produce reflections towards the mobile unit. Near to the mobile unit, a second scattering mechanism takes over and the incident reflections are re-scattered by local obstructions surrounding the mobile unit. The model gives rise to a mixture type of statistical distribution which is expressed in integral form. He then suggests that the total signal is given by a process which is multiplicative with regard to the number of distinct paths in the first scattering process and yet is additive with regard to the local scattering. Suzuki justifies his model because it agrees with measured results and explains the transition from quasi-Rayleigh statistics to log-normal statistics.
2.9 DIVERSITY

2.9.1 INTRODUCTION

Diversity reception in its theoretical description and practical application is a vast subject; a full analysis is beyond the scope of this Thesis [1]. However, because diversity is of prime importance for any mobile radio communication system, its implications must be addressed. The approach taken here is to outline the different ways in which diversity may be effected, examine the difficulties which would occur implementing these at 55 GHz, and summarize how simple measurements may be used to assess the potential of diversity in a millimetre wave radio system. The application of diversity reception is addressed again in Chapter 5 where the improvement to the bit error rate (BER) performance of a digital receiver with the application of a simple non-coherent combination scheme is considered.

Diversity techniques alleviate the effects of multipath fading by combining the responses of two or more independently fading channels. Their principle of operation is based on the assumption that independently fading channels are unlikely to suffer simultaneously, a fade deep enough to cause a break down in communications. Channels, in this Thesis, are assumed to be independent if their squared envelope correlation coefficient is 0.5 or less. Diversity systems can operate with correlations greater than this, but their performance becomes progressively more degraded (Ref: Jakes W C Jr, 1974, pp 324).

Sections 2.9.2 through 2.9.5 discuss the different classes of diversity system. These are classified based on the following factors: the scale size over which they operate, Section 2.9.2; the combination system they employ, Section 2.9.3; and the mechanism employed to produce independently fading channels. Any mechanism which provides uncorrelated input channels may be used for diversity reception. In this Thesis, space, frequency and polarization diversity schemes are investigated. The properties of these diversity mechanisms are described respectively in Sections 2.9.5.1, 2.9.5.2, and 2.9.5.3, and a summary is provided in Section 2.9.5.4. The improvement in the performance achieved by diversity reception is related to the change in the probability structure of the output of the diversity receiver compared to that of its inputs. Section 2.9.4 illustrates the

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1 A great wealth of literature is available on the theory and application of diversity reception and many references can be found in the standard works on mobile radio of Lee (Ref: Lee W C Y, 1982) and Jakes (Ref: Jakes W C, 1974).
modification to the probability structure of the data effected by selection diversity. Section 2.9.7 considers the difficulties of implementing diversity systems at millimetre wave frequencies and shows how the simple measurements summarized can be used to assess the potential of a simple diversity system.

2.9.2 MICRO-DIVERSITY AND MACRO-DIVERSITY

Micro-diversity is designed to combat short term multipath (fast) fading and requires only a single base station. Macro-diversity is designed to combat fading due to large obstructions. This type of system requires multiple base stations to provide alternate paths around obstructions such as hills or large buildings which would otherwise prevent any communication from taking place. Macro-diversity is a solution of last resort due to the extra cost incurred by duplication of base stations. In a practical situation, coverage problems are usually effected by tailoring cell geometry. A micro-cellular system with its very small cells based on street geometry would not require macro-diversity. Thus, since we are only concerned with the use of micro-diversity, only measurements from a single base station are required.

2.9.3 COMBINATION SCHEMES

Diversity combination schemes can broadly be divided into selection diversity schemes which select the input with the best performance, and combination diversity schemes which combine all the inputs together. Selection and combination schemes are discussed in Sections 2.9.3.1 and 2.9.3.2 respectively and their respective properties are summarized in Section 2.9.3.3.

2.9.3.1 Selection Diversity

Selection diversity operates by continually selecting the channel which has the best signal to noise ratio (SNR) at a particular time. Usually selection is made on the basis of \( S + N \), since SNR is difficult to measure. However, the process is sometimes impractical to implement and performance is degraded by switching transients resulting from rapid switching between channels.
Two similar processes, switched diversity and scanning diversity, can be used to reduce these problems. Switched diversity initially selects the channel with the highest SNR and then holds it until the signal falls below a threshold value, at which point the channel with the highest SNR is again selected. Scanning diversity is effected in a similar way to switched diversity. However, instead of selecting the channel with best SNR, it scans through the channels in a fixed sequence, selecting the first channel which has an S + N above a threshold value. When this channel subsequently falls below threshold, the scanning process is resumed. Switched and scanning diversity have a slightly lower theoretical performance than selection diversity. However, Table 2.3 can be used as an approximate guide to their performance.

2.9.3.2 Combination Diversity

The fundamental combination diversity algorithm is Maximal Ratio Combination. The scheme operates by bringing all the channels to a common phase, weighting them according to their SNR and adding them together coherently. This process has a significant advantage over the other diversity schemes because the signals of the various channels are added coherently which results in a dramatic enhancement of the signal combined with a reduction in the noise due to cancellation. In ideal circumstances it is the most effective diversity combination system. However, the scheme is difficult to implement due to the necessity of continually altering the weight of each channel and the need to bring each channel to a common phase.

A similar, but less complicated, system is Equal Gain Combination. The process brings each channel to a common phase and coherently sums them without weighting them. However, care must be taken to prevent inclusion in the summing process of a channel whose input has fallen below the noise threshold of its receiver; such a channel consists of highly amplified noise. Its addition would considerable degrade the receiver performance. This problem is usually avoided by completely suppressing the output of a channel which has fallen below its noise threshold.

---

1 The channels initially are brought to a common amplitude by separate automatic gain control amplifiers (AGC).
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2.9.3.3 Summary of Combination Schemes

Table 2.3 summarizes the combination schemes considered for constructing the output of a diversity receiver from its uncorrelated inputs. The Table indicates the relative performance of the different schemes and the schemes which require co-phasing of their inputs. It should be noted that the performance of the coherent combination techniques is enhanced if combination is effected pre-detection. The improvement arises because the increase in pre-detection SNR effectively extends the detector threshold (Ref: Jakes W C Jr, 1974, pp 397).

<table>
<thead>
<tr>
<th>Type of Scheme</th>
<th>Coherent</th>
<th>Improvement in Performance Relative to System without Diversity</th>
</tr>
</thead>
<tbody>
<tr>
<td>Selection</td>
<td>No</td>
<td>10 dB 99% reliability level [1] 1.5 x SNR[2]</td>
</tr>
<tr>
<td>Maximal Ratio</td>
<td>Yes</td>
<td>11.5 dB 99% reliability level 2 x SNR</td>
</tr>
<tr>
<td>Equal Gain</td>
<td>Yes</td>
<td>10.5 dB 99% reliability level 1.8 x SNR</td>
</tr>
</tbody>
</table>

Table 2.3 Comparison of Diversity Combination Schemes
(Ref: Jakes W C Jr, 1974)

2.9.4 JOINT PROBABILITY DISTRIBUTION (SELECTION DIVERSITY)

This Section illustrates how diversity reception improves receiver performance by modifying the output probability structure (Ref: Jakes W C Jr, 1974, pp 313). Selection diversity operates by selecting the channel possessing the best SNR in a continual process. As far as the statistics of the output are concerned, it is immaterial where the selection is done. The antenna signals themselves could, for example, be sampled and the best one sent to the receiver, or selection could be carried out after the various channels had been detected by separate receivers.

Assume that the signals in each of M diversity branches are uncorrelated and Rayleigh distributed. The density function of the signal envelope is given in Equation 2.34:

\[ P(y) = \frac{y^{\gamma-1} e^{\frac{-y}{\Gamma}}}{\Gamma^{\gamma} y} \]

1 Given as dB difference between the respective 99% points on the cumulative distribution curves of \( \gamma / \Gamma \) versus probability ordinate > abscissa for a one and a two branch diversity system.

2 This is the improvement in the mean SNR compared to that of the input channels.
where \( r_i \) is the signal in the \( i \)th branch. We are interested in the SNR. Thus, it is convenient to introduce two new variables, the "instantaneous SNR, \( \gamma_i \), and the mean SNR in each branch, \( \Gamma \). The local (averaged over one RF cycle) mean signal power per branch is \( r_i^2/2 \). The mean noise power per branch is \( \mathbb{E}[n_i^2] \), we assume that this is the same for all channels, \( \mathbb{E}[n_i^2] = N \). Thus, we obtain:

\[
\gamma_i = \frac{r_i^2}{2N}, \quad \Gamma = \mathbb{E}\left[\frac{r_i^2}{2N}\right] = \frac{\sigma^2}{N} \tag{2.122}
\]

and the distribution of the instantaneous SNR in each channel is given by:

\[
p(\gamma_i) = \frac{1}{\Gamma} e^{-\gamma_i/\Gamma} \tag{2.123}
\]

The probability that the SNR, \( \gamma_i \), is given by:

\[
P[\gamma_i \leq \gamma] = \int_0^{\gamma_i} p(\gamma_i) d\gamma_i = 1 - e^{-\gamma/\Gamma} \tag{2.124}
\]

The joint probability that \( \gamma_i \) in all \( M \) branches is simultaneously less than or equal to \( \gamma \) is then

\[
P[\gamma_1 \ldots \gamma_M \leq \gamma] = \left(1 - e^{-\Gamma/\Gamma}\right)^M = P_M(\gamma) \tag{2.125}
\]

This is the distribution of the best signal selected from the \( M \) branches. \( P_M(\gamma) \) is plotted in Figure 2.18 for diversity systems with 1, 2, 3, 4, 5 and 6 branches. The Figure is reproduced from Jakes (Ref: Jakes W C Jr, 1974, pp 315). The potential savings in power offered by diversity are immediately obvious: 10.5 dB for two-branch diversity at the 99% reliability level, for example, and 16 dB for four branches. The savings offered by coherent combination schemes are greater than this, as indicated in Table 2.4.
2.9.5 DE-CORRELATION MECHANISMS

2.9.5.1 Space Diversity

Space diversity requires no extra sophistication at the transmitter. It relies upon the random spatial distribution of the signal to produce uncorrelated fading envelopes when two receiver antennae are separated in space. The minimum distance required to accomplish de-correlation is defined by the spatial auto-corre-
lation function of the received signal. In terms of the squared envelope auto-correlation function, a fall in the correlation coefficient to 0.5 is usually sufficient to produce effectively uncorrelated envelopes.

The elevated position of the base station makes it relatively uncluttered by nearby scatterers, in marked contrast to the almost uniform angular distribution of scatterers surrounding the mobile. As a result, the angular distribution of received waves at the base station is narrower than that at the mobile unit. In Section 2.5.5.1 the width of the spatial auto-correlation function is shown to be dependent upon the angular arrival distribution of the received signal components; ie a narrow angular distribution gives rise to a broad auto-correlation and vice versa. Thus, greater antenna separations are required at the base station than at the mobile; typically ~0.5λ is required at the mobile and ~100λ at the base station (Ref: Jakes W C Jr, 1974). The greater space requirement at the base station can prove inconvenient. At millimetre wave frequencies, 100λ corresponds to only 50 cm which would not necessarily cause a problem for implementation. However, even 50 cm could prove inconvenient if "base stations" for micro-cellular radio were mounted on lampposts (Ref: Steele R, 1983).

2.9.5.2 Frequency Diversity

Frequency diversity has already been mentioned in regard to the coherence bandwidth of the mobile radio channel, B_c, in Section 2.6. Frequency diversity requires no antenna separations at either base or mobile. In fact, co-located antennae may be used which provides for the use of a compact system. However, the frequency separation between the various channels must be greater than B_c in order for them to be effectively uncorrelated, and every diversity channel requires its own carrier frequency. The major disadvantage of frequency diversity is the greater bandwidth required. In addition, the transmitter is more complex to implement in comparison to a space diversity system which only transmits on a single channel.

2.9.5.3 Polarization Diversity

In Section 2.6.5, the mechanism by which orthogonally polarized channels become depolarized is discussed. Polarization diversity, like frequency diversity, may be accomplished using co-located antennae to provide a compact system implementation. In addition, the method requires only a single transmission frequency, thus
avoiding the problem of additional bandwidth required by frequency diversity. Polarization may be split at the receiver simply by using a linear polarization antenna at 45° to the vertical. The method has the disadvantage, compared with space diversity, that the power in each channel is reduced by 3 dB. The most serious limitation of the technique is that only two orthogonal polarized channels exist and only two branch diversity may be implemented if this technique is used exclusively.

2.9.5.4 Summary of Diversity Mechanisms

Table 2.4 summarizes the properties of the three de-correlation mechanisms investigated in this Thesis: space, frequency, and orthogonal polarization diversity. The Table indicates that space diversity requires different antenna separations at the base station and the mobile, frequency diversity requires at least twice the bandwidth of the other methods, and polarization diversity requires a power splitter or two separate transmitters. In addition, in the case of polarization diversity, because only two orthogonal polarizations may be defined, only two branch diversity may be implemented. The different schemes have properties which make them more or less applicable for different situations depending on the available space at the base and the mobile locations, the available bandwidth, and the available power.

<table>
<thead>
<tr>
<th>Diversity Scheme</th>
<th>Specification for De-correlation</th>
<th>Comments</th>
</tr>
</thead>
<tbody>
<tr>
<td>Space</td>
<td>Antenna separation at mobile ~1λ</td>
<td>No extra power or bandwidth required</td>
</tr>
<tr>
<td></td>
<td>Antenna separation at base station ~100λ</td>
<td>-</td>
</tr>
<tr>
<td>Frequency</td>
<td>Frequency Separation &gt; Coherence bandwidth $B_c$</td>
<td>At least double the bandwidth needed, 3 dB more power required</td>
</tr>
<tr>
<td>Polarization</td>
<td>Polarization vectors orthogonal</td>
<td>Only two diversity channels available, 3 dB loss in power</td>
</tr>
</tbody>
</table>

Table 2.4 De-Correlation Mechanisms
2.9.6 DIVERSITY MEASUREMENTS

This Section briefly indicates how the potential for the various de-correlation mechanisms above may be assessed from simple measurements of the channel properties. Section 2.9.4 indicates how the joint probability distribution, ie the distribution for the output of the diversity combiner, is used to calculate the amount by which transmitter power can be reduced to achieve the same power at a given reliability level compared with the no-diversity case.

Space diversity can be very simply examined from measurement of the received signal envelope in a micro-cell as a function of distance. The normalized auto-correlation function of the signal envelope will provide a measure of the minimum required antenna separation, and the joint probability distribution provides a measure of the expected increase in performance.

Frequency diversity can be examined by considering the properties of the received signals when two transmissions are made simultaneously at different carrier frequencies. The normalized cross-correlation function, as a function of frequency, will provide a measure of the minimum frequency separation required between the channels, and the joint probability distribution provides a measure of the expected increase in performance.

Finally, polarization diversity may be assessed in a similar way by examination of the properties of the received signals when two transmissions are made simultaneously on orthogonal polarizations. The normalized cross-correlation coefficient indicates the feasibility of the technique, and the joint probability distribution provides a measure of the expected improvement in performance due to its application.

2.9.7 DIVERSITY AT MILLIMETRE WAVE FREQUENCIES

Implementation of a diversity reception in the millimetre wave frequency band is not straightforward due to the limitations of available technology. The limitations of the receiver used in this work are examined with regard to the type of diversity systems which could be implemented. In the longer term, however, most of these limitations will be soluble with the application of improved millimetre wave technology.
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The experimental equipment used in this work does not provide a stable frequency or phase at the transmitter. Thus, after the rapid phase fluctuations imposed by multipath fading, phase recovery at the receiver becomes a virtual impossibility. This rules out the use of a systems which require co-phasing of the diversity branches before combination; systems which are tolerant of both phase and frequency errors are favoured. However, a differential phase coherent system could be possible.

The cost of millimetre wave components is substantially greater than that of baseband or IF components and as far as can be predicted this will remain so in the future. Thus, schemes which combine signals at baseband or IF [1] are favoured against schemes which employ multiple RF receivers. In addition, the use of square law detection rules out the possibility of using any form of coherent post-detection diversity.

On the positive side, the pre-detection bandwidth of the receiver, 1 GHz (Table 2.5), is sufficient to contain at least 10 independently fading channels provided that the experimental coherence bandwidth is not greater than 100 MHz, Section 2.7.2. The separation of the channels could be accomplished by IF filters and only one set of millimetre wave receiving equipment would be required.

However, both polarization and space diversity schemes would be difficult to implement without the use of multiple receivers which would be an expensive option. The problem could be circumvented by sampling the RF signal at the antenna terminals and using some sort of selection diversity. However, this option would require the use of expensive millimetre wave hardware in order to handle the sampling and switching operation. Alternatively, a polarization diversity system could be implemented by introducing a small frequency offset between the orthogonal polarizations to allow the two channels to be separated at IF. The two channels could then be combined (selected) incoherently after detection. In this case, the frequency separation need only be sufficient to contain the modulation on each channel, ie it can be substantially smaller than the ~100 MHz coherence bandwidth. Thus, in summary, the most suitable types of diversity system, given the constraints of the experimental equipment, appear to be post-detection, frequency diversity or frequency/polarization diversity schemes, utilizing either selection or incoherent combination algorithms.

---

1 Intermediate Frequency (IF).
2.10 CHARACTERIZING REAL SIGNALS

In the preceding sections, the properties of the received signal are considered from a theoretical standpoint. Here, we address the problem of choosing a set of measurable parameters for characterizing the received signal experimentally. The aim of the analogue measurements is to characterize both the short term (fast) and long term (slow) variations of the signal in a way which is useful for predicting the performance of a future communication system. The choice of parameters is moderated by the bandwidth constraints of the experimental equipment which make it necessary to find indirect ways for estimating the coherence bandwidth and delay spread of the channel, Section 2.10.1. In addition, assumptions outlined in Section 2.5.1 led to the prediction of a Rayleigh/Rice distribution for the transmission loss distribution. A re-assessment of these assumptions in Section 2.10.2 results in the replacement of the Rayleigh/Rice distribution with an empirically fitted Weibull distribution.

2.10.1 EXPERIMENTAL BANDWIDTH LIMITATIONS

At this point we consider the limitations of our measuring apparatus for probing the time delay spread and coherence bandwidth of the mobile radio channel. Table 2.5, below, summarizes the bandwidth limitations of the various stages of the receiver.

<table>
<thead>
<tr>
<th>Device</th>
<th>Comment</th>
<th>Bandwidth</th>
</tr>
</thead>
<tbody>
<tr>
<td>RF Mixer</td>
<td>Single sideband</td>
<td>1 GHz (3dB)</td>
</tr>
<tr>
<td>Square Law Detector</td>
<td>Back diode</td>
<td>30 MHz (3dB)</td>
</tr>
<tr>
<td></td>
<td>Schottky diode</td>
<td>50 KHz (3dB)</td>
</tr>
<tr>
<td>Logarithmic Amplifier</td>
<td>High signal level</td>
<td>80 KHz (3dB)</td>
</tr>
<tr>
<td></td>
<td>Low signal level (&lt;-80 dBmV)</td>
<td>1 KHz (3dB)</td>
</tr>
<tr>
<td>Tape Recorder</td>
<td>Maximum tape speed 60 ips</td>
<td>20 KHz (1dB)</td>
</tr>
</tbody>
</table>

Table 2.5 Receiver Bandwidth Limitations

In Section 2.7.2 path differences between the direct and reflected ray in an urban micro-cell are estimated to be of the order of 6 m which results in a delay time of $2 \times 10^{-4}$ seconds. To measure this delay directly requires a bandwidth of at least 50 MHz. Examination of Table 2.5 shows that it is impossible to achieve this bandwidth using the available experimental equipment. However, in Section 2.6.1, it is shown that the coherence bandwidth and delay spread may be estimated indirectly from the cross-correlation coefficient between the received envelopes of two simultaneously transmitted signals at different carrier fre-
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frequencies. In this case, the bandwidth which we are concerned with is the maximum fluctuation frequency of the envelopes. This is approximately 1 KHz and is defined by the maximum Doppler frequency. Thus, the bandwidth requirement for the indirect measurements are easily met by the tape recorder and, in practice, a tape speed of only 15 ips was used (5 KHz bandwidth) to increase the available recording time.

2.10.2 RE-ANALYSIS OF "FAST FADING"

In a real experimental scenario the assumptions made in Section 2.5 that the received signal possesses a very large number of scattered components, a uniform angle of arrival distribution and independent delay and angle of arrival distributions, are not met. In the general case, there is a limited number of received waves, a non-uniform angle of arrival distribution, and cross-coupling between the angle of arrival and delay time distributions. The failure of these assumptions distorts the probability structure of the received signal from its predicted Rayleigh/Rice character.

To quantify the actual distribution of the received signal envelope, a new quantity, the loss deviation (LD), is used. Loss deviation is defined as the decibel difference between the 50% and 90% points on the cumulative distribution of the small area signal strength. In addition, the Rayleigh/Rice statistical model is replaced by a Weibull statistical model (Ref: VT Special Issue, 1988) with distribution and cumulative distribution given respectively by:

\[ p(r) = \frac{k}{\theta} \left( \frac{r}{\theta} \right)^{k-1} e^{-(r/\theta)^k} \]  \hspace{1cm} 2.126

\[ P(r \leq R) = 1 - e^{-(r/\theta)^k} \]  \hspace{1cm} 2.127

where \( \theta, k, r > 0 \).

The Weibull distribution has the Rayleigh distribution as a special case when \( \theta = \sqrt{\sigma} \) and \( k = 2 \). It can also be made to approximate distributions from the negative exponential to the normal distribution. There is no theoretical justification for using the Weibull distribution as there is in the case of the Rayleigh, Rice, or even the log-normal distributions. However, the distribution is widely used empirically to model the extreme values of distributions (Ref: D'Agostino R B, Stephens M A, 1986).
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2.10.3 PARAMETER CHOICE

This Section is a collation of the quantities, defined in the preceding sections of this Chapter, which are used for characterizing the received signal in the mobile radio environment. Figure 2.19 summarizes the conceptual approach adopted for the analysis and Table 2.6 summarizes the quantities which are used to effect this analysis. Before the Figure and the Table are presented, a brief synopsis of the preceding Chapter is given.

The most important factor to be considered for assessing the feasibility of a mobile radio system is the disposition of the received signal power throughout the micro-cell because, in simple terms, if the received power is insufficient to drive the receiver, the communication link will break down. However, because the envelope of the received signal strength is a non-stationary random variable, due to the complex relationship between the propagation path and the disposition of scatterers in the micro-cell, it cannot be quantified in a straightforward manner (see Section 2.1).

The basic assumption made about the signal, in order to deal with its non-stationary character, is that it may be modelled as a product type non-stationary process. The approach is formalized in Sections 2.4.2 and 2.4.3 which define the product model, and "fast fading" and "slow fading" respectively. In Section 2.8.4.1 we illustrated, using Figure 2.17, how the received signal strength may be divided phenomenologically into variations which occur over different scale sizes described by the following quantities: signal versus distance power law, location variability, and on the smallest scale the transmission loss distribution ("fast fading"). Together, these quantities form a complete description of the first order statistics of the signal.

In order to examine the potential of diversity reception, we are interested in the properties of the joint probability distribution function of the "fast fading" envelope for two channels which are de-correlated due to space separation, frequency separation, or of orthogonal polarization. When this quantity is known, following Jakes (Ref: Jakes W C Jr, 1974), the improved SNR may be calculated using an approach similar to that in Section 2.9.4.

In addition to the distribution of signal power, however, we are also interested in the power spectrum and auto-correlation of the signal envelope, introduced in Sections 2.5.4 and 2.5.3 respectively, because they give an insight into the propagation process.
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The spatial distribution of the scatterers in the micro-cell gives rise to a range of time delays in the components of the received signal, resulting in a finite coherence bandwidth for the channel, Section 2.6. The experimental apparatus does not have the bandwidth to directly measure this coherence bandwidth or to resolve the very short time delays between the direct and reflected rays (see Section 2.10.1). However, coherence bandwidth and delay spread can be estimated from measurements carried out simultaneously at two separate frequencies or from swept frequency measurements (see Section 2.6.1).

Finally, the different scattering properties of the micro-cellular environment for horizontal and vertical polarization waves gives rise to de-correlation (see Section 2.6.5). The cross-correlation coefficient for simultaneously transmitted vertical and horizontal polarization waves provides a measure of this.

Figure 2.19 presents the logical structure adopted to characterize the received signal in the mobile radio environment. Some of the quantities presented are inter-related are derived from the same experimentally measured data set.
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Measured Data: Time Series; Frequency Series; Dual Frequency/Polarization

First Order Statistics

- Large Scale Variations
  - Signal Versus Distance Power Law

- Medium Scale Variations
  - Location Variability

- Small Scale Variations
  - Transmission Loss Distribution

Second Order Statistics

- Auto-Correlation
- Spectrum

Joint First Order Statistics (Statistics of the Combined Signal Envelope)

- Increase in Mean SNR after Combination
- Increase in 90% Reliability Level of the SNR

Joint Second Order Statistics

- Frequency Cross Correlation
  - Coherence Bandwidth
  - Delay Spread
- Polarization Cross Correlation

Figure 2.19 Conceptual Approach to Data Analysis
Chapter 2

The Table indicates the types of experiment required to measure these quantities and the parameters used to describe them. The description of the experimental results in Chapter 4 follows the approach presented in Figure 2.19

<table>
<thead>
<tr>
<th>Measurement</th>
<th>Measured Quantity and Its Parameters</th>
</tr>
</thead>
<tbody>
<tr>
<td>Wide Area Probing</td>
<td>Signal Versus Distance i Power Law ii Comparison to Free Space Loss</td>
</tr>
<tr>
<td>Single Channel Probing</td>
<td>Location i Mean (dB) ii Standard Deviation (dB)</td>
</tr>
<tr>
<td></td>
<td>Transmission Loss Distribution i Mean (dB) ii Loss Deviation (dB) iii Power Spectrum iv Auto-Correlation</td>
</tr>
<tr>
<td>Dual Frequency Probing</td>
<td>Joint Transmission Loss Distribution i Increase in Mean SNR after Combination ii Increase in 90% Reliability Level SNR after Combination iii Cross-Correlation iv Coherence Bandwidth (MHz) v Delay Spread (µS)</td>
</tr>
<tr>
<td>Swept Frequency Probing</td>
<td>Squared Envelope of Transfer Function i Delay Spread (µS) ii Coherence (MHz) Bandwidth</td>
</tr>
<tr>
<td>Dual Polarization Probing</td>
<td>Joint Transmission Loss Distribution i Increase in Mean SNR after Combination ii Increase in 90% Reliability Level SNR after Combination iii Cross-Correlation</td>
</tr>
</tbody>
</table>

Table 2.6 Parameter Choice
Chapter 3 ANALOGUE EXPERIMENTAL EQUIPMENT

3.1 INTRODUCTION

A variety of different types of experiment were performed using the equipment described in this Chapter: single frequency (basic system), two frequency, two polarization, wide area probing, and swept frequency measurements. To facilitate this, the equipment was built in modular form to allow easy re-configuration. The equipment consists of four main functional blocks: transmitter, receiver, data collection, and data acquisition and analysis. The approach adopted for describing the experimental equipment is, first, to describe the basic experimental configuration, then, using this as a starting point, to describe the other more complex experimental configurations with reference to it.

Section 3.2 describes the four main functional blocks in the basic experimental configuration; the functional blocks are illustrated in Figure 3.1 and photographs of the experimental equipment are included in Section 3.2.1. Sections 3.3 and 3.4 examine the details of the experimental equipment; Section 3.3 describes the properties of the Gunn oscillators used to generate the transmitted signal and local oscillator; and Section 3.4 describes the antenna design for both transmitter and receiver installations for the various experimental configurations.

The power budget is the most fundamental property of a radio link; it determines whether it will work or not. Section 3.5 describes the quantities which make up the power budget calculation: effective radiated power (ERP), propagation loss (PL), and the minimum detectable signal of the receiver (MDS). The power budget calculation for the basic experimental system is summarized in Table 3.2. The experimental configurations required for the other measurements are described in Sections 3.6 through 3.9; each Section describes the effect which the different system configurations have on the link power budget.

Finally, Section 3.10 presents a summary of the various experimental configurations. Their individual requirements are summarized in Table 3.3 and their respective fading margins, at 30 m and 300 m receiver transmitter separations, are presented in Table 3.4.
3.2 THE BASIC SYSTEM

Figure 3.1 illustrates the basic system. The transmitter is a 90 mW Gunn oscillator which generates a 55 GHz output and the transmitter antenna is a 25 dB standard gain horn which has a beamwidth of 10°. This beamwidth was chosen to provide adequate illumination of the streets used for the propagation study, whilst also maintaining a relatively high gain. More details of the transmitter assembly are presented in Section 3.2.2.
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The receiver is mounted on the roof of a car. The scattered and reflected energy from the transmitter is received by a 120° scalar horn. The receiver uses a 2 GHz bandwidth balanced mixer with Gunn local oscillator to down convert the signal to an intermediate frequency. After the down-conversion, the signal is amplified and the desired signal is selected from noise and interference by a 15 MHz bandpass filter. The signal is then square law detected and passed through a logarithmic amplifier. The logarithmic amplifier compresses the dynamic range of the signal which facilitates recording and digitization prior to processing the results. More details of the receiver assembly are presented in Section 3.2.3.

The log-converted signal is recorded together with the car speed information onto magnetic tape. An FM tape recorder is used for this purpose since it has a frequency response which extends to DC. The speed information is obtained from a sensor attached to the car speedometer. This provides a pulse every 12.5 cm of distance travelled. Further details of the data collection process are given in Section 3.2.4.

Data acquisition is performed by sampling the recorded information with a 12-bit analogue-to-digital converter controlled by an IBM PC-AT computer. The resulting data files are then processed by various computer programs to produce estimates of the statistical properties of the signal, mean, variance, correlation coefficient and power spectra.

3.2.1 PHOTOGRAPHS OF THE EQUIPMENT

Figure 3.2 is a photograph of the transmitter assembly. The photograph illustrates the configuration used for some of the two polarization measurements before the dual polarization horn was constructed. On the front of the box, the top ammeter indicates the charge state of the batteries; this can be switched back and forth between the two batteries by the switch on the left of the meter. The red line indicates the adequate charge level. On the immediate left of the meter is the fuse which protects the Gunns in case of short circuit. The two lower ammeters indicate the bias voltage to each of the Gunns. The switches on the left of each meter allow each Gunn to be turned on and off independently. The LED's on the left of the switches indicate that the Gunns are operating. The multi-turn recessed potentiometers below the LED's are used to adjust the bias voltage to each Gunn. On each side of the box are the 6 V, 6 AH batteries used to power each Gunn. In the far corner of the box the voltage regulation circuit for both Gunns are mounted on a heatsink. Running across the centre of the box are the two transmitter assemblies which consist of: heatsink, Gunn oscillator, isolator and standard gain...
Chapter 3

horn antenna. The nearest transmitter contains a 90° waveguide twist to rotate the plane of polarization. In the Figure, the nearest horn is transmitting vertical polarization and the furthest, horizontal polarization. In the centre of the box is a cooling fan which provides increased circulation of air over the heatsinks for operation in very hot conditions. Also visible, mounted on the fan, is a large smoothing capacitor which minimizes supply line ripples caused by the fan. The right hand wall of the box is made of expanded polystyrene. This provides protection from the elements, but allows unimpeded transmission, due to its near unity dielectric constant. During experiments the transmitter assembly has a perspex lid.

Figure 3.3 is a photograph of the RF section of the receiver assembly. This configuration is used for the polarization measurements. However, the difference between this and the configurations used for the other experiments is slight. In the foreground is the 120° beamwidth scalar horn antenna and behind it is the polarization selection switch. After the polarization switch, the signal is fed into a circular to rectangular waveguide transition which only passes vertically polarized signals. Thus, whether the receiver is sensitive to vertical or horizontal polarization depends on the setting of the polarization switch. Beyond the polarization switch is the large bulk of the precision rotary vane attenuator used to calibrate the system. To the left of the attenuator lies the heatsink on which the voltage regulation circuitry is mounted. Visible at the rear of the assembly is the wave meter used to check the local oscillator frequency. At the far right hand corner of the assembly is the local oscillator's heatsink. The waveguide assemblies for the RF mixer and level setting attenuator are largely obscured by the attenuator in the foreground of the picture. These are more clearly illustrated in the block diagram of the receiver assembly.
Figure 3.2 Photograph of Transmitter Assembly

Figure 3.3 Photograph of Receiver Assembly
Figure 3.4 shows the basic configuration for the transmitter. The 90 mW Gunn oscillator type CME919 AC, manufactured by Alpha Industries, has a nominal centre frequency of 55 GHz. The output of the Gunn is fed via a 30 dB isolator to the transmitter antenna which is a 25 dB standard gain horn (see Section 3.4). The isolator prevents the frequency of the Gunn being pulled by varying mis-matches in the vicinity of the antenna. Mis-matches are caused by reflections from moving cars and also when transmitting through a glass window. More details of the properties of the oscillator are given in Section 3.3.

The transmitter is powered either by a 6 V, 6 AH battery which makes the transmitter completely portable, or by an external stabilized power supply unit. The operation of the Gunn oscillator is monitored by an LED and a volt meter. The LED provides a quick indication of battery failure and the volt meter is used to aid setting up the Gunn bias voltage. The battery allowed about 3 hours' continuous transmission. If battery voltage drift were found to be a problem during the experiments, it could be avoided, at the expense of portability, by using
a stabilized power supply. Frequency stability is further increased with this arrangement because more time is available to allow the Gunn to reach thermal equilibrium as battery discharging is no longer a problem.

The Gunn oscillator requires a voltage of 3.3 V at a current of about 1 A. The voltage is controlled by a voltage regulator. In addition, the Gunn is protected from excess voltage by a crowbar IC. This shorts the Gunn's power supply lines when the voltage exceeds 4.5 V causing the supply line fuse to blow. Details of both these circuits are contained in RS [1] data sheets 8818, July 1988 and 3396, March 1985 respectively. The designs are not duplicated here.

The portable transmitter was mounted on a sturdy tripod in a weather proof box during data collection. The tripod, manufactured by Gitza, is designed to carry a professional 5" x 4" camera and as such it could easily cope with the 4 Kg weight of the most complex transmitter configuration. The use of tripod mounting allowed the transmitter to be positioned either within buildings or in the street where it could be fully extended to give a 2.6 m elevation. The tripod head also allowed accurate positioning of the head in azimuth, elevation and horizontal tilt.

3.2.3 THE RECEIVER

Figure 3.5 shows the basic configuration for the receiver. This can be considered in two sections: the RF circuitry comprises one section, and the detection and baseband circuitry comprises the other. The RF section is mounted on the roof of the car and the detection-baseband section is located inside the car. The two are connected by a coaxial cable to pass the signal and a 6-way connector for the various power supply voltages. Military specification connectors are used since other types were found to be too weak to withstand the strain of field operation. The RF circuitry consists of antenna, calibration attenuator, mixer, local oscillator and IF amplifier. All the components, apart from the amplifier, are connected by short lengths of waveguide. The functional blocks of the detection-baseband section are interconnected with BNC cables.

---

1 Radio Spares.
3.2.3.1 The RF Sub-Section

The receiver uses a 120° beamwidth scalar horn antenna with 6 dB gain. The details of this antenna and the reasons for its choice are discussed in detail in Section 3.4.2. The received signal is passed through a precision rotary vane attenuator for calibration of the whole receiver system. The signal is combined with a local oscillator in a 2 GHz bandwidth balanced mixer type 9601V05HZ,
manufactured by Alpha Industries, to give an intermediate frequency of about 150 MHz. The output of the mixer is amplified and fed via a coaxial cable to the detection-baseband circuitry located in the car.

The RF section of the receiver is contained in a weather proof box so that measurements can be taken in any weather conditions. This, in turn, is mounted on a rotary mount which also allows for adjustment of the antenna elevation.

3.2.3.2 The Local Oscillator

The local oscillator (LO) is of the same type as the transmit oscillator and uses an identical voltage regulator to provide a smooth bias voltage. The LO output passes through an isolator to a level setting attenuator which sets the power for optimum mixer performance. The manufacturers' data sheet specifies an LO power of +7 dBm. However, in practice, LO power was set by observing a spectrum analyser display and adjusting the power to maximize the signal to noise ratio (SNR). The LO power was fed to the mixer via a 10 dB coupler which had a wavemeter and crystal detector on the coupled arm for monitoring LO frequency.

3.2.3.3 The Detection-Baseband Sub-Section

The amplified signal from the RF sub-section is fed to the detection-baseband sub-section via a coaxial cable. The intermediate frequency (IF) is selected using tuneable VHF filters. These span the range 100 - 175 MHz, 200 - 375 MHz and 400 - 750 MHz to cover the centre frequency and sidebands for experiments in which the Gunn oscillator is narrow band frequency modulated, Section 3.6. Three types of filter were used which have 15%, 10% and 1% 3dB bandwidths respectively. These filters define the receiver noise bandwidth and so are very important in defining receiver sensitivity. The output signal is then square law detected to produce an output which is proportional to the squared envelope of the received signal.

The dynamic range of the signal at this point is very large. The signal in the mobile radio environment is subject to rapid fades of the order of 20 - 30 dB which, when added to the change in average propagation path loss over a range of approximately 300 m, result in an input dynamic range of the order of 40 dB. The dynamic range is too great to facilitate easy recording, due to the problem of noise at low signal levels and distortion at high levels. In addition, digitization of the
signal also presents a problem as the A/D converter only has a 12-bit resolution which is equivalent to ~ 32 dB. To avoid these problems, the square law detected signal is passed through a logarithmic amplifier type 4358, manufactured by Teledyne Philbrick. This circuit produces an output which conforms to a logarithmic input-output relationship with 1% error over an input dynamic range of 40 dB. It has a total input dynamic range of 80 dB. The log-converted signal is then passed to the tape recorder for recording and subsequent processing.

3.2.3.4 Power Supply

The oscilloscope and spectrum analyser are fed from up-converted mains voltage. The control unit and up-converter are driven by separate 63 AH sealed lead-acid batteries sufficient for approximately 3 hours' experimentation. The exact duration is dependent on the precise system configuration. Two ammeters, mounted in the car display the battery voltage and the current being drawn by the experimental equipment. A third ammeter is used to display the Gunn bias voltage. The control unit contains a DC/DC converter which provides the 15 - 0 - 15 voltage used for the logarithmic amplifiers and the IF amplifier on the car roof. The inputs and outputs of the logarithmic amplifiers are terminated in BNC connections on the front of the control unit.

3.2.4 DATA COLLECTION

Figure 3.6 illustrates the data collection process and car speed sensor. The tape speed used for the majority of experiments is 15 ips giving a frequency response which is flat to within 1 dB from 0 to 5 KHz and a recording time of 48 minutes for a 1100 m tape.

The 5 KHz frequency response gives a large safety margin for recording the maximum fading frequency of ~1 KHz expected in the mobile environment for the car speed and wavelength used in these measurements (Ref: Clarke R H, 1968). The car speed is obtained from a sensor attached to the car speedometer cable. This produced 10 pulses per revolution of the cable. After calibration over a known distance, the distance between pulses was calculated to be 12.5 cm which corresponds to ~20λ at 55 GHz. The car speed pulses were recorded together with the received signal information.
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Figure 3.6 Data Collection Process
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3.2.5 DATA ACQUISITION AND PROCESSING

Figure 3.7 Data Acquisition - Processing Stage

Figure 3.7 is a block diagram of the data acquisition and processing stage. Data acquisition is performed by sampling the recorded information using a 12-bit A/D converter. The experiments in this Thesis use sample rates of 10 KHz, 5 KHz, and 1 KHz which give bandwidths of 5 KHz, 2.5 KHz and 500 Hz respectively.

To minimize the effect of noise and spurious signals greater than the Nyquist frequency being folded into the sample bandwidth, a low-pass filter was used. This provided 24 dB attenuation at the Nyquist frequency increasing by 12 dB per octave above it. The attenuation combined with the very low residual power at the Nyquist frequency resulted in aliasing errors being smaller than errors due to noise. To verify this, the data was sampled at 20 KHz and the spectra for the different sampling rates were compared. If aliasing were a problem, the two spectra would have been different due to the different folding frequency.
However, they were identical, allowing for experimental error. Filtering reduced the effective 3 dB bandwidth at the different sample rates to 2 KHz for the 10 KHz sample rate, 1 KHz for the 5 KHz rate, and 200 Hz for the 1 KHz sample rate.

The output from the A/D converter consists of 12-bit binary numbers equivalent to the integers 1 to 4096. In the computer, these numbers are converted to equivalent real voltage levels using a calibration routine which makes use of the calibration levels recorded before every experimental run.

A major source of non-linearity, ie non-conformity to a logarithmic transfer function in the receiver system, is due to DC voltage error imposed on the input to the logarithmic amplifier. The prime source of this DC offset is the square law detected and averaged wideband noise from the mixer and IF amplifier stages. This noise is essentially flat across the 15 MHz IF filter bandwidth and the square-law detection and filtering procedure produces a constant DC term corresponding to its mean power. The square law detection process produces, in addition, a filtered convolved version of the input noise extending up to 2 KHz; a varying DC term corresponding to the average signal power, and two spectral components, extending up to 2 KHz, representing the convolution of the signal spectra and the interaction of the noise and signal spectra (Ref: Davenport W B Jr, Root W L, 1958). If a correction is not made for the DC offset on the logarithmic amplifier input, the output dynamic range of the receiver is compressed. It is possible to remove this DC noise term because it is effectively a constant. However, the varying spectral components caused by the noise cannot be removed and represent residual errors, albeit of small amplitude.

The computer used to control the operation of the A/D converter has a storage limitation of 32 K points. Thus, the sample time is limited to approximately 3.2 seconds for a single channel, or 1.6 seconds for two channels at a sample rate of 10 KHz. This time corresponds to a distance of 6.4 m, or 1163X at a car speed of 4 ms\(^{-1}\) (approximately 10 mph) which is a typical speed in the mobile environment in the presence of traffic.

### 3.3 Gunn Oscillator Properties

There are a number of factors which govern Gunn oscillator performance. In particular, frequency stability under changes of temperature, voltage and load mis-match are critical. These effects are described below.
3.3.1 FREQUENCY STABILITY WITH TEMPERATURE

Gunn oscillators are susceptible to frequency variations due to temperature changes. The dependence arises predominantly from the thermal expansion of the oscillator cavity and is aggravated by the inefficiency of the Gunn device. The typical efficiency of a Gunn oscillator is about 2%. Thus, a 100 mW device will generate 5 W of heat. To allow heat generated by the Gunn device to be dissipated without giving rise to too large a rise in temperature, the oscillators are fabricated from a solid block of metal which provides a good path for heat dissipation. In addition, the Gunns were mounted on large fan cooled heatsinks.

3.3.2 POWER STABILITY

The change in the oscillator power with the variations in temperature and bias voltage encountered experimentally is very small, typically less than 1 dB, and can be ignored.

3.3.3 FREQUENCY STABILITY WITH VOLTAGE

Figure 3.8 Gunn Oscillator Frequency Versus Bias Voltage Curve

Figure 3.8 shows the typical relationship between frequency and Gunn bias voltage. The curve is obtained at a room temperature of 21° using device CME919AC. The Gunn is usually operated near to the peak of the curve as this
yields the maximum power and also reduces the sensitivity of the Gunn to small variations in bias voltage. As can be seen from the Figure, the practice of tuning a Gunn by bias voltage has disadvantages because the Gunn device will be operated in the region of greater voltage-frequency slope which increases the amount of residual oscillator noise converted into random AM and FM modulation (see Section 3.3.5).

Initially operation of the cooling fans caused FM modulation of the Gunn oscillator with approximately 10 KHz deviation. The sensitivity of the Gunn oscillator to bias voltage pushing is \( \approx 290 \text{ MHz V}^{-1} \), so a deviation of 10 KHz requires only 0.034 mV change in bias voltage. The line voltage regulation of the 388 K chip is 0.005%. Thus, a 0.034 mV change in output requires change in input of 0.68 V. Each cooling fan draws an average current of 0.55 A with a peak current of 0.9 A and the approximate resistance of the wire connecting the fan to the battery in the car was 0.2 x 2 = 0.4 \( \Omega \) (wire resistance \( \approx 0.016 \Omega \) ft.\(^{-1}\)). Thus, the fans can cause a voltage drop of \( \approx 0.7 \text{ V} \). This effect was avoided by providing separate voltage lines to supply the fans. Thus, the voltage drop at the fan is not impressed onto the supply to the Gunn voltage regulator.

### 3.3.4 FREQUENCY PULLING

An isolator with an insertion loss of approximately 1 dB and a reverse loss of 30 dB was used at the output of the Gunn oscillator. This reduced the effects of varying mis-matches on the Gunn frequency to a negligible level.

### 3.3.5 NOISE

Two types of noise are present in the oscillator output: AM and FM noise. The effects of AM noise were not found to be significant during these experiments and were ignored. Figure 3.9 shows the measured FM spectrum of the IF output signal of the experimental receiver. The spectrum is the combination of the TX Gunn and LO Gunn FM spectra. The main cause of FM noise in Gunn oscillators is flicker noise. The spectrum of flicker noise is concentrated in the audio frequency range and is approximately a function of \( 1/f \). However, it can have a very large amplitude which causes high order inter-modulation products with the carrier frequency resulting in a very wide spectrum (Ref: Davis R G, Lazarus M J, 1986).
The conversion of flicker noise into random FM modulation is a function of the frequency pushing characteristic of the oscillator. Thus, it can be minimized by choosing a voltage operating point which minimizes the slope of the frequency pushing curve (see Figure 3.9). Conversely, excessive use of bias voltage tuning will degrade flicker noise.

![RF Noise Spectrum of Gunn Oscillator](image)

**Figure 3.9 Gunn Oscillator Noise Spectrum**

**3.3.6 FREQUENCY TUNING CHARACTERISTICS**

The oscillators are mechanically tuneable over approximately 100 MHz. Tuning of the oscillator can be effected by altering the bias voltage to the Gunn. This can be used to vary the frequency over a range of approximately 200 MHz. However, as mentioned in Section 3.3.5, this leads to a degradation in the noise performance of the oscillator. Bias tuning is, however, very useful for identifying the correct signal on the spectrum analyser screen when spurious frequencies are present.

In addition, two of the oscillators contained varactors to allow electronic tuning of the Gunn oscillators over a 200 MHz range. The varactors change capacitance with applied voltage which changes the electronic size of the cavity altering the
resonant frequency. The varactors could also be used to FM-modulate the Gunn oscillators. This was used to generate two frequency coherent sidebands for the two frequency measurements, a wide frequency chirp for the swept frequency measurements and also to provide frequency-shift-keying for the digital experiments.

3.3.7 SUMMARY

The mobile experiments described in this Thesis were carried out over relatively short periods of time and it was found to be unnecessary to provide elaborate forms of temperature control, such as Peltier devices. Peltier devices can be used to minimize long term drift of oscillator frequencies by controlling operating temperature (Ref: Medeiros Filho F C, 1981).

Experiments were usually made after allowing approximately 5 minutes for the oscillators to approach thermal equilibrium. Subsequent shifts in frequency were adjusted between experiments. The resulting frequency drift with time during experiments was found to be \(-400 \text{ kHz min}^{-1}\) when the Gunns were battery powered, and \(-200 \text{ kHz min}^{-1}\) when the Gunns were mains powered. The initial rate of drift on switch on was \(-2 \text{ MHz min}^{-1}\). Fluctuations in power were very small, less than 1 dB, and their effects could be ignored in comparison to the experimentally measured fading of 20 - 30 dB.

Frequency drift with bias voltage which occurred as the batteries discharged, made it very difficult to carry out the dual polarization measurements as these require two Gunns to be operated with a very small frequency separation. However, this difficulty was avoided by powering the two Gunns from a mains supply and leaving a considerably longer time for the oscillators to gain thermal equilibrium.

3.4 ANTENNA DESIGN

The directivity of both receive and transmitter antennas are critical elements in the design of a mobile radio system. The function of both antennae is to maximize the transfer of energy between the receiver and transmitter in all propagation situations. Clearly, for a free space point-to-point link, the solution is to increase the antenna gain as far as practical given the constraints of cost and alignment error. However, in the mobile radio environment, the situation is not so obvious.
Chapter 3

The antennae must allow transmission between the transmitter and receiver even when there is an obstruction to the line of sight path. To try to fulfil this objective, the antennae must make use of all available scatterers to provide alternative signal paths to the mobile unit in the absence of a line of sight path.

It must be noted, however, that under some propagation conditions there may be no suitable paths from transmitter to receiver able to provide adequate signal strength for communication. Normally, in a cellular structure, the mobile unit would then be passed on to a different base station. If this is not the case, then a second spatially separate antenna must be used to provide an alternative signal path. This system is termed macro-diversity, Section 2.9.2.

3.4.1 TRANSMITTER ANTENNA DESIGN

The aim of the transmitter antenna is to provide an antenna radiation pattern which is in some way matched to the spatial scattering characteristics of the environment. This is equivalent to the idea of using a matched filter to provide optimum reception of a signal, but expressed in the space domain.

The solution adopted here is to consider the geometry of the urban micro-cell as an open ended box with lossy walls. The task of the transmitter antenna is to illuminate this box, ideally producing uniform signal strength throughout. One solution is to illuminate the micro-cell from an antenna suspended above the micro-cell with its radiation pattern shaped to match the dimensions of the micro-cell. This is similar to the illumination provided by the very tall down lights used at major road junctions, or lights which are suspended on wires over urban streets. The advantage of this scheme is that it may be incorporated into existing structures in the micro-cell.

However, this solution of vertical illumination is not available. The solution adopted instead is to illuminate the micro-cell with a relatively narrow beamwidth antenna mounted at one end of the street; antenna heights from 2.6 m up to 40 m are used. The beamwidth is chosen to illuminate the side walls of the street at a distance of 50 m because this is where the obstruction of the line of sight signal by high sided vehicles starts to become a problem for a typical transmitter height of 15 m.
3.4.2 RECEIVER ANTENNA DESIGN

The main constraint of the receiver antenna design is to collect as much as possible of the transmitted radiation. Most radiation is incident in a cone facing towards the transmitter, but some radiation is scattered to arrive from all possible azimuth angles at the receiver. Thus, one choice of radiation pattern for the receiver antenna is a pattern omni-directional in azimuth with a directional elevation response tailored to the distribution of elevation angles. This is the usual radiation pattern adopted for mobile units. However, this type of antenna only responds to one polarization and the receiver antenna is required to be able to probe the channel at all polarizations.

In order to probe all polarizations, the antenna must have equal radiation patterns when excited by vertical and horizontal polarizations and it must have a low cross polar response; any polarization can be made up from a combination of two orthogonal polarizations. To achieve this, the distribution of the E and H-field across the antenna aperture must be symmetrical [1]. In general, this is not the case for an aperture antenna because the boundary conditions for the E and H fields at the edges of the aperture, approximately those of a perfect conductor, are different. The field distributions may be equalized by altering the boundary conditions so that the antenna supports a balanced hybrid mode which approximates a cosine squared distribution for the E and H fields in both vertical and horizontal planes. Figure 3.10 illustrates various antenna structures in which this mode is excited. First is a corrugated horn, second is a conical horn loaded with dielectric material (Ref: Clarricoats P J B, Salema C E R C, 1972), and third is a Potter horn.

Of the three horns illustrated, the conical horn is the most versatile. It has the best cross-polar response and the widest operational bandwidth, but it also appears the most complex to construct. However, if we consider the constraint that the antenna should receive the greatest possible amount of the scattered radiation, the problem of manufacturing the conical horn is greatly simplified. If the flare angle of the horn is made equal to 90°, the antenna becomes a disc and the corrugations become radial grooves. This design gives the antenna a wide beamwidth which allows it to receive most of the scattered radiation.

---

1 The radiation patterns for vertical and horizontal polarisations are equivalent to the E and H plane radiation patterns of an antenna excited with a single polarisation.
Figure 3.10 Antenna Structures With Equal E and H-Plane Radiation Patterns
Figure 3.11 Receiver Antenna Design For Equal E and H-Plane Radiation Patterns

The dimensions of the horn are calculated from James (Ref: James G L, 1977). The slot thickness was $0.05\lambda = 0.275 \text{ mm}$ and the slot depth $0.26\lambda = 1.43 \text{ mm}$. Figure 3.11 shows the final form of the receiver antenna. The antenna has a $120^\circ$ beamwidth with 6 dB gain. The antenna is fed by a short length of circular waveguide which connects to a circular-to-rectangular transition which acts as a
low cross-polar exciter for the antenna. For the polarization measurements the antenna output passes through a polarization switch before passing through the mode transition.

Figure 3.11 shows the resultant radiation pattern. Sidelobe levels and cross-polar levels are less than 30 dB below the main beam. The antenna provides a good match to the distribution of scattered radiation available in the micro-cellular environment. Measurements in this Thesis, Chapter 4, show that the difference in received signal level between the antenna facing forward and backward is greater than 25 dB. This confirms that most of the radiation in the micro-cell is confined to a cone of angles in the direction towards the transmitter and that the receiver antenna collects most of the scattered radiation.

Figure 3.12 Receiver Antenna E-Plane Co-Polar Radiation Pattern
3.4.3 DUAL POLARIZATION TRANSMITTER ANTENNA

To carry out the initial orthogonal polarization correlation experiments, two spatially separated standard gain horns may be used. However, this arrangement is unsatisfactory for three reasons. First, the E and H plane patterns of the two antennae are different which gives rise to antenna pattern diversity. Secondly, the antennae may point in slightly different directions resulting in antenna pointing diversity. Thirdly, the antennae are spatially separate which will result in antenna space diversity. These effects may not be very large. However, they introduce an unwanted uncertainty into the measurements.

To circumvent the difficulties described above, an antenna which could transmit two polarizations simultaneously with equal beamwidths and from the same location in space is required. In addition, the antenna should also have gain approximately equal to that of the standard gain horn antenna currently used so that the power budget calculations for the link are not disturbed. The solution adopted is to use the 90° flare angle scalar horn antenna designed for the receiver with a dielectric lens antenna to increase the gain from 6 dB to ~25 dB.

3.4.3.1 Dielectric Lens Antenna

A perspex plano-convex lens with a polystyrene quarter wavelength matching layer on the flat surface was constructed to increase the directivity of the scalar horn antenna from 6 dB to ~25 dB and thus make it comparable with that of the standard gain horn antenna. The ideal gain, $G$, of a lens of area, $A$, and radius, $r$, assuming 100% efficiency and uniform illumination is:

$$G = \frac{4\pi A}{\lambda^2} = \frac{4\pi^2 r^2}{\lambda^2}$$  \hspace{1cm} (3.1)

Thus, to obtain a gain of 25 dB, the radius should be 1.6 cm. However, there are several factors which reduce the gain, $G$, of a lens antenna: attenuation in the lens medium, reflection from surfaces, spillover loss and non-uniform amplitude distribution in the aperture plane.

i) The loss tangent of perspex is very small (0.0001) and, for the thickness of the lens used here, this loss mechanism is ignored.
ii) Reflection from the surfaces was minimized by using a quarter wavelength matching layer on the flat surface of the lens. The matching layer increases the transfer of power into the lens and although it does not stop reflections from the curved surface of the lens, it does reduce multiple reflections which are a prime source of sidelobes.

iii) Spill-over loss was minimized by making the lens oversized compared to the illuminating antenna beamwidth. This results in a non uniform amplitude distribution across the lens.

iv) The effect of the amplitude taper across the lens reduces the effective aperture of the lens. However, it also has the benefits of reducing spill-over loss and sidelobe levels because it reduces the abrupt change in field strength at the edges of the aperture.

The effective aperture was calculated by assuming that the aperture illumination provided a linear amplitude taper across the lens. This first order approximation would tend to over estimate the gain. The effective area, $A_E$, is given by summing the increments of the antenna area, from $r = 0$ to $r = r_{max}$, the maximum radius, each weighted by the tapering factor. This is illustrated in Equation 3.2. Table 3.1 shows the gain for uniform and tapered illumination of the lens aperture using Equation 3.1 and 3.2 respectively.

$$A_E = \int_0^{r_{max}} 2\pi r dr \frac{(r_{max} - r)}{r_{max}} = \frac{1}{3} \pi r_{max}^2$$

Table 3.1 Lens Antenna Gain Versus Illumination

<table>
<thead>
<tr>
<th>Lens Radius, r</th>
<th>Gain Versus Illumination</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Uniform</td>
</tr>
<tr>
<td>40 mm</td>
<td>33 dB</td>
</tr>
<tr>
<td>30 mm</td>
<td>30 dB</td>
</tr>
<tr>
<td>20 mm</td>
<td>27 dB</td>
</tr>
</tbody>
</table>

The required radius is thus ~30 mm. From the assumption that the aperture illumination provided by the scalar horn antenna is a linear taper, we find that the 3 dB point of the illumination must coincide with a point half way from the centre of the antenna to the edge of the lens. Thus, for $\theta$, the half-angle beamwidth of the
horn equal to 60° from simple trigonometry, the focal length, \( f \), of the lens is 8.66 mm. The lens is designed so that the path length from the source to a plane in front of the lens is equal.

Figure 3.13 shows a section through the lens. The contour of the lens is determined by the hyperbolic curve in Equation 3.3.

\[
y = \sqrt{(n^2 - 1)x^2 + (2n - 1)fx}
\]

where: \( f \) = focal length of lens, \( n \) = refractive index of lens, and the co-ordinates \((x, y)\) are defined in Figure 3.13 (Ref: Jasik H, Johnson R C, 1984).

To simplify the manufacture of the lens, the contour was approximated to a spheroid for the centre of the lens and a cone for the edge of the lens which is also shown on Figure 3.13. This procedure is acceptable provided that the resultant path difference error compared to the true contour is not greater than \( \lambda/10 \). The lens was manufactured from perspex which has a dielectric constant of 2.5 at microwave frequencies. To reduce reflections at the surface of the lens, a \( \lambda/4 \) matching layer of solid polystyrene, dielectric constant 1.6, was used. This dielectric constant is conveniently close to the ideal value for a matching layer of
\sqrt[n]{1.58} \text{. The dielectric lens antenna was mounted on a threaded dielectric cylinder to allow fine adjustments to be made to the lens position. This is illustrated in the schematic diagram of the finished lens in Figure 3.14. The horn-to-lens separation was adjusted to give maximum gain with minimum sidelobe levels.}

---

![Figure 3.14 Schematic Diagram of Lens Antenna](image)

**Figure 3.14 Schematic Diagram of Lens Antenna**
3.4.3.2 Radiation Patterns

Figure 3.15 shows the E and H-plane radiation pattern for the antenna together with their cross-polar patterns. The beamwidth is ~10° in both planes which gives a directivity of ~25 dB. The lens has a higher insertion loss than expected. This was most likely due to reflection losses caused by the large angle of incidence of rays towards the edges of the lens which could be avoided if a narrower beamwidth horn were used to illuminate the lens.

3.4.3.3 Cross Polar Response

The cross polar response of the antenna, as can be seen from Figure 3.13, was about -30 dB down on the main beam. This was found to be adequate during the polarization measurements.

3.4.3.4 Sidelobe Levels

The major sidelobes were -20 dB down on the main beam. It was initially thought that the high sidelobe levels were due to reflections between the lens surface and the horn. However, when an annular disc of radar absorbing material was introduced into the cavity between the lens and the horn, there was no major change in sidelobe level. This suggests that sidelobes were due to multiple reflections within the lens itself.
Figure 3.15 E and H-Plane Radiation Patterns for Dielectric Lens Antenna
3.5 LINK POWER BUDGET

The link power budget is given in Equation 3.5. The fading margin \( F_{\text{MAR}} \) is defined as the excess of the effective radiated power (ERP) less the propagation loss over the link (PL) over the minimum detectable signal of the receiver (MDS). The ERP is defined to include the gain of both transmitter and receiver gains. If the difference of these quantities is positive, the link will operate. If not, the received signal will be submerged in noise and the link will fail.

\[ F_{\text{MAR}} = ERP - PL - MDS, \]  

where: \( ERP = \) effective radiated power (see Equation 3.6), \( PL = \) propagation loss (see Equation 3.7), \( MDS = \) minimum detectable signal (see Equation 3.8), \( F_{\text{MAR}} = \) fading margin.

The terms in Equation 3.5 are defined in Sections 3.5.1 through 3.5.3. The fading margin determines both the maximum range of operation for a given quality of reception and the performance of the system during a severe fade. In terms of channel probing, if the signal falls below the noise floor no statement can be made about it apart from that it is below some specified level. In addition, as the signal approaches the noise floor, the random errors in estimating signal strength increase.

In Section 3.5.4 the link budget for the basic receiver transmitter configuration is calculated at the minimum and maximum expected ranges of 30 m and 300 m respectively. The elements of the calculation are summarized in Table 3.2. The link budgets for the other experimental configurations are different from that of the basic system. Sections 3.6 through 3.9 describe each experimental configuration and indicate where these differences in link budget arise. Table 3.4, in Section 3.10, provides a quick summary of the link parameters for all of the experimental configurations with reference to those of the basic system.

3.5.1 EFFECTIVE RADIATED POWER (ERP)

The effective radiated power, ERP, is usually defined as the power which would have to be fed to the terminals of a hypothetical, lossless, omni-directional antenna to achieve the same radiated power density as that obtained by a directional antenna on its principal lobe. However, in the definition used here, we also include the gain of the receiving antenna.

\[ ERP = P_{Tx} - L_{w_g} + G_{Tx} + G_{Rx} \]  

where: \( P_{Tx} = \) transmitted power, \( L_{w_g} = \) weighted loss, \( G_{Tx} = \) transmitter gain, \( G_{Rx} = \) receiver gain.
where: ERP = effective radiated power (dBm), $P_{T_x}$ = oscillator output power (dBm), $L_{w_g}$ = waveguide loss (dB), $G_{T_x}$ = transmit antenna gain (dB), $G_{R_x}$ = receiver antenna gain (dB).

### 3.5.2 PROPAGATION LOSS (PL)

In order to estimate the propagation loss in a micro-cell we assume that the dominant propagation mechanism is free space spreading loss (see Section 2.8.2), and add on a correction to account for the effects of atmospheric absorption discussed in Section 2.8.3. If we neglect precipitation absorption which is only significant for about 1 hour per year, Section 2.8.3.2, and consider only the effect of oxygen absorption which gives an attenuation of 5.5 dB Km$^{-1}$, Section 2.8.3.1, then the propagation loss in dB is:

$$PL = 20 \log\left(\frac{4\pi r}{\lambda}\right) + 5.5 \frac{r}{1000}$$

where: $PL =$ propagation loss, $r =$ transmitter-receiver separation (m), $\lambda =$ free space wavelength.

### 3.5.3 MINIMUM DETECTABLE SIGNAL (MDS)

The minimum detectable signal (MDS) is defined as the signal which is exactly equal to the noise power of the receiver. Thus, a receiver output, with the MDS as its input, will be 3 dB greater than the noise power. The noise power of a receiver in dB is given by (Ref: Southworth G C, 1950):

$$MDS = N_T + 10 \log(B) + NF_{R_x}$$

where: $N_T = -114$ dBm is the thermal noise in a 1 MHz bandwidth at 300$^\circ$C, $B$ is the pre-detection bandwidth of the receiver in MHz, and $NF_{R_x}$ is the noise figure of the receiver.
3.5.3.1 Receiver Noise Figure (\(NF_{\text{Rx}}\))

For frequencies greater than 1 MHz the 1/f noise of the mixer diode can be neglected (Ref: Southworth G C, 1950) and the receiver noise figure is given by:

\[
NF_{\text{Rx}} = L_C + NF_{\text{IF}} + L_{\text{wg}}
\]

where: \(NF_{\text{Rx}}\) = noise figure of the receiver, \(NF_{\text{IF}}\) = noise figure of the IF amplifier, \(L_C\) = mixer conversion loss in dB, \(L_{\text{wg}}\) = waveguide losses in the receiver.

3.5.4 POWER BUDGET FOR BASIC SYSTEM

In this Section, the power budget for the basic system is calculated at the minimum and maximum expected transmitter-receiver separations of 30 m and 300 m respectively. The calculation is summarized in Table 3.2. The calculation brings together the separate elements of the power budget, presented in Sections 3.5.1 through 3.5.3, and provides a basis for the power budget calculations for the other experimental configurations described in Sections 3.6 through 3.9.

The power budget calculation is summarized in Table 3.2. The effective radiated power of the transmitter (ERP), including both transmitter and receiver antenna gain, is 50 dBm [1], Column 1, Table 3.2. The output power of the Gunn oscillator is taken from the data sheet supplied by the manufacturer and this value was checked using a calibrated spectrum analyser at Hirst Research Centre. Antenna gains are measured in comparison with 25 dB standard gain horns. The propagation loss (PL) of the mobile link is estimated as 97 dB at 30 m and 118.4 dB at 300 m, Column 2, Table 3.2. The minimum detectable signal (MDS) is -93.1 dBm, Column 4, Table 3.2. The noise figure of the receiver is 9.1 dB, Column 3, Table 3.2. The fading margin, at the minimum separation of 30 m and maximum separation of 300 m, is respectively 45.7 dB and 27.4 dB, Column 5, Table 3.2.

---

1 Power relative to 1 mW.
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Table 3.2 Fading Margin Calculation for Basic System

<table>
<thead>
<tr>
<th>Effective Radiated Power ERP dBm</th>
<th>Propagation Loss PL dB @ 30m &amp; 300m</th>
<th>Receiver Noise Figure NF_r dB</th>
<th>Minimum Detectable Signal MDS dBm</th>
<th>Fading Margin F_m dB</th>
</tr>
</thead>
<tbody>
<tr>
<td>P_r</td>
<td>19.5</td>
<td>300m</td>
<td>118.4</td>
<td>L_w</td>
</tr>
<tr>
<td>G_r</td>
<td>25.0</td>
<td>30m</td>
<td>97.0</td>
<td>N_F</td>
</tr>
<tr>
<td>G_{fa}</td>
<td>6.0</td>
<td></td>
<td></td>
<td>L_w_g</td>
</tr>
<tr>
<td>L_{wq}</td>
<td>-0.5</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>ERP</td>
<td>50.0</td>
<td>PL 300m</td>
<td>118.4</td>
<td>NF_r</td>
</tr>
<tr>
<td></td>
<td></td>
<td>PL 30m</td>
<td>97.0</td>
<td></td>
</tr>
</tbody>
</table>

The fading margin was found to be sufficient during the experiments. Fades are of the order of 20 dB at the maximum range which result in a margin of ~10 dB between signal and noise. Nearer to the transmitter the fades are nearly 30 dB because there are fewer reflections to fill in the nulls. However, at this range the received signal is greater and the resulting margin between signal and noise is again ~10 dB.

The predicted fading margin was confirmed to within experimental error of ~2 dB by a measurement of the receiver minimum detectable signal. The error in the measurement arises because the output power of the Gunn oscillator used as the signal source is not accurately known. Only limited access was available to the equipment to measure this quantity directly, and the power level was not necessarily constant.

3.5.4.1 Measurement of Minimum Detectable Signal

The measurement of minimum detectable signal is made by measuring the input power required to double the output of the receiver compared to its noise only output power. This is accomplished in two stages. First, the square law detected receiver noise power is measured. Secondly, the input power required to exactly double the output power of the receiver is measured.

The receiver noise power is measured by comparison with the square law detected output of a calibrated VHF signal generator. The output of the signal generator is adjusted, using a precision attenuator, until the two powers are equal. Equality is tested using a high gain setting on an oscilloscope. The precision attenuator is then adjusted to double the output power of the signal generator. Use of the

1 Thermal noise in a 1 MHz bandwidth at 300°K.
precision attenuator to set the power doubling point eliminates one source of error because the doubling point may then be judged by a simple equality of level test rather than taking an accurate reading from a scope or meter.

The MDS is then measured by feeding a signal into the input of the receiver from the 55 GHz transmitter oscillator. The power level is adjusted, using a 55 GHz precision attenuator, until the receiver output is exactly equal to the output of the signal generator, ie the input power is exactly sufficient to double the receiver output. The minimum detectable signal is then given by the output power of the 55 GHz oscillator, minus waveguide losses, minus the setting on the precision attenuator.

3.6 TWO FREQUENCY MEASUREMENTS

The functional block diagram of the transmitter and receiver is shown in Figure 3.16. The transmit Gunn oscillator is narrow band FM modulated by a frequency synthesizer to produce two phase coherent sidebands at frequencies $f_c + f_M$ and $f_c - f_M$, where $f_c$ is the carrier frequency and $f_M$ the modulation frequency. The receiver separates the two down converted sideband frequencies using separate bandpass filters. The IF frequencies are determined by the side band separation and are limited by the 3 dB IF amplifier bandwidth of 600 MHz. The sidebands are then individually square law detected and passed through logarithmic amplifiers. The signals are recorded together with the car speed information onto magnetic tape.

The minimum frequency separation which the receiver can resolve is determined by the slope of the filter characteristic at the band edge. This was measured at 5 dB per MHz at ~100 MHz. Thus, for 30 dB suppression of the unwanted signal the minimum separation is 12 MHz. In practice, the suppression of the unwanted signal is nearer to 20 dB due to the drift of the transmitted signal and the local oscillator during experiments. Measurements were carried out up to a modulation frequency of 160 MHz, giving a sideband separation of 320 MHz. The results show that 320 MHz separation reduces the correlation between the two signals to a low value.
Transmitter

Varactor Tuned Gunn Oscillator

Standard Gain Horn Antenna

RF Frequency Synthesizer

Gunn Oscillator

Precision Attenuator

Gunn Oscillator

Isolator

IF Amplifier

Scaler Horn Receiver Antenna

Receiver

Spectrum Analyser

Log Amps

Bandpass Filters

Output to Tape Recorder

Square Law Detectors

Figure 3.16 Two Frequency Measurement Block Diagram
3.6.1 FADING MARGIN

The main effect to consider is the effective reduction in the transmitter output power caused by modulating the Gunn oscillator. The power used to calculate the link budget for the two frequency measurements is that contained in the two FM sidebands. This power is at least 4 dB less than the normal, un-modulated, output of the Gunn and the ratio drops rapidly as the modulation frequency increases; at 160 MHz modulation frequency the sidebands contain ∼20 dB less power. In addition, the receiver pre-detection bandwidths, defined by the VHF filters which separate the channels, are different to that used in the basic configuration.

Taking the worst case modulation as an example (320 MHz sideband separation, 160 MHz modulation frequency) the sidebands are 20 dB below the un-modulated carrier power. For a lower IF of 100 MHz and an upper of 420 MHz, using a 15% and a 1% filter respectively, the pre-detection bandwidths for each sideband are 15 MHz and 4.2 MHz. The fading margin for the maximum and minimum separations of 300 m and 30 m, for the line of sight case, are respectively 7.7 dB and 25.7 dB for the lower sideband and 13.2 dB and 31.2 dB for the upper sideband.

3.7 TWO POLARIZATION MEASUREMENTS

Figure 3.17 shows the schematic diagram of the experimental equipment for the polarization measurements. This is similar to that for the two frequency measurements because the two orthogonal polarizations are separated in the receiver by transmitting them with a small frequency separation. The frequency separation ranged from 15 MHz to 20 MHz. This is large enough to be resolved by the receiver, Section 3.6, yet is sufficiently small to ensure that any de-correlation effects result from the difference in polarization and not from frequency separation. The maximum frequency separation is set by the coherence bandwidth of the path which is estimated from two frequency correlation measurements as ∼100 MHz, Section 4.5.1.1.

The transmitter uses two Gunn oscillators to produce the two orthogonal polarizations. The outputs are combined by an ortho-mode transducer which combines them without disturbing the orientation of their respective polarization vectors. The cross polarization resulting from this device, according to the manufacturer, is better than -30 dB. The transmitter antenna is a 120° beamwidth scalar horn antenna with a dielectric lens which has the same directivity as the standard gain
horn. However, its gain is lower by ~6 dB due to inefficiency in illumination and reflection losses. The design and performance of this antenna is given in Section 3.4.

The receiver uses a Faraday rotation polarization switch to sample the vertical and horizontal polarization components of the receiver antenna output. The sampled output consists of four components: the horizontal cross-polar component at $f_1$, the vertical co-polar component at $f_1$, the horizontal co-polar component at $f_2$ and the vertical cross-polar component at $f_2$. These components are separated in frequency using band pass filters and in polarization by rapidly switching between the received vertically and horizontally polarized signals. A switching rate of 2.7 KHz was chosen because this allowed the polarization switch sufficient time to switch between states, producing an approximate square wave whilst also sampling the waveform sufficiently quickly to allow successive samples to be compared directly.

A control experiment was conducted to check whether the use of two separate Gunn oscillators to generate the vertical and horizontal components at the transmitter was, in itself, sufficient to cause the two received signals to become de-correlated. The experimental configuration is identical to that used above, but the outputs of the two Gunn oscillators are combined using a bi-directional coupler in place of an ortho-mode transducer. To carry out the experiment, the dual polarization transmitter antenna is configured to transmit two vertical polarization signals from different Gunn oscillators identified by a small frequency separation of approximately 20 MHz. The measurement was carried out along Torrington Place, Location 4 in Figure 4.1. The measured time series showed an apparently high correlation assessed visually on a dual trace oscilloscope and analysis of five 1.6 second samples had correlation coefficients which ranged between 0.77 and 0.91. This is comparable with the range of correlation coefficients measured at 25 MHz separation during the two frequency measurements. The spread of correlation coefficient values at 25 MHz separation is illustrated in Figure 4.27, Section 4.5.1.1. Thus, the origin of the de-correlation in the use of orthogonal polarizations is confirmed.
Chapter 3

Figure 3.17 Two Polarization Measurement Block Diagram
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3.7.1 FADING MARGIN

The fading margin of the system is reduced relative to the basic configuration because the polarization switch introduces an extra insertion loss of 0.5 dB, increasing the receiver noise figure to 9.6 dB, and the transmitter gain is reduced by 6 dB, reducing the effective radiated power to 44 dB. The resulting fading margin at the maximum and minimum separation of 300 m and 30 m is then respectively 21.2 dB and 39.2 dB.

3.8 WIDE AREA PROBING

The schematic diagram for the wide area probing measurements is given in Figure 3.18. The transmitter uses a single Gunn oscillator and the down-converted received signal is fed directly into a spectrum analyser which acts as a narrow bandwidth, 30 KHz, receiver. The output of the spectrum analyser is a spike with amplitude representing signal power and time delay, relative to the sweep origin, representing frequency. It would not be possible to use such a narrow band filter conventionally due to the instabilities in both the transmitter and the local oscillator frequencies without fairly sophisticated signal processing. However, since the spectrum analyser sweeps over all frequencies in the region of the IF frequency, the frequency drift due to instability is converted into a slight delay in the output of the spectrum analyser. The narrow bandwidth considerably reduces the minimum detectable signal which allows field strength probing to be carried out beyond the boundaries of the micro-cell. Thus, the power leaking into streets perpendicular and parallel to the micro-cell can be investigated.

The output of the spectrum analyser is recorded on magnetic tape together with the synchronization pulse of the spectrum analyser and the car speed information. In addition, written comments are made noting the reading of the tape counter in relation to road junctions and other landmarks to allow experimental runs along different roads to be identified.

When the tape is replayed, the envelope of the recorded signal is reconstituted by peak detecting and filtering. The envelope detecting circuit is reproduced from a circuit given in Horowitz and Hill (Ref: Horowitz P, Hill W, 1988, Figure 3.38, pp 119) which is modified by incorporating an FET switch to short out the integration capacitor. The switch circuit also being reproduced from Horowitz and Hill (Ref: Horowitz P, Hill W, 1988, Figure 6.33, pp 242).
The end points of each section of road are determined from the written comments made when the data is recorded and the number of speed pulses between the end points is counted. This information is used to sample the data from each section of road at constant distance intervals. Car speed pulses occur every 12.5 cm and results are averaged over eight readings to give a reading of average power every metre. However, because the speed of the car is varying, samples occur at irregular intervals of time which produces "spiky" data. A moving average routine is used to filter the data. This calculates a new value for each data point as a weighted summation of the values of the points adjacent to it. It is an approximation to the convolution of the data with a filter function and thus it approximates the operation of a low pass filter.

The moving average routine used to sample the data used 11 points, 5 before the data point and 5 after it. The distance between each point is 1 m. The weighting function used is Gaussian with a 3 dB half width of approximately 1.5 points. This is equivalent to a length of 1.5 m. Thus, the 3 dB point in the spatial frequency domain is $2/3 \text{ m}^{-1}$.

Calibration is straightforward because the linearity and wide dynamic range of the spectrum analyser allows a simple conversion factor to be applied to its output. This factor was experimentally determined to be 200 dB V$^{-1}$. The received signal is calibrated directly in path loss in dB, thus making the results independent of transmitted signal, by subtracting 102.1 dB from the measured spectrum analyser reading. This takes account of the +10 dB scale used on the spectrum analyser, the +67 dB gain of the IF amplifier, the 4.9 dB conversion loss of the RF mixer, the 19.5 dBm power of the transmitter, the combined 1 dB wave guide loss of the transmitter and receiver, and the combined receiver and transmitter antenna gains. Certain sections of the data recorded near to the transmitter had to be attenuated to prevent the high signal levels overloading the spectrum analyser input. However, it is a simple matter to readjust the recorded signal levels when the data is calibrated.
Figure 3.18 Wide Area Probing Measurement Block Diagram
3.8.1 FADING MARGIN

The reduction in pre-detection bandwidth from 15 MHz to 30 KHz reduces the minimum detectable signal, MDS, by 27.7 dB to -120.1 dBm. The fading margin, at the maximum and minimum separation of 300 m and 30 m is respectively 54.7 dB and 72.7 dB.

3.9 SWEPT FREQUENCY MEASUREMENTS

The schematic diagram for the swept frequency measurements is shown in Figure 3.20. The configuration is very similar to that for the wide area probing measurements. However, the transmitter uses a varactor tuned Gunn oscillator which is swept by a voltage ramp. The amplitude of the voltage ramp is 2 V which results in a sweep of 66 MHz. The sweep repetition rate is 10 KHz.

The down-converted received input signal is fed directly to the spectrum analyser. This is used as a very narrow bandwidth receiver. The spectrum analyser scans over a bandwidth of 200 MHz every 50 ms. Each time the spectrum analyser scan coincides with the transmitter chirp it produces a voltage spike. The envelope of these spikes is detected by a peak detector followed by a low pass filter. This envelope is recorded together with the scan synchronization pulse and the car speed information on magnetic tape. This results in a record of successive snapshots of the mobile radio transfer function $T(f,t)$, Section 2.3.

The envelope detector is a modified version of that used for the wide area probing measurements, Section 3.8. However, to allow the envelope detector to follow the envelope of the pulses from the swept frequency measurements which occur at a far higher rate than for the wide area probing measurements due to the 10 KHz sweep rate, the rate of discharge of the charge storage capacitor in the envelope detector was increased by connecting a 1 MΩ resistor across it. The value of this resistor was adjusted according to the sweep rate of the spectrum analyser. The output of the envelope detector was then smoothed using a simple RC network.
Figure 3.19 Swept Frequency Measurement Block Diagram
3.9.1 FADING MARGIN

The fading margin for the swept frequency measurements is the same as that for the wide area probing measurements; 54.7 dB and 72.7 dB at 300 m and 30 m respectively. The effect on the power budget caused by sweeping the transmitter oscillator is very small, less than 1 dB, and is ignored.

3.10 SUMMARY

This Section briefly summarizes the requirements and performances of the experimental configurations described above; the configurations are summarized in Table 3.3 and the performances, at 30 m and 300 m receiver transmitter separations, are summarized in Table 3.4.

<table>
<thead>
<tr>
<th>Measurement</th>
<th>Transmitter Requirements</th>
<th>Receiver Requirements</th>
<th>Fig Ref</th>
</tr>
</thead>
<tbody>
<tr>
<td>Coverage/fading margin</td>
<td>Gunn device, (SGH) antenna</td>
<td>IF Amp, 1 BPF</td>
<td>3.1</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Square law detector, logamp</td>
<td></td>
</tr>
<tr>
<td>Two frequency coherence</td>
<td>VHF modulator, varactor tuned Gunn, SGH antenna</td>
<td>IF Amp, 2 BPF</td>
<td>3.14</td>
</tr>
<tr>
<td></td>
<td></td>
<td>2 square law detectors, 2 logamps</td>
<td></td>
</tr>
<tr>
<td>Polarization diversity</td>
<td>Two Gunns, ortho-mode transducer, lens antenna</td>
<td>IF Amp, polarization switch, 2 BPF</td>
<td>3.15</td>
</tr>
<tr>
<td></td>
<td></td>
<td>2 square law detectors, 2 logamps</td>
<td></td>
</tr>
<tr>
<td>Swept frequency</td>
<td>Single varactor tuned Gunn, voltage ramp generator, SGH antenna</td>
<td>IF Amp, spectrum analyser</td>
<td>3.16</td>
</tr>
<tr>
<td>Wide area probing</td>
<td>Single Gunn, SGH antenna</td>
<td>IF Amp, spectrum analyser</td>
<td>3.17</td>
</tr>
</tbody>
</table>

Table 3.3 Analogue Experimental Requirements
Table 3.4 Analogue Experimental Equipment Performance

Table 3.4 demonstrates that the various experimental configurations are suited for different purposes. The experiments which use a spectrum analyser as the detector, the wide area probing and swept frequency measurements, have a very large fading margin. However, they have a very slow response time illustrated in the maximum rate of change column in Table 3.4. The experiments which use a VHF filter plus square law detector have faster response times, but their fading margins are lower and, in the case of the two frequency measurements, frequency modulation of the transmitter oscillator reduce the fading margin still further.

The performance of the experimental equipment is by no means optimum. However, significant improvement in performance could only be effected by providing some means of phase locked frequency stability on the transmitter and receiver Gunn oscillators. If this were available, the fading margins realized using the spectrum analyser could be achieved together with a fast response time.
Chapter 4  ANALOGUE RESULTS

4.1  INTRODUCTION

This Chapter presents the results of the analogue measurements. The theoretical structure on which the analysis is based is presented in Chapter 2 and the experimental equipment used for the measurements is described in Chapter 3.

An overview of the different environments in which the propagation measurements were undertaken is presented in Section 4.1.1. The locations in which the urban propagation measurements are taken are illustrated in Figures 4.1a and 4.1b which are maps of the local area surrounding the College in Torrington Place and that surrounding the transmitter location in High Holborn, reproduced from the A-Z of London. The details of the local geography of the urban propagation area are illustrated in Figures 4.2 - 4.5 which are photographs showing the approximate view from each of the transmitter sites.

An introduction to the observed time series behaviour of the received signal envelope is presented in Section 4.1.2. Comparison is made with time series recorded in urban and in open rural areas, and with time series predicted by a simple deterministic model. Typical examples of time series are presented which illustrate the "fast fading" and "slow fading" processes described in Section 2.4.2 and the variation in character of the received signal with the mobile vehicle velocity. Together these observations confirm the statements made in the opening paragraphs of Chapter 2, about the rapid and extreme variation of the signal strength over small distances and the slower variation in the mean value over larger distances which led to postulation of a product type non-stationary model for the mobile radio channel in Section 2.4.3.

The mobile radio experiments generate a vast amount of data, more than can be processed with the available computing facilities in a reasonable amount of time. Section 4.1.4 describes the data selection process by which characteristic portions of recorded time series are chosen for sampling and analysis.
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Following the empirical justification of the product type non-stationary model of the mobile radio channel in Section 4.1.2, the Chapter is laid out under four main headings which follow the conceptual approach summarized in Section 2.10, Figure 2.21. These are: First Order Statistics, Section 4.2; Second Order Statistics, Section 4.3; Joint First Order Statistics, Section 4.4; and Joint Second Order Statistics, Section 4.5. Section 4.6 summarizes the results. The results of some of these experiments have been presented in published papers (Ref: Thomas H J et al, 1986; Siqueira G L et al, 1987; Siqueira et al, 1988; Cole R S et al 1988).

4.1.1 PROPAGATION ENVIRONMENT

The measurements presented in this Thesis are obtained mostly in urban environments. However, measurements were also taken in an "open-country" area at the University College London sports ground at Shenley, in Hertfordshire. This environment provides a comparison with the measurements made in the urban environment so that the effect of multiple reflections from buildings and traffic in the urban environment may be assessed. Only the road surface and a few trees, set back 10 m from the sides of the drive, act as scatterers. The transmitter was tripod mounted at 2.6 m elevation and radiated from one end of the drive of the sports ground.

The urban environment consists of the streets in the vicinity of the Department of Electrical and Electronic Engineering in Malet Street, WC1. The streets range in width from 10 - 30 m and are lined by buildings with average heights of four to five storeys with a few extending to 20 or more. Some streets are also lined with trees. Figures 4.1a and 4.1b illustrate the region in the vicinity of the College where measurements were obtained and indicate the location of the various transmitter sites. Table 4.1 summarizes the different transmitter locations, the area which they illuminate, and the photographs which show the approximate view from the transmitter.
<table>
<thead>
<tr>
<th>Transmitter Location</th>
<th>Height</th>
<th>Direction</th>
<th>Fig No</th>
</tr>
</thead>
<tbody>
<tr>
<td>1) 10th floor of Engineering Building, Malet Place</td>
<td>40 m</td>
<td>Byng Place</td>
<td>4.2</td>
</tr>
<tr>
<td>2) 2nd floor of Anthropology Building, Malet Place</td>
<td>12 m</td>
<td>Malet Street</td>
<td>4.3</td>
</tr>
<tr>
<td>Street level outside Anthropology Building, Malet Place</td>
<td>2.6 m</td>
<td></td>
<td>4.4</td>
</tr>
<tr>
<td>3) Ground floor of Quaker International Institute, Byng Place, and on pavement outside it</td>
<td>2.6 m</td>
<td>Byng Place</td>
<td>4.2</td>
</tr>
<tr>
<td>4) 4th floor of Warburg Institute Library, Torrington Square</td>
<td>25 m</td>
<td>Torrington Place</td>
<td>4.4</td>
</tr>
<tr>
<td>5) 4th floor office of Convention Travel, High Holborn</td>
<td>20 m</td>
<td>High Holborn</td>
<td>4.5</td>
</tr>
</tbody>
</table>

Table 4.1 Map of Urban Transmitter Locations (A-Z of London)
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Figure 4.1a Map of Urban Propagation
Area Adjacent to UCL

Figure 4.1b Map of Urban Propagation
Area Along High Holborn
Figure 4.2 View of Byng Place from the 10th Floor of the Engineering Building

Figure 4.3 View of Malet Street from the 10th Floor of the Engineering Building
Figure 4.4 View of Torrington Place From the 4th Floor of the Warburg Institute

Figure 4.5 View of High Holborn from the 4th Floor Offices of Convention Travel
Figure 4.2 is a photograph taken from the 10th floor of the Engineering Building in Malet Place, Location 1 in Figure 4.1. It shows the merging of Torrington Place, in the bottom right hand of the photograph, into Byng Place, Gordon Square, Tavistock Square and Torrington Place. In the centre foreground is the rear view of the Quaker International Centre which is transmitter Location 3 in Figure 4.1. In the centre right is the Warburg Institute, the top floor of which is the Courtauld Gallery; this is transmitter Location 4 in Figure 4.1.

Figure 4.3 is a photograph taken from the 10th floor of the Engineering Building in Malet Place along Malet Street. This is similar to the view from Location 2, the Anthropology Building in Malet Place, but is from a higher vantage point.

Figure 4.4 is a photograph showing the view from the 4th floor of the Warburg Institute, Location 4 in Figure 4.1, along Torrington Place. In the foreground of the picture is the junction of Torrington Place with Malet Street outside the Engineering Building. It is apparent from the Figure that quite a large number of vehicles with heights two to three times that of a normal car use the street. In particular, illustrated in the picture, are delivery vans, lorries, single and double decker busses. The picture also illustrates the concentration of traffic at road junctions.

Figure 4.5 is a photograph showing the view from the 4th floor offices of Convention Travel in High Holborn, transmitter Location 5 in Figure 4.1. It shows the view along High Holborn towards Holborn Circus and the City of London. The buildings lining the street here are more than nine storeys high and there are many high sided vehicles in the general traffic in the street.

### 4.1.2 TIME SERIES AND VELOCITY PROFILES

This Section presents an overview of the observed character of the received signal before the detailed results of time series are presented in Sections 4.2 through 4.5. Qualitative comparison of the signal envelope measured in an urban environment is made with that measured in a rural area and with predictions of a simple deterministic model. In addition, typical samples of the time series observed when two frequencies, and two polarizations, are transmitted simultaneously are presented. To illustrate the small scale fluctuations of the signal in a clear manner, portions of the two frequency measurements, plotted on a UV chart recorder with a frequency response extending to 1000 Hz, are shown on an expanded time scale.
4.1.2.1 Comparison of a Velocity Profile and a Time Series

Figure 4.6a is a 100 second portion of an experimental run along Torrington Place, Location 4 in Figure 4.1. Figure 4.6b, below, is the recorded variation of the car speed during the run. The speed of the mobile vehicle in an inner-city area is dictated by the local traffic conditions as is clearly illustrated in Figure 4.6b. Comparison of the curves in Figure 4.6a shows the effect of the car velocity on the character of the received signal. Sections A and C of the time series show large fluctuations in the intensity of the received signal envelope, whilst the fluctuations of the signal envelope in Section B are far smaller. The results support the assumption, outlined in the opening paragraphs of Section 2.1, that the "fast fading" of the signal arises from the motion of the vehicle through a spatial interference pattern. The envelope at B is not constant because movement of other scatterers in the environment modifies the received signal strength even when the vehicle is stationary.

4.1.2.2 Comparison of Urban and Rural Time Series

Figure 4.7a is a 20 second portion of an experimental run along Malet Street, Location 2 in Figure 4.1. Figure 4.7b is a 20 second portion of an experimental run conducted in the open rural environment at Shenley in Hertfordshire. Comparing the traces in Figure 4.7a the effect of increased multipath propagation in an urban environment can clearly be seen. Figure 4.7b, taken in the "open country" environment at Shenley, resembles a simple interference pattern. This might be expected since the road surface is the only significant scatterer. Figure 4.7a, taken in the built-up urban environment of Malet Street is very different, the signal being far more random in character.
Figure 4.6a Experimental Time Series in an Urban Environment

Figure 4.6b Velocity Profile for the Experimental Time Series
Figure 4.7a Experimental Time Series in an Urban Environment (Malet Street)

Figure 4.7b Experimental Time Series in an Open Rural Area (Shenley Sports Ground)
4.1.2.3 The Deterministic Time Series

Figure 4.8 is the deterministically calculated variation of received signal power with transmitter-receiver separation. This was calculated in Section 2.7 using a simple model which assumes a received signal made up from four components: a direct line of sight component; a specular component reflected from the road surface; and two specular components reflected from the buildings lining each side of the street. The equivalent time duration of the trace for an assumed constant car velocity of 12 mph is 56 seconds. This is a typical velocity for the mobile vehicle in an urban area. When comparison is made with experimental time series, it must be noted that the speed of the mobile unit is not constant during the experimental runs. This has the effect of compressing or stretching the time duration between successive nodes in the experimental time series. Also, when the mobile unit is stationary, or moving very slowly, the received signal strength fluctuates due to the motion of other scatterers.

Figure 4.8 Calculated Variation of Received Signal Versus Distance

The comparison between the deterministic time series in Figure 4.8 and the other time series illustrated in Figures 4.6a and 4.6b, and 4.7a and 4.7b, shows that the deterministic model predicts the character of some of the observed variations in the received signal strength quite well. In particular, the behaviour of the signal envelope near to the transmitter in both curves of Figure 4.8 is particularly well
modelled. This is because, near to the transmitter, the received signal power is determined by the interference of the line of sight signal with strong reflections, predominantly from the road surface, which make the simple deterministic model particularly well suited to modelling the signal power. Further from the transmitter the number of scattered reflections increases, the line of sight component becomes less dominant, and the deterministic model does not predict the character of the signal fluctuations so well. The character of the received power variation predicted by the deterministic model, in terms of the depth and duration of fades, is intermediate in character between the experimental time series measured in the rural and urban environments. This is to be expected from the increasing number of scatterers which influence the shape of each curve. In the rural environment only a single scatterer is present. For the deterministic model, reflections from the three principal scatterers are considered, whilst in the urban environment there are potentially a vast number of scatterers. The details of each individual time series depend on the disposition of the scatterers in the micro-cell at the time of the measurement and are not predicted by the deterministic model.

4.1.2.4 25 MHz and 320 MHz Separation Dual Time Series

Figure 4.9a shows an 8 second portion of a 25 MHz separation dual frequency run along Torrington Place, Location 4 in Figure 4.1. The time series for the two channels are displaced vertically from each other to avoid confusion between the traces. In addition, there is a small horizontal displacement between the two traces caused by the separation of the pens on the chart recorder. Figure 4.9b shows a small portion of the two time series traces on an expanded time scale. The trace was obtained using a UV chart recorder which can track variations in the signal envelope which occur at frequencies up to 1000 Hz, 50 times higher than that obtainable from the chart recorder which was used to produce the traces in Figure 4.9a. The traces are displaced vertically from each other to avoid confusion, but there is no horizontal displacement between traces which arises in Figure 4.9a due to mechanical separation of chart recorder pens.

Comparison of the traces in Figure 4.9a shows a high correlation with nearly all the peaks and troughs of the traces lining up with each other after taking the small mechanical displacement between the chart recorder pens into account. The high correlation between the traces is even more marked in Figure 4.9b where the expanded time axis and absence of horizontal displacement between the traces
makes visual comparison easier. The high correlation between the two traces suggests that the correlation bandwidth of the channel must exceed 25 MHz during this experiment.

Figure 4.9a Dual Frequency 25 MHz Separation Time Series Along Torrington Place
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In addition, if the fast fluctuations in the two time series shown in Figure 4.9a are ignored and an imaginary line is drawn through the mid-points of the fades, the reader can estimate the variation in the local average signal strength in each trace. When this is done, it can be seen that the apparent variation of the mean signal strength is similar in both traces.

![Relative Power dB](image)

**Figure 4.9b UV Trace of Dual Frequency Time Series (25 MHz Separation)**

Figure 4.10a shows an 8 second portion of a typical 320 MHz separation dual frequency run along Torrington Place, Location 4 in Figure 4.1. Figure 4.10b shows a portion of the same time series, obtained using a UV chart recorder, on an expanded time scale. It must be noted, as in the case of the 25 MHz separation time series in Figures 4.9a and 4.9b, that there is a small horizontal displacement between the two traces in Figure 4.10a due to mechanical separation of the chart recorder pens, whilst in Figure 4.10b the two traces are exactly aligned. Comparison of the traces in both Figures shows a low correlation. It is difficult to identify fades which are common to both channels and sometimes fades in one channel are transposed with peaks in the other. In the portion of the time series on the expanded time scale, Figure 4.10b, the two time series appear to be completely uncorrelated, or even there may be some degree of negative correlation between them. In addition, if the variation of the local mean signal strength in both traces of Figure 4.10a is estimated by visualizing a line through the mid-point of all the fades, it can be seen that whilst there is some similarity in the variation of the
mean signal strength in each trace, there are also some marked differences between the variations. In particular, in the mid-point of the Figure, the average signal level of Channel 1 is high, and that of Channel 2 is low.

Figure 4.10a Dual Frequency Time Series (320 MHz Separation)
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Examination of the typical time series segments illustrated in Figures 4.10a and 4.10b shows that the correlation bandwidth of the mobile radio channel is below 320 MHz during the experiment. Comparison of the results of the two frequency measurements at 25 MHz and 320 MHz frequency separations indicate that from these typical measurements the coherence bandwidth of the channel lies somewhere between 25 MHz and 320 MHz. Full analysis of the two frequency experiments is presented in Section 4.5.1 where the average coherence bandwidth of the channel is shown to be approximately 100 MHz. However, different locations in the micro-cell exhibit a range of coherence bandwidths from 17 MHz up to 150 MHz.

4.1.2.5 Dual Polarization Time Series

Figure 4.11 shows a typical 2.5 second portion of one of the dual polarization experimental runs along Torrington Place, Location 4 in Figure 4.1. The frequency separation between the two traces is 20 MHz. The trace is produced on a chart recorder with a maximum frequency response of 2 Hz. However, the speed of the tape recorder is reduced from 15 ips for recording to 15/16 ips for playback. Thus, the trace has an equivalent frequency response of 32 Hz. The mechanical separation of the chart recorder pens gives rise to a horizontal displacement of the two
traces equivalent to 0.025 seconds on the time axis. The two traces are displaced vertically from each other by 10 dB to avoid confusion between them. Comparison of the traces shows that there is a low correlation between them. As in the case of the two frequency measurement with a frequency separation of 320 MHz, it is difficult to identify fades which are common to both channels, and sometimes fades in one channel are transposed with peaks in the other. The transposition of peaks and fades between the traces during parts of the experiment indicates that the traces may have a negative correlation. From an examination of the mean signal level of each trace, produced by visualizing a line through the mid-point of the fades, it is apparent that whilst there is some similarity between mean signal levels there are also differences between them.

Comparison of the dual polarization time series with the 25 MHz separation time series in Figure 4.10 suggests that the de-correlation observed is due solely to the difference in polarization between the two channels and not to the 20 MHz frequency separation between the two channels. This is confirmed by the analysis of the two frequency experimental results in Section 4.5.1. Thus, examination of Figure 4.11 suggests that a millimetre wave mobile radio system could use orthogonal polarizations as the basis for a diversity reception system. Further
evidence supporting this suggestion is presented in Section 4.4 where a two branch diversity system is simulated using experimental data and in Section 4.5.4 where full analysis of the dual polarization experiments is presented.

4.1.3 SUMMARY OF OBSERVED TIME SERIES BEHAVIOUR

The measured results show that the character of the received signal is dependent on the speed of the mobile vehicle. The received signal is more varied at higher car speeds. However, the variation of the received signal when the mobile is stationary indicates that moving scatterers have significant effect on propagation.

The influence of the multiplicity of the scatterers in the urban environment is illustrated by comparison of a time series recorded in a rural environment with a time series recorded in an urban environment. A simple interference pattern is observed in the rural environment, whilst the time series in the urban environment is far more random in character. Comparison of a simple model based on interference between a line of sight component and reflections from the principal scatterers with experimental data indicates that the observed variations of the received signal strength in the urban environment at small transmitter-receiver separations are due to interference of a very small number of scattered waves. At larger transmitter-receiver separations, in the urban environment, the model is not accurate, indicating that there is a greater influence from randomly scattered components.

Examination of time series recorded with frequency separations of 25 MHz and 320 MHz shows that the properties of the channel are highly correlated at 25 MHz and apparently de-correlated at 320 MHz separation. This suggests that the correlation bandwidth of the channel is in the range of 25 - 320 MHz. Comparison of the apparent change in the mean signal level in the traces separated by 320 MHz indicates that the variation in the mean signal level is also partially de-correlated.

Examination of time series recorded from signals transmitted simultaneously on orthogonal polarizations shows that the received signals are essentially de-correlated or even that a small degree of negative correlation exists between the time series. In addition, the apparent variation in the mean signal level of the two traces is partially de-correlated.

A simple visual examination of the time series presented in Figures 4.10 and 4.11 shows that both frequency diversity and polarization diversity have a potential for use in diversity systems at 55 GHz.
4.1.4 DATA SELECTION PROCEDURE

It is in the nature of the propagation measurements conducted here, that a very large quantity of data is collected during each measurement. To analyse this data it is necessary to carry out data selection procedures.

To illustrate the quantity of data available for analysis, the amount of data in a set of two frequency runs is calculated.

The duration of a typical run is from 60 ~ 240 seconds and represents a distance travelled of between 200 m and 500 m depending on car speed and experiment location. Thus, a typical two channel measurement sampled at 10 KHz will generate between \(1.2 \times 10^6\) and \(4.8 \times 10^6\) two byte integers representing respectively 2.4 Mb and 9.6 Mb of data. After calibration, the data is stored as 4 byte real numbers. Thus, the data storage requirement is increased to between 4.8 and 19.2 Mb. In addition, each measurement is repeated between four and six times. Thus, a single set of measurements would potentially generate between 28.8 and 114.2 Mb of data. This amount of data must be contrasted with the hard disk storage area of 20 Mb of the IBM PC-AT used for data processing and the floppy disc size of 1.2 Mb per disc. In addition, the IBM PC-AT can only sample 32 K two byte numbers at a time. Thus, to sample \(1.2 \times 10^6\) and \(4.8 \times 10^6\) two byte integers, the number of data points generated by each run would require respectively 37 and 146 separate sampling operations. Also, when the data is processed to calculate the local average variation and the normalized data, the storage required is tripled, so requiring a storage capacity of between 86.4 and 345.6 Mb. In addition, there are approximately twenty different experiments to consider for different frequency separations and different polarizations.

The first stage in the data selection process is to observe the recorded signals on a dual trace oscilloscope and to produce a paper chart recorder trace of each experiment. The level of correlation between the recorded signals may then be assessed visually. The oscilloscope traces follow all signal variations up to the maximum fading frequency, whilst the chart recorder shows the variation of the average signal [1]. Portions of data where the IF frequency has drifted out of the pass band of the receiver or where the tape recorder of the receiver has failed to latch into record mode can then be disregarded.

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1 If the tape is slowed down, the effective frequency response of the chart recorder can be increased to 32 Hz which records most of the fast signal variations as well.
Sections of the run where the signal levels are not too low which would lead to noise problems and where the mobile unit is moving with approximately constant velocity are then selected. Initially, three of these data sections from each run which are identified as suitable for analysis are sampled and processed. If more data is required for analysis to back up a result, then more of these sections are sampled for data processing.

4.2 FIRST ORDER STATISTICS

The character of the variation of the received signal envelope in the urban environment is different over different scale sizes. Thus, to facilitate the description of the experimental results, this Section is divided under three main headings: Section 4.2.1 describes the average signal variation which occurs over large distance scale sizes; Section 4.2.2 describes the fine structure of the average signal variation which occurs over medium distance scale sizes; and Section 4.2.3 describes the variation of the signal envelope which occurs over small distance scale sizes.

4.2.1 SIGNAL VARIATION OVER LARGE SCALE SIZES

The analysis is based on the assumption of a micro-cell centred on Malet Street with the base station sited at Location 2 in Figure 4.1a. The transmitter antenna location and bearing is fixed for all of the measurements presented here. The results are presented as a plot of the propagation loss versus distance along Malet Street and as an approximate propagation-loss contour map along the roads surrounding UCL. The measurements provide an estimate of the signal versus distance power law, illustrate the coverage which a single base station would provide, and indicate the potential for interference between adjacent micro-cells.

The measurements are carried out using the wide area probing experimental configuration described in Section 3.8. This arrangement has the greatest receiver sensitivity and can faithfully record the variation of the average signal strength over the whole of the micro-cell and into the adjacent streets. The slow response time, ~1 Hz, of the equipment is immaterial for the measurement of the general trend of the received signal with distance.
The micro-cell used for these measurements extends the whole 320 m length of Malet Street. Figure 4.3 illustrates the approximate field of view of the transmitter which, from the Anthropology Building, has an almost unimpeded view down the street, but for a row of trees and, for the case discussed here, a number of delivery vans near to the transmitter. Measurements were carried out along Malet Street and along the following adjoining streets: Malet Place; Torrington Place; Montague Place; and Gower Street.

It is important to note that the receiver antenna is not omni-directional and had to be rotated for each of the different runs. However, its orientation was always either directly along the direction of motion or at 180° to it.

### 4.2.1.1 Malet Street, Malet Place and Montague Place

Figure 4.12 illustrates the variation of propagation loss between two hypothetical isotropic antenna along Malet Street, Malet Place and Montague Place. The scale on the right of the Figure is received power in dBm. Malet Street forms the main part of the micro-cell. The data for this Figure was gathered from two runs which are typical of the results obtained; the first starting 80 m away from the transmitter and driving along Malet Street towards Montague Place; and the second starting from Montague Place and driving back along Malet Street and into Malet Place.

The minimum propagation loss is 100 dB which occurs at a transmitter separation of 30 m. This is close to the predicted value of 97 dB from free space propagation. The difference between the two values may be explained by the displacement of the mobile unit from the main lobe of the transmitter antenna at small separations. This gives rise to an apparent reduction in the transmitter antenna gain. The propagation loss drops relatively slowly along the 350 m length of Malet Street. The variation in propagation loss between the junction of Malet Street and Torrington Place, near to the transmitter, and the junction with Montague Place, at the end of the micro-cell, is ~40 dB which is larger than the 21 dB expected due to $r^{-2}$ propagation loss. The relationship between signal strength and distance is discussed in Section 4.2.1.4. The presence of the lorries in Malet Street appears to create a small disturbance to the fall off 150 m along the street which is approximately 250 m from the origin of the distance axis of Figure 4.12 in Malet Place. However, this disturbance does not have a major effect on the propagation loss along the rest of the street. Thus, the results would be similar if the lorries were not there. The propagation loss rises very quickly as the mobile unit leaves...
Malet Street at right angles to the LOS path. The propagation loss rises 30 dB in only 30 m when the mobile vehicle turns into Montague Place and it rises 40 dB in 80 m when it moves from Malet Street into Malet Place, i.e., behind the transmitter.

![Figure 4.12 Received Power (dBm) Along Malet Street (The Micro-Cell)](image)

4.2.1.2 Torrington Place

Torrington Place runs at right angles to the main body of the micro-cell along Malet Street, along which the transmitter points, at the site of the transmitter. There are a number of large buildings at the junction of the two roads which scatter power into Torrington Place at right angles to the direct path along Malet Street. The increase in propagation loss along Torrington Place is more severe than that along Malet Street. However, the propagation loss is less than 130 dB for approximately 150 m of its length which compares favourably with the propagation loss along Malet Street illustrated in Figure 4.12. Thus, a micro-cell would be able to extend quite some distance along this street.
4.2.1.3 Gower Street

Gower Street runs parallel to Malet Street and is separated from it by a block of five storey buildings. The two streets are connected in three places: by Torrington Place and Montague Place which cross it at each end; and by Keppel Street which crosses it midway along its length. In addition, there is an enclosed passageway which runs through the School of Hygiene and Tropical Medicine building joining the two streets. The propagation loss increases very abruptly by 30 dB as the mobile vehicle moves into Gower Street from Torrington Place. The propagation loss remains below -170 dB [1] until the vehicle passes the enclosed passageway through the School of Hygiene and Tropical Medicine. Here the signal level rises and the propagation loss falls to 146 dB. Continuing along Gower Street, the signal level rises again as the vehicle passes Keppel Street, and the propagation loss falls to 156 dB. The measurement shows that very little signal is transferred between the two streets. Gower Street could not be used as part of the same micro-cell as Malet Street using the same base station. However, the signal which leaks through from Malet Street might possibly cause co-channel interference. The severity of any possible co-channel interference would depend upon the particular system design implemented.

4.2.1.4 Average Signal Versus Distance Power Law

The relationship between the average signal strength and the distance between transmitter and receiver is an important parameter for estimating the operating range of a communication system. This relationship is usually described in terms of a power law:

\[ P(r) = Kr^{-n} \]  

where: \( r \) = receiver transmitter separation, \( P(r) \) = received power in dBm, \( k \) = constant of proportionality, \( n \) = power law.

---

1 The maximum effective path loss which the system can measure is -170 dB. This figure is arrived at by subtracting the antenna gains (25 dB and 6 dB) and transmitter power (19.5 dB less 0.5 dB waveguide loss) from the minimum detectable signal which is -120 dBm for this experimental configuration.
The power law is determined by measuring the slope of the curve when $P(r)$, in dBm, is plotted against $10 \log(r)$. Figure 4.13 illustrates the propagation loss variation along Malet Place and Malet Street plotted against $10 \log(r)$ where $r$ is the transmitter-receiver separation. The maximum signal level occurs at a transmitter-receiver distance of 30 m. As the mobile unit moves towards Malet Place the signal strength falls while the transmitter-receiver distance decreases. This is because the mobile unit is moving out of the main beam of the transmitter antenna. At the minimum transmitter-receiver separation, the transmitter and receiver are in line and displaced 16 m to the side of each other and by 12 m in height. As the mobile moves into Malet Place the transmitter-receiver separation rises again and the signal power falls.

**Figure 4.13 Log-Log Plot of Propagation Loss Along Malet Street**
(Right hand scale shows power in dBm)
From Figure 4.13, the power law is approximately $r^{-3.6}$ for propagation along the main body of the micro-cell. This is higher than $r^{-2}$ law predicted by free space propagation. It is also higher than the $r^{-2.3}$ reported by McGeehan for propagation loss over an open grass surface (Ref: McGeehan J P, 1985). The power law is less rapid than the $r^{-4}$ law which results from propagation over a plane earth at large transmitter-receiver separations, Section 2.8.4.2. It appears that the buildings surrounding the micro-cell prevent radiation from escaping, but there is no indication of channelling along Malet Street leading to an increase in received signal. The increase in the propagation loss above that for free space propagation appears to be determined by the increased chance of signal obstruction as transmitter-receiver separation increases.

From Figure 4.13, the power law for propagation around corners is approximately $r^{-1.0}$. It should be possible to model the fall off in power around corners in terms of knife edge diffraction.

### 4.2.1.5 Propagation Loss Contour Map

The propagation loss distribution resulting from propagation by a single fixed transmitter mounted on the 2nd floor of the Anthropology Building in Malet Place, Location 2 in Figure 4.1, and pointing along Malet Street, is illustrated in Figure 4.14.
The propagation loss values are referred to a 0 dB point which occurs at a transmitter-receiver separation of 30 m. The propagation loss at this point obtained from Figure 4.12 is 100 dB [1]. Thus, the values on Figure 4.14 may be converted into propagation loss values by subtracting 100 dB. Superimposed on the Figure is a map of the area in the immediate vicinity of the transmitter. Malet Street defines dimensions of the micro-cell. The Figure illustrates the containment

1 Free space propagation loss prediction at 30 m is 97 dB.
of the transmitted power within the micro-cell. Only a very small amount of power escapes into the streets which cross the micro-cell, and the signal level in the street which runs parallel to the micro-cell is more than 50 dB [1] below that in the micro-cell itself in almost all circumstances.

4.2.2 SIGNAL VARIATION OVER MEDIUM SCALE SIZES

This Section presents the results of the study of the "slow fading" introduced in Section 2.4.2 to describe the variation of the "local" mean of the signal from place to place within the micro-cell. The "slow fading" is characterized in terms of location variability defined as the standard deviation of the median propagation loss, expressed in dB, Section 2.8.4.3.

The local mean signal strength, in the absence of an ensemble average, is estimated by a local time (or distance) average of the received time series over a few tens of wavelengths. Lee shows that the estimate of the local mean is not very sensitive to the distance over which the average is computed and that averaging lengths in the range 10 - 40λ are suitable, (Ref: Lee W C Y, 1982). The local mean is estimated using a moving average routine which forms the weighted average of N points on either side of the input point. A triangular weighting function is used to reduce the possibility of the sidelobes of the equivalent lowpass filter of the weighting function in the Fourier domain distorting the data. The length averaging window between its 3 dB points is 40λ, giving an overall length for the triangular window of 80λ. This length is chosen to average out the "fast fading" yet be short enough not to obliterate the "slow fading" of the signal envelope, Section 2.4.4, Table 2.1. The number of points used for the 80λ length of the averaging window is calculated using the measured speed information for each experimental run.

To ensure a simple relationship between the time and the distance intervals between the sampled points, data is selected from portions of the run where the velocity of the mobile is constant. Each digitized segment of 16 K two byte integers represents 1.6 seconds of data at 10 KHz sample rate, Section 3.2.5.

Figure 4.7b is a typical speed profile of an experimental run along the micro-cell and illustrates that although normalizing the data with respect to the speed of the mobile is difficult to accomplish, there are many sections of the run where speed is

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1 Equivalent to a propagation loss of 150 dB referred to the transmitter.
approximately constant for time durations of 1.6 seconds or longer. Thus, the restriction of analysing data sections with constant velocity does not unduly affect the number of data segments available for analysis.

A 1.6 second segment of digitized data represents only 7.1 m of an experimental run at the maximum car speed of \( \sim 16 \) mph [1] and, for transmitter-receiver separations greater than 60 m, the difference in \( r^{-2} \) propagation loss between the two ends of the data segment is less than 1 dB. Thus, to all intents and purposes all the variation in the average signal level in the data segment may be attributed to location variability. The location variability of each data segment is estimated from the dB difference between the 50% and 84% points of the cumulative distribution of the local mean [2]. The data for the location variability calculation may be obtained from the single channel probing, two frequency probing, and two polarization probing measurements.

### 4.2.2.1 Location Variability

A sample of the large number of location variability measurements taken on a particular urban path is presented in Table 4.2, Appendix I, contains an index of the measurements shown in Figure 4.2. Each measurement is obtained from a 1.6 second sample of data. The Table shows that the location variability changes even within a single type of propagation environment. The measurements are taken from experimental runs along Torrington Place, Location 4 in Figure 4.1. A spread of location variabilities from 0.64 dB to 7.2 db is observed with an average value of 3.2 dB. Comparison with results obtained at 900 MHz by other workers where the location variability is 4.46 dB, suggests that the result at 55 GHz is similar to that at 900 MHz measured in a suburban environment (Ref: IEEE VT Special Issue, 1988). The results illustrate the extreme variability of statistical estimates obtained in the micro-cellular environment. This variability is expected due to the essential line of sight character of the propagation.

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1 The car speed is severely restricted by the density of traffic in the mobile radio environment.

2 This follows from the shape of the normal distribution the deviation where the 50% \( \rightarrow \) 84% deviation is equal to \( \sigma \).
<table>
<thead>
<tr>
<th>Experiment Reference in Appendix I</th>
<th>Location Variability</th>
<th>Experiment Reference in Appendix I</th>
<th>Location Variability</th>
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<td>2b(25)-D</td>
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<td>1-B</td>
<td>5.04 dB</td>
</tr>
<tr>
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<td>3.79 dB</td>
<td>2-B</td>
<td>6.0 dB</td>
</tr>
<tr>
<td>8b(100)-G</td>
<td>2.27 dB</td>
<td>1a(200)-B</td>
<td>6.0 dB</td>
</tr>
<tr>
<td>9a(100)-G</td>
<td>4.8 dB</td>
<td>1a(100)-B</td>
<td>0.94 dB</td>
</tr>
<tr>
<td>9b(100)-G</td>
<td>5.2 dB</td>
<td>1b(100)-B</td>
<td>0.64 dB</td>
</tr>
<tr>
<td>10a(100)-D</td>
<td>1.64 dB</td>
<td>1a(320)-C</td>
<td>5.04 dB</td>
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<td>7.2 dB</td>
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<td>2a(25)-D</td>
<td>1.96 dB</td>
<td>-</td>
<td>-</td>
</tr>
</tbody>
</table>

Table 4.2 Variation of Urban Location Variability

Figures 4.15 and 4.16 are produced from the same data set (Reference 1a(25)-D in Appendix I) and illustrate a typical time series, the variation of its local mean and the cumulative distribution of the local mean on a log-normal probability plot. Figure 4.15 shows a comparison of a 1.6 second portion of a time series with the estimated variation of its local mean.
Figure 4.15 Time Series Compared with Local Average Variation
The time series is taken from a measurement along Torrington Place, Location 4 in Figure 4.1, at a transmitter-receiver separation of ~120 m. The time series is sampled at 10 KHz which allows frequencies of up to 5 KHz to be reproduced. However, the 16 K points produced by the sampling operation is too large a number to be illustrated on a single plot. To circumvent this difficulty, 30 points are averaged to produce a single point on the traces in Figure 4.15. This operation is referred to as decimation by 30. After the decimation, only frequencies of up to 170 Hz are reproduced. Figure 4.16 shows the cumulative log-normal probability plot of the local mean data. The location variability of the data is 4.8 dB which is near to the mid-point of the values summarized in Table 4.2. The fit of the data to a log-normal distribution is good, evidenced by the near straight line fit of the points in Figure 4.16. The curve is typical for other measurements taken in the vicinity of the College.

![Figure 4.16 Log-Normal Probability Plot of Local Average Distribution](image-url)
Chapter 4

4.2.3 SIGNAL VARIATION OVER SMALL SCALE SIZES

The results of the measurement of the small scale size signal strength distribution are presented in two formats: as standard probability density function histograms; and as Weibull probability plots.

To examine the small scale signal distribution using data from a typical experimental time series, the variation in the mean level throughout the time series must be removed. This is accomplished by dividing the signal by its local mean value which is calculated from time series using the moving average routine described in Section 4.2.2. After normalization, the data is approximately ergodic. Attempts to estimate the statistical properties of the signal from time/distance averages are subject to distortion due to the modulation of the mean value of the data.

The probability density of the small scale signal distribution is characterized by its loss deviation, defined as the spread between the 50% and 90% probability levels on the cumulative probability distribution of the data. The loss deviation is used because it is a more sensitive measure of the signal distribution at low signal levels than the standard deviation of the signal, Section 2.10.2.

4.2.3.1 Probability Histograms

Initially, analysis of the probability distribution of the received signal envelope was made via a study of the probability density function. Subsequently, analysis was switched to study probability plots of the cumulative distribution of the data which is the generally accepted approach for this analysis because the distribution of the signal at low powers is shown more clearly (Ref: IEEE VT Special Issue, 1988). However, the two estimates of the probability density function of the received signal envelope are shown here because they illustrate clearly how the character of the received signal statistics is different at different locations within the micro-cell.
Figure 4.17a Probability Density Histogram at 100 m Tx-Rx Separation

Figure 4.17b Probability Density Histogram at 190 m Tx-Rx Separation
Figures 4.17a and 4.17b are the probability density functions (PDF) of the received signal measured at 100 m and 190 m transmitter-receiver separations respectively in Malet Street, Location 2 in Figure 4.1. Overlaid on the probability histograms respectively are the Rayleigh and Rice probability distributions. The curves are fitted using the estimated mean and standard deviation of the experimental data and, in the case of the Rician distribution, by estimating the size of the line of sight component in proportion to the scattered components.

The shape of the distribution changes quite noticeably between the two locations. At 100 m separation, Figure 4.17a, the curve is closer to a Rician than a Rayleigh distribution in nature, thus indicating a significant line of sight component. However, at 190 m transmitter-receiver separation, Figure 4.17b, the curve is apparently Rayleigh distributed, at least on a plot with linear scales, indicating that the line of sight component is no longer dominant.

The change in the character of the probability density function of the signal between the two locations is to be expected due to the change in the nature of the propagation mechanism from essentially line of sight near to the transmitter to a more random distribution of scattered waves at larger separations. This change is also reflected in the observed character of urban time series which shows close agreement with a prediction based on interference between direct and specular reflected components at small transmitter-receiver separations, but less agreement as the separation increases, Section 4.1.23. There is not a definite relationship between distance and received signal probability density function. The distribution could well be Rician distributed at very large separations if the LOS component happened to be strong in comparison with the randomly scattered component. However, there is a general tendency for the distribution to become more Rayleigh distributed at large transmitter-receiver separations.

4.2.3.2 Loss Deviation

Table 4.3 summarizes the results of the measurement of loss deviation \(^1\) on a particular urban path; each measurement is obtained from a 1.6 second sample of data. The results are obtained along Torrington Place, Location 4 in Figure 4.1. Appendix I contains an index of the measurements presented in Table 4.3. The Table shows that the loss deviation is variable even within a single propagation environment. A spread of loss deviations from 0.9 dB to 9.3 dB is observed with a

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\(^1\) Loss deviation is defined as the 50% - 90% spread on the cumulative distribution of the small area signal distribution.
mean value of 4.7 dB. The results show a similar spread in values to that observed in the location variability in Section 4.2.21. Comparison with results obtained at 900 MHz by other workers where a spread in location variabilities from 0.94 dB to 7.54 dB with a mean value of 4.32 dB is observed, shows that similar results are obtained for 900 MHz measurements in suburban environments (Ref: IEEE VT Special Issue, 1988).

Shepherd presents experimental results, measured at 900 MHz, in an urban environment which show loss deviations in the range of 3 dB to 30 dB (Ref: Shepherd N H, 1977). Shepherd suggests that the higher loss deviation values occur when reflections from high sided buildings are of comparable amplitude to the LOS signal. This situation is quite common in Dallas where Shepherd conducted his experiments because Dallas has a very large number of tall buildings.

The results underline the extreme variability of statistical estimates in the micro-cellular environment which was seen in the location variability measurements in Section 4.2.21. This variability is expected due to the essential line of sight character of propagation in a micro-cell which means that the distribution of the received signal will change radically if any object obstructs the line of sight signal or if specular reflections become comparable in amplitude with the LOS component.

<table>
<thead>
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<td>1b(25)-D</td>
<td>6.9 dB</td>
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<td>8a(100)-G</td>
<td>9.3 dB</td>
<td>2a(25)-D</td>
<td>4.6 dB</td>
</tr>
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<td>8b(100)-G</td>
<td>4.5 dB</td>
<td>2b(25)-D</td>
<td>3.4 dB</td>
</tr>
<tr>
<td>9a(100)-G</td>
<td>2.9 dB</td>
<td>1-B</td>
<td>3.9 dB</td>
</tr>
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<td>9b(100)-G</td>
<td>3.1 dB</td>
<td>2-B</td>
<td>4.0 dB</td>
</tr>
<tr>
<td>10a(100)-G</td>
<td>6.6 dB</td>
<td>1a(200)-B</td>
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</tr>
<tr>
<td>-</td>
<td>-</td>
<td>1b(100)-B</td>
<td>1.0 dB</td>
</tr>
</tbody>
</table>

Table 4.3 Variation of Urban Loss Deviation

Weibull probability plots show more clearly the distribution of signal power at low signal levels. On a linear scale this portion of the signal distribution is compressed. However, it is this portion of the distribution which is important for predicting the performance of a communications system.
Figure 4.18 Time Series Compared with Normalized Time Series
Figure 4.18 shows a comparison of a 1.6 second portion of a typical time series before and after division by the local mean of the data. The data is obtained from a measurement along Torrington Place, Location 3 in Figure 4.1, at a transmitter-receiver separation of 120 m (Reference 1b(25)-D in Appendix I). As with the time series illustrating the variation in the local mean of the data in Section 4.2.2.1, the data is decimated [1] by a factor of 30 which limits the maximum frequency reproduced in Figure 4.17 to 170 Hz. It is apparent that in some sections of the data the signal variation is reduced as the average signal strength rises, indicating that the rise in the average signal level is due to a non-fading, ie line of sight, component.

Figure 4.19 Weibull Probability Plot of Normalized Signal Distribution

Figure 4.19 shows the cumulative Weibull probability plot of the normalized data, shown in Figure 4.18. Also shown on the Figure is the cumulative Rayleigh

1 NB Every point on the curve is the average of 30 data samples.
distribution which is a straight line on a Weibull probability plot with a loss deviation of 8.2 dB. The fit of the data to a Rayleigh distribution is not good, evidenced by the difference in slope of the two curves. The loss deviation of the data, 3.6 dB, is significantly smaller than the 8.2 dB loss deviation expected for Rayleigh distributed data. A Rayleigh distributed signal envelope arises when a large number of randomly phased components are added, given that no component dominates the others. Thus, departure from a Rayleigh distribution is not surprising given the essential line of sight character of propagation in a micro-cell.

The similarity with experimental results obtained at 900 MHz in a suburban area, also noted in Section 4.2.2.1 with regard to the location variability of the data, suggests that the suburban propagation mechanism at 900 MHz is the same as the urban propagation mechanism in a micro-cell at 55 GHz (Ref: IEEE VT Special Issue, 1988). This proposition seems quite reasonable because in both environments there is a high probability of line of sight between the base station and mobile since there are relatively few scatterers which can obstruct the signal.
4.2.4 RESULTS FROM "OPEN AREA" MEASUREMENTS

This Section presents the results of the first order statistics measured in an "open area" rural environment at Shenley in Hertfordshire, Section 4.1.1.

Figure 4.20 is a typical probability histogram of the received signal distribution obtained at Shenley at a transmitter-receiver separation of 100 m (Reference A-1 in Appendix I). The distribution is apparently Gaussian. Comparison with the probability density functions measured in the urban environment, shown in Figure 4.17, show that the line of sight component here is even more significant in the rural environment. This is expected given the undisturbed line of sight path throughout the whole of the run. The loss deviation of the data is ~1 dB. The location variation parameter is very close to 0 dB due to the constant line of sight propagation conditions in the rural environment.

Figure 4.20 Probability Histogram for Open Area Data (100 m Tx-Rx Separation)
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4.2.5 SUMMARY OF FIRST ORDER STATISTICS

This Section brings together the results of the received signal envelope characterization over different scale sizes.

The results of the large scale size measurements in Section 4.2.1 graphically illustrate the disposition of power within the micro-cell and its adjoining streets, and confirm the assumption that the buildings surrounding the micro-cell act as "hard boundaries" which determine the extent of a particular micro-cell. Thus, the argument presented in Chapter 1 which describes the differences between conventional cellular and micro-cellular radio is shown to be valid. The power law which describes the fall off in signal strength as a function of transmitter-receiver separation, $r$, is shown by a plot of propagation loss in dB versus $10 \log r$ in Section 4.2.1.4, Figure 4.13, to be approximately $r^{-3.4}$. The fall off in power outside the bounds of the micro-cell, also shown in Figure 4.13, follows an approximate $r^{-10}$ law and appears to be governed by a diffraction process. However, since the power levels outside the main cell are very low, and subject to error in their estimation, no attempt is made to model the precise relationship.

The medium scale size measurements in Section 4.2.2 show that the location variability [1] has a range of values from 0.6 dB to 7.2 dB with a mean value of 3.2 dB and the distribution of the local average signal conforms to a log-normal distribution. The small scale size measurements in Section 4.2.3 show that small scale cumulative distribution is usually not Rayleigh distributed. The loss deviation [2] has a range of values from 1 dB to 9.3 dB with a mean value of 4.7 dB. Comparison of the medium and small scale signal distributions to measurements reported at 900 MHz in a low density urban environment (Ref: IEEE VT Special Issue, 1988) show a marked degree of accord. The mean location variability and loss deviation are very similar, as is the spread of loss deviation values observed. From the agreement of the results at 55 GHz and 900 MHz, it is concluded that the type of environment presented by a low density urban environment at 900 MHz in a conventional cellular scheme is similar to that presented by a heavily built up urban environment at 55 GHz in a micro-cellular scheme. Both scenarios exhibit a very high probability of a direct line of sight path existing between base and mobile combined with a relatively small number of scatterers.

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1 The standard deviation of the local average decibel signal level, ie the difference between 50% and 84% probability levels of the cumulative distribution.

2 The difference between the 50% and 90% probability levels of the cumulative distribution.
Chapter 4

It is suggested that one cause of the variation in both the location variability and loss deviation is explained by the change in the relative strength of the line of sight component. Where the line of sight component is constant in amplitude and large in comparison with the scattered components, both location variability and loss deviation are suppressed. This is the situation found in the rural environment at Shenley and for small Tx-Rx separations in the urban environment. It is evidenced respectively by the apparently Gaussian and Rician distributions measured in these environments, Figures 4.20 and 4.17. If the line of sight component fades in and out over the set of locations used to calculate location variability, it produces a large modulation of the mean value and consequently a large location variability. If, however, the line of sight component is small or is of comparable magnitude to specular reflections from the road surface, then the loss deviation will be large. The argument is supported by the change in the probability distribution between different transmitter-receiver separations in an urban micro-cell, Figure 4.17. At larger separations, the probability that the distribution is dominated by the line of sight component becomes less and the loss deviation of the distribution tends to increase.

The quantities used to describe the distribution of the signal power over different scale sizes provide an overall framework within which the highly varied nature of the received signal envelope may be described. The framework appears to agree with the observed character of the received signal envelope at 55 GHz at least as well as it does at 900 MHz. However, it must be noted that the framework is only approximate and, in reality, there are no clear divides between the different scale sizes.

The most striking feature of the received signal strength is the variability of the estimates of the statistical quantities used to characterize it within a single propagation environment.

4.3 SECOND ORDER STATISTICS

This Section presents the results of the calculation of the auto-correlation and power spectrum of the normalized squared envelope of the received signal. The measured results in an urban environment are compared to the predictions in Section 2.5 derived from the theoretical models of Clarke and Aulin (Ref: Clarke RH, 1968; Aulin T, 1979) and to the predictions of a simple deterministic model derived in Section 2.7. In addition, the power spectra measured in urban and rural environments are compared.
The second order statistics are estimated from a time average because an ensemble average is not available. To reduce the bias to the estimate caused by the non-stationary variation of the signal mean, the data is normalized with respect to its local mean value, thus providing an approximately ergodic data set.

### 4.3.1 AUTO-CORRELATION

The normalized auto-correlation of the squared envelope, $x$, as a function of the time increment, $j$, is defined in terms of the sampled data by the sum:

$$R_x(j) = \frac{1}{N-j} \sum_{i=0}^{N-j} x_i x_{i+j} - \bar{x}^2$$

$$j = 0 \rightarrow M$$

where: $N =$ number of samples, $M =$ number of lags, $j =$ lag number, $x_i =$ the $i$th sample of the squared envelope $r^2$, $\bar{x} =$ the mean of $x$.

The auto-correlation is defined in terms of time lags, $j$. However, by assuming that the disposition of scatterers remains constant and that, as a result, the mobile vehicle is moving through a spatial interference pattern with no time variation, the auto-correlation may be represented as a function of the distance, $d$, traversed by the mobile unit between two times $t_2$ and $t_1$, instead of as a function of the time lag $(t_2 - t_1)$ between the two times, ie $d = (t_2 - t_1)V$ where $V$ is the velocity of the vehicle.

The auto-correlation of the squared envelope of the received signal expressed as a function of distance may be used to predict the minimum antenna separation on the mobile unit which would be required for a space diversity system. To do this we consider the auto-correlation function not as the correlation of the signal with itself, but as the cross-correlation function between the signal received by two spatially separated antennae. Ideally, the inputs to a diversity system are uncorrelated, ie their correlation coefficient is zero. Thus, we are interested in the predicted spatial separation between two hypothetical antenna which will reduce the cross-correlation coefficient between their two received envelopes to zero. This distance is defined as the de-correlation distance.
Figure 4.21 shows the estimated squared envelope, normalized, auto-correlation function calculated for a typical experimental run along Torrington Place, Location 4 in Figure 4.1 (Reference 10a(100)-G, Appendix I). From Figure 4.21 it can be seen that an antenna separation of $7\lambda$ is required to reduce the correlation between the two antennae to zero. Also shown on the Figure is the auto-correlation function calculated from the data before the normalization process. It is clear from comparison of the two curves that the estimation from the data, before the normalization operation, over-estimates the distance.

![Figure 4.21 Experimental Squared Envelope Auto-Correlation Function](image)

In Section 2.5.5.1 a prediction is made of the expected shape of the squared envelope auto-correlation function of the received signal. The prediction is based on the theoretical model of Aulin (Ref: Aulin T, 1979). It takes account of the $120^\circ$ directional antenna used by the receiver which truncates the angle of arrival distribution and it also considers the effect of a line of sight component. The prediction is based on an Aulin type angle of arrival distribution which is uniform over the beamwidth of the receiver antenna and has a maximum elevation arrival angle of $30^\circ$. The predicted de-correlation length in the absence of a line of sight
component is $2\lambda$. However, if a line of sight component is included in the angle of arrival distribution, the predicted width of the angle of arrival distribution can be increased almost indefinitely, depending on the proportion of the power contained in the line of sight component in relation to that in the scattered component. In the limit where all the power is contained in the line of sight component, the angle of arrival distribution will be a delta function in azimuth and the auto-correlation function will be flat.

Table 4.4 shows a number of de-correlation measurements taken in a particular urban environment. The measurements are obtained in Torrington Place, Location 4 in Figure 4.1. Also shown in the Table is the estimated strength of the line of sight component in relation to the power in the scattered component. This is estimated using Equation 2.69 from Section 2.5.7:

$$\alpha/P = (L'/L)^2 - 1$$  \hspace{1cm} 2.69

where: $\alpha$ = power of LOS component; $P$ = total power of scattered component; $L'$ = experimentally measured de-correlation length (length to first zero of the auto-correlation function); and $L$ = predicted width of auto-correlation for a $120^\circ$ beamwidth horn.

<table>
<thead>
<tr>
<th>Reference (Appendix I)</th>
<th>Width of Auto-Correlation</th>
<th>Est Rate $\alpha/p$ (Eqn 2.69)</th>
<th>Reference (Appendix I)</th>
<th>Width of Auto-Correlation</th>
<th>Est Rate $\alpha/p$ (Eqn 2.69)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1a(25)-D</td>
<td>3.6</td>
<td>2.2</td>
<td>7a(P)-G</td>
<td>2.5</td>
<td>0.6</td>
</tr>
<tr>
<td>1b(25)-D</td>
<td>3.0</td>
<td>1.3</td>
<td>6a(P)-G</td>
<td>2.5</td>
<td>0.6</td>
</tr>
<tr>
<td>2a(25)-D</td>
<td>3.6</td>
<td>2.2</td>
<td>5a(P)-G</td>
<td>17.8</td>
<td>78.2</td>
</tr>
<tr>
<td>2b(25)-D</td>
<td>4.2</td>
<td>3.4</td>
<td>4a(P)-G</td>
<td>13.8</td>
<td>46.6</td>
</tr>
<tr>
<td>1a(100)-G</td>
<td>4.0</td>
<td>3.0</td>
<td>4b(P)-G</td>
<td>13.3</td>
<td>43.2</td>
</tr>
<tr>
<td>1b(100)-G</td>
<td>4.36</td>
<td>3.6</td>
<td>1(P)-G</td>
<td>10.0</td>
<td>24.0</td>
</tr>
<tr>
<td>2a(100)-G</td>
<td>17.5</td>
<td>76.0</td>
<td>7a(100)-G</td>
<td>8.7</td>
<td>17.9</td>
</tr>
<tr>
<td>3a(100)-G</td>
<td>2.9</td>
<td>1.1</td>
<td>10a(100)-G</td>
<td>2.7</td>
<td>0.8</td>
</tr>
<tr>
<td>3b(100)-G</td>
<td>4.36</td>
<td>3.6</td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Table 4.4 Selection of De-Correlation Lengths for a Series of Measurements in an Urban Environment (Appendix I contains an index of the measurements presented in Table 4.4)

---

1 The width of the auto-correlation is defined as the distance in wavelengths to the first zero of the function.
The values in the Table range from a minimum of 2.5$\lambda$ up to a maximum of 22$\lambda$. The mean value of the de-correlation lengths in Table 4.4 is 6.4$\lambda$. The estimated ratio between the power in the direct and the scattered component of the received signal also varies over a very large range of values. The lowest ratio between line of sight power and scattered power from the values in Table 4.4 is 0.6 and the highest ratio is 78. The ratio between the direct and scattered radiation for the average de-correlation length of 6$\lambda$ is 8.

The results show that the received signal contains a very large and highly variable line of sight component. This is reasonable given the geometry of a micro-cell which exhibits few major obstructions to the line of sight signal. The result is also supported by the measurements of the probability distribution of the signal, summarized in Section 4.2.5, which indicate the presence of a strong non-fading component.
4.3.2 POWER SPECTRUM

The comparison of square law power spectra measured in an open area rural environment and an urban environment, at identical car speeds, is presented in Figure 4.22. The reduced level of high frequency content in the open area power spectrum is marked in comparison with the power spectrum derived from the urban measurements.

This result is explained by the presence of only one major scatterer in the rural environment in contrast to the multiplicity of scatterers found in the urban environment. The paucity of scatterers restricts the range of Doppler components at the receiver. Thus, the spectrum of the square law detected envelope is very narrow, whilst the relatively large number of scatterers in the urban environment results in a larger spread of Doppler shift components at the receiver.
Typical Power Spectrum for an Urban Environment

Typical Power Spectrum for a Rural Environment
(with no moving scatterers)

Figure 4.22 Comparison of Urban and Rural Squared Envelope Power Spectra
A large number of power spectra were measured in the vicinity of UCL. Figure 4.23 illustrates several typical results of these measurements compared with the theoretical predictions, evaluated in Section 2.5, based on the models of Aulin and Clarke (Ref: Aulin T, 1979; Clarke R H, 1968). Both models assume a receiver antenna beamwidth of 120° and an angle of arrival distribution which contains a line of sight component. The relative power in the line of sight component in relation to the scattered power is adjusted to give the best fit to the experimental points. These results are published in Siqueira G L et al, 1988. The experimental points in the pair of Figures 4.23a and 4.23c are the same as the two spectra in Figures 4.23b and 4.23d. Each spectrum is an average of three spectra obtained from three adjacent 1.6 second portions of data from the same experimental run. Spectrum in Figures 4.23a and 4.23c is obtained from an experiment carried out along Malet Street, Location 2 in Figure 4.1, with the transmitter mounted on a tripod at 2.6 m elevation. The other spectrum in Figures 4.23b and 4.23d is obtained from an experiment carried out along Torrington Place, Location 4 in Figure 4.1, with the transmitter mounted at 25 m elevation. The maximum Doppler frequency, \( f_m \), calculated from the car velocity and free space wavelength [1] is 2000 Hz in both spectra.

Comparison of the experimental points with the theoretical predictions in Figure 4.23 illustrates that the Clarke model provides a better fit to the data obtained from a transmitter antenna with 2.6 m elevation, whilst the Aulin model provides a better fit to the data obtained from a transmitter antenna with 25 m elevation. The differences between the predictions are particularly marked at low frequencies. It is suggested that the agreement of theory with prediction in the two cases arises from the different assumptions made about the distribution of elevation arrival angles made by each model. The Clarke model assumes a horizontal angle of arrival distribution; a condition which it is expected will be met in the experimental scenario with 2.6 m elevation. The Aulin model, however, assumes a narrow vertical distribution of elevation angles in addition to an azimuthal distribution. This situation will be expected to occur in the experimental scenario with 25 m elevation.

\[ f_m = \frac{2V}{\lambda}, \text{ Where V is the speed of the car.} \]
Figure 4.23a Comparison of Urban Squared Envelope Power Spectrum (2.6 m Tx Antenna Elevation) with Theoretical Prediction of Clarke Model

Figure 4.23b Comparison of Urban Squared Envelope Power Spectrum (25 m Tx Antenna Elevation) with Theoretical Prediction of Clarke Model
Figure 4.23c Comparison of Urban Squared Envelope Power Spectrum (2.6 m Tx Antenna Elevation) with Theoretical Prediction of Aulin Model

Figure 4.23d Comparison of Urban Squared Envelope Power Spectrum (25 m Tx Antenna Elevation) with Theoretical Prediction of Aulin Model
A deterministic model is developed in Section 2.7 to investigate the effect of the line of sight and specular components. Figures 4.24a and 4.24b show the spectrum generated by this model compared with experimental spectra and the predictions of the models of Clarke and Aulin respectively. Figure 4.24a shows a comparison with experimental data obtained with 2.6 m base station elevation, whilst Figure 4.24b shows a comparison with experimental data obtained with 25 m base station antenna elevation. The deterministic spectrum shows similar behaviour to the experimental spectra at low frequencies. However, the fall off in spectral content at higher frequencies is much more rapid than that measured experimentally or than that predicted by the models of Aulin and Clarke.

The experimental spectra are intermediate between the predictions of the deterministic model and those of the theoretical models of Clarke and Aulin. The differences between experiments and the various theoretical predictions are easily explained in terms of the spread of Doppler frequencies in each case. The smaller the spread in Doppler frequency values, the narrower the square law power spectrum will be. This effect is used in Section 2.5.7 to derive an approximate relationship between the relative strength of the LOS component and the width of the auto-correlation function to the first zero. The deterministic model assumes a very narrow angular distribution of received waves, of the order of 10°. The theoretical models assume an omni-directional distribution of arrival angles truncated to 120° by the beamwidth of the receiver antenna. Thus, the effective width of the distribution of arrival angles is between 10° and 120°.

The theoretical curves in Figure 4.24 represent the best visual fit to the measured data points which is obtainable by adjusting the strength of the LOS component in relation to the scattered component. That the theoretical curves in Figure 4.24 do not fit more accurately to the experimental points, suggests that the assumed shape of the distribution of the scattered component is incorrect. This indicates that the angular distribution of the randomly scattered component is not uniform across the beamwidth of the receiver antenna.

The results of the power spectrum measurements are in agreement with expectations for propagation in an urban micro-cell. First, there is strong line of sight component present which is expected due to the small number of objects present in the micro-cell which are capable of obstructing the direct line of sight path. Secondly, the angular distribution of the randomly scattered part of the received signal is not uniform which is expected due to the small number of scatterers present in the micro-cell in comparison with the assumption of an omni-directional continuum of scatterers which is assumed by the models of Clarke and Aulin.
Figure 4.24a Comparison of Urban Squared Envelope Power Spectrum (2.6 m Tx Antenna Elevation) with Deterministic Spectra and with the Theoretical Prediction of Clarke

Figure 4.24b Comparison of Urban Squared Envelope Power Spectrum (25 m Tx Antenna Elevation) with Deterministic Spectra and with the Theoretical Prediction of Aulin
4.3.3 SUMMARY OF SECOND ORDER STATISTICS

The results of both the auto-correlation and power spectra measurements on the squared envelope of the received signal indicate that the received signal contains a substantial, but highly variable line of sight component and that the angular distribution of the randomly scattered part of the received signal is not uniform in azimuth. Indeed, it would have been surprising if this were not found to be the case considering the distribution of scatterers within the micro-cell and the essential line of sight propagation conditions.

The results of the auto-correlation measurements indicate that a space diversity system would typically require an antenna separation of $6\lambda$ at the mobile unit to provide un-correlated input channels. However, the separation required can range from $2.5\lambda$ to $22\lambda$. This is equivalent to a separation of 1.4 cm to 12 cm between the receiver antennae at the mobile unit which is feasible to implement even in a hand-held system. In addition, the measured width of the auto-correlation function is used to provide an estimate of the ratio of the power contained in the LOS component over that contained in the randomly scattered component of the received signal. Estimates range from a ratio of 0.6 up to a ratio of 78 illustrating the extreme variability of the propagation in a micro-cell.

Comparison of the experimental squared envelope power spectrum with theory suggests that the distribution of scattered waves arriving at the mobile terminal is approximately horizontal when the base station and mobile unit antenna heights are equal. However, when the base station elevation is 25 m in comparison with a mobile terminal height of 1.5 m, the distribution of received waves of the mobile has a significant distribution in vertical arrival angles. In addition, the experimental power spectra are shown to lie midway between the theoretical predictions of the models of Clarke and Aulin and that of a simple model based on the interference of a direct ray with reflections from three principal reflectors in the micro-cell. This suggests that the assumption of a uniform angular distribution of the randomly scattered component assumed by the Clarke and Aulin models is incorrect.

Thus, taken together, the results agree with the expected nature of propagation within a micro-cell, that is the propagation is heavily influenced by the special geometry of the micro-cell which results in a high probability of line of sight transmission and a non-uniform angular distribution of the randomly scattered part of the received signal. The results are in accord with the results of the first order statistics summarized in Section 4.2.5 which also indicated the strong influence of line of sight propagation in the urban micro-cell.
4.4 JOINT FIRST ORDER STATISTICS

A diversity system combats multipath fading by combining two or more uncorrelated channels to obtain an output which has fewer deep fades. The performance of a diversity system may be assessed by comparing the distributions of the input channels with the distribution of the output. Usually only comparison with one of the inputs is necessary because their statistical distribution is the same. The two parameters commonly used to characterize the difference between the input and the output distributions are the increase in the level of the median signal level and the increase in the level which the signal exceeds for 99% of the time. Both of these parameters are easily displayed on a probability plot of the input and output distribution of the signal.

This Section assumes a simple additive, post square law detection, combination process to generate the combiner output signals. To effect the combination, the signals are converted from units of relative power in dB to linear units of relative power $[1]$. After combination, the signal is re-converted back to units of relative power in dB. The input branches can be derived from the single channel probing measurements by using a time separation between the two channels to simulate antenna space separation, by using two frequency measurements to simulate frequency diversity, or by using two polarization measurements to simulate polarization diversity.

The enhancement to system performance due to diversity reception is normally based on assumption of uncorrelated input channels, ie channels with a correlation coefficient of zero. If the correlation coefficient between the input channels of a diversity combiner is greater than zero, the enhancement to system performance is reduced. Jakes (Ref: Jakes W C Jr, 1974, pp 326) examines the reduction in the performance of a two branch selection diversity receiver for Rayleigh fading input channels with envelope correlation coefficients of 1, 0.8, 0.6, 0.3 and 0, using the example of a selection diversity receiver. To illustrate the reduction in performance, Jakes plots curves of the percentage of time for which the signal to noise ratio (SNR) of the channel is greater than a particular signal level referred to the mean signal to noise ratio of the channel. The performance of the diversity receiver for the different correlation coefficients is indicated by the increase in the signal level at the 99% reliability level $[2]$ in comparison with the no diversity case. A correlation coefficient of 1 between the input channels yields no improvement in

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1 The anti-log is taken.

2 This is the signal level which is exceeded for 99% of the time.
performance as could be expected. At 0.8 correlation coefficient the improvement is 5.8 dB, at 0.6 correlation coefficient it is 8.7 dB, at 0.3 correlation coefficient it is 10 dB, and at a correlation coefficient of 0 the improvement is 10.5 dB. Thus, for correlation coefficients less than 0.3 the performance enhancement is within 0.5 dB of the predicted improvement for uncorrelated channels. Even when the correlation coefficient rises to 0.6, the performance enhancement is still within 1.8 dB of the predicted value for uncorrelated channels.

Two examples of simulated diversity reception using portions of experimental time series are presented here. The first example uses two portions of time series transmitted on orthogonal polarizations (Reference 1(P)-G, Appendix I). The experimental apparatus for these measurements is described in Section 3.7. The estimated cross-correlation coefficient between these time series is -0.3. The second example uses two portions of time series transmitted with 100 MHz frequency separation (Reference 10(100)-G and 5b-G, Appendix I). The experimental apparatus for these measurements is described in Section 3.6. The estimated cross-correlation coefficient between these two time series is 0.4.

The first example should provide an example of ideal diversity receiver performance. Indeed, since the correlation is slightly negative, the system may perform slightly better than prediction, based on zero correlation between the channels. The negative correlation indicates that the probability of peaks in one waveform lining up with troughs in the other is greater than it is for zero correlation between the input channels. The second example should provide an example of performance which is slightly inferior to the prediction for uncorrelated input channels. The correlation of 0.4 between the input channels increases the probability that deep fades in each trace will coincide.
Chapter 4

4.4.1 JOINT PROBABILITY DISTRIBUTION

The first example presented is for combination of two time series from an orthogonal polarization experiment, the correlation between the time series is -0.3. Figure 4.25 illustrates the two input time series to the combination process. They are 3 second portions taken from an orthogonal polarization experiment carried out along Torrington Place, Location 4 in Figure 4.1. The top trace is for vertical polarization and the bottom for horizontal polarization. The correlation coefficient between the two time series is -0.3. The time series are sampled at 5 KHz and decimated by a factor of 30 [1] which reduces the maximum frequency reproduced in the Figure from 2.5 KHz to 83 Hz.

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1 Every point in the Figure is the average of 30 sampled data points.
Figure 4.25 Time Series Inputs to Simulated Diversity Combiner
(Orthogonal Polarization Correlation Coefficient - 0.3)
Figure 4.26 shows the output time series after the combination operation. Comparison between Figures 4.25 and 4.26 clearly shows that the majority of fades are suppressed by the combination operation. Figure 4.27 shows the comparison between one of the input distributions to the combiner with the combiner's output distributions on a log-normal probability plot. Only one of the input distributions is illustrated because they are similar. The combination operation increased the mean signal level by 1 dB and increases the signal level at the 99% reliability level by 9 dB. The loss deviation is reduced from 3.8 dB for the input channels to 1 dB for the output channels. This can be seen from the change in the 50% to 90% deviations between the input and output distributions in Figure 4.27.
Comparison with a theoretical model, based on selection diversity applied to Rayleigh fading channels predicts an increase in the mean SNR of 1.7 dB and an increase in the SNR of 10.5 dB at the 99% reliability level, Figure 2.18, Section 2.9.4 (Ref: Jakes W C, 1974). Thus, the measured improvement is inferior to the theoretical prediction, but nonetheless provides a significant improvement over the no-diversity case. The discrepancy between theory and experiment does not arise from the correlation between the two inputs to the diversity combiner. Indeed, as stated in Section 4.4, the small negative correlation between the two time series should result in an enhanced performance in comparison with the combination of two time series with zero correlation. Instead, the discrepancy arises from the assumption of Rayleigh fading inputs to the combiner in the theoretical prediction. Rayleigh fading channels are characterized by a loss
deviation of 8.2 dB [1]. After two branch diversity combination, the loss deviation resulting from combination of two uncorrelated Rayleigh fading channels is reduced to 4.6 dB, Figure 2.18, Section 2.9.4 (Ref: Jakes W C, 1974). The loss deviation of the experimentally measured input channels to the diversity combiner (3.2B) is already smaller than that of the output of a two branch diversity combiner with Rayleigh fading inputs (4.6 dB). Therefore, it is not surprising that the simulated performance of the diversity combiner in increasing the 50% and 99% reliability levels is inferior to that of a theoretical system based on the combination of Rayleigh fading inputs. However, the discrepancy between the Rayleigh fading prediction and the simulation from experimental data is relatively small, 1.5 dB at the 99% reliability level.

The second example illustrated is for combination of two time series from a dual frequency experiment. The correlation between the time series is 0.4. Figure 4.28 illustrates the two input time series to the combination process. They are 3 second portions taken from a 100 MHz separation two frequency experiment carried out along Torrington Place, Location 4 in Figure 4.1. The time series are sampled at 5 KHz and decimated by a factor of 30 [2]. Thus, the maximum variation frequency represented on Figure 4.28 is 83 Hz. The correlation between the two time series is evident from visual examination. Each time series undergoes similar peaks and troughs. However, the deep fades in each time series do not coincide exactly. Thus, improvement in system performance is expected from combination of these time series.

1 Loss deviation is the spread between the 50% and 90% points on the small area cumulative distribution of the signal.

2 Every point in the figure is the average of 30 sampled data points.
Figure 4.28 Time Series Inputs to Simulated Diversity Combiner (100 MHz Frequency Separation Correlation Coefficient 0.4)
Figure 4.29 shows the output time series after the combination operation. Comparison between Figures 4.28 and 4.29 clearly shows that the majority of deep fades are suppressed by the combination operation even though there is a correlation of 0.4 between the two input time series. However, comparing Figure 4.29 with Figure 4.26, the combiner output time series produced from inputs with -0.3 correlation shows that the suppression of fading is reduced. This is expected from the positive correlation of 0.4 between the two input time series which are combined to produce Figure 4.29. Figure 4.30 shows the comparison between one of the input distributions to the combiner with its output distribution on a log-normal probability plot. Only one of the input distributions is illustrated because they are similar. The combination operation increases the level of the mean signal by 1 dB and increases the signal level which the data exceeds for 99% of the time by 7 dB. The loss deviation is reduced from 3.5 dB for the input
channels to 1.9 dB. This can be seen from the reduction in the spread between the 50% and 90% points on the cumulative distribution of the output in comparison with that of the input distribution, Figure 4.30.

Figure 4.30 Comparison of Simulated Diversity Combiner’s Input and Output Distributions (100 MHz Separation Between Input Channels, Correlation Coefficient 0.4)

Comparing Figure 4.30 with Figure 4.27, the combiner output distribution produced from combination of inputs with -0.3 correlation, also illustrates the reduction in the performance of the simulated diversity combiner when its inputs have a positive correlation. The shape of the input distribution is similar in both cases, but the slope of output distribution is significantly steeper when the input correlation is -0.3 than when it is 0.4. The increase in the mean signal level (~1 dB) is similar in both cases. However, the increase in the signal level at the 99% reliability level is reduced from 9 dB for the case where the correlation between the inputs is -0.3 and to 7 dB where the correlation between the inputs is 0.4.
A theoretical model, based on selection diversity applied to Rayleigh fading channels, predicts an increase in the mean SNR of 1.7 dB and an improvement in the SNR of 10.5 dB at the 99% reliability level for uncorrelated input channels, Figure 2.18, Section 2.9.4. The reduction in the enhancement to the signal at the 99% reliability level, as the correlation between the input channels increases, is of a similar order to that observed for the theoretical predictions based on Rayleigh fading inputs.

4.4.2 SUMMARY OF FIRST ORDER STATISTICS

The results show that the performance of a simulated diversity combiner operating on experimental data is similar to that predicted theoretically for a diversity combiner operating on Rayleigh fading channels. The discrepancy between the performances, 1.5 dB at the 99% reliability level, is small enough to allow a first order approximation of the performance of a diversity system to be made on the basis of the Rayleigh assumption. A more accurate prediction would have to account for the variation of the actual signal distribution at every location throughout the micro-cell. This procedure would be difficult to accomplish due to the variability of the distribution, evidenced by the results of the location variability measurements and loss deviation measurements in Sections 4.2.2.1 and 4.2.3.2 respectively, and it is beyond the scope of this Thesis.

Therefore, on the basis of the assumption of a Rayleigh fading model for the channel, it is possible to make first order approximations of the expected enhancement due to two branch diversity reception for space, frequency and polarization diversity systems using the measured value of the correlation coefficient between the respective sections of time series in question.

4.5 JOINT SECOND ORDER STATISTICS

This Section presents the results of the estimation of the correlation properties of the mobile radio channel from an analysis of the two frequency, two polarization and swept frequency measurements.

The aim of both two frequency measurements and swept frequency measurements is the estimation of the coherence bandwidth, $B_c$, and the delay spread, $\sigma_T$, of the channel. A direct measurement of either coherence bandwidth or delay spread would require a very wide bandwidth measurement to be made. Geometrical
analysis of expected path delays in Section 2.7 suggests coherence bandwidths in the range of 30 - 70 MHz, and 57 GHz urban experiments reported by Violette (Ref: Violette EJ et al, 1988) suggest coherence bandwidths in the range of 20 - 200 MHz in an urban environment with no moving vehicles. However, two frequency and swept frequency measurements circumvent the requirement for a very large bandwidth. The two frequency measurements build up a picture of the average behaviour of the channel by estimating the correlation coefficient between a number of sections of time series at discrete carrier frequency separations, whilst the swept frequency measurements provide a quasi-snapshot of the channel response over a time interval which is short in comparison with the movement of scatterers in the environment. The configuration of the experimental equipment used for these measurements is described respectively in Sections 3.6 and 3.9.

The polarization cross-correlation measurements provide an indication of the potential for polarization diversity at millimetre wave frequencies. Transmission of orthogonal polarizations with a small frequency separation allows the receiver to discriminate between the four components of the polarization matrix \( \Gamma \) in the micro-cell. The components are: \( \Gamma_{vv} \) vertical transmission and vertical reception; \( \Gamma_{hh} \) horizontal transmission and horizontal reception; \( \Gamma_{vh} \) vertical transmission and horizontal reception; and \( \Gamma_{hv} \) horizontal transmission and vertical reception. The configuration of the experimental equipment used for this measurement is described in Section 3.7.

4.5.1 TWO FREQUENCY MEASUREMENTS

The two frequency measurements are used to provide an estimate of the coherence bandwidth, \( B_c \), and delay spread, \( \sigma_T \), of the channel in two ways: 1) plotting the normalized correlation coefficient between sections of time series as a function of frequency separation, \( \Delta_f \), and estimating the average \( B_c \) from a smooth curve drawn through the scatter of points; 2) estimating the delay spread of the channel \( \sigma_T \) from individual two frequency measurements and using this to estimate the different correlation bandwidths observed within a micro-cell \( B_c \). With regard to the second method of estimating delay spread, there are two ways of estimating delay spread from a two frequency measurement which are described in Section 2.6.3. The first method estimates \( \sigma_T \) directly from the correlation coefficient. The second method forms the coherency function which is the analogue of the correlation function expressed as a ratio of the cross-spectrum and co-spectrum of the data, and then estimates \( \sigma_T \) from the coherency function.
4.5.1.1 Estimate of the Coherence Bandwidth from a Set of Two Frequency Measurements

As stated in Section 4.5, the introduction to this Section, the two frequency measurements provide an average picture of the coherence bandwidth of the channel. Figure 4.31 shows the plot of the correlation coefficient between the two square law detected signals from each two frequency measurement as a function of frequency separation. Each point on the graph is an estimate of the correlation coefficient obtained from two 1.6 second samples of data, at frequency separation Ω, each containing 16 K points. Three or more estimates of the correlation coefficient at each frequency separation are plotted on the Figure. In total, Figure 4.31 represents 37 sets of experimental data points. Two frequency measurements were carried out at frequency separations of: 15 MHz, 25 MHz, 37.5 MHz, 50 MHz, 62.5 MHz, 75 MHz, 100 MHz, 125 MHz, 150 MHz, 200 MHz, 250 MHz and 320 MHz. Although there is a scatter of points in Figure 4.31, there is a clear tendency for the correlation coefficient to fall as the frequency separation increases. Superimposed on the Figure is a smooth curve representing the visually estimated trend of the points. Appendix I contains an index of measurements presented in Figure 4.31.
The average coherence bandwidth, $B_c$, is estimated from the smooth curve superimposed on the experimental points graph by reading off the frequency separation corresponding to a value of 0.5 for the correlation coefficient. The estimated value of the average coherence bandwidth is 100 MHz. However, it is clear from the scatter of the experimental points on the curve that although the average coherence bandwidth of the channel is 100 MHz, the coherence bandwidth for a particular section of a time series can vary over a large range of values.
For example, at 62.5 MHz separation, there are correlation coefficients ranging in value from 0.45 to 0.88; the smallest value suggests that the correlation bandwidth is less than 60 MHz and the largest suggests a coherence bandwidth greater than 100 MHz. The delay spread $\sigma_T$ is estimated from the coherence bandwidth, $B_c$, using the relationship in Equation 2.85; ie $\sigma_T = 1/2\pi B_c$, where $B_c$ is in Hertz. Thus, a 100 MHz $B_c$ suggests a delay spread of 1.6 ns.

The average coherence bandwidth value of 100 MHz is greater than the value predicted by the deterministic model in Section 2.7 which is in the range of 30 MHz to 70 MHz. However, the deterministic model gave equal weight to the direct and specular reflected signals and this is unlikely to be the case. If a greater weight is given to the line of sight signal, then the predicted value of the coherence bandwidth will increase. This is the dual effect to that observed in Section 4.3.1 whereby the increase of the power in the line of sight component, in relation to the power in the randomly scattered component, caused an effective narrowing of the width of the Doppler spectrum leading to a broadening of the auto-correlation function. Here the relative strength of the line of sight component causes an effective narrowing of the spread of the time delay distribution which leads to a broadening of the frequency correlation function and, thereby, an increase in the coherence bandwidth $B_c$. The results of the first order statistics, summarized in Section 4.2.5, also support the presence of a strong line of sight signal.

4.5.1.2 Estimate of the Coherence Bandwidth from a Single Measurement of the Channel Response at Two Frequencies

The analysis of sections of individual time series from single two frequency measurements provides an alternative way of viewing the coherence bandwidth of the channel to that presented in Section 4.5.1.1. The scatter of the points in Figure 4.31, Section 4.5.1.1, indicates that the coherence bandwidth of the channel fluctuates from time to time and from location to location throughout the micro-cell. The approach followed in Section 4.5.1.1 illustrates the general trend of the correlation coefficient to decrease, averaged over all propagation conditions as frequency separation increases. From this trend the average value of the coherence bandwidth is estimated. Conversely, estimation of the coherence bandwidth from individual two frequency measurements gives the value of $B_c$ for individual sets for propagation conditions averaged over the length of the data samples taken from each measurement. This approach emphasizes the variability of the coherence bandwidth more strongly.
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It has been shown in Section 2.6.3 that the coherence bandwidth and delay spread of the channel may be estimated from a measurement of the channel response at two carrier frequencies with separation $\Delta$. The estimate relies on the assumption that the relationship between the delay spread and coherence bandwidth in Equation 2.85 is correct. The estimate is accomplished in two ways. However, both methods use the same input data which is used to produce Figure 4.27 in Section 4.5.1.1.

The equations which relate the squared envelope correlation coefficient and the value of the squared envelope coherency squared function to the coherence bandwidth, Equations 2.92 and 2.97, are respectively:

$$\hat{B}_c = \frac{1}{2\pi \sigma_T} = \Delta \left[ \frac{1}{\hat{\rho}} - 1 \right]^{\frac{1}{2}} \quad 2.92,$$

$$B_c = \frac{1}{2\pi \sigma_T} = \Delta \left[ \frac{1}{\sqrt{\gamma^2}} - 1 \right]^{\frac{1}{2}} \quad 2.97,$$

where: $\hat{B}_c = $ estimated coherence bandwidth, $\sigma_T = $ delay spread, $\Delta = $ frequency separation, $\hat{\rho} = $ estimated correlation coefficient, and $\gamma$ = estimated coherency squared.

In the first method, the average value of the correlation coefficient is calculated and substituted into Equation 2.92 to yield an estimate of the coherence bandwidth, $B_c$. In the second, the coherency squared function is evaluated and its average value is substituted into Equation 2.97 to yield an estimate of the coherence bandwidth.

The coherency function is obtained by first calculating the power spectra of the two time series and then calculating the cross spectrum between them. The coherency squared is the ratio of the squared modulus of the cross spectrum divided by the product of the two power spectra. The estimate of the coherency squared $\hat{\gamma}()$ [1] is approximately flat, as a function of the fading frequency $f$, until the power spectrum estimates fall close to the noise floor at the higher fading frequencies and errors become very large. The value of $\hat{\gamma}()$ used to make the estimate of $B_c$, is taken from the approximately flat part of the function at the lower fading frequencies.

---

1 The theoretical prediction is that the coherency function should be flat for all Doppler frequencies where the power spectra are non-zero, with a value equal to the square of the envelope correlation coefficient.
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The methods are similar from a theoretical standpoint. Table 4.5 summarizes the estimated values of $B_c$ and $\sigma_T$ from the two frequency measurements using both methods of estimation. Appendix I contains an index of the measurements presented in Table 4.5.

<table>
<thead>
<tr>
<th>Frequency Separation MHz</th>
<th>Correlation $\rho_r$</th>
<th>$B_c$ MHz</th>
<th>$\sigma_T$ ns</th>
<th>Coherency $\gamma^2$</th>
<th>$B_c$ MHz</th>
<th>$\sigma_T$ ns</th>
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</table>

Table 4.5 Estimates of Coherence Bandwidth and Delay Spread from Two Frequency Measurements (The correlation and coherency coefficients shown here are averages obtained from the data samples at each frequency separation which are used to produce the points illustrated in Figure 4.31)

Both methods give roughly similar estimates for the coherence bandwidth ($B_c$) and the delay spread ($\sigma_T$) at each frequency separation for frequency separations up to 100 MHz. These range from 55 MHz up to 153 MHz with corresponding delay spread estimates in the range of 1–2 ns. The differences between the estimates up to this frequency separation arise from noise problems in the estimates based on the coherency squared function. This estimation method is more susceptible to noise problems since it forms the ratio of two very small quantities which become progressively smaller at higher fading frequencies, as evidenced by the shape of the power spectra illustrated in Figure 4.23.

Above 125 MHz carrier frequency separation, the reduction in the system fading margin for the two frequency measurements, Section 3.6.1, gives rise to an increase in noise levels which dramatically increases the error in the estimates of coherence bandwidth from either of the two methods.
However, apart from the increase in the error with frequency separation, there does not appear to be any particular relationship between the frequency separation and the estimate of $B_c$ or $\sigma_f$ in Table 4.5. The results in Table 4.5 broadly agree with the estimation of the average coherence bandwidth ($B_c$) in Section 2.7 which was estimated from a visually fitted smooth curve through the scattered correlation coefficient points in Figure 4.31.
4.5.2 SWEPT FREQUENCY MEASUREMENTS

This Section presents the results of an initial investigation into the use of swept frequency probing on the 55 GHz channel link. A very simple experimental configuration, described in Section 3.9, which uses a spectrum analyser as a receiver and varactor swept Gunn oscillator as the transmitter is used for the measurements. Each experiment provides an estimate of the squared envelope of the channel transfer function, $T(f,t)$, described in Section 2.3, over a band of frequencies, $\Delta$, in the region of a particular time, $t'$. However, because the duration of each estimate is short in comparison with the time varying properties of the channel, it is assumed that the measurement represents the properties of the channel at a single instant of time $t'$. Having made this assumption, the measurements allow direct estimation of the correlation function of the channel versus carrier frequency separation for all separations from zero up to $\Delta$. The correlation versus carrier frequency is given by the auto-correlation, $R(\Delta)$, of the measured response of the squared envelope of the received power versus frequency. The delay distribution $W(T)$ is related to the auto-correlation $R(\Delta)$, by an inverse Fourier transform, Section 2.6, Equation 2.88. This relationship is the dual of that between the auto-correlation function $R(\tau)$ and the Doppler power spectrum $S(f)$, Section 2.6. The coherence bandwidth, $B_c$, and delay spread, $\sigma_\tau$, are easily estimated from $R(\Delta)$ and $W(T)$. The swept frequency technique for wideband probing of the mobile radio channel in the 400 - 900 MHz bands has been reported widely in the literature, for example Ref: Molkdar D, Mathews P A, 1988.

Figure 4.32 illustrates three typical, un-processed, experimental results. The vertical axis on the Figure is calibrated in relative power in dB. The horizontal axis is the frequency sweep axis of the spectrum analyser output. The trace consists of repeated sweeps across a 66 MHz bandwidth interspersed with dead periods while the trace returns to the sweep origin.

During the measurement, the mobile vehicle is stationary and the sweep duration of 0.1 second is short compared to the rate of change of the received signal due to the movement of scatterers. Thus, the properties of the channel can be assumed to be constant for the duration of the sweep and each sweep is equivalent to a snapshot of the squared envelope of the time dependent channel transfer function, $T(f,t')$, Section 2.3.
Figure 4.32 Three Examples of the Received Power Versus Frequency Characteristic Along Gordon Square, Location 1 in Figure 4.1 (Swept Frequency)
The three traces illustrated in Figure 4.32 illustrate different aspects of the channel characteristic. In the first sweep, the power shows a sloping characteristic as the swept frequency increases with no deep nulls in the trace. This indicates that the received signal is being influenced by a delay time which is substantially shorter than the reciprocal of the sweep bandwidth (15 ns) and that the coherence bandwidth during these sweeps will be greater than 66 MHz. However, moving to the left of the Figure, the average signal power is decreasing and nulls start to appear with periods less than 66 MHz. This suggests that the line of sight signal may be decreasing and that signals with larger delays are having greater influence on the signal which will result in a decrease in coherence bandwidth. The second trace shows a characteristic which is essentially flat on the right of the Figure, but which exhibits successively deeper fades becoming more like an interference pattern moving to the right of the Figure. The average power is quite large and is approximately the same for all of the traces in the Figure which suggests that the line of sight signal is approximately equal for all the traces. This suggests that the signal power in specular reflected components with delay times of the order of 15 ns is increasing towards the left of the trace which will bring a decrease in coherence bandwidth from greater than 66 MHz to approximately 20 MHz (estimated from the delay corresponding to the frequency separation between nulls $B_c = 1/2\pi\sigma_f$) towards the right of the trace. The third trace shows similar behaviour to the second, the traces exhibits a ripple which is small in magnitude compared to the average magnitude with a period of approximately 20 MHz with a magnitude which increases towards the right of the Figure.

4.5.2.1 Frequency Correlation Function

A single swept frequency measurement provides an estimate of the power versus frequency characteristic for all frequencies in the range $f_c$ to $f_c + \Delta$; where $f_c =$ carrier frequency and $\Delta =$ maximum sweep deviation. The auto-correlation function of the swept frequency power response provides an estimate of the envelope squared correlation function for all frequency separations from zero up to the maximum sweep deviation $\Delta$.

A large number of the swept frequency power response estimates are either flat over the 66 MHz bandwidth of the sweep or have a sloping characteristic. This suggests that the received signal is dominated by signals with a delay spread of less than $1/66$ MHz second (15 ns) which results in the frequency separation of the major nulls in the swept frequency measurement being spaced further than 66 MHz apart, Figure 4.32a. This indicates that the coherence bandwidth at the
particular time $t'$ at which the sweep was obtained is greater than 66 MHz. However, processing a sweep which is flat or has a linear slope across the sweep bandwidth does not yield any more information than this. Therefore, a sweep which is not flat across the 66 MHz bandwidth is selected for analysis. This allows an estimate of the coherence bandwidth at that time (location) to be made. The trace selected is the last complete sweep in Figure 4.32b. Although the sweep selected is atypical, it provides information about the lower limit of coherence bandwidth. The sweep was taken from an experimental run along Gordon Square about 100 m distant from the transmitter site at Location 1 in Figure 4.1.

![Figure 4.33 Correlation Coefficient Versus Frequency Separation (Swept Frequency)](image)

Figure 4.33 is the auto-correlation function of the swept frequency data sample indicated in Figure 4.32a. The coherence bandwidth, $B_c$, of the channel estimated from the 0.5 coherence value of Figure is 17 MHz which implies a delay spread, $\sigma_T$, of 9.4 ns. This coherence bandwidth is lower than the value estimated from the two frequency measurements which is in the range 80 - 150 MHz. However, the received power in Figure 4.33a is relatively high which indicates that the LOS path is not obstructed. Also, the geometry of the path at the location where the measurement was made, close to a high elevation base station, makes it unlikely
that there would be any object present to block the LOS path. Therefore, the small coherence bandwidth suggests that there are specular components present in the received signal distribution which are approximately equal in strength to the LOS component. The estimated standard deviation of delays of 9.4 ns indicates that the path difference between the direct and reflected ray is 5.6 m on the assumption that the power is equally distributed between a LOS component and a single specular reflected ray. This is of a similar order to the path differences predicted from geometrical calculations based on the micro-cell's dimensions in Section 2.7.2 which are in the range 1.5 - 3 m. The difference between the two distances may be explained by the greater transmitter antenna elevation in this case (40 m) compared with that assumed for the geometrical predictions in Figure 2.12, Section 2.7 (15 m).

The sample waveform used to calculate the estimate is not typical of all the observed waveforms. A large number of the waveforms are flat across the whole 66 MHz bandwidth of the sweep, indicating a coherence bandwidth of more than 66 MHz, Figure 4.32b. However, the sweep bandwidth is too small to enable an accurate estimate of coherence bandwidth to be made from these results. Thus, whilst the results show that at some locations within the micro-cell the coherence bandwidth is as small as 17 MHz, examination of the general trend of the power versus frequency characteristic suggests that the typical coherence bandwidth is greater than 66 MHz.
4.5.2.2 Delay Distribution Function

The delay distribution function, $W(T)$, is equal to the inverse Fourier transform of the frequency correlation function, $R(\Delta)$, with respect to frequency separation $\Delta$, Section 2.6.1, Equation 2.88. Figure 4.34 is the estimate of the delay profile $W(T)$ obtained from the inverse Fourier transform of the auto-correlation function shown in Figure 4.33.

![Figure 4.34 Estimated Delay Distribution W(T) (Swept Frequency)](image)

The delay spread $\sigma_T$, is the standard deviation of the delay distribution $W(T)$. The estimated value of $\sigma_T$, calculated from the data set which is plotted in Figure 4.34, is approximately 15 ns. This is at the limit of the resolution of the measurement defined by the reciprocal of the sweep bandwidth. This estimate is larger than that estimated from the two frequency measurements in Section 4.5.1 (1.6 ns). However, as stated in Section 4.5.2.1, the data segment analysed to produce this estimate is atypical. It was chosen because it appeared that it would yield an estimate within the resolution of the swept frequency measurement. Examination of a large number of the power versus swept frequency characteristics shows that they are flat across the 66 MHz sweep bandwidth, an example is shown in Figure 4.32b. This indicates that the delay spread is often less than 2.4 ns. It is not
possible to produce an accurate estimate of either delay spread or coherence bandwidth when the sweep is flat over the measurement bandwidth. However, if it is assumed that the average distance between nulls in the swept frequency measurement is twice the sweep bandwidth, an assumption which is compatible with the observation of a flat sweep, then it indicates a coherence bandwidth of approximately 130 MHz and a delay spread of 1.2 ns. These values are in accord with the estimates presented in Section 4.5.1.

4.5.3 SUMMARY OF COHERENCE BANDWIDTH AND DELAY SPREAD ESTIMATION

The results of the coherence bandwidth and delay spread estimates are presented in two formats. In the first, all of the correlation coefficients obtained from the two frequency measurements are plotted as a function of frequency separation and a smooth curve representing the average behaviour in all propagation conditions is drawn through the scattered points. In the second, the average correlation coefficient from each experimental run at a particular frequency separation is used to provide an estimate of the channel coherence bandwidth at that location. The estimate of the coherence bandwidth from the swept frequency measurements falls into this second category. However, the swept frequency estimate represents the "instantaneous" value of the coherence bandwidth rather than an estimate over 1.6 seconds or more of sampled data.

The results confirm the suggestion that the mobile radio channel is dominated by a strong, but highly variable, line of sight component. The strength of the LOS component is evidenced by the very large average coherence bandwidth of 100 MHz estimated from the two frequency measurements in Section 4.5.1.1. Its variability is evidenced by the scatter of the correlation coefficient estimates in Figure 4.31, by the different estimates of coherence bandwidth in Table 4.5 and by the small coherence bandwidth estimate from the swept frequency. In addition, the delay spread of 1.6 ns which corresponds to the estimated average coherence bandwidth of 100 MHz would suggest an average path difference of 45 cm if the distribution of scattered radiation at the antenna were uniform. Clearly this is incompatible with the range of average path delays estimated from geometrical considerations in Section 2.7 which range from 0.7 m to 1.5 m. Thus, the simplest explanation is that the path delay is dominated by a very strong, but highly varied line of sight component.
4.5.4 DUAL POLARIZATION MEASUREMENTS

A variety of measurements were made of the correlation coefficient between the squared envelopes of two signals transmitted on orthogonal polarizations. The initial experimental results were obtained using two standard gain horns mounted a few wavelengths apart. The results show that the received envelopes for vertical and horizontal polarization are un-correlated. However, it is not apparent what portion of the de-correlation effect is due to the separation of the antennae, the different E and H-plane radiation patterns of the antenna or the small difference in bearing of each antenna. To eliminate the possibility that the de-correlation is due to these effects, an antenna capable of simultaneously transmitting two orthogonal polarizations and with equal E and H-plane radiation patterns was designed and built. The design is described in Section 3.4.3.

The transmission of two signals on orthogonal polarizations gives rise to four possible components at the receiver. These may be described in terms of a polarization matrix \( \Gamma \) which describes the power in each component.

\[
\Gamma = \begin{pmatrix}
\Gamma_{vv} & \Gamma_{vh} \\
\Gamma_{hv} & \Gamma_{hh}
\end{pmatrix}
\]

where: \( \Gamma_{vv} \) describes the coupling coefficient from vertical transmission to vertical reception, \( \Gamma_{hh} \) describes the coupling coefficient from horizontal transmission to horizontal reception and the two terms \( \Gamma_{hv} \) and \( \Gamma_{vh} \) are the cross coupling components respectively between horizontal transmission and vertical reception.

The experimental configuration used for the polarization measurements allows all four components to be distinguished and recorded separately, Section 3.7.

The first experiment described here summarizes the results of a measurement of the four components polarization matrix \( \Gamma \) and estimation of the cross-correlation between the coupling terms \( \Gamma_{vv}, \Gamma_{hh} \) and the coupling terms \( \Gamma_{vh}, \Gamma_{hv} \).

The second experiment described here uses the observation from the first experiment that \( \Gamma_{vv} \gg \Gamma_{vh} \) and \( \Gamma_{hh} \gg \Gamma_{hv} \) to measure the cross-correlation between \( \Gamma_{vv} \) and \( \Gamma_{hh} \) in a simpler manner than that required to measure all four components of the polarization matrix \( \Gamma \). This approach allows a large number of measurements to be processed in the propagation environment along Torrington Place, Location 4, Figure 4.1a, and along High Holborn, Location 5, Figure 4.1b.
4.5.4.1 Measurement of the Polarization Matrix $\Gamma$

The polarization matrix is made up of the four power coupling components $\Gamma_{vv}$, $\Gamma_{vh}$, $\Gamma_{hv}$, and $\Gamma_{hh}$ which were described in Section 4.5.3. These measurements required a relatively sophisticated technique to discriminate between the pairs of components $\Gamma_{vv}$, $\Gamma_{vh}$, and $\Gamma_{hv}$ which are received on the same frequency. The pairs of components are separated in time by the sampling operation of the polarization switch during data collection. To separate them during data acquisition they must be sampled using a clock which is synchronous with the polarization switching signal. To enable this to be done, the polarization switching frequency is recorded onto the same magnetic tape as the data signals during data collection. Sampling is accomplished using a phase locked loop locked onto twice the recorded polarization switch rate. The phase locking operation is accomplished by a Hewlett Packard Function Generator type 3314A.

Figures 4.35a and 4.35b illustrate the recorded time series from a typical measurement of the four components of the polarization matrix $\Gamma$. The four traces in Figure 4.35 represent an 8 second portion of a measurement along Torrington Place, Location 4 in Figure 4.1. The top trace in Figure 4.35a is the $\Gamma_{vv}$ coupling term which represents the power transmitted and received on vertical polarization. The bottom trace in Figure 4.35a is the $\Gamma_{vh}$ coupling term which represents the power transmitted on vertical polarization and received on horizontal polarization. Comparison of the traces in Figure 4.35a shows that the cross-coupling term is 16-20 dB smaller than the co-coupling term. Similar behaviour is observed from comparison of the traces in Figure 4.35b, the $\Gamma_{hv}$ and $\Gamma_{hh}$ coupling terms, which shows that the cross coupling term is 12-14 dB smaller than the co-coupling term. Four other sets of results (Reference 1(4P)-G to 4(4P)-G Appendix I) also showed this behaviour; in particular, $\Gamma_{vv} \gg \Gamma_{vh}$ and $\Gamma_{hh} \gg \Gamma_{hv}$.
Figure 4.35a Comparison of Experimentally Measured Vertical to Vertical Polarization Coupling Coefficient with the Vertical to Horizontal Polarization Cross-Coupling Coefficient
Figure 4.35b Comparison of Experimentally Measured Horizontal to Horizontal Polarization Coupling Coefficient with the Horizontal to Vertical Polarization Cross-Coupling Coefficient
The observation that $\Gamma_{vv} \gg \Gamma_{vh}$ and $\Gamma_{hv} \gg \Gamma_{hv}$ suggests that the proportion of the received signal, due to multiple scattering, is very small. If the influence of multiple scattering were greater, and the polarization vectors had been completely randomized, the distribution of power between the four coupling components would be more even.

The calculated value of the cross-correlation coefficient between the data in the top traces of Figures 4.35a and 4.35b, i.e., between $\Gamma_{vv}$ and $\Gamma_{hv}$, is -0.259. The average value of the correlation coefficient from the four sets of results analysed is -0.097. This strongly suggests that $\Gamma_{vv}$ and $\Gamma_{hv}$ are de-correlated and confirms that the argument of Lee, presented in Section 2.6.5 which explains the depolarization of the co-polar coupling coefficients in terms of the different reflection coefficients exhibited by surfaces to waves incident on horizontal and vertical polarizations, may be applied in a micro-cell. The calculated value of the cross-correlation coefficient between the data in the bottom traces Figures 4.35a and 4.35c, i.e., between the polarization cross coupling terms $\Gamma_{vh}$ and $\Gamma_{hv}$, is 0.001. The average value of the correlation coefficient from the four sets of results is 0.043. These values strongly suggest that the two cross coupling terms are also uncorrelated.

In Section 2.6.5 an argument, reproduced from Lee (Ref: Lee W C Y, 1982), is put forward to predict the expected correlation properties of the channel for the transmission of orthogonal polarizations. Lee uses the expected differences between the reflection coefficients for waves incident on surfaces with polarization vectors horizontal and vertical with respect to those surfaces to predict that the co-polar coupling coefficients $\Gamma_{vv}$ and $\Gamma_{vh}$ will be uncorrelated. An argument is also put forward to suggest that the two cross-coupling coefficients $\Gamma_{vh}$ and $\Gamma_{hv}$ will also be uncorrelated. The argument is based on the assumption that multiple reflections are needed to produce cross coupling terms (Ref: Ruck G T et al, 1970) and that scattering from vertical polarization to horizontal polarization will usually arise from different scattering events to the converse process. However, the model is based on assumption of a uniform distribution of scattered waves at the mobile which is clearly not the case given the results of the first and second order statistics summarized in Sections 4.2.5 and 4.3.3 respectively. Thus, experiments must be conducted to see whether these arguments may be applied to propagation within a micro-cell.

Comparison of the experimentally measured cross-correlation coefficient between $\Gamma_{vv}$ and $\Gamma_{hv}$, and between $\Gamma_{hv}$ and $\Gamma_{vh}$, with Lee's measured results reported at
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900 MHz, shows that similar behaviour is observed at 55 GHz in a micro-cell to that which is observed in a conventional cellular environment at 900 MHz (Ref: Lee W C Y, Yeh Y S, 1975).

Comparison of the bottom trace in Figures 4.35a with that in Figure 4.35b, ie comparing $\Gamma_{vh}$ with $\Gamma_{hv}$, suggests that $\Gamma_{hv} \approx \Gamma_{vh}$. However, the difference between the mean level of the received signal in $\Gamma_{vv}$ and $\Gamma_{hh}$ suggests that a large part of this difference is due to different transmission powers on each polarization. Thus, a more reasonable estimate of the difference in relative power between the cross coupling coefficients would be to compare the difference between $\Gamma_{vv}$ and $\Gamma_{vh}$ with that between $\Gamma_{hh}$ and $\Gamma_{hv}$. This process indicates that $\Gamma_{hv}$ is approximately 4 dB greater than $\Gamma_{vh}$. The observed behaviour at 55 GHz that $\Gamma_{hv} > \Gamma_{vh}$ is in agreement with the observation of channel behaviour in a conventional cell at 900 MHz. From experiments carried out at 900 MHz, Lee observed that the cross coupling component $\Gamma_{hv}$ is considerably greater than the cross-coupling component $\Gamma_{vh}$ (Ref: Lee W C Y, Yeh Y S, 1974). He ascribed this difference to the preponderance of vertical scattering surfaces in an urban environment where a large part of the scattering takes place from the vertical sides of buildings, an argument which is also valid for scattering in a micro-cell.

4.5.4.2 Measurement of the Correlation between $\Gamma_{vv}$ and $\Gamma_{hh}$

The second experiment uses the observed result of the first experiment that $\Gamma_{vv} \approx \Gamma_{vh}$ and $\Gamma_{hh} \approx \Gamma_{hv}$ to simplify the analysis procedure undertaken on the data from the polarization measurements. The two components $\Gamma_{vv}$ and $\Gamma_{hv}$ are added together, as are the two components $\Gamma_{hh}$ and $\Gamma_{hv}$. The two resulting data signals are then sampled by a computer controlled clock which is asynchronous with the polarization switching clock. The two signals which would usually be received by a polarization diversity receiver are $\Gamma_v = \Gamma_{vv} + \Gamma_{hv}$ and $\Gamma_h = \Gamma_{hh} + \Gamma_{vh}$, ie the two components on equal polarization are added together. However, with the system described here, it is the components received on the same frequency [1] which are added together, ie $\Gamma_v = \Gamma_{vv} + \Gamma_{vh}$ and $\Gamma_h = \Gamma_{hh} + \Gamma_{hv}$. Thus, the order of the cross polar components is switched. However, since $\Gamma_{vv} \approx \Gamma_{vh}$ and $\Gamma_{hh} \approx \Gamma_{hv}$ this does not result in large errors and the two signals are representative of the signals which would be received by a receiver using a polarization diversity system.

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1 The vertically and horizontally polarized signals at the transmitter are transmitted on two slightly different carrier frequencies.
A series of measurements were conducted along Torrington Place and High Holborn, Locations 4 and 5 respectively, in Figures 4.1a and 4.1b. Observation of the time series from the two polarization experiments on a dual trace oscilloscope and of chart recorder traces show that the correlation between the signals received on orthogonal polarizations is very low which is in accord with the estimation of the polarization matrix $\Gamma$ in Section 2.6.5. Table 4.6a shows the results of 16 estimates of the polarization cross-correlation coefficient along Torrington Place.

<table>
<thead>
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<th>Correlation Coefficient</th>
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</tr>
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<td>9(P)-G</td>
<td>0.471</td>
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<tr>
<td>2(P)-G</td>
<td>-0.021</td>
<td>10(P)-G</td>
<td>0.228</td>
</tr>
<tr>
<td>3(P)-G</td>
<td>-0.310</td>
<td>11(P)-G</td>
<td>-0.008</td>
</tr>
<tr>
<td>4(P)-G</td>
<td>-0.224</td>
<td>12(P)-G</td>
<td>-0.043</td>
</tr>
<tr>
<td>5(P)-G</td>
<td>-0.077</td>
<td>13(P)-G</td>
<td>-0.041</td>
</tr>
<tr>
<td>6(P)-G</td>
<td>-0.028</td>
<td>14(P)-G</td>
<td>0.110</td>
</tr>
<tr>
<td>7(P)-G</td>
<td>-0.219</td>
<td>15(P)-G</td>
<td>-0.012</td>
</tr>
<tr>
<td>8(P)-G</td>
<td>-0.106</td>
<td>16(P)-G</td>
<td>-0.080</td>
</tr>
</tbody>
</table>

Table 4.6a Estimates of Cross-Correlation Coefficient between Signals Received on Orthogonal Polarizations along Torrington Place

Table 4.6b shows the results of 14 estimates of the polarization cross-correlation coefficient along High Holborn. An index of the results presented in Tables 4.6a and 4.6b is presented in Appendix I.

<table>
<thead>
<tr>
<th>Experiment Number</th>
<th>Correlation Coefficient</th>
<th>Experiment Number</th>
<th>Correlation Coefficient</th>
</tr>
</thead>
<tbody>
<tr>
<td>1(P)-H</td>
<td>0.011</td>
<td>8(P)-H</td>
<td>0.136</td>
</tr>
<tr>
<td>2(P)-H</td>
<td>0.192</td>
<td>9(P)-H</td>
<td>-0.008</td>
</tr>
<tr>
<td>3(P)-H</td>
<td>-0.059</td>
<td>10(P)-H</td>
<td>0.045</td>
</tr>
<tr>
<td>4(P)-H</td>
<td>-0.021</td>
<td>11(P)-H</td>
<td>0.133</td>
</tr>
<tr>
<td>5(P)-H</td>
<td>-0.178</td>
<td>12(P)-H</td>
<td>-0.127</td>
</tr>
<tr>
<td>6(P)-H</td>
<td>-0.001</td>
<td>13(P)-H</td>
<td>-0.140</td>
</tr>
<tr>
<td>7(P)-H</td>
<td>-0.152</td>
<td>14(P)-H</td>
<td>0.067</td>
</tr>
</tbody>
</table>

Table 4.6b Estimates of Cross-Correlation Coefficient between Signals Received on Orthogonal Polarizations along High Holborn

The average of the correlation coefficient estimates along Torrington Place is -0.005 and the average of the correlation coefficients along High Holborn is 0.006.
Chapter 4

The range of correlation coefficients along Torrington Place is quite high, from -0.43 to +0.47, in comparison with the range of correlation coefficients along High Holborn, -0.178 to 0.192. The difference between the two ranges of values in the two environments is probably due to the difference in propagation conditions between the two environments. The initial part of the run, in Torrington Place, as the mobile moves away from the transmitter sited on the fourth floor of the Warburg Institute, is over a relatively large open area which is used as a car park. This provides relatively few scatterers, particularly when the mobile is near to the transmitter site. In contrast, along High Holborn, there are a large number of scatterers near the mobile unit even at small transmitter receiver separations. This is evidenced by the photograph in Figure 4.5 which shows a large number of vans, lorries, lampposts, etc. The number of scatterers is important for determining correlation properties because if signals are transmitted with no obstruction or interference there is no clear mechanism which will give rise to polarization de-correlation. In contrast, the presence of a large number of scatterers [1] gives rise to a clear mechanism for polarization de-correlation. The model put forward by Lee (Ref: Lee W C Y, 1982) assumes a large number of scattered components are present and that the de-correlation between orthogonal polarizations arises due to the different reflection coefficients experienced by each component. If only a few components are present, Lee’s model implies that the components received on orthogonal polarizations will not necessarily be totally de-correlated.

4.5.5 SUMMARY OF DUAL POLARIZATION MEASUREMENTS

The results are in agreement with the model put forward from Lee based on the observation of channel properties at 900 MHz (Ref: Lee W C Y, Yeh Y S, 1974) which predicts that the orthogonal polarizations of the received signals will be un-correlated. The results also agree with the observation of Lee from measurements at 900 MHz that the two components $\Gamma_{vh}$ and $\Gamma_{hv}$ are un-correlated and that $\Gamma_{hv} > \Gamma_{vh}$ (a difference of approximately 4 dB is observed in measurements at 55 GHz). Lee explains the difference between the two coupling coefficients in terms of the preponderance of vertical scatterers in an urban environment.

The results of the measurement of the polarization matrix $\Gamma$ show that the co-polar polarization coupling coefficients $\Gamma_{vv}$ and $\Gamma_{hh}$ are much bigger than the cross coupling coefficients $\Gamma_{vh}$ and $\Gamma_{hv}$ by respectively 16 - 20 dB and 12 - 14 dB. This

---

1 Each with different reflection coefficients for vertically and horizontally polarized waves.
suggests that multiple scattering plays only a small rôle in micro-cellular propagation because cross coupling polarization results from multiple events (Ref: Ruck G T et al, 1970).

Comparison of experimental results obtained in Torrington Place with those obtained in High Holborn show that the variation of the cross-correlation coefficient is higher in Torrington Place. The difference is ascribed to the relative paucity of scatterers in Torrington Place in comparison with the numbers present in High Holborn particularly close to the base station transmitter site. This results in the partial failure of the depolarization mechanism which relies on a received signal composed by the addition of a number scattered components.

### 4.6 SUMMARY OF RESULTS

Analysis of the results suggests four principal observations:

i) The buildings in the mobile radio environment act as natural boundaries to the urban micro-cell. This is indicated graphically by the contour map of the propagation loss in the street surrounding the College, Figure 4.13, Section 4.2.1.5.

ii) The power law which governs the general trend of the signal to fall with distance is $r^{-3.6}$. This lies between the expected value for free space propagation and propagation over a plane earth, Section 4.2.1.4.

iii) The depth of fades of the average signal level throughout a micro-cell 300 m in length is less than 40 dB.

iv) The statistical parameters which describe the signal are extremely varied from location to location and from time to time throughout the micro-cell. A large part of this variation is ascribed to the essential line of sight propagation conditions encountered in the micro-cell and to the various obstructions which modulate the strength of the line of sight component at the mobile terminal.

The analysis of the statistics of the signal is undertaken over three different scale sizes in order to deal with the non-stationary nature of the mobile radio channel's properties. Variations over the largest scale size describe the general trend of the signal level to fall with increased transmitter receiver separation (this is summarized in ii above). Variations over the medium scale size describe the distribution
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of the local mean signal level throughout the micro-cell. Finally, variations on the smallest scale size describe the distribution of the "fast fading" envelope of the received signal.

The distribution of the local mean signal strength is approximately log-normally distributed. The distribution has a location variability [1] which ranges from 1 dB to 7.2 dB throughout the micro-cell with an average value of 3.2 dB. The distribution of the "fast fading" envelope of the signal is highly variable throughout the micro-cell. The value of its loss deviation [2] varies from 0.9 dB to 9.3 dB with a mean value of 4.7 dB. Therefore, it is generally not well described by a Rayleigh distribution [3]. These results show a broad similarity with measurements reported at 900 MHz in a low density urban environment (Ref: IEEE VT Special Issue, 1988). This indicates that the propagation in each environment is similar in both cases; that is, it exhibits a very high probability of a line of sight path combined with a relatively small number of scatterers.

The results of both auto-correlation and power spectrum measurements on the squared envelope of the received signal support the evidence from the first order statistics that the received signal contains a substantial, but highly variable, line of sight component. In addition, they suggest that the angular distribution of the randomly scattered part of the received signal is not uniform in azimuth.

In the context of a space diversity system, the auto-correlation measurements indicate that a separation between receiver antennae mounted on the mobile unit of between \(2.5\lambda\) and \(22\lambda\) is required to provide un-correlated inputs. However, even a separation of \(22\lambda\) only equates to a 12 cm separation between the diversity antenna on the mobile unit which is still feasible in a hand-held system.

It is suggested that the estimated antenna separation varies between different locations in the micro-cell due to changes in the strength of the LOS component in relation with the strength of the randomly scattered component of the received signal. A simple model, based on this assumption, is applied to the experimental data to estimate the ratio of the power in the LOS component to that in the scattered component. The ratio varies from 0.6 up to 78 and underlines the essential line of sight nature of propagation within a micro-cell.

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1 Location variability is the standard deviation of the log-normal distribution; the 50% to 84% spread on the cumulative distribution of the local mean.

2 Loss deviation is defined as the 50% to 90% spread of the cumulative distribution of the small area signal distribution.

3 The Rayleigh distribution is the common assumption for the distribution of the mobile radio signal statistics, it is characterised by a loss deviation of 8.2 dB.
The experimental power spectra are intermediate between the theoretical predictions of Clarke and Aulin (Ref: Clarke RH, 1968; Aulin T, 1979) and those of a simple deterministic model based on the interference of a direct ray with reflections from principal scatterers in the micro-cell. This suggests that whilst the angular distribution of received waves at the mobile terminal is not uniform in azimuth, it is broader than a model based on specular reflection would suggest. Comparison of experimental spectra with the Clarke and Aulin models suggests that the received waves are horizontally distributed at the receiver when the transmitter and receiver antenna are at the same elevation and that it has a significant vertical angle of arrival distribution when the elevation of the transmitter antenna is greater than that of the receiver.

The potential for diversity reception in the micro-cellular mobile radio environment is investigated by simulating a simple two branch diversity system using measured data. The results of the simulation are compared with theoretical predictions based on the assumption of Rayleigh fading channels (Ref: Jakes W C, 1974). It is found that even though the input channels are not Rayleigh distributed, the enhancement predicted by the simulation is similar to that for the Rayleigh fading channels. The discrepancy between the performances at the 99% reliability level [1] is 1.5 dB. This is small enough to allow first order approximations of system performance to be made using a simple Rayleigh fading model. A more accurate prediction would require extensive computer simulation and is beyond the scope of this work. The results suggest that diversity systems based on spatial separation of antennae, frequency separation and on orthogonal polarization all have significant potential for improving received signal quality at a mobile terminal in a micro-cell.

The average correlation bandwidth of the channel is 100 MHz, estimated from scatter of the correlation coefficient estimates versus carrier frequency separation in Figure 4.31, Section 4.5.1.1. However, the correlation bandwidth is highly varied throughout the micro-cell with values ranging from 17 MHz to 153 MHz. The delay spread values (σₜ) corresponding to the coherence bandwidths of 17 MHz and 153 MHz are 9.4 ns and 1 ns respectively. The respective average path length in each case is 2.8 m and 45 cm. The 45 cm path length is incompatible with average path length estimates based on micro-cell geometry (0.7 m to 1.5 m), but the 2.8 m path delay is compatible with these figures given the greater transmitter height used experimentally than assumed for the model (40 m as

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1 The 99% reliability level is the signal level which is exceeded for 99% of the time.
opposed to 15 m). Thus, the results suggest that the high coherence bandwidth values correspond to large LOS signal strength which dominates the delay spread giving an apparently very small average path difference.

It is proposed that the spread in coherence bandwidth values is due to variation in the strength of the LOS component. The 17 MHz value appears to be the result of the delay spread of the received signal being determined by interference of a LOS component with a specular reflection of comparable amplitude, whilst the 153 MHz value is the result of a delay spread dominated by a strong LOS component. This result agrees with that of Violette (Ref: Violette et al, 1988) who conducted 57 GHz propagation experiments, using narrow beamwidth antennae, in an urban environment with no other vehicles present, and reported typical bandwidths of 200 MHz which were reduced to 10–20 MHz during fades in the received signal strength.

Together these observations provide strong evidence that the predominant propagation mode in an urban micro-cell is by line of sight. The variability of the coherence bandwidth throughout the micro-cell makes the choice of the maximum bandwidth which the mobile link could handle without the occurrence of inter-symbol interference, and the minimum frequency separation which would be required by a frequency diversity system, difficult to estimate. The actual frequency chosen would depend on the type of system implemented on the link. In particular, diversity reception reduces the apparent depth of fading in a system which tends to increase the coherence bandwidth, since the bandwidth tends to be lower in deep fades where signal components with longer delay times have a greater influence on the signal.

Comparison of the results of the orthogonal polarization measurements with results reported by Lee show that similar behaviour is observed at 55 GHz as that which is observed at 900 MHz (Ref: Lee W C Y, Yeh Y S, 1974). In particular, the two co-polar polarization coupling components $\Gamma_w$ and $\Gamma_H$ are shown to be uncorrelated, the two cross-polar coupling components $\Gamma_{VH}$ and $\Gamma_{HV}$ are shown to be uncorrelated, and $\Gamma_{HV} > \Gamma_{VH}$. The range in the values of the cross-correlation coefficient along Torrington Place suggests that the mechanism responsible for producing the de-correlation starts to break down when there are few scatterers present. However, all the correlation coefficient estimates are less than 0.5. Therefore, even in the worst case with a cross coupling coefficient of approximately 0.5, a diversity system based on orthogonally polarized inputs would be expected to yield an improvement in the signal level at the 99% reliability level of approximately 7 dB.
Chapter 5 DIGITAL THEORY

5.1 INTRODUCTION

There is a fundamental limitation in assessing the behaviour of a communication channel using a particular experimental system, that is, the experimental results of the measurements are specific to that one system. This Chapter addresses the difficult question of how to generalize the results from one experimental system, with known imperfections, to predict the results which would be obtainable using an "ideal" system. "Ideal" is used here to indicate a system with no imperfections, using a matched filtered phase shift keyed (PSK) modulation system [1]. The analysis assumes that the noise on the mobile radio channel is Gaussian distributed and that the fading envelope of the signal is Rayleigh distributed. A further simplification is made that the recovery of the timing information is perfect. In effect, this means that errors are assumed to occur only when the signal becomes indistinguishable from noise. This assumption is justified because it permits a simple analysis to be made. A full analysis, including timing errors, is outside the scope of this Thesis. In practice, there are complex non-linear effects which occur due to the apparent "threshold" behaviour of the phase locked loops used for timing recovery and binary "word" recovery (Ref: Gardiner FM, 1979). In addition, a threshold rather than matched filter detection is assumed. These effects are discussed in Section 6.4 with reference to the differences between the predicted and the measured performances of the digital receiver.

The primary aim of the analysis is to obtain predictions of bit error rate (BER) for a given average signal to noise ratio (SNR) for a variety of de-modulation schemes when the received signal envelope has a Rayleigh amplitude distribution. These predictions are used in Chapter 6 to assess the performance of the receiver in the laboratory. In Chapter 7, the predictions are used to assess the performance of the link in the mobile radio environment and, in conjunction with the analogue results in Chapter 4, to make approximate predictions of the improvement in system performance which would result from implementing two branch diversity on the link.

1 The PSK system provides the best obtainable error rate for a given SNR.
5.2 BIT ERROR RATE VERSUS SIGNAL TO NOISE RATIO

This Section presents an analysis of the expected BER for a given received signal level using a non-coherent FSK decoding scheme with a set decision threshold. The results are compared to those predicted for other detection schemes. The analysis is based on the probability that noise will cause the receiver to make an incorrect decision as to whether a transmitted bit is either a 0 or a 1. Section 5.2.1 is concerned with the case of a flat fading situation, i.e., in the absence of multipath propagation. Section 5.2.2 considers the case of a mobile fading environment where the received signal is subject to random fades described by a Rayleigh statistical density function. The Rayleigh distribution function is not valid for all fading situations. However, it is the best all round model for describing mobile radio statistics because of its versatility and mathematical simplicity. The analysis presented here follows that of Lee (Ref: Lee W C Y, 1982).

5.2.1 NON-FADING CASE

The non-coherent FSK selection scheme uses envelope detection rather than phase detection. A pair of bandpass filters tuned to "mark" and "space" frequencies are used to distinguish between two signal envelopes $r_1$ and $r_2$, respectively. Here, the case is examined where one filter passes a signal ‘A’ and the other filter passes only noise. The aim of the following argument is to calculate the probability that the signal ‘A’ is not detected. The probability distribution function of the envelope, $r_1$, which contains the signal ‘A’ is a Rician function given by:

$$p(r_1) = \frac{r_1}{\sigma^2} e^{-\frac{(r_1^2 + A^2)}{2\sigma^2}} I_0 \left[ \frac{r_1 A}{\sigma^2} \right], \quad 0 \leq r_1 < \infty \quad 5.1$$

where $I_0$ is a zero order modified Bessel function. At the same instant of time, the other filter passes only noise and the PDF of the envelope, $r_2$, is:

$$p(r_2) = \frac{r_2}{\sigma^2} e^{-\frac{r_2^2}{2\sigma^2}}, \quad 0 \leq r_2 < \infty \quad 5.2$$
Errors will occur when the envelope $r_1$ is less than the envelope $r_2$. Thus, the probability error can be expressed:

$$P_e = 	ext{prob}(r_1 < r_2) = \int_{r_1=0}^{\infty} p(r_1) \int_{r_2=r_1}^{\infty} p(r_2) dr_2 dr_1$$

$$= \frac{1}{2} e^{-\lambda^2 \sigma^2}$$

$$= \frac{1}{2} e^{-\gamma^2} \quad 5.3$$

where $\gamma$ is the signal to noise ratio (SNR).

The detection system employed in this project does not compare the envelope of $r_1$ to the noise envelope of $r_2$, but to a fixed decision threshold $V_T$. Errors occur when the envelope of $r_1$ is less than $V_T$ and when the envelope of $r_2$ is greater than $V_T$. In one case, a 1 is mistaken for a 0 and in the other a 0 for a 1. Assuming that the probability of transmitting a 1 or a 0 is equal, the combined average error rate is:

$$P_e = \frac{1}{2} \int_0^{V_T} \frac{r}{\sigma^2} e^{-\frac{r^2}{2\sigma^2}} I_0 \left[ \frac{rA}{\sigma^2} \right] dr +$$

$$\frac{1}{2} \int_{V_T}^{\infty} \frac{r}{\sigma^2} e^{-\frac{r^2}{2\sigma^2}} dr, \quad 0 < r < \infty \quad 5.4$$

The first term cannot be evaluated by simple means. Numerical techniques or a series approximation must be used. However, for large SNR's, where $V_T \gamma^2/2\sigma^2 \gg 1$ and $A \gg V_T - A$, a first order approximation for this term is given by Skolnik (Ref: Skolnik M I, 1981) and the combined error rate is:

$$P_e = \frac{1}{2} \left( 1 - \text{erf} \left( \frac{A - V_T}{\sqrt{2} \sigma} \right) + e^{V_T^2/2\sigma^2} \right) \quad 5.5$$
5.2.2 FADING CASE (MOBILE RADIO ENVIRONMENT)

If the time delay spread of the mobile radio environment is small in comparison with the inverse of the signalling bandwidth, the received signal will only be corrupted by fading. The SNR is a varying quantity on account of the multipath fading. The instantaneous SNR $\gamma$ is proportional to the square of the Rayleigh fading envelope $r^2$ which can be obtained, following the approach taken by Lee (Ref: Lee W C Y, 1982), by letting:

$$\gamma = \frac{r^2}{2\sigma^2}$$  \hspace{1cm} 5.6

Using the expression for the probability of $r$:

$$p_r(r) = \int_0^\pi P_{r,\theta}(r, \theta) d\theta = \frac{r}{\sigma^2} e^{-r^2/2\sigma^2}, \quad r \geq 0$$  \hspace{1cm} 5.7

We can obtain an expression for the probability of the SNR:

$$p_\gamma(\gamma) = \frac{1}{\gamma_0} e^{-\gamma_0 \gamma}, \quad \gamma_0 = <\gamma>$$  \hspace{1cm} 5.8

and the average error rate can be obtained from:

$$<P_e> = \int_0^\infty p_\gamma(\gamma) P_e(\gamma) d\gamma$$  \hspace{1cm} 5.9

For non-coherent FSK the average error rate is:

$$<P_e> = \int_0^\infty \frac{1}{\gamma_0} e^{-\gamma_0} \frac{1}{2} e^{-\gamma^2} d\gamma$$

$$= \frac{1}{2 + \gamma}$$  \hspace{1cm} 5.10
For non-coherent FSK, using a fixed decision threshold $V_T$, the average error rate using the large SNR approximation presented, above, is given by:

$$< P_e > = \int_0^\infty \frac{1}{\gamma_0} e^{-\gamma_0} \frac{1}{2} \left( 1 - \text{erf} \left( \frac{A - V_T}{\sqrt{2} \sigma} \right) + e^{-\gamma_0 \sigma^2} \right) d\gamma \tag{5.11}$$

Figure 5.1 shows a plot of the expected BER versus SNR for a conventional non-coherent FSK system and also a non-coherent FSK system using a fixed decision threshold set 3 dB above the noise floor. For comparison, the plot also shows the BER-SNR relationship for coherent FSK and PSK in both fading and non-fading conditions. The plots for a fading scenario and a non-fading scenario are given. The trace of BER versus SNR for a bi-phase-shift-keying (BPSK) detection represents the best achievable performance. It can be seen from the Figure that multipath propagation causes severe degradation for all detection systems. In the fading scenario, non-coherent FSK detection using a fixed threshold is approximately 3 dB worse than non-coherent FSK detection which itself is approximately 6 dB worse than coherent PSK detection. Thus, in conclusion, we can see that the error performance of the receiver used here is about 9 dB worse than an optimum system using BPSK.
Chapter 5

Figure 5.1 Expected BER Versus SNR for a Conventional FSK System, an FSK System Using a Fixed Design Threshold and PSK Systems
5.3 DIVERSITY RECEPTION

Diversity reception and the various means by which it may be accomplished are summarized in Section 2.9. In this Section we examine the improvement in BER effected by an M branch diversity reception scheme which combines the signals on each channel after square law detection. This follows an analysis by Jakes (Ref: Jakes W C Jr, 1974, pp 526). Specific results are presented for a two branch diversity system. After squaring and combining at baseband, the output of the combiner is considered as the sum of M, squared, independent, Rayleigh variables. Following Jakes, the probability distribution function of the signal envelope, \( r \), which contains a signal 'A' is:

\[
p_1(r) = \frac{r^{M-1}}{(M-1)!(A^2+2\sigma^2)^M} e^{-r(A^2+2\sigma^2)}
\]

and the probability distribution of the envelope, \( r \), which contains only noise is:

\[
p_0(r) = \frac{r^{M-1}}{(M-1)!(2\sigma^2)^M} e^{-r/\sigma^2}
\]

Let the decision threshold be \( V_T \), that is the binary data is a 1 if \( r > V_T \) and a 0 if \( r < V_T \). Following Section 5.2.1, the combined error probability that a 1 will be mistaken for 0 and vice versa, assuming an equal probability that a 1 or a 0 will be transmitted, is:

\[
P_e = \frac{1}{2} \int_0^{V_T} p_1(r) dr + \frac{1}{2} \int_{V_T}^{\infty} p_0(r) dr
\]

where the first term represents the probability of a 1 being mistaken for a 0 and the second the probability of a 0 being mistaken for a 1. The solution of Equation 5.14 is not straightforward and must be effected by numerical methods. Jakes reproduced two numerical solutions to Equation 5.14, one for the case of an optimum fixed threshold \( V_T \) based on the average received power and the other for the case of a moving threshold based on the instantaneous received power. Figure 5.2, reproduced from Jakes (Ref: Jakes W C Jr, 1974, pp 520), shows the performance of the two branch ON-OFF AM system described here for the fixed and variable threshold strategies, curves 5A) and 5B) respectively, in comparison to four other diversity systems. These are three pre-detection systems: 1) phase shift keying PSK; 2) differential phase shift keying DPSK; 3) frequency shift keying FSK; and 4) post-detection differential phase shift keying DPSK. ON-OFF AM
with a fixed threshold requires about 5 dB more power than DPSK to produce an error rate of $10^{-4}$; with a moveable threshold only 2.5 dB more power is required. If the transmitter peak power is limited, the FSK and PSK systems have an additional 3 dB power advantage since an ON-OFF AM transmits only half the time. Thus, additional powers of 8 dB and 5.5 dB are required respectively by the two systems.

This analysis presented here is used in Chapter 8 to estimate the improvement that a simple two branch diversity reception system would yield in BER performance.

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**Figure 5.2 Expected BER Versus SNR for Two Branch Diversity Systems**
(Ref: Jakes W C Jr, 1974)
Chapter 6 DIGITAL EXPERIMENTAL EQUIPMENT

6.1 INTRODUCTION

A simple digital receiver was constructed to provide initial measurements of bit error rate (BER) versus signal to noise ratio (SNR) on a mobile link at data rates from 10 - 20 Mb/s. The experimental configuration is similar to the basic system used for the analogue measurements described in Section 3.2 with the addition of a means for encoding digital data onto the transmitted carrier and decoding it at the receiver.

Figure 6.1 shows the experimental arrangement used for the field trials. The system uses frequency shift modulation to encode the digital data onto the carrier frequency. This is accomplished by varactor tuning of the Gunn oscillator. Detection is performed by narrow band filtering the received signal, followed by square law detection and comparison of the signal envelope to a fixed decision threshold.

The transmitter is modulated by a pseudo-random-bit-sequence (PRBS) $2^{10} - 1$ bits long with non-return-to-zero (NRZ) encoding, produced by a Hewlett Packard Data Generator type 3762A. The properties of PRBS's are discussed briefly in Section 6.2.1. This type of modulation suppresses the clock component in the transmitted signal. To recover the clock frequency at the receiver, the detected signal is passed through a non-linear circuit element. The non-linear element produces a discrete component at the clock frequency which is used as the reference for a phase-locked-loop (PLL). The details of these circuits are discussed in Sections 6.3.2 and 6.3.4.2.

The received signal is passed through an automatic gain control amplifier (AGC) and brought to binary voltage levels by an ECL comparator circuit. The output from the digital receiver is a binary data stream together with the recovered clock, both of which are at ECL voltage levels. These are fed to the error detector, a Hewlett Packard Error Detector type 3763A which produces a local PRBS sequence which is identical to the transmitted sequence and performs the code synchronization process, Section 6.3.4.3. The differences which occur between the sequences are due to errors. These errors are counted for a set gating period and are used to calculate the received BER, Section 6.3.5. The error detector produces an output voltage pulse proportional to the exponent of the BER, Section 6.3.5. The BER output from the error detector is recorded together with the AGC voltage level.
which provides a measure of the received signal level, and the car speed information on magnetic tape. The recorded results are then processed to yield curves of distance versus BER and signal level versus BER. The error rate during measurements fluctuated from $\sim 10^{-6}$ to unsynchronized.

In Section 6.4 the predicted BER versus average SNR performance of the receiver, Section 5.2, is compared to its measured performance in the laboratory.

### 6.2 TRANSMITTER

Figure 6.1 shows a block diagram of the transmitter. The data generator clock rate is set by an external clock, Section 6.2.2. The binary output of the data generator passes through a low pass filter to attenuate the higher harmonics of the signal. These sidebands are not used by the receiver which is band limited, and represent a waste of available energy. The bit stream is fed to the varactor modulated Gunn oscillator. The spectrum of the FM modulated IF signal at the receiver is displayed on a spectrum analyser and the amplitude and offset voltage of the bit stream fed to the transmit Gunn are adjusted until the spectrum is symmetrical.
Digital Transmitter

Chapter 6

Varactor tuned Gunn

Hewlet Packard Data Generator

External Clock

Digital Receiver

AGC loop filter

AGC amplifier

IF Selection Filter

Hewlet Packard Error Detector

ECL comparator circuit

Clock recovery PLL

Hewlet Packard Error Detector

IBM PC-AT

ADC

Speed Input

FM Taperecorder

Figure 6.1 Digital Transmitter and Receiver Block Diagram
Chapter 6

6.2.1 PSEUDO-RANDOM-BIT-SEQUENCE GENERATION

The Hewlett Packard Data Generator uses a shift register to produce various pseudo-random-bit-sequences (PRBS) with $2^{10} - 1$, $2^{15} - 1$ and $2^{23} - 1$ bits respectively. The sequences are generated by a shift register which is switchable to length 10, 15 and 23 bits respectively. The PRBS's produced are maximal length sequences. The data generator can also produce 10 and 16 bit sequences of binary digits which are set by keys on the front panel.

6.2.2 EXTERNAL CLOCK

The clock rate of the data generator is set by an external clock, a Racal Dana frequency synthesizer. The clock rate is variable from 1 KHz to 25 MHz. Measurements were carried out at 10, 15 and 20 Mb/s in the field and bench measurements were carried out at 30 Mb/s.

6.3 RECEIVER

The block diagram of the receiver is shown in Figure 6.1. The received signal is brought to a constant level by an automatic gain control amplifier, Section 6.3.1; filtered by a 20 MHz bandwidth VHF filter; video detected by a fast rise time back diode detector and amplified by a fast OP-amp circuit, Section 6.3.2; brought to binary voltage levels by an ECL comparator acting as a Schmitt trigger, Section 6.3.3; synchronized by clock and code recovery loops, Section 6.3.4. Then the errors present in the received signal are detected, Section 6.3.5, and a voltage corresponding to the exponent of the BER is recorded on magnetic tape together with the AGC voltage and the car speed detector output, Section 6.3.6. Finally, the recorded signal is digitized by the computer, Section 6.3.7, and processed to yield curves of BER versus average SNR, Section 6.3.7.1.
6.3.1 AUTOMATIC GAIN CONTROL (AGC)

The automatic gain control circuit detects the amplitude of the received signal and automatically adjusts the gain to maintain a constant output signal. The circuit has four elements: a detector, a variable gain amplifier, a low pass filter and a difference amplifier. Figure 6.2 shows a functional block diagram of the AGC loop including the circuit diagram of the AGC loop filter. The gain control function is carried out by an Avantek AGC 553 chip. This is a voltage controlled attenuator which has a gain control range of 45 dB. The received voltage is sensed at the output of the video amplifier. This voltage is fed to the active loop filter which compares the received voltage to a reference. The difference between the two voltages is averaged to provide the control voltage for the attenuator.

The loop bandwidth is an important factor because the loop must be fast enough to track out deep fades, yet it must not be subject to modulation by noise fluctuations. The maximum fading frequency is approximately 2.5 KHz for the wavelength and speed used and for an omni-directional antenna. The fade frequency reduces to ~1 KHz for a 120° horn. The desired natural frequency is more than 2000 Hz and ~10 MHz. An exact analysis of the loop parameters is not attempted because the variation of loop gain with input voltage gives rise to a non-linear transfer function where loop bandwidth varies with input signal. The values of the components in the loop filter are estimated by using Equation 6.1 for the natural frequency and damping factor of an ordinary phase locked loop as a guide and then adjusting the loop components on a trial and error basis. The loop natural frequency is given by the product of the IF amplifier gain, the AGC gain and the detector gain over the time constant set by CR, (Ref: Gardiner F M, 1979). The final values of the loop components, 20 KΩ and 0.1 μF, give a loop bandwidth of ~3800 Hz for a mid-point AGC gain of 20 dB and detector conversion efficiency of 1. The main restriction on loop performance is noise saturation of the AGC control voltage at low signal levels. The dynamic range of the loop is ~30 dB.
6.3.2 VIDEO DETECTION

A back diode detector type 2085-6013-00, manufactured by Omni Spectra, is used as the video detector. It has a rise time of 3 ns and a maximum bandwidth of ~30 MHz which is achieved by minimizing its video output impedance; 80 Ω compared to 1 - 2 kΩ for a Schottky barrier. However, minimizing the impedance makes the diode very susceptible to electrical and mechanical damage. In
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particular, damage by static electricity is a great danger. All handling operations were carried out at an anti-static work station to minimize the possibility of damage to the devices.

The signal output from the video detector is at a low level and must be amplified for further processing to be carried out. Video amplification is carried out by a high speed operational amplifier type NE5539, supplied by Radio Spares and constructed on their propriety printed circuit board. The circuit is built to the standard configuration illustrated in RS data sheet Number 5601, March 1988. A 3 pF compensation capacitor, marked \( C_{cc} \) on the circuit diagram on the data sheet, is used to minimize ringing on the amplifier output. The chip has a 1200 V per \( \mu \) second slew rate and a full power bandwidth of 45 MHz. A wideband amplifier is used to prevent distortion of the digital signal.

6.3.3 DATA CONDITIONING (ECL COMPARATOR)

The output of the video detector is brought to binary logic levels by an ECL comparator type SP9687, manufactured by Plessey, configured as a Schmitt trigger with \( \sim 20 \) mV of hysteresis to minimize the possibility of multiple transitions occurring due to noise at the threshold voltage. The circuit is configured, as indicated in the data sheet of the device in the Plessey High Speed Logic Data Book (Ref: Plessey Semiconductors, 1987), using double sided printed circuit board with copious use of de-coupling capacitors. The circuit compares the raw data input signal to a reference voltage set by a potentiometer. The reference is set approximately at the mid-point voltage of the received signal eye diagram. The reference is set by adjusting the potentiometer to minimize the BER.

6.3.4 SYNCHRONIZATION

The receiver must acquire two types of synchronization for a measurement of BER to be made: clock synchronization and code synchronization. Clock synchronization is needed to provide timing information for decoding a digital bit sequence. However, efficient data pulse streams contain no component at the clock frequency, as such a component represents a waste of energy. For example, the pseudo-random-bit-sequence with non-return to zero (NRZ) encoding which is used here, has a spectral null at the clock frequency. Clock synchronization involves two stages: recovering a discrete clock component, and generating a stable local clock from this component. Code synchronization is needed so that
the receiver knows where it is in the transmitted bit sequence. In a communication system each bit sequence would represent a block of information or data word and incorrect sync would give garbage output. A coherent receiver would also require carrier synchronization.

6.3.4.1 Clock Regeneration

A discrete clock component is recovered by passing the NRZ signal through a non-linear circuit element (Ref: Gardiner FM, 1979). The non-linearity used is an over-driven amplifier. When an amplifier is driven into dipping it produces a "hard" non-linearity generating a wide range of harmonics.

Figure 6.3 shows the circuit diagram of the clock regeneration circuit. The circuit consists of two Radio Spares 560 C RF amplifiers. The first 560 C which performs the clipping function operates in high gain mode with pin 5 disconnected. Potentiometer $P_1$ is used to adjust the current to the first stage transistors on the chip and controls the clipping level. $L_1, C_1$ provide a selective element to reduce the noise level and unwanted harmonics. The second 560 C provides a buffer to drive the phase detector of the PLL. The circuit is relatively broad-band and will regenerate a clock from a bit stream at data rates from 9 Mb/s to 25 Mb/s. The RF voltage level at the output of the first 560 C is approximately a 5 V square wave and the circuit is enclosed in a metal box to prevent spurious radio emission.

![Figure 6.3 Clock Regeneration Circuit](image-url)
6.3.4.2 Clock Phase Locked Loop (PLL)

The clock component, recovered by the circuit described above, only exists when a pulse is transmitted. It is also subject to timing jitter due to noise. A phase locked loop (PLL) is used to generate a continuous local clock and to minimize timing jitter. Figure 6.4 shows the functional block diagram of the PLL including the loop filter circuit diagram. A phase detector (type RDP-1, manufactured by Mini-Circuits) is used to detect the phase difference between the recovered clock signal and the local clock from the VCO (a Hewlett Packard Function Generator type 3314A). An active loop filter with a Miller type capacitor is used to average the phase error signal from the phase detector and to provide the correction voltage to the VCO. The loop contains a feedback resistor R_f, which limits the DC gain of the OP amp. This stops the VCO frequency drifting away from the clock frequency when lock is lost during severe fades. If this resistor is not included, the VCO frequency drift during a severe signal fade will make the lock re-acquisition time excessively long. The resistor has the disadvantage of reducing the ultimate pull-in characteristics of the loop, but this is not a problem because the VCO frequency is set very near to the clock frequency. Consequently, the frequency error after loss of lock is very small and re-acquisition is fast.

Two important parameters of a phase locked loop are its natural frequency and damping factor which are given by:

\[ \omega_n = \left( \frac{K_0 K_d}{R_1 C} \right)^{1/2}, \quad \zeta = \frac{R_2 C \omega_n}{2} \]

where: \( K_0 \) = PSD gain factor \((3 \times 10^{-3} \text{ Vrad}^{-1})\), \( K_d \) = VCO gain factor \((700 \times 10^3 \text{ HzV}^{-1})\), \( C \) = filter capacitor \((0.47 \mu\text{F})\), \( R_1 \) = resistor \((16k)\), \( R_2 \) = resistor \((15k)\), \( \omega_n \) = loop natural frequency \((500 \text{ Hz})\), \( \zeta \) = loop damping factor \((0.5)\).

The values of \( K_0 \) and \( K_d \) were found by experiment; \( K_0 \) by measuring the change in VCO frequency for a given change in voltage; \( K_d \) by measuring the change in PSD output voltage for a given change in VCO phase whilst the loop is in lock. The VCO phase is changed by adjusting the offset voltage in the loop. The two parameters define the apparent loop SNR, \( \text{SNR}_L \), which is given by:

\[ \text{SNR}_L = \frac{P_s}{2B_w W_i} = \frac{P_s}{\omega_n W_i} \]

where: \( P_s \) = signal power, \( W_i \) = noise power density.
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As the loop signal to noise ratio $SNR_L$ approaches 0 dB, cycle slipping starts to occur and at $SNR_L = 0$ dB the loop loses lock. Lock is only regained at $SNR_L \sim 3 - 6$ dB. The functional relationship between $SNR_L$ and the SNR of the received signal is highly non-linear. An exact analysis is beyond the scope of this Thesis. Section 6.4 describes the measured performance of the receiver.

Figure 6.4 Diagram of Clock Recovery PLL Including Circuit of Loop Filter
6.3.4.3 Code Synchronization

Synchronization is acquired by feeding a set number of cycles of the received, binary converted signal into the local PRBS register. This sets the initial condition for the register. The number of errors is checked. If it is below 10 errors in the first 90 clock cycles, the loop is assumed to be in code lock. If it is greater than this, the process is repeated. When the loop is in lock and at any time the error rate exceeds 10,000 errors in 90,000 clock periods, the loop is assumed to have lost lock and the lock acquisition process begins again. The acquisition relies on the auto-correlation properties of PRBS’s to achieve lock. The system finds the correct point of lock because the pseudo-random sequences will only lock at integer multiples of the sequence length.

6.3.5 ERROR DETECTION

Error detection is performed by a Hewlett Packard Error Detector type 3763A which compares the locally generated PRBS with the received data using exclusive OR gates. A counter circuit records the number of clock cycles elapsed until either 10 or 100 errors have occurred. A reading of the BER is then given which is determined from the reciprocal of the counter reading. The variance of the BER reading depends on whether 10 or 100 errors are counted. For normally distributed errors the variance is proportional to the square root of the number of errors. The error rates in the mobile experiments are quite high and the detector is set to record 100 errors.

The output from the error detector is an analogue voltage pulse with amplitude, proportional to the exponent of the received BER which is produced at the end of every count of 10 or 100 errors. Separate voltage levels are used to indicate signal loss and loss of code lock. The exponent of the BER is a relatively coarse measure of the BER. However, a simple measurement of the exponent is sufficient for the preliminary digital probing carried out on the mobile link.

6.3.6 DATA COLLECTION

Data collection is performed using the same procedure as for the analogue measurements, Section 3.2.4. The BER voltage pulses are recorded on magnetic tape together with the AGC voltage which is calibrated to give a reading of the received signal strength and the car speed information.
6.3.7 DATA ACQUISITION AND ANALYSIS

Data acquisition is performed by sampling the recorded data AGC voltage and the voltage representing the BER exponent using the same procedure as for the analogue measurements, described in Section 3.2.5. Data analysis requires two separate types of calibration: one to obtain the received signal level and SNR from the AGC voltage; the other to convert the voltage pulses representing exponent of the BER into average BER. The calibrated data is processed to provide a measure of the BER versus receiver SNR and BER versus distance.

The calibration between received signal level and AGC voltage is effected by plotting the AGC voltage versus received signal power in the laboratory and using the input-output relationship obtained to provide a direct calibration of receiver SNR in dB. The SNR of the receiver is obtained by measuring the root mean square (RMS) noise voltage of the receiver, $V_n$, at the output of the first IF amplifier under no-signal conditions. The SNR of the receiver for an arbitrary receiver output voltage, $V_{OUT}$, is then given by:

$$SNR = \frac{V_{OUT}^2 - V_n^2}{V_n^2}$$  \hspace{1cm} 6.3$$

BER calibration is performed by identifying the voltage levels which correspond to the various BER's and distinguishing these from the voltage levels indicating loss of lock and signal loss. Using this information, the digital output from the analogue to digital converter (ADC) is divided into bins. The mid-point of each bin is the voltage corresponding to each BER exponent from the Hewlett Packard Error Detector. The frequency of occurrence in each bin for a particular signal level is summed and the resulting frequency distribution used to provide an estimate of the expected value of BER.
6.3.7.1 Data Processing

The main parameter of interest is the expectation rate of the BER for a given receiver SNR or transmitter-receiver separation. The expectation of the BER at a given signal level A is:

\[
\langle BER \rangle = \sum_{BER} BER \times P(BER, A)
\]

where: BER = bit error rate, P(BER, A) = probability of a given BER at amplitude A.

The probability distribution of the BER is calculated from the experimentally measured number of events with a particular BER at a given signal level, divided by the total number of events at that signal level. This is given by:

\[
P(BER, A) = \frac{f(BER, A)}{\sum_{BER} f(BER, A)}
\]

The expectation of the exponent of the BER, E[EXP] at a given signal is given by:

\[
E[\text{EXP}_{BER}] = \log \left( \sum_{\text{EXP}} 10^{\text{EXP}P(BER, A)} \right)
\]

where EXP = exponent of BER.

The data was processed to yield a measurement of BER versus average signal by dividing the received signal strength into bins and calculating the expectation value of BER for each bin. The results were obtained by averaging across all data recorded at similar car speeds. To do this, the computer program has the facility to edit out the portions of the runs where the car was stationary.

6.3.8 COMPARISONS WITH ANALOGUE CONFIGURATION

Table 6.1 summarizes the requirements of the digital experimental equipment configuration. This Table provides a comparison with the requirements of the various analogue experimental configurations summarized in Section 3.10, Table 3.4.
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<table>
<thead>
<tr>
<th>Measurement</th>
<th>Transmitter Requirements</th>
<th>Receiver Requirements</th>
<th>Fig Ref</th>
</tr>
</thead>
<tbody>
<tr>
<td>BER versus average signal strength</td>
<td>Single varactor tuned Gunn, PRBS generator, external clock</td>
<td>IF AMP, AGC loop, 1 BPF Video detector, video amp, clock recovery, code sync, error detection</td>
<td>6.1</td>
</tr>
</tbody>
</table>

Table 6.1 Digital Experimental Requirements

6.4 SYSTEM PERFORMANCE

To assess the function of the digital link, the relationship between receiver SNR and BER and the points at which lock is lost and re-acquired were measured in the laboratory. The measurements are for the whole system and thus also involve the performance of the AGC Loop. Figure 6.5 shows the plot of BER versus SNR measured in the laboratory compared to the theoretical prediction derived in Section 5.2.

All SNR's are specified with reference to the 20 MHz noise bandwidth and taking into account all transmitted power. The bandwidth of the clock recovery loop is ~1 KHz. Thus, specifying SNR with respect to 20 MHz gives rise to negative SNR.

The code recovery loop loses lock at an SNR of about 7.7 dB. Due to the nature of the code acquisition process, this point is not well defined and the loop jumps in and out of lock around this SNR. This can be is seen in Figure 6.5 where the experimental error rate is greater than the theoretical error rate near to this threshold.

The clock recovery loop loses lock at an SNR of -4.5 dB and recovers lock quickly when the SNR increases to -2.5 dB. The code recovery loop only recovers lock if the VCO frequency error is very small. In practice, this is not a problem because the data clock rate is well defined and Doppler shifts in frequency are very small.
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6.4.1 DATA RATE

Measurements were carried out at 10 MHz and 15 MHz data rates. The system would operate at 20 MHz data rate, but the code and clock recovery loops were very unstable at this data rate and it was not possible to carry out useful measurements.
6.4.2 RANGE

The fading margin of the digital receiver is lower than that of the analogue receiver. This restricted the range over which useful measurements could be made to ~200 m. In addition, the receiver has problems with code synchronization. These problems arise because the synchronization process is not designed to cope with the severe multipath fading encountered in the mobile radio environment. Section 6.4.3 details the shortcomings of the synchronization process.

6.4.3 SYNCHRONIZATION PROBLEMS

The code acquisition system has a severe drawback for operation in a mobile radio environment. The received receiver loses lock if more than 10,000 errors are counted in 90,000 clock periods. This occurs every time the signal fades below the receiver threshold of -82 dBm for more than $1.7 \times 10^{-4}$ seconds at a 15 MHz clock rate. Fades of this duration are not uncommon (Ref: Lee W C Y, 1982). When lock is lost the re-acquisition process feeds the received, binary converted output into the receiver PRBS shift register. However, if the receiver's input is below threshold, its output consists of random bits bearing no relation to the transmitted pseudo random sequence. Thus, the receiver loses all knowledge of the code phase information.

A more robust code acquisition loop for a mobile digital receiver would maintain an estimate of the most current value of the pseudo random sequence phase and would hold this during the signal fade. Hence, the phase difference between the transmitted and received pseudo random sequences would be very small when the signal returned, allowing very fast re-acquisition of lock after a fade (Ref: Gardiner F M, 1979).

6.4.4 SUMMARY

Table 6.2 summarizes the performance of the various elements of the digital link. In a system employing a more sophisticated modulation system, the loop threshold level for the code recovery loop would be almost 0 dB.
Basic Receiver Characteristics

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Comments</th>
</tr>
</thead>
<tbody>
<tr>
<td>Noise Figure</td>
<td>12.1 dB</td>
<td>Basic value is 9 dB. However we reject 3 dB signal power in the IF selection filter.</td>
</tr>
<tr>
<td>Minimum Detectable Signal</td>
<td>-90 dBm</td>
<td>Basic value is - 93 dBm in a 15 MHz bandwidth. However, we reject 3 dB IF selection filter.</td>
</tr>
</tbody>
</table>

Code Recovery Loop Performance

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Rx Level</th>
<th>Comments</th>
</tr>
</thead>
<tbody>
<tr>
<td>Loop Threshold</td>
<td>7.7 dB SNR</td>
<td>-82.3 dBm</td>
<td>SNR in 15 MHz BW loop loses lock when number of errors more than 10,000 in 90,000 clock cycles</td>
</tr>
</tbody>
</table>

Clock Recovery Loop Performance

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Rx Level</th>
<th>Comments</th>
</tr>
</thead>
<tbody>
<tr>
<td>Loss of Lock</td>
<td>-4.5 dB SNR</td>
<td>-94.5 dBm</td>
<td>SNR in 15 MHz BW</td>
</tr>
<tr>
<td>Acquire Lock</td>
<td>-2.5 dB SNR</td>
<td>-92.5 dBm</td>
<td>SNR in 15 MHz BW</td>
</tr>
<tr>
<td>Loop Bandwidth</td>
<td>~1 KHz</td>
<td>-</td>
<td></td>
</tr>
<tr>
<td>Clock Range</td>
<td>9 - 25 MHz</td>
<td>-</td>
<td>Achieved by manual adjustment of local VCO centre frequency</td>
</tr>
<tr>
<td>NRZ Clock Regeneration</td>
<td>-</td>
<td>-</td>
<td>Over-driven amplifier ~ hard clipping</td>
</tr>
</tbody>
</table>

Automatic Gain Control Loop Performance

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Comments</th>
</tr>
</thead>
<tbody>
<tr>
<td>Loop Bandwidth</td>
<td>~3.8 KHz</td>
<td>Bandwidth depends on gain level, specified as 20 dB gain, detector efficiency ~1</td>
</tr>
<tr>
<td>Dynamic Range</td>
<td>~30 dB</td>
<td></td>
</tr>
</tbody>
</table>

Miscellaneous Parameters

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Comments</th>
</tr>
</thead>
<tbody>
<tr>
<td>Errors per BER Measurement</td>
<td>10 or 100</td>
<td>The number of errors is switch for &quot;fast&quot; or &quot;slow&quot; error detection</td>
</tr>
<tr>
<td>Bits per Word</td>
<td>16 or 1023</td>
<td>Manually set or PRBS</td>
</tr>
<tr>
<td>Clock Rate</td>
<td>15 MHz</td>
<td></td>
</tr>
</tbody>
</table>

Table 6.2 Digital Receiver Characteristics
Chapter 7 DIGITAL RESULTS

7.1 INTRODUCTION

The experimental equipment used for the digital measurements is described in Chapter 6. The measurement system is designed to measure the speed of the mobile unit throughout the run, the bit error rate (BER), and the received signal level. From these values it is possible to measure the instantaneous signal strength and average signal to noise ratio (SNR). These results are presented in Section 7.2.

Section 7.3 summarizes the work carried out in this study of digital transmission in the mobile environment in the millimetre wave frequency band. It makes suggestions for future work which would be useful to prove the feasibility of a digital mobile radio system operating at these frequencies. From the measurements described in this Chapter, it is possible to infer how a more sophisticated receiver would perform. The modulation system employed uses frequency shift keying to perform a version of ON-OFF keying. Experimental results are compared with predictions for ON-OFF keying and with optimum systems based on bi-phase shift keying [11]. The results of using a two branch diversity system on the link are predicted. The result of this prediction is used to indicate the form which a future 55 GHz micro-cellular system could take and an indication of its potential performance.

7.2 EXPERIMENTAL RESULTS

Measurements were made on a path in close proximity to the College. The road concerned, Malet Street (Location 2 on Figure 4.1), is about 30 m wide with buildings ranging in height between four and seven storeys. The transmitter was located on the second floor of a building overlooking the length of Malet Street. Measurements were carried out at a variety of car speeds which depended primarily on traffic conditions.

1 Bi-phase shift keying (BPSK) provides the greatest immunity to noise because the difference between a one and a zero is maximised, the two being equal in magnitude, but 180° out of phase with each other.
Figure 7.1 Average Signal Strength Versus Distance Compared with Average BER Versus Distance, Example 1

Figure 7.1 shows a record of average signal strength and BER versus distance for a typical run. The average speed of the mobile receiver in this case was 5 mph. The top curve shows the variation of average received signal level as a function of distance from the transmitter. The average signal level is increasing as the receiver approaches the transmitter. The lower curve shows the measured BER for this run. Clusters of high errors occur around deep fades in the average received signal strength and the correlation between the two traces, from visual inspection, is very high. However, in some places the BER is high even where the signal level in Figure 7.1 is not in a deep fade. This is explained by the behaviour of the BER during a deep fade. Even though the average signal in Figure 7.1 is high, there is a possibility of a deep fade occurring. This will produce only a marginal change in average signal level, but will result in a very large number of errors. The high error rate pulses occurring in this Figure are of a longer duration than would be
expected from the SNR versus BER characteristic measured in the laboratory, illustrated in Figure 6.5, Section 6.4. This is due to the code lock acquisition system used by the Hewlett Packard Error Detection Receiver, Section 6.4.3.

Figure 7.2 shows a run where lock was maintained and clearly, in this case, the maximum BER of $10^{-3}$ is sufficiently low to allow the 15 Mb/s mobile link to carry a large number of speech channels without using error correction.

Figure 7.3 shows the BER as a function of the average SNR computed over ten experimental runs. All SNR's on the Figure are referred to that of a channel with no diversity. This curve is comparable to that expected for a single port receiver operating in a multipath "Rayleigh" fading environment using ON-OFF keying modulation (Ref: Jakes W C Jr, 1974). However, the received signal does not have
a true Rayleigh distribution due, for example, to the presence of a direct line of sight component and the limited number of scatterers. Hence, occasionally, the performance is superior to that indicated by the theoretical prediction.

The similarity of the experimental results to the theoretical prediction for a Rayleigh fading channel suggests that it is feasible to estimate the potential improvement to the system performance obtainable by using diversity reception, using a model based on un-correlated Rayleigh fading inputs to a diversity combiner. The approximation is partially justified by comparison of the performance of a simulated diversity combiner using experimental time series with theoretical predictions based on Rayleigh fading channels, Section 4.4.1. However, an accurate prediction of the expected increase in performance must take account of the actual joint statistics of the input signals to the combiner. If contributions to the error rate from inter symbol interference and random FM are ignored (Ref: Lee W C Y, 1982), the error rate performance of the link can be predicted by integrating the curve of BER versus SNR measured in the laboratory, Figure 6.5, Section 6.4.1 [1], over the measured distribution of the instantaneous SNR of the combined signal envelope. However, the distribution of instantaneous SNR is a highly variable quantity both in terms of the shape of the distribution and in terms of the variation of the average value of the SNR. The predicted diversity improvement would have to be applied piece-wise at every location within the micro-cell. This is a complex operation and is beyond the scope of this work. Thus, although the assumption of Rayleigh statistics is not accurate, it provides a simple and accessible way to investigate the potential of diversity reception in a micro-cell.

Shown in Figure 7.3, for comparison with the experimental data, are two curves representing the average BER versus SNR performance of a two branch diversity receiver in a Rayleigh fading environment using respectively ON-OFF keying with a fixed decision threshold [2], and an optimum system based on PSK modulation with coherent maximal ratio combination (Ref: Jakes W C Jr, 1974). The PSK system would be difficult to implement due to the problem of phase recovery in a system at 55 GHz, but it provides an indication of the ultimate performance of the link.

1 The BER versus SNR performance of the receiver measured in the laboratory in this circumstance is interpreted as a curve of instantaneous BER versus SNR.

2 This is equivalent to the incoherent FSK system with a fixed decision threshold which is used experimentally.
Chapter 7

From the top curve in Figure 7.3, it can be seen that for an average SNR of 20 dB at the mobile terminal, the average BER is 10^-2 for the no diversity case, and 4 x 1^-4 for the two branch diversity, using ON-OFF keying. For the PSK system, the average BER is 5 x 10^-3. Thus, at an average SNR of 20 dB, two branch diversity improves the average BER by nearly two orders of magnitude using the same modulation scheme and by nearly three if an optimum modulation system is used.

![Graph showing BER vs SNR for different diversity schemes and modulation methods.]

Figure 7.3 Experimental BER Versus SNR in a Fading Environment Compared with Theoretical Predictions for No Diversity and Two Branch Diversity

To gain an idea of what these numbers mean in a practical system, it is useful to estimate the approximate range at which the average SNR falls to 20 dB. This may be done by using the plot of propagation loss versus distance obtained along Malet Street, Figure 4.12, Section 4.2.1.2. The minimum detectable signal (MDS) for the digital receiver is -90 dB, Table 6.2, Section 6.4.4, and the effective radiated power of the transmitter (ERP) is 50 dBm which is made up of a transmitter power of 19.5 dBm, a transmitter antenna gain of 25 dB, a waveguide loss of 0.5 dB and a receiver antenna gain of 6 dB. Thus, the SNR will fall to 20 dB at a propagation loss of -120 dB. [Propagation loss = -90 (MDS) - 50 (ERP) +20 (SNR) = -120 dB].

From Figure 4.12, the propagation loss is less than 120 dB for a range of approximately 100 m along Malet Street. Thus, using two branch diversity, the system is predicted to operate at an error rate less than 10^-3 for a range of approximately
100 m along the Street. If an optimum diversity system is used, the error rate is
less than $5 \times 10^{-4}$ for a range of approximately 100 m (20 dB SNR) and less than
$10^{-3}$ for a range of approximately 150 m (10 dB SNR), Figure 7.3.

The estimated range of the system can be increased in two ways:

i) by increasing the transmitter power, and

ii) by reducing the receiver bandwidth.

For example, if the system were modified to use a transmitter power of 900 mW (a
10 dB increase), its fading margin would be increased by 10 dB. Thus, it would
have an SNR of 20 dB at 130 dB propagation loss and an SNR of 10 dB at 140 dB
propagation loss opposed to 120 dB and 130 dB respectively. Using these figures
in conjunction with Figure 4.11, an optimum two branch diversity system using
maximal ratio combining and PSK modulation is predicted to operate along the
whole length of Malet Street (350 m) with an error rate of less than $10^{-3}$ and an
error rate of less than $5 \times 10^{-5}$ along the majority of the street.

Section 4.4 reports initial results of a simulation of a two branch frequency
diversity system using input channels de-correlated by a 100 MHz and by a
320 MHz frequency separation. The cross-correlation coefficient measured be­
tween channels transmitted on orthogonal polarizations and the predicted fall in
the correlation coefficient for antenna separated by more than $8 \lambda$ suggests that
similar results would be obtained with channels de-correlated by the use of a small
time (space) separation or the use of orthogonal polarizations. Thus, the predic­
tion of BER improvement for frequency diversity can also be applied to space or
polarization diversity.
7.3 SUMMARY OF RESULTS

The initial results which have been processed show marked similarities to the theoretical predictions for a Rayleigh fading channel, top curve, Figure 7.3. On the basis of this similarity, and supported by results of simulation of diversity reception using experimental data in Section 4.4, predictions are made of the enhancement to link performance which would be expected if two branch reception diversity were implemented in an urban micro-cell.

The predictions suggest that, for a range of 100 m along Malet Street from the transmitter site in Malet Place, (Location 2, Figure 4.1) the average error rate would be reduced from $10^{-2}$ to $10^{-3}$ by two branch diversity. If an optimum system [1] were implemented, the predicted error rate is reduced to less than $5 \times 10^{-4}$ for a 100 m range and the range for an error rate less than $10^{-3}$ is increased to 150 m. In addition, it is predicted that if a modified receiver with 10 dB greater sensitivity and employing the optimum described in [1] were implemented, it would provide an error rate less than $10^{-3}$ for the whole length of Malet Street.

Thus, initial results suggest that using a diversity system with a more sophisticated receiver and modulation system, a performance comparable with a 900 MHz system, but with a far greater channel bandwidth and operating over a few hundred metres, should be obtainable at millimetre wave frequencies.

---

1 Using bi-phase shift keying plus two branch maximal ratio combination.
Chapter 8  CONCLUSION

A study has been made of the propagation characteristics of 55 GHz millimetre waves between a fixed transmitter site and a mobile receiver terminal in an urban environment. Transmitter-receiver ranges up to 400 m are used and measurements are obtained in the presence of high traffic densities.

The estimated values of the statistical quantities used to describe the received signal envelope are found to be highly varied both from time to time and from location to location within the micro-cell.

A large part of the variation of the statistics of the channel is seen to follow from the essential line of sight propagation conditions in a micro-cell. A received signal, dominated by line of sight propagation, will give rise to one type of distribution. However, a small perturbation to the LOS path, for example a tree, lamppost or another vehicle, will dramatically change the signal distribution as the strength of the LOS component is momentarily reduced.

The statistical distribution of the received signal envelope is analysed over three different scale sizes. A general trend, approximated by a power law, describes the average decrease in received power with increasing transmitter receiver separation. A "slow fading" process describes the non-stationary variation of the local average signal level from location to location throughout the micro-cell and a "fast fading" process describes the signal variations over a small distance scale.

i) On the largest scale size, path loss along the micro-cell follows an approximate $r^{-3.6}$ power law, whilst along intersecting streets the power law is approximately $r^{-10}$, Section 4.2.1.4. The signal power is largely contained within the micro-cell by the buildings surrounding it, Figure 4.14, Section 4.2.1.5, with the result that power leakage between parallel streets is very low, suggesting that micro-cells based along them could use the same frequency allocations.

ii) The distribution of the "slow fading" process is approximately log-normal. The distribution has a location variability [1] which ranges from 0.64 dB to 7.2 dB from place to place throughout the micro-cell,

---

1 Location variability is the standard deviation of the distribution in dB. It may be estimated from the spread between the 50% and 84% probability levels of the cumulative distribution of the local average signal.
Chapter 8

Section 4.2.2.1. The average location variability (3.2 dB) is comparable with that measured in a suburban area at 900 MHz (4.5 dB) (Ref: IEEE VT Special Issue, 1988).

iii) The distribution of the "fast fading" process has a loss deviation [1] which ranges from 0.9 dB to 9.3 dB, Section 4.2.3.2. The average value of the loss deviation is 4.7 dB which is comparable with that measured in a suburban environment at 900 MHz (4.32 dB) (Ref: IEEE VT Special Issue, 1988). The range of location variability values [2] shows that the distribution is only rarely described by a Rayleigh distribution.

The agreement between experimental results obtained in an urban micro-cell at 55 GHz with those obtained in a suburban cell at 900 MHz suggests that the predominant propagation mode in both environments is similar. It appears that the mode of propagation in both environments is characterized by a very high probability of a line of sight signal occurring combined with a relatively limited angular distribution of scatterers in comparison with the omni-directional continuum of scatterers assumed in theoretical models of the channel (Ref: Clarke R H, 1968; Aulin T, 1979). This suggestion is supported by comparison of experimentally measured squared envelope auto-correlation and power spectra with theoretical predictions.

i) The measured results of the auto-correlation function of the squared envelope are incompatible with the assumption of a uniform distribution of received waves. A simple model developed to estimate the ratio of the power in the line of sight component over that in a uniformly scattered component indicates that the ratio ranges from 0.6 to 78. This illustrates the extreme variation of the propagation conditions throughout the micro-cell and the domination of those conditions by the line of sight signal strength.

ii) The experimental squared envelope power spectrum is compared with the theoretical predictions of Clarke and Aulin (Ref: Clarke R H, 1968; Aulin T, 1979) and with those of a simple model based on interference of a direct ray with reflections from principal scatterers in the environment. The results of the comparison indicate that the received signal contains a substantial line of sight component and that the angular distribution of

---

1 The loss deviation is defined as the spread between the 50% and 90% probability levels on the cumulative distribution of the small area signal distribution.

2 A Rayleigh distribution is characterised by a loss deviation of 8.2 dB.
the randomly scattered component of the received signal is between 10° and 120°, that is it is intermediate between the angular width of the distribution assumed by the models of Clarke and Aulin (120° [1]) and that assumed by the deterministic model (10°).

Further corroboration is found for the suggestion of the dominant role of line of sight propagation in the micro-cell from analysis of the coherence bandwidth of the channel. Estimates of the variation in the coherence bandwidth [2] of the channel vary from 17 MHz to 150 MHz; the larger values being associated with a dominant LOS signal, the lower values being associated with a LOS component which is either comparable in magnitude with specularly reflected components or with a LOS component that is obstructed. This is far higher than the corresponding coherence bandwidth of 25 KHz measured in the urban environment at 900 MHz or than that measured in the suburban environment where it is approximately 640 KHz (Ref: Gans M J, 1972; Cox D C, 1973).

The variation of the coherence bandwidth throughout the micro-cell suggests that the value of bandwidth chosen for system design calculations will depend on whether a calculation is being made of the maximum bandwidth which may be transmitted across the link without suffering inter-symbol interference or whether a calculation is being made of the minimum separation to be provided between two channels in order for them to be sufficiently de-correlated for use in a frequency diversity system. In the first case, a conservative estimate of bandwidth would be chosen to minimize the introduction of errors due to inter-symbol interference occurring. In the second, a large value of coherence bandwidth would be chosen to maximize the possibility that two diversity channels would be de-correlated.

The performance of a digital link over the mobile channel has been investigated. Section 7.3 has shown that at 15 MHz data rate, the performance of the current system is limited by fading and that inter-symbol interference has very little effect [3]. This is in accordance with the results of the coherence bandwidth

---

1 The models assume uniform azimuthal distribution of reduced waves which is truncated to 120° by the receiver antenna.

2 Coherence bandwidth (Bc) is defined as the frequency separation required between two transmitted carriers in order to reduce the correlation between their two respective squared envelopes to 0.5 (Ref: Jakes W C, 1974).

3 However, if the receiver sensitivity were higher, the importance of inter-symbol interference during deep fades would increase. Currently the receiver simply loses lock during these periods. If lock were maintained, lower amplitude signal components with longer delay times would have a greater influence on the received signal.
measurements, Section 4.5.2, and with the theoretical prediction of BER versus signal to noise ratio (SNR) presented in Section 5.2. However, the error rate of the system was found to be very high. There are two reasons for this:

i) Inefficient detection. The system is 9 dB worse in terms of BER for a given SNR than the optimum detection system (see Section 5.2).

ii) Poor code recovery loop. Section 6.4.3 discusses the short-comings of the code recovery loop.

Neither of these reasons for poor error performance of the loop is impossible to remedy and thus does not represent a major obstacle to developing a communication system in the millimetre waveband (Section 6.4.3). The main problem is the severe fading which is present in the environment.

It seems likely that any digital mobile radio system operating at these frequencies would employ some type of diversity reception. The results for the two frequency measurements and polarization diversity measurements indicate that a diversity system based on a frequency separation of the order 50-100 MHz, spatially separated antennae, or on orthogonal polarizations, would considerably reduce the received bit error rate (BER). A simulation of a two branch post detection frequency diversity system gave a performance improvement similar to that predicted for diversity applied to Rayleigh fading channels. Using this result, a two branch diversity system using the same modulation scheme is predicted to increase average BER from $10^{-2}$ to $10^{-4}$ at an average SNR of 20 dB. For 90% of the time, at 20 dB SNR, the error rate of the system is less than $10^{-3}$, whereas in the case of two branch diversity, for 90% of the time the error rate of the system will be reduced to less than $10^{-5}$.

Results suggest that the exact disposition of power in the micro-cell will be substantially influenced by the height of the other vehicles using the micro-cell in comparison with the height of the mobile vehicle's antenna because with the essential line of sight propagation conditions present in the micro-cell it is the other vehicles which are primarily responsible for causing obstruction to the line of sight signal. This situation could be improved by illuminating the micro-cell from antenna suspended above it. This would reduce the length of the shadows cast by obstructions in the micro-cell.

The results of the digital measurement suggest that a simple post detection diversity system using input channels de-correlated by space, frequency or polarization mechanisms would be able to operate in a micro-cell with an average error rate of approximately $10^{-3}$ at 15 Mb/s data rate. This approximation is made
on the assumption that a diversity system would yield nearly the same improvement to the digital error rate as a system operating in Rayleigh fading channels. Support for this assumption is provided by the results of a simulation of diversity reception using experimental channels in Section 4.4.

The conclusion which can be drawn from the work is that there is a real possibility of constructing a mobile communication system in the millimetre wave frequency band. The results from both the analogue and digital experiments show, however, that implementing such a system would not be a simple task and a working system would require some form of diversity reception to be employed to reduce the effects of multipath fading.
Chapter 9

Chapter 9  SUGGESTIONS FOR FUTURE WORK

There are a number of different questions, raised by the work carried out here, which are beyond the scope of this work and are unable to be addressed fully. The experiments carried out here generate a very large amount of data which the currently used computing facilities could not really deal with in a reasonable amount of time. In particular:

i) The simulation of diversity reception throughout the whole of a micro-cell would require very large amounts of data to be processed (approximately 100 Mb). This data could be used to predict the improvement in the BER performance of the link, with application of two branch diversity, throughout the micro-cell. This calculation would be very computer intensive as it would require integration of the BER versus SNR characteristic of the receiver over the distribution of the received signal at each location in the micro-cell.

ii) The experimental results indicate the extreme variability of the various statistical parameters used to describe the signal. For example: location variability, loss deviation, coherence bandwidth, the predicted antenna separation required to decorrelate the signal from two spatially separated antenna and the shape of the signal distribution at various locations throughout the cell. It would be interesting plot out the variation of each of these parameters throughout the cell. This would enable any trends in the channel parameters which are currently swamped by random variation to be identified. Of particular interest would be a plot of the estimated variation of the coherence bandwidth throughout the micro-cell against the strength of the LOS signal component estimated from the auto-correlation. However, such extensive analysis would require multiple data sets of approximately 100 Mb to be analysed.

In order to investigate the properties of the radio channel further, it would be instructive to construct a receiver capable of directly probing the channel with a bandwidth of approximately 1000 MHz. This could be accomplished in two ways:

i) The measurement of the channel response using the swept frequency measurement could be developed by increasing the sweep bandwidth, thus increasing the time resolution of the system.

ii) Alternatively, a digital spread spectrum system could be employed. Systems of this type have been used to probe channel responses at 30 GHz in urban environments (Ref: Violette et al, 1988).
An experimental apparatus of this description would allow the strength in the various LOS and specular components to be resolved and would, in addition, allow the time variation in the value of the coherence bandwidth to be determined more accurately. However, construction of such a system would entail considerable expense.
APPENDICES

APPENDIX I  INDEX OF EXPERIMENTAL RESULTS

This Appendix presents an index of the experimental results reported in this Thesis. It relates the locations on the magnetic tapes where the measurements are stored to the type of experiment, and for the two channel measurements it also gives the cross-correlation coefficient between the two channels. Four tables are presented:

i) The first presents an index of the two frequency measurements, reported in Section 4.5.1.

ii) The second presents an index of the two polarization measurements obtained along Torrington Place, reported in Section 4.5.4.

iii) The third presents an index of the two polarization measurements obtained along High Holborn, reported in Section 4.5.4.

iv) The fourth presents an index of the other measurements undertaken on the link, not summarized in the other tables.

The purpose of these tables is to enable future workers to identify the location of the data from which the results in this Thesis are derived. Only the data which is presented in the Thesis is indexed. Other experiments are indexed in the data books which were used at the time when the data was initially gathered.
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Table 1 - Index of Tape Locations for the Two Frequency Measurements Reported in Section 4.5.1.
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Table 2 - Index of Dual Polarization Measurements Along Torrington Place Reported in Section 4.5.4.

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<tr>
<td>Single Channel Signal Probing</td>
<td>1-B</td>
<td>1701</td>
<td></td>
</tr>
<tr>
<td></td>
<td>2-B</td>
<td>1713</td>
<td></td>
</tr>
<tr>
<td>Miscellaneous Two Frequency Measurements</td>
<td>B</td>
<td>2100</td>
<td>1(100)-B</td>
</tr>
<tr>
<td></td>
<td>B</td>
<td>2200</td>
<td>2(100)-B</td>
</tr>
<tr>
<td></td>
<td>B</td>
<td>2700</td>
<td>1(200)-B</td>
</tr>
<tr>
<td>Digital Experiment</td>
<td>L</td>
<td>1377-3364</td>
<td>A total of 16 runs</td>
</tr>
<tr>
<td></td>
<td>L</td>
<td>0-1160</td>
<td></td>
</tr>
<tr>
<td></td>
<td>M</td>
<td>1377-3364</td>
<td></td>
</tr>
<tr>
<td>Wide Area</td>
<td>M</td>
<td>1160</td>
<td></td>
</tr>
<tr>
<td></td>
<td>M</td>
<td>1825</td>
<td></td>
</tr>
<tr>
<td>Swept Frequency Measurements</td>
<td>O</td>
<td>0-244</td>
<td>32,000 individual sweeps</td>
</tr>
<tr>
<td>Twin Polarization Measurements, Measurement of all Four Coefficients</td>
<td>1(4P)-G</td>
<td>267</td>
<td>-</td>
</tr>
<tr>
<td></td>
<td>2</td>
<td>270</td>
<td></td>
</tr>
<tr>
<td></td>
<td>3</td>
<td>285</td>
<td></td>
</tr>
<tr>
<td></td>
<td>4(4P)-G</td>
<td>296</td>
<td></td>
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</table>

Table 4 - Miscellaneous Table. Summary of Tape Locations for the Measurements not Described in Tables 1, 2 and 3.
APPENDIX II COMPUTER PROGRAM LISTINGS

This Section describes the operation of the computer programs written by the author for the analysis of the experimental data obtained in this work [1]. The programs to calculate the power spectrum and to retrieve the car speed information are described in the Thesis presented by my colleague Dr G L Siqueira who wrote these programs in the early part of this work (Ref: Siqueira G L, 1989). A brief description of the aim of each program is presented followed by a simple introduction to its use. The programs are written to be "user-friendly" and so, in use, their application should not be too difficult. The data sampling programs are written around a "Burr-Brown Personal Computer Intelligent Instrumentation System" (Ref: Burr-Brown, 1985). This is an analogue to digital conversion board which is supported by its own applications language which is itself referenced from programs written by the author in "C".

i DATA SAMPLING

This program samples multiple segments of data from the tape recorder using the Burr-Brown data acquisition system. The aim of the program is to circumvent, in some measure, the limited sampling capacity of the IBM PC-AT computer which is limited to sampling data segments of a maximum size of 64 Kb. The sampled data segments are not contiguous because the computer takes a finite time to store each data segment. However, the delay is only about 3 seconds, depending on which drive is used to store the data. This is because the virtual drive "D:" is faster than the hard disk drive "C:" which in turn is faster than the floppy drives "A:" and "B:". In order to run the program, the computer must be initialized by running the "PCI20K-2" program which sets up the Burr-Brown data acquisition system. The program requests that you enter the sampling rate, the number of input channels and template names for storing the output files. The program creates a separate directory for storing the output files. Each request is presented to the user with the default option most commonly selected, displayed in order to facilitate easy use of the program.

1 To present the programs in this Thesis, the "\" delimeter used for indicating escape sequences in the "C" programming languages is replaced by a "/" delimeter.
/* M_recO: This program samples analog signals using the high speed acquisition mode. Consecutive segments are stored as the program proceeds together with time and speed information.

Written by H. Thomas - December 1987*/

#include "pci20k.h" /*for access to pci2000 calls */
#include "stdio.h" /*for standard i/o eg. files */
#include "conio.h"
#include "stdlib.h"
#include "string.h" /*for string manipulation */
#include "io.h"
#include "malloc.h" /*for dynamic memory allocation */
#include "direct.h" /*for directory control */
#include "process.h" /*for DOS system calls */

int quit_p (label)
char *label ;
{
  puts (label);
  system ("C:");
  exit (1);
}

int get_path ()
{
  int result;
  char pname[80], buf[80], c ;
  char *req, *path ;

  path = pname ;
  printf ("THE MAXIMUM NUMBER OF FULL DATA SEGMENTS ");
  printf ("WHICH CAN BE STORED IN D: IS %d", (896100/65536));
  printf ("DO YOU WANT TO CHANGE WORKING DIRECTORY TO D: ?");
  printf ("IF YES, TYPE 'Y' or 'y'.");
  c = getch () ;
  puts ("");
  if (c == 'Y'|c == 'y')
  {
    if (system ("D:")) == -1)
      quit_p ("system() FAILED") ;
    if (getcwd (buf, 5) == NULL)
      quit_p ("getcwd() FAILED") ;
    if ((result = mkdir (path)) == -1)
      quit_p ("MAKING DIRECTORY FAILED %s%s",buf, path);
    if ((result = chdir (path)) == -1)
      quit_p ("CHANGING TO NEW DIRECTORY FAILED");
    if (getcwd (buf,80) == NULL)
      quit_p ("getcwd() FAILED") ;
    printf ("NEW WORKING DIRECTORY IS %s", buf);
  }
  req = "/ENTER NAME OF NEW DIRECTORY TO STORE THE DATA FILES. /
  USE ONLY FORWARD SLASH DELIMITERS IN THE PATHNAME. /NB. ONLY /
  THE LAST NAME CAN BE A NEW DIRECTORY NAME/";

  if ((result = get_f (path,req)) == NULL)
    quit_p ("get_f() FAILED") ;
  if (getcwd (buf,18) == NULL)
    quit_p ("getcwd() FAILED") ;
  if ((result = mkdir (path)) == -1)
    quit_p ("/nMAKING DIRECTORY FAILED %s%s",buf, path);
  if ((result = chdir (path)) == -1)
    quit_p ("/nCHANGING TO NEW DIRECTORY FAILED") ;
  if (getcwd (buf,80) == NULL)
    quit_p ("/ngetcwd() FAILED") ;
  printf ("/nNEW WORKING DIRECTORY IS %s
", buf);
int get_f (name, request)
char name[];
char *request;
{
    int i, c_count, flag;
    char c;

    puts (request);
    flag = 1;
    while (flag)
    {
        gets (name);
        c_count = strlen (name);
        if (c_count > 20)
            cputs ("FILENAME TOO LONG/n");
        else
            flag = 0;
    }
}

error_routine (error_string)
char *error_string;
{
    int error_code;
    if ((error_code = errsys ()) != 0)
    {
        printf ("%s %d/n", error_string, error_code);
        quit_p("n");
    }
}
askint(s)
char s[];
{
    int i, num;
    gets (s);
    for (num = 0, i = 0; s[i] >= 'O' && s[i] <= '9'; i++)
        num = 10*num + s[i] - 'O';
    return (num);
}
data [32767];
main()
{
    int vect, segmt, rchn, chans[4], cnts[2], nchan, fsam;
    int mode, enable, nfile, i, j, nseg, q, radix = 10;
    float tt;
    int ct_chn, reset, dist[3], delta_S, S_dist;
    unsigned int nsam, count, ct_cnt;
    char c, s[20], fname[80], buffer[8];
    char *template, *result, *new_f;
    FILE *f1;

    vect = 0x60;
    setvec (vect);
    sysinit (vect);
    segmt = 0xC000;
    init (segmt);
    error_routine ("ERROR DURING INITIALIZATION");

    if ((i = get_path ()) == NULL)
        quit_p ("/nGET PATH FAILED");

    puts ("/n/" /*<<<<<<<<<<<<<<<<<<<<<<<<<<<<<<<<<<<<<<<<<<<*/);
    puts ("/n/" /*ENTER NUMBER OF CHANNELS IN EACH SEGMENT/n");
    nchan = askint(s);
    printf ("Number of channels = %d", nchan);

    puts ("/n/" /*ENTER NUMBER OF SAMPLES PER CHANNEL */);
    printf ("THE MAXIMUM NUMBER IS %u/n", (unsigned)(32768/nchan));
    count = askint(s);
nsam = nchan*count;
printf ("Number of samples per segment = %u/n", nsam);
if (nsam > 32768)
    quit_p ("/nTHE NUMBER OF SAMPLES IS TOO BIG");

rchn = 0;
puts ("/n/\nENTER SAMPLING FREQUENCY IN Hz/n");
fsam = askint(s);
cnts [1] = 4000000/(fsam*nchan);
printf("fsam = %d cts[1] = %d/n", fsam, cts [1]);
cnts [2] = 2;
con_rg (rchn, cts [1], cts [2]);
error_routine ("ERROR DURING CONFIGURATION OF RG");

enable = 1;
write_ch (RG_TYP, rchn, enable);
error_routine ("ERROR DURING WRITE RG");
for (i = 0; i < nchan; i++)
    chans [i] = i;
chans [nchan] = -1;

rchn = -1;
con_hs (rchn, cts, chans);
error_routine ("ERROR DURING CON_HS");
reset = 0;
ct_cnt = 0;
ct_chn = 0;
con_cntr (ct_chn, ct_cnt, reset);
error_routine ("ERROR DURING CON_CNTR");
write_ch (CT_TYP, ct_chn, reset);

r = "/n/\nENTER FILE TEMPLATE/n";
if ((template = calloc(nchan,sizeof(fname))) == NULL)
    quit_p ("/nCalloc FAILED");
puts ("/n/\nNOW ASSIGNING VALUES TO TEMPLATE; ");
puts ("ENTER ONE TEMPLATE FOR EACH CHANNEL");
for (i = 0; i < nchan; i++)
{
    if ((get_f (template[i])) == NULL)
        quit_p ("/nget_f FAILED");
    template = template + 80;
}
template = template - nchan*80; /*RESET POINTER TO BEGINNING OF ARRAY*/

puts ("/n/\nENTER NUMBER OF SEGMENTS TO BE SAMPLED/n");
printf ("THE MAXIMUM IS %d IF DRIVE D: IS USED/n/n", 448050/nsam);
nseg = askint(s);
printf ("Number of segments = %d", nseg);

puts ("/n/\nCONNECT RATE GENERATOR OUTPUT TO COUNTER CHANNEL 3 INPUT./n");
puts ("/n/To start conversion type 'Y' or 'y'/n/n");
c = getch();
if (c == 'Y' || c == 'y')
{
    puts ("/n/\nWAIT APPROXIMATELY %7.1f sec FOR ", tt);
    printf ("CONVERSION AND STORAGE./n");
    q = 1;

    reset = 0;
    ct_chn = 0;
    dist[0] = read_ch (CT_TYP, ct_chn, reset);
for (; q++)
{
    if (q == (nseg + 1))
        break;
    printf ("\nSAMPLING SEGMENT %d ", q);

    ct_chn = 0;
    dist[1] = read_ch (CT_TYP, ct_chn, reset);

    run_hs (rchn, data, count);
    error_routine ("error occurred during data acquisition");
    puts ("\nSAMPLING FINISHED");

    ct_chn = 0;
    dist[2] = read_ch (CT_TYP, ct_chn, reset);

    printf ("\n{dist 0} = %d", dist[0]);
    printf ("\n{dist 1} = %d", dist[1]);
    printf ("\n{dist 3} = %d", dist[2]);

    S_dist = dist[0] - dist[1];
    printf ("\nS_dist = (%d - %d) = %d", dist[0], dist[1], S_dist);
    delta_S = dist[1] - dist[2];
    printf ("\ndelta_S = (%d - %d) = %d", delta_S);

    itoa (q, buffer, radix);
    r = strcat (buffer, ".dat");
    printf ("\nSAVING SEGMENT %d/n", q);
    for (j = 0; j < nchan; j++)
    {
        result = strcpy (fname, template);
        new_f = strcat (result, r);
        if ((fl = fopen (new_f, "wb")) == NULL)
            quit_p ("\nI can't open %s\n", new_f);
                exit (1);
        rewind (fl);
        putw (S_dist, fl);
        putw (delta_S, fl);
        if (ferror (fl))
            perror ("putw failed");
            clearerr (fl);
        } /* SAVING CHANNEL j IN FILE new_f*/
        for (i = j; i < nsam; i = i + nchan)
        {
            putw (data [i], fl);
            if (ferror (fl))
            {
                perror ("putw failed");
                clearerr (fl);
            }
        }
        template = template + 80;
        fclose (fl);
    }
    template = template - nchan*80;
}

enable = 0;
rchn = 0;
ct_chn = 0;
write_ch (CT_TYP, ct_chn, enable);
write_ch (RG_TYP, rchn, enable);
system ("C: ");
This program calibrates the sampled data. This is output from the analogue to digital converter as a series of 12-bit numbers. The program relates these integer values to real numbers corresponding to the recorded signal power. The program has a number of different options for calibrating the data in a linear manner, converting the data from linear units into decibel units, or re-calibrating data output from a logarithmic amplifier into decibel units. The program has the facility to calibrate multiple data files with similar names by entering file names with wild-card characters, for example the "?" and "*" characters. In addition to producing a calibrated file with a " .CBR" extension name, it also produces a decimated file with a " .DEC" extension name. This file can be plotted using normal graphic packages because the number of data points is averaged down from in excess of 16 K to of the order of 500. The program makes simple requests for the file name, the directory in which to store the output files and the sample frequency. If the re-calibration of data from a log-amplifier option is selected, the program requests the various parameters of the log-amplifier conversion.

/* CONVERT2.C Program to calibrate binary data files from the A/D converter into real voltage levels. The program provides a facility for calculating the natural log of the data, and also, a facility for re-calibrating log format data from an analogue log conversion IC.

NB the name of the file to be calibrated may be entered on the command line to save time.

Written by H Thomas - November 1989*/

#include "stdio.h"
#include "dos.h"
#include "errno.h"
#include "conio.h"
#include "math.h"
#include "string.h"
#include "ctype.h" /*lower case character conversions */
#include "process.h"
#include "alloc.h"
#include "string.h"
#include "io.h"
#include "dir.h"

int get_pause (r)
char *r ;
{
    void clear_key(void);
    clear_key();
    if (r != NULL) puts (r);  
    if (isspace (getch())) return(1); 
    return (0);
}

void get_question(info_request)
char *info_request ;
{
    char *request ;
    request = "/

void get_info (info, length, name)
float *info;
char *name;
long length;
{
    int test;
    char *info_request, *request;
    void get_question(char *question);
    int get_pause(char *r);
    float a, b;

    info_request = (char *)malloc(1000* sizeof (char));
    get_question(info_request);
    clrscr();
    printf(" You are entering information for processing %s/n", name);
    puts("/n/n/
ENTER A/D BOARD SENSITIVITY /n/n/ 
DEFAULT VALUE IS +/- 5 V /n");
    puts(info_request);
    if(get_pause(NULL)) info [0] = 2.0;
    else {
        clrscr();
        printf(" You are entering information for processing %s/n", name);
        test = 1;
        do {
            puts("/n/n/
ENTER: 
1 FOR +/- 2.5 V/n/
2 FOR +/- 5.0 V/n/
3 FOR 0 -> 5 V/n/
4 FOR 0 -> 10 V/n/");
            scanf("%f", &info [0]) ;
            if (info[0] != 1 && info[0] != 2 && info[0] != 3 && info[0] != 4)
                puts("ENTER 1, 2, 3 OR 4 !!");
            else test = 0 ;
        } while(test);
    }
    a = 5.0;
    b = 2.5;
    if(info[0]==2.0) {a = 10.0; b = 5.0;)
    if(info[0]==3.0) {a = 5.0; b = 0.0;)
    if(info[0]==4.0) {a = 10.0; b = 0.0;)
    info[0] = a ;
    info[1] = b ;
    clrscr();
    printf(" You are entering information for processing %s/n", name);
    puts("/n/n/
ENTER GAIN AND OFFSET VALUES
DEFAULT VALUES ARE:
GAIN = 1
OFFSET = 0
puts(info_request);
if (get_pause(NULL)) {
    info[2] = 1.0;
    info[3] = 0.0;
} else {
    clrscr();
test = 1;
    while(test) {
        printf("You are entering information for processing %s/n", name);
        puts("ENTER GAIN VALUE");
        scanf("%f", &info[2]);
        clrscr();
        if(info[2]==0.0) puts("ENTER NON-ZERO GAIN VALUE !!");
        else test = 0;
    }
    puts("ENTER OFFSET VALUE");
    scanf("%f", &info[3]);
    clrscr();
    printf("You are entering information for processing %s/n", name);
    puts("Do you want dB conversion? /
DEFAULT IS LINEAR CONVERSION
puts(info_request);
if(get_pause(NULL)) info[4] = 0.0;
else info[4] = 1;
clrscr();
if(info[4]) info[5] = 0;
else {
    clrscr();
    printf("You are entering information for processing %s/n", name);
    puts("Do you want to calibrate log data? /
DEFAULT IS LINEAR CONVERSION
puts(info_request);
if(get_pause(NULL)) info[5] = 0;
else {
    info[5] = 1;
    clrscr();
    printf("You are entering information for processing %s/n", name);
    puts("The assumed form of the log conversion function is:
Vout = A.log(Vin + C) + B
Where
A = log conversion factor
B = input offset to log amp
C = output offset to log amp
THE CALIBRATION ROUTINE IS THE INVERSE OF THIS FUNCTION");
test = 1;
while(test) {
    puts("ENTER <A>");
    scanf("%f", &info[6]);
    if(info[6]>=0.0) puts("ENTER NEGATIVE NON-ZERO <A> VALUE !!");
    else test = 0;
}
puts("ENTER <B> VALUE");
scanf("%f", &info[7]);
test = 1;
clrscr();
while(test) {

- 301 -
printf("You are entering information for processing %s/n", name);  
puts("/n/n/nENTER THE <0 dB> CALIBRATION VALUE");  
scanf("%f", &info[9]);  
puts("/n/n/nENTER THE <-40 dB> CALIBRATION VALUE");  
scanf("%f", &info[8]);  
if(info[8]<info[9])  
  puts("THE <0 dB> VALUE IS BIGGER THAN THE <-40 dB> VALUE");  
else  
  test = 0;  
}  
info[8] -= info[7];  
info[8] /= info[6];  
info[8] = pow((double)10, (double)info[8]);  
/*This is the value C in the equation above it represents  
rectified HF noise DC offsets on the logamp input etc.*/  
info[9] -= info[7];  
info[9] /= info[6];  
info[9] = pow((double)10, (double)info[9]);  
info[9] -= info[8];  
info[9] = 10*log10((double)info[9]);  
}  
clrscr();  
printf("You are entering information for processing %s/n", name);  
request ="/n/n/n/ENTER NUMBER OF POINTS FOR DECIMATION FILE /n/n/DEFAULT IS %d /n/n/";  
printf(request, (int)(length/2/100));  
puts(info_request);  
if(get_pause(NULL)) info[10] = 100;  
else  
  clrscr();  
  printf("You are entering information for processing %s/n", name);  
  printf(request, (int)(length/2/100));  
  scanf("%f", &info[10]);  
  info[10] = (int)(length/2/info[10]);  
}  
clrscr();  
printf("You are entering information for processing %s/n", name);  
request ="/n/n/n/ENTER SAMPLE FREQUENCY /n/THIS IS USED TO CALIBRATE SCALE ON DEC FILE /n/n/DEFAULT IS %d /n/n/";  
printf(request, (int)info[11]);  
puts(info_request);  
if(get_pause(NULL)==0)  
  clrscr();  
  printf(request, (int)info[11]);  
  scanf("%f", &info[11]);  
}  
clrscr();  

float cal(float info, float data)  
float *info;  
int data;  
{  
  return (((4095 - data)*info[0]/4096 - info[1])/info[2] - info[3]);  
}  

void nl_cal(float info, float data)  
float *info;  
float *data;  
{
data = Vout
info[6] = dB conversion gain <A>
info[7] = input offset <B>
info[8] = ~noise value <C>
info[9] = dB offset dB
*/
*data -= info[7];
*data /= info[6];
*data = pow((double)10, (double)*data);
*data -= info[8];
if(*data<0.001) *data = -3.0;
else *data = 10*log10((double)*data);
*data -= info[9];
}
main(argc, argv)
int argc;
char *argv[];
char *name,  *request, *old_name, *current_file;
int i, k, num, loop=0, *input, repetition=0;
long length, position;
float *info, max=-100, min=100, *calibrated, value;
double av=0.0, sq=0.0;
FILE *fl, *f2, *f3, f4;
void get_info(float *info, long length, char *name);
void get_f(char *name, char *request);
float cal(float *info, int input);
void nl_cal(float *info, float *calibrated);
int matherr(struct exception *e);
struct ffblk file;
drive = (char *)calloc(5,sizeof(char));
dir = (char *)calloc(80,sizeof(char));
drive_out = (char *)calloc(5,sizeof(char));
dir_out = (char *)calloc(80,sizeof(char));
name = (char *)calloc(80,sizeof(char));
request = (char *)calloc(80,sizeof(char));
namel = (char *)calloc(80,sizeof(char));
ext = (char *)calloc(5,sizeof(char));
calibrated = (float *)calloc(8192,sizeof (float));
info = (float *)calloc(15,sizeof(float));
input = (int *)calloc(8192,sizeof(int));
old_name = (char *)calloc(80,sizeof(char));
k = 1;
if(argc[1]!=NULL) strcpy(name, argv[1]); k=0;

/*Loop back to here to begin processing on the next file*/
NEXT_FILE:
if(repetition) {
  if(findnext(&file)!=0) {
    if(erno==ENMFILE || erno==ENOENT)
      goto FINISH ;
exit(1);
  }
  fnsplit(old_name, drive, NULL, NULL, ext);
  strset(name, '/0');
  strset(namel, '/0');
  fnsplit(file.ff_name, NULL, NULL, namel, ext);
  fnmerge(name, drive, dir, namel, ext);
  fl = fopen(name, "rb");
  if(fl==NULL) exit(1);
  goto START ;
}
do {
  clrscr();
}
if(k)
{
    request = "\nEnter drive, path, and file name of input data /n/
/leave no spaces !!/n";
    get_f(name, request);
}
fnsplit(name, drive, dir, name1, ext);
findfirst(name, &file, FA_DIREC);
fnmerge(name, drive, dir, file.ff_name, NULL);
fnsplit(name, drive, dir, name1, ext);
fnmerge(name, drive, dir, name1, ext);

f1 = fopen(name, "rb");
if(f1==NULL) puts("RE-TYPE, OR TRY ANOTHER FILE NAME !");
k = 1;
} while(f1==NULL);

/*Begin processing subsequent data files*/
START:
if((length = filelength(fileno(f1)))==-1) exit(1);
clrscr();
strcpy(old_name, name);
current_file = "File being processed is ";
printf("%s %s/n", current_file, old_name);
do
{
    if(repetition) {
        puts("/
/
***************
ENTER DRIVE/n/
FOR STORING OUTPUT FILES /n")
        scanf("%s", drive_out);
        puts("/
/
ENTER DIRECTORY/n/
FOR STORING OUTPUT FILES /n "
        scanf("%s", dir_out);
        fnmerge(name, drive_out, dir_out, name1, ".cbr");
        if((f2 = fopen(name, "wb")) == NULL) puts("RE-TYPE DRIVE SPECIFICATION !")
        if((f3 = fopen(name, ".dec")) == NULL) puts("RE-TYPE DRIVE SPECIFICATION !")
        if((f4 = fopen(name, ".inf")) == NULL) puts("RE-TYPE DRIVE SPECIFICATION !")
    } while(f2==NULL || f3==NULL || f4==NULL);
    if(!repetition) {
        printf("/n/file is %s/nis %ld bytes long/n", old_name, length);
        get_info(info,length,old_name);
    }
    clrscr();
    loop = 0;
    while((long)loop*8192 < length/2) {
        print("/n/file is %s/nis %ld bytes long/n", old_name, length);
        print("/n/reading data section %d/t from file position %ld"");
        print("/n/process data section %d/n", loop);
        for(i=0, value=0, k=1; i<length; i++)
        {
            convert binary data to real numbers (input voltage)*/
            calibrated[i] = cal(info, input[i]);
            /*find maximum and minimum values of input voltage*/
            if(calibrated[i]<min) min = calibrated[i];
            if(calibrated[i]>max) max = calibrated[i];
            /*calculate average and variance of data*/
            av += calibrated[i];
        }
    }
}
calibrated[i] = cal(info, input[i]);
/*find maximum and minimum values of input voltage*/
if(calibrated[i]<min) min = calibrated[i];
if(calibrated[i]>max) max = calibrated[i];
/*calculate average and variance of data*/
av += calibrated[i];
sq += calibrated[i]*calibrated[i];
/*convert data to dB*/
/*re-calibrate data from analogue log amp*/
if(info[4])
calibrated[i] = 10*log10((double)calibrated[i]);
else
if(info[5]) nl_cal(info, &calibrated[i]);

/*calculate value for decimated data file*/
if(info[4] && info[5])
value += pow((double)10,(double)(calibrated[i]/10));
else value += calibrated[i];

/*write decimated data value to output file*/
if(i/info[10]==k) {
value /= info[10];
if(info[4] && info[5])
value = 10*log10((double)value);
fprintf(f3, "%+9.4f %+9.4f/n", ((float)i+loop*8192)/info[11], value);
value = 0;
k++;
}
/*write trial values to screen to provide visual check on
program operation*/
for(i=1000; i<8001 && i<num; i+=1000)
printf("DATA[%ld] = %+9.6f", (long)loop*8192 + i, calibrated[i]);

/*position file output pointer*/
position = lseek(fileno(f2), (long)loop*8192*sizeof(float), SEEK_SET);
printf("WRITING DATA SECTION %d/n", loop);

/*write calibrated data to output file*/
num = fwrite((char *)calibrated, sizeof(float), num, f2);
loop++;
}

/*generate processed data information file*/
printf("PROCESSING %s FINISHED NUMBER %d/n", old_name, repetition);
av /= (float)length;
sq /= (float)length;
clrscrO;
printf(" %s %s"
current_file, old_name);
printf("n/THE FOLLOWING QUANTITIES ARE FOR/n/THE DATA BEFORE ANY LOG CONVERSION/n/n/
THE AVERAGE IS %f/n/
THE VARIANCE IS %f/n/
THE MINIMUM VALUE IS %f/n/
THE MAXIMUM VALUE IS %f/n/
", av, sq, min, max);
fprintf(f4, "THIS FILE CONTAINS THE PROCESSING INFO ON FILE %s/
/n/THE FILE IS %ld BYTES LONG/n", old_name, length);
fprintf(f4,"/n/THE FOLLOWING QUANTITIES ARE FOR/n/THE DATA BEFORE ANY LOG CONVERSION/n/n/
THE AVERAGE IS %f/n/
THE VARIANCE IS %f/n/
THE MINIMUM VALUE IS %f/n/
THE MAXIMUM VALUE IS %f/n/
", av, sq, min, max);
request = "THE CODE FOR THE INFORMATION ARRAY IS/n/
info[0] = A/D rangeinfo[6] = Log gain <A>/n/
fprintf(f4, "%s", request);
request = "THE VALUES OF THE INFORMATION ARRAY ARE\n";
fprintf(f4, "%s", request);
request = " Info[%d] = %f Info[%d] = %f\n";
for(i=0; i<6; i++)
  fprintf(f4, request, i, info[i], i+6, info[i+6]);
if(closeall()==EOF) puts("FILE ALLOCATION ERROR ON closeall()\")
  repetition ++;
  goto NEXT_FILE;
FINISH:
puts("ALL FILES PROCESSED\")
putch('a')
return(1)
This Section describes the computer programs which were used to analyse the statistics of the local average of the signal strength and of the "fast fading" envelope of the signal. The description of the signal in terms of these two processes was presented in Chapter 2.

The first computer program (AVERAGE2.C), described here, calculates the probability density function of the signal and its cumulative distribution. It also produces a decimated version of the input data file to allow the data to be processed by a standard computer graphics package. The next stage of the program calculates the local average of the signal using a moving average algorithm. The probability density function and cumulative distribution of the local average signal are then calculated. The final stage of the program divides the input data by the estimate of the local average to produce a "normalized" signal which is representative of the "fast fading" of the signal envelope. The program then produces the probability density function and cumulative distribution of the "fast fading" of the signal. The program also produces decimated time series of the local average and the "normalized" signals.

/* AVERAGE2.C Program to calculate the moving average of data files calibrated in relative dB's and generate probability functions.
Written by H Thomas - November 1989*/

#include "stdio.h"
#include "errno.h"
#include "conio.h"
#include "math.h"
#include "string.h"
#include "ctype.h" /*lower case character conversions */
#include "alloc.h"
#include "dir.h"
#include "dos.h"
#include "alloc.h"
#include "stdlib.h"
#include "io.h"

main(argc, argv)
int argc;
char *argv[];
{
  int i, j, k, l, m, max_point;
  int no_bin, sample_frequency, point_num1, decimation;
  int file_type, N, iii ;
  int L, new_max, filter;
  long int L2 ;

  char *string[2], *request[2], *r, *exp_no1, *date1, *output_drive ;
  char *extension[100], *name, *name2, *n, *output_dir ;

  char *name3, *c, *drive, *dir, *ext ;
  int num, array[4], len ;
  struct ffblk file ;
float far *X, data, xdata, speed, Xmax, Xmin, X50, X90, X84;
float step, level, dynamic_range, distance, loss_deviation;
float time, location_variability, *ma;
double average;

FILE *p[10], *pe, *temp;
void *vector(int, int);
void get_pause(char *r);
void get_domain(int dom[], int length, int points);
void get_f(char *r, char *name);
float get_info(char *r);
int choice(void);
drive = (char*)malloc(sizeof(char)*3);
dir = (char*)malloc(sizeof(char)*40);
output_dir = (char*)malloc(sizeof(char)*40);
output_drive = (char*)malloc(sizeof(char)*40);
ext = (char*)malloc(sizeof(char)*5);
exp_nol = (char *)calloc(20, sizeof(char));
date1 = (char*)malloc(sizeof(char)*20);
name = (char*)malloc(sizeof(char)*80);
name2 = (char*)malloc(sizeof(char)*80);
name3 = (char*)malloc(sizeof(char)*12);

/*This is the main processing loop run once for every file in <name>*/

iii = 0;
for(;;) {

/*This is the section which requests the file information and the outline file name which may include question marks*/

if(iii==0) {

/*This chooses whether the input file is BINARY or ASCII*/

while(1) {
    r = "ENTER FILE FORMAT, TYPE/n/n/n/
    /  'F' FOR FORMATTED/n/
    /  'B' FOR BINARY/n";
    clrscr();
    puts(r);
    i = getch();
    if(i=='F'| i==') {file_type=0; break ;}
    if(i=='B'| i=='b') {file_type=1; break ;}
}

puts("ENTER DRIVE FOR STORING OUTPUT DATA");
gets(output_drive);
puts("ENTER DIR FOR STORING OUTPUT DATA");
gets(output_dir);

printf ("/n file_type ....... %d", file_type);

/*This sets up the format string for fopen() accordingly*/
if (file_type) string[0] = "rb";
else string[0] = "r";
string[1] = "w";

/*Select the filter function shape required the direct calculation of the triangular shape is very slow. The procedure repeating the square filter operation twice is more efficient*/
clrscr();

t = "/

ENTER THE FILE SHAPE REQUIRED DEFAULT = 2/n/n/
ENTER ...... 0 FOR TRIANGULAR SHAPE/n/
ENTER ...... 1 FOR SQUARE SHAPE/n/
ENTER ...... 2 FOR 2*SQUARE SHAPE(=TRIANGULAR)/n/";
while(1) {
  puts(r);
  filter = getche();
  filter = '0';
  if(filter==0 | filter==1 | filter==2) break;
  clrscr();
}
c1rscr();
/*This sets up the maximum number of points to be analysed from the
input file*/
r = "ENTER NUMBER OF POINTS TO ANALYSE/nDEFAULT = 16384";
max_point = (int)get_info(r);
clrscr();
/*This checks if a file name has been entered on a command line*/
if(argc>1 && argv[1]!=NULL) strcpy(name, argv[1]);
/*This requests a file name if none was entered on the command line*/
else {
  puts("/nENTER FILE NAME INCLUDING ?")
  gets(name);
}
/*This section requests the outline file name
and checks whether it exists */
fn split(name, drive, dir, name2, ext);
while(findfirst(name, &file, FA_DIREC)==-1) {
  request[0] = "/nERROR OCCURRED, I CAN'T FIND ";
  strcat(request[0], file.ff_name);
  perror(request[0]);
  puts("/nENTER RLE NAME INCLUDING ? CHARACTERS")
  gets(name);
  fn split(name, drive, dir, name2, ext);
}
/*End of file name exits loop
/*This code gets the characters from the input file which
are to remain fixed in the output file*/
printf("/nTHE FILE NAME IS %s", file.ff_name);
len = strlen(file.ff_name) - strlen(ext);
/*Write the file name on the screen with character numbers beneath it*/
puts("/n");
p uts(file.ff_name);
for(i=0; i<len; i++) printf("%c", i + '0');
/*Read in characters to be kept fixed*/
puts("/nCHOOSE FOUR CHARACTERS TO KEEP THE SAME IN THE FILE NAME/
/nINDICATE CHARACTERS BY ARRAY NUMBER");
for(i=0; i<4;) {
  num = getche();
  if(isdigit(num)) {array[i] = num - '0'; i++ ;}
  clrscr();
/*Generate base for output file name, other bits will
be appended onto this to create the various output
file names as they are required*/
c = "c";
strcpy(name3, c);
for(i=0; i<4; i++) strcat(name3, file.ff_name+array[i], 1);
/*Enter number of bins for calculating PDF*/
r = "ENTER NUMBER OF SIGNAL LEVEL BINS, DEFAULT = 45";
no_bin = (int)get_info(r);
if (!no_bin) no_bin = 45;

/*Enter decimation factor for calculating plotting files*/

r = "ENTER DECIMATION FACTOR FOR TIME SERIES OUTPUT = 50";
decimation = (int)get_info(r);
if (decimation) decimation = 50;

/*Enter approximate speed of mobile vehicle*/

r = "SPEED OF VEHICLE, DEFAULT = 4 m/s";
speed = get_info(r);
if (!speed) speed = 4;

/*Enter sampling frequency for calculating time axes on plots*/

r = "SAMPLING FREQUENCY, DEFAULT = 10,000 /s";
sample_frequency = (int)get_info(r);
if (!sample_frequency) sample_frequency = 10000;

/*Enter number of wavelengths to be used for moving average*/

r = "NUMBER OF WAVELENGTHS FOR MOVING AVERAGE, DEFAULT = 40";
N = (int)get_info(r);
if (!N) N = 40;

/*Assign space for storing input and output information*/

if((max_point>16383) || (max_point==0)) max_point = 16383;
X = (float *)malloc(max_point*sizeof(float));
ma = (float *)malloc(max_point*sizeof(float));

/*L = length of the moving average filter */
L = (int)N*5.5*(sample_frequency/(1000*speed));

/*write file names and loop increment to a file for examination
of program function*/

pe = fopen("info.inf", "a");
if(ferror(pe)!=0) exit(1);
fprintf(pe,"/niii = %d/t/tname = %s", iii, file.ff_name);

/*Generate bases for output file names subsequent file names
output file name base is stored as !!!!<name3>!!!*/
if(iii>0) {
if(findnext(&file)!=0) {
  perror("/nEXIT PROGRAM FROM <findnext()>");
  if(errno==ENMFILE) {
    puts("/nERRNO == ENMFILE/n/nNO MORE FILES!");
    goto finish;
  }
}
exit(1);
}
strcpy(name3, c);
for(i=0; i<4; i++) strcat(name3, file.ff_name+array[i]);

fnmerge(name2, drive, dir, file.ff_name, NULL); p[0] = fopen(name2, string[0]);
if (ferror(p[0])!=0) exit(1);

/*SET THE FILE POINTERS TO THE START VALUE*/
if (!file_type) 
  fscanf(p[0], "%s %s %d", exp_no1, date1, &point_num1);
/*The position of the pointer after reading the file information
is offset by 33 units from the origin */

clrscr();
printf("/n/n/n/n/n/n/

* 
* LOADING DATA FROM */n/
* */n/

/n%n%s", file.ff_name);

/*Read data according to file type*/

if(file_type) fread((void *)X, sizeof(float), max_point, p[0]);
else
  for(i=0; i<max_point; i++)
    if (fscanf(p[0], "%f", &X[i]) == EOF) exit(1);
    fclose(p[0]);

if(iii==0) {
  /*NB the calibrated signals are sometimes input
   as positive attenuations thus the actual signal
   levels are equal to minus X[i].*/
  r = "/n/
IS DATA ENTERED AS POSITIVE ATTENUATION VALUES OR/n/
AS RELATIVE POWER IN dB ?/n/n/
FOR RELATIVE POWER TYPE ........... 0/n/
FOR ATTENUATION TYPE ............. 1/n/
IF NOT SURE TYPE ........... 2/n"
  while(1) {
    clrscr();
    puts(r);
    num = getche();
    num -= '0';
    if(num==0 | num==1 | num==2) break;
  }
  if(num==2) {
    clrscr();
    for(i=0; i<15; i++) {
      printf("X[%d] = %f / t X[%d] = %f / t X[%d] = %f/n", i, X[i], i+10, X[i+10], i+20, X[i+20]);
    }
    r = "/n/
IS DATA ENTERED AS POSITIVE ATTENUATION VALUES OR/n/
AS RELATIVE POWER IN dB ?/n/n/
FOR RELATIVE POWER TYPE ........... 0/n/
FOR ATTENUATION TYPE ............. 1/n"
  while(1) {
    puts(r);
    num = getche();
    num -= '0';
    if(num==0 | num==1) break;
    clrscr();
  }
}
clrscr();
if(num) for(i=0; i<max_point; i++) X[i] *= -1;

clrscr();
printf("/n/n/n/n/n/n/

* 
* CONVERTING DATA TO LINEAR */n/
* LINEAR UNITS & CALCULATING */n/
* AVERAGE */n/
* */n/
* */n/

/n%n%s", file.ff_name);

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/*This converts data from dB’s to linear units*/
for(i=0, average=0; i<max_point; i++) {
    X[i] = pow((double)10, (double)X[i]/10);
    average += X[i];
} 
average /= max_point;

/*open output files*/
exension[0] = "inf"; /*text record of file information*/
exension[1] = "pdf"; /*probability density function*/
exension[2] = "cpd"; /*cumulative probability density*/
exension[3] = "tim"; /*signal level vs time*/
exension[4] = "dis"; /*signal level vs distance*/
puts("/


");
strset(name2, '/O ');
fnmerge(name2, output_drive, output_dir, name3, extension[0]);

p[0] = fopen(name2, "a");
if(ferror(p[0])!=0) {
    printf ("I can't open %s
",name2);
    exit(1);
}

for (i=1; i<5; i++) {
    strset(name2, '/O ');
    fnmerge(name2, output_drive, output_dir, name3, extension[i]);
    p[i] = fopen (name2, string[1]);
    if(ferror(p[i])!=0) {
        printf ("I can't open %s/n",name2);
        exit(1);
    }
}

c1rscrO;
c1rscrO;


*/NORMALIZING WITH RESPECT TO */
*/AVERAGE*/
*/


*/


*/


*/divide by the average value of the whole data set*/
data = 0;
for(i=0, Xmin=100, Xmax=-100, k=1; i<max_point; i++) {
    X[i] /= average;
data += X[i];
    if (i/decimation == k) {
data /= decimation;
time = (float)decimation*k/sample_frequency;
k++;
    fprintf(p[3], "%+9.6f %+9.6f/n",data, time);
    fprintf(p[4], "%+9.6f %+9.6f/n",data, time*speed);
data = 0;
}
}

if (X[i] < Xmin) Xmin = X[i]; /*calculate dynamic range*/
if (X[i] > Xmax) Xmax = X[i];

Xmax = 10*log10((double)Xmax);
Xmin = 10*log10((double)Xmin);
dynamic_range = Xmax - Xmin;
step = dynamic_range/no_bin;
distance = ((float)max_point/sample_frequency)*speed;
fclose(p[3]);
fclose(p[4]);

fprintf(p[0], "/n/n/
aver age = %fdB/n/
dynamic range = %f dB/n",
average, dynamic_range );

fprintf(p[0], "%f dB/n/Xmin = %f dB/n/Xmax = %f dB/
/tcar speed = %f m/s/n/sample frequency = %d Hz/
/distance travelled = %f m/n",
    Xmin, Xmax, speed, sample_frequency, distance);
clrscr0;

printf ( " /
/ 
/ 
/ 
/ 
/ 
%
/* WRITING DATA TO * /
*/

/temporary storage * /
%
/* *
/ 
/ 
/ 
/ 
/ 
/
%s", file.ff_name);

*/Temporary storage of normalized linear data in D:TEMP.DAT*/
 temp = fopen("D:temp.dat", "wb");
if(ferror(temp)!=0) exit(1);
fwrite((void *)X, sizeof(float), 16383, temp);
fclose(temp);
clrscr0;

printf ("/n/n/n/n/n/n/
/ 
/ 
/ 
/ 
/ 
/
%s", file.ff_name);

*/This converts data back to dB's here*/
for(i=0; i<max_point; i++)
    X[i] = 10*log10((double)X[i]);
clrscr0;

printf ("/n/n/n/n/n/n/
/ 
/ 
/ 
/ 
/ 
/
%s", file.ff_name);

*/CALCULATING PDF & CPD*/
for (m=0; m<no_bin; m++) {
    level = m*step + Xmin;
    for (i=0, k=0, l=0; i<max_point; i++)
        if (X[i] < (level + step))
            k++;
        if (X[i] >= level) l++;
    xdata = (float)k/max_point;
    if (xdata < 0.1) X90 = level;
    if (xdata < 0.16) X84 = level;
    if (xdata < 0.5) X50 = level;
    fprintf (p[1], "%+9.6f %+9.6f/n", (float)l/max_point, level);
    fprintf (p[2], "%+9.6f %+9.6f/n", xdata, level);
}

loss_deviation = X50 - X90;
location_variability = X50 - X84;
fprintf(p[0], "/loss deviation = %f/n", loss_deviation);
fprintf(p[0], "/location variability = %f/n", location_variability);
for(i=1; i<5; i++) fclose(p[i]);
printf("/n/READING DATA FROM /n/TEMPORARY STORAGE /n/
/n%s", file.ff_name);

/*Read data back into array X after temporary storage*/
temp = fopen("D:temp.dat", "rb");
if(ferror(temp)!=0) exit(1);
fwrite((void*)X, sizeof(float), 16383, temp);
cfclose(temp);

clrscr();
printf("/n/"CALCULATING MOVING AVERAGE /n/
/n%s", file.ff_name);

/*This routine calculates the moving average of the data
using a triangular weighting function with overall
length L2. The quantity abs(j-i)/L2 is the weight of
each element. The external for(;;) loop scans the whole
data set, whilst the internal for(;;) loop calculates the
moving average at each location.*/

if(filter==0) {
    clrscr();
    printf("/n/"TRIANGULAR FILTER /n/
/n%s", file.ff_name);
    
    /*Triangular filter*/
    L2 = (long)2*L*(L+1);
    for(i=0; i<max_point; i++) {
        for(j=i-L, average=0; i<i+L; j++) {
            if(i>max_point) j = (long)2*max_point - j ;
            average += X[j]*abs(j-i);
        }
        ma[i] = average/L2 ;
    }
}
if(filter>0) {
    /*Square filter*/
    clrscr();
    printf("/n/"1st SQUARE FILTER /n/
/n%s", file.ff_name);
    
    /*Calculates the value of the first moving
    average, the integration is folded about the zero point
    from -L/2 to L/2*/
    
    /*Calculate first value of moving average*/
    for(i=(-L/2), average=0; i<L/2; i++)
        average += X[abs(i)];
ma[0] = average/L;

for(i= -(L/2+1; i<max_point; i++) {
    average = 0;
    average -= X[abs(i)];
    /*This folds the data over at end of file*/
    if(i+L>max_point) average += X[max_point-i-L+max_point];
    else average += X[i+L];
    average /= L;
    ma[i+L/2] = average + ma[i+L/2 - l];
}

if(filter>1) {
    clrscrO;
    printf("/n/READING DATA FROM /n/
    READING DATA FROM /n/"
    file.ff_name);

    /*Swapping the information in the data array <X>
    for that in the moving average array <ma>* /
    for(i=0; i<16383; i++) X[i] = ma[i];

    /*This loop repeats the moving average operation with
    another one giving an overall filter with a triangular
    filter function; is the result of the convolution of
    two square functions.*/

    /*Note the filter folds around the origin so that the
    no points are lost at the beginning or end of the data
    file. Thus, output file length is the same as the input
    file length*/

    /*Assign storage area to calculate the moving average
    of the moving average data*/

    /*Calculate first value of moving average*/
    for(i=(L/2), average=0; i<L/2; i++) average += X[abs(i)];
    ma[0] = average/L;
    for(i=0; i<max_point; i++) {
        average -= X[abs(i)];
        if(i+L>max_point) average += X[(long)2*max_point-(i+L)];
        else average += X[i+L];
        ma[i+L/2] = (average/L + X[i+L/2 - l]);
    }
    clrscrO;
    printf("/n/OPEN OUTPUT FILES*/
    temp = fopen("D:temp.dat", "rb");
    if(ferror(temp)!=0) exit(1);
    fread((void*)X, sizeof(float), 16383, temp);
    fclose(temp); 
/*open output files*/
extension[1] = "tma"; /*moving average vs time*/
extension[2] = "dma"; /*moving average vs distance*/
extension[3] = "t-m"; /*normalized signal vs time*/
extension[4] = "d-m"; /*normalized signal vs distance*/

for (i=1; i<5; i++) {
    strset(name2, '/O ');
    fnmerge(name2, output_drive, output_dir, name3, extension[i]);
    p[i] = fopen(name2, string[1]);
    if (ferror(p[i])!=0) {
        printf("I can't open %s/n",name2);
        exit(1);
    }
}
clrscr();
printf("/n/n/n/n/n/

^ /n/

* REMOVING MOVING AVERAGE */n/

* */n/

*******************************************************************************/n/
/n%s" file.ff_name);

for(i=0, xdata=data=0, k=1; i<max_point; i++) {
    X[i] /= ma[i];
    xdata += ma[i];
    data += X[i];
    if(i/decimation == k) {
        k++;
        xdata = 10*log10(double)xdata/decimation);
        data = 10*log10(double)data/decimation);
        time = (float)i/sample_frequency;
        fprintf(p[1],"%+9.6f %+9.6f/n", xdata, time);
        fprintf(p[2],"%+9.6f %+9.6f/n", xdata, time*speed);
        fprintf(p[3],"%+9.6f %+9.6f/n", data, time);
        fprintf(p[4],"%+9.6f %+9.6f/n", data, time*speed);
        xdata = data = 0;
    }
}
for(i=1; i<5; i++) fclose(p[i]);

/*open output files*/

extension[1] = "pdm"; /*probability density function*/
extension[2] = "cdm"; /*cumulative probability density*/
extension[3] = "pd-"; /*signal level vs time*/
extension[4] = "cd-"; /*signal level vs distance*/
extension[5] = "dma"; /*moving average signal vs distance*/
extension[6] = "d-m"; /*normalized signal vs distance*/

for (i=1; i<5; i++) {
    strset(name2, '/O ');
    fnmerge(name2, output_drive, output_dir, name3, extension[i]);
    p[i] = fopen(name2, string[1]);
    if (ferror(p[i])!=0) {
        printf("I can't open %s/n",name2);
        exit(1);
    }
}

n = strdup(name3);

for (i=5; i<7; i++) {
    strcpy(n, name3);
    strcat(n, extension[i]);
    strset(name2, '/O ');
    fnmerge(name2, output_drive, output_dir, n, ext);
    p[i] = fopen(name2, "wB");
    if (ferror(p[i])!=0) {
        printf("I can't open %s/n",name2);

exit(1);
}
}
clrscr();
printf("/n/n/n/n/n/
*******************************************************************************/n/
  */n/
  CONVERTING DATA TO */n/
  dB UNITS */n/
  */n/
*******************************************************************************/n/
/n%s", file.ff_name);

/*The data and its moving average are now reconverted back to dB's after the
calculation and removal of the moving average*/
for(i=0; i<max_point; i++) {
  X[i] = 10*log10((double)X[i]);
  ma[i] = 10*log10((double)ma[i]);
}
clrscr();
printf("/n/n/n/n/n/
*******************************************************************************/n/
  */n/
  CONVERTING DATA TO */n/
  dB UNITS */n/
  */n/
*******************************************************************************/n/
/n%s", file.ff_name);

/*CALCULATING PDF & CPD OF MOVING AVERAGES*/
for (m=0; m<no_bin; m++) {
  level = m*step + Xmin;
  for (i=0, k= 0, l=0; i<max_point; i++) {
    if (ma[i] < (level + step)) {
      k++;
      if (ma[i] >= level) l++;
    }
  }
  xdata = (float)k/max_point;
  if (xdata < 0.1) X90 = level;
  if (xdata < 0.16) X84 = level;
  if (xdata < 0.5) X50 = level;
  fprintf(p[1], "%+9.6f %+9.6f/n", (float)l/max_point, level);
  fprintf(p[2], "%+9.6f %+9.6f/n", xdata, level);
}
loss_deviation = X50 - X90;
location_variability = X50 - X84;
fprintf(p[0], "moving average loss deviation = %f/n",
loss_deviation);
fprintf(p[0], "moving average location variability = %f/n",
location_variability);
clrscr();
printf("/n/n/n/n/n/
*******************************************************************************/n/
  */n/
  CONVERTING DATA TO */n/
  dB UNITS */n/
  */n/
*******************************************************************************/n/
/n%s", file.ff_name);

for (m=0; m<no_bin; m++) {
  level = m*step + Xmin;
  for (i=0, k=0, l=0; i<max_point; i++) {
    if (X[i] < (level + step)) {
      k++;
      if (X[i] >= level) l++;
    }
  }
}
xdata = (float)k/max_point;
if (xdata < 0.1) X90 = level;
if (xdata < 0.16) X84 = level;
if (xdata < 0.5) X50 = level;
fprintf (p[3], "%+9.6f %+9.6f/n", (float)/max_point, level);
fprintf (p[4], "%+9.6f %+9.6f/n", xdata, level);
}
loss_deviation = X50 - X90;
location_variability = X50 - X84;
fprintf(p[0], "/tnormalized data loss deviation = %f/n",
loss_deviation);
fprintf(p[0], "/tnormalized data location variability = %f/n",
location_variability);
c1rsr();
new_max = 16384-max_point;

for(i=0; i<max_point; i++) {X[i] *= -1; ma[i] *= -1;}
fwrite ((char*)ma, sizeof(float), max_point, p[5]);
if (new_max)
{lseek(fileno(p[5]), (long)(max_point)*sizeof(float), SEEK_SET);
fwrite ((char*)ma, sizeof(float), new_max, p[5]);
}
fwrite ((char*)X, sizeof(float), max_point, p[6]);
if (new_max)
{lseek(fileno(p[6]), (long)(max_point)*sizeof(float), SEEK_SET);
fwrite ((char*)X, sizeof(float), new_max, p[6]);
}
if (fcloseall()==EOF) perror("/nFILE STORAGE ERROR OCCURRED/n");
c1rsr();
printf("/n/a %s PROCESSING FINISHED/n", file.ff_name);
li++;
}
finish:
puts("/n/n/n/n/ DATA ANALYSIS FINISHED HIT ANY KEY TO CONTINUE/n");
printf("/a");
puts("/n/n/n/n/THE FORMAT OF THE DATA <m data[m]> FOR DG IS 9");
return(1);
The second program (LOGNORM.C) presented is used to process the cumulative probability density functions produced by the first program so that the respective distributions may be plotted on lognormal and Weibull probability plots. To do this, the program uses an algorithm which is the inverse of the lognormal and Weibull distributions respectively. On its own probability plot each distribution is a straight line. Thus, this procedure allows easy comparison of the experimental distributions with theoretical models.

/* LOGNORM.C Program to calculate the non-linear axes for probability plots from the experimental cumulative distributions.

Written by H Thomas - November 1989*/

#include "stdio.h"
#include "conio.h"
#include "math.h"
#include "string.h"
#include "ctype.h" /* lower case character conversions */
#include "process.h"
#include "alloc.h"

main()
{
    int i, no_files, iii, length, flag ;
    char *request[2], *nameA[100], *path, c ;
    FILE *p[10] ;

    void *vector(int, int) ;
    void get_f(char *r, char *name) ;
    int log_normal_plot(char *name, char *path) ;
    int weibull_plot(char *name, char *path) ;
    int choice(void) ;
    int matherr(struct exception *) ;

    request[0] = "/
    ENTER NAME OF PATH FOR OUTPUT FILES/n" ;
    path = (char *)calloc(80, sizeof(char)) ;
    flag = 0 ;
    clrscr() ;
    do {
        if(flag) {
            printf("/n %s IS AN INVALID PATH/n/n", strupr(path)) ;
            flag = 1 ;
            get_f(request[0], path) ;
            clrscr() ;
            length = strlen(path) ;
            c = path[length-1] ;
            if(c == '/') c == ':'
                flag = 0 ;
        } while (flag) ;

    no_files = 0 ;
    while (choice()) {
        clrscr() ;
        nameA[no_files] = (char *)calloc(80, sizeof(char)) ;

        request[0] = "/
        ENTER NAME OF INPUT FILE/n/t" ;

        if(no_files>0) {
            puts("files entered so far are :-/n") ;
            for(i=0; i<no_files; i++)
                printf("%s /n", nameA[i]) ;
        }
    }
get_f (request[0], nameA[no_files]);
while ((p[0] = fopen (nameA[no_files], "r")) == NULL) {
    clrscr();
    printf ("I can't open %s/n", nameA[no_files]);
    if(no_files>0) {
        puts("Files entered so far are :-/n");
        for(i=0; i<no_files; i++)
            printf("%s/n", nameA[i]);
    }
    get_f (request[0], nameA[no_files]);
} fclose(p[0]);
no_files++; }

for(iii=0; iii < no_files; iii++)
{
    clrscr();
    puts ("/n/n/n/n/n/n/
          ******************************************/n/
          *
          * CALCULATING LOGNORMAL    */n/
          * DATA SET                */n/
          * */n/
          ******************************************/n");
    log_normal_plot(nameA[iii], path);
    clrscr();
    puts ("/n/n/n/n/n/n/
          ******************************************/n/
          *
          * CALCULATING WEIBULL     */n/
          * DATA SET                */n/
          * */n/
          ******************************************/n");
    weibull_plot(nameA[iii], path);
}
    clrscr();
    puts ("/n/n/n/n/n/DATA ANALYSIS FINISHED/n");
    printf ("/a");
    puts ("/n/n/n/n/THE FORMAT OF THE DATA <m data[m]> FOR DG IS 9,2,9");
    return(1);
iv SIMULATION OF DIVERSITY RECEPTION

This program is designed to simulate diversity reception from experiments carried out using two frequency or two polarization measurements. The program functions by converting the two input signals from logarithmic (decibel) scales to linear scales, adding the two signals and then converting the resultant signal back to a logarithmic (decibel) scale. The input to the program is two calibrated sections of time series and its output is a combination of the two inputs. The output of the program may be analysed in terms of the increase in signal to noise ratio if the addition of the two noise powers of the inputs is taken into account.

The program requests a file name with wild-card arguments to specify the different input files.

/* COMBINE.C Program to calculate the combination of two input files in order to estimate diversity reception.

NB the name of the file to be calibrated may be entered on the command line to save time.

Written by H Thomas - January 1990*/

#include "stdio.h"
#include "conio.h"
#include "math.h"
#include "string.h"
#include "ctype.h" /*lower case character conversions*/
#include "process.h"
#include "alloc.h"
#include "string.h"
#include "io.h"
#include "dir.h"
#include "dos.h"
#include "errno.h"

int get_pause (r)
char *r;
{
    while(kbhit() != 0) getch();
    if (r != NULL) puts (r);
    if (isspace (getcn())) return (1);
    return (0);
}

void get_question(info_request)
char *info_request;
{
    char *request;

    request = "/
    /t/t/* **************************** /n/
    /t/t/*                         /n/
    /t/t/* TYPE ENTER             /n/
    /t/t/*                         /n/
    /t/t/* FOR DEFAULT VALUES     /n/
    /t/t/*                         /n/
    /t/t/ **************************** /n/n/n/
    /t/t/ **************************** /n/
    /t/t/                         /n/
    /t/t/ PRESS ANY OTHER KEY TO CHANGE SETTINGS /n/
    /t/t/                         /n/
    /t/t/****************************",";
    strcpy(info_request, request);
void get_info (info, length)
float *info ;
long length ;
{
    char *info_request ;
    void get_question(char *question);
    int get_pause(char *r);

    info_request = (char *)malloc(1000*sizeof(char)) ;
    get_question(info_request);

clrscr();
puts("/n/n/n /
IS DATA IN LOG FORMAT? /n/n/
DEFAULT IS LINEAR /n/n/n") ;
puts(info_request); 
if (get_pause(NULL)) info[0] = 0.0;
else info[0] = 1;
clrscr();
}

main(argc, argv)
int argc ;
char *argv[] ;
char *name, *name1, *name_out, *request ;
char *drive, *dir, *ext, *out_ext ;
int i, k, num, loop=0, test, clear=0 ;
long length, long_num, number ;
float *info, *input, *input2 ;
float max_value, min_value ;
FILE *f1, *f2, *f3 ;
struct ffblk file ;
void get_info(float *info, long length) ;
void get_f(char *name, char *request);
int matherr(struct exception *e);
drive = (char *)malloc(5*sizeof(char));
dir = (char *)malloc(80*sizeof(char));
request = (char *)malloc(200*sizeof(char));
ext = (char *)malloc(5*sizeof(char));
out_ext = (char *)malloc(5*sizeof(char));
name = (char*)calloc(80, sizeof(char));
namel = (char*)calloc(80, sizeof(char));
name_out = (char*)calloc(80, sizeof(char));
info = (float *)calloc(4,sizeof(float));
input = (float *)calloc(1200,sizeof(float));
input2 = (float *)calloc(1200,sizeof(float));
clrscr();

out_ext = ".add" ;
k = 1 ;
if(argc>1 & & argv[1]!=NULL) {strcpy(name, argv[1]); k = 0;}
do {
if(k) {
    .request = "/nENTER DRIVE, PATH, AND FILE NAME OF INPUT DATA /n"
    strcat(
        request,
        "TYPE ? FOR THE CHARACTER WHICH IDENTIFIES THE SEPARATE CHAN­
NELS/n") ;
    strcat(request,"LEAVE NO SPACES !!/n") ;
    puts(request);
    gets(name);
} 
k = 1 ;
}
strcat(name_out, file.ff_name);
f1 = fopen(name_out, "rb"); 
if(ferror(f1) || f1==NULL) 
  perror("I can't open file 1");
test = -1;
} else printf("/nFILE 1 %s OPENED SUCCESSFULLY/n", name_out);

if((test!=-1)) {
  if(findnext(&file)!=0) { 
    perror("/nFILE 2 DOES NOT EXIST");
    test = -1;
  } else {
    fnmerge(name_out, drive, dir, NULL, NULL);
    strcat(name_out, file.ff_name);
    f2=fopen(name_out, "rb");
    if(ferror(f2) || f2==NULL) {
      perror("/nCANNOT OPEN FILE 2");
    } else printf("/nFILE 2 %s OPENED SUCCESSFULLY/n", name_out);
    if(test==-1) {
      clrscrO;
      puts("RE-TYPE, OR TRY ANOTHER FILE NAME!");
      test = -1;
    } while(test==-1);
    length = filelength(fileno(f1));
    if(length != filelength(fileno(f2))) {
      perror("file lengths are not equal");
      exit(1);
    }
    fnsplit(name_out, drive, dir, name1, ext);
    while (1) {
      puts("/n
                                 » » » < » » » » » » » » » » » » » ♦ » » » » * » » » » » » » » » » » » » » » » » /n
                                 ENTER DRIVE FOR STORING OUTPUT FILE /n
                                 TYPE /"/" FOR NO ENTRY !! /n");
      scanf("%s", drive);
      puts("/n
                                 ENTER DIRECTORY FOR STORING OUTPUT FILE /n
                                 TYPE / / FOR NO ENTRY !! /n");
      scanf("%s", dir);
      fnmerge(name_out, drive, dir, name1, out_ext);
      f3 = fopen(name_out, "wb");
      if(ferror(f3)==0) break;
      clrscrO;
      puts("/nRE-TYPE DRIVE SPECIFICATION !/n");
    }
    get_info(info,length);
    clrscrO;
    long_num = 0;
    while((number = length/sizeof(float)-long_num)>0) {
      if(number>1024) number = 1024;
      lseek(fileno(f1), long_num*sizeof(float), SEEK_SET);
      num = fread((void*)input, sizeof(float), number, f1);
      lseek(fileno(f2), long_num*sizeof(float), SEEK_SET);
      if(fread((void*)input2, sizeof(float), number, f2) != num) {
        perror("ERROR DURING DATA READING NUM1 != NUM 2");
        exit(1);
      }
      for(i=0, max_value=-40.0, min_value=0.0; i<num; i++) {
        if(input[i]>max_value) max_value=input[i];
        if(input[i]<min_value) min_value=input[i];
      }
      if(loop == clear) {
        clrscrO;
      }
    }
  }
}
- 323 -
clear += 3;
}
printf("\nPROCESSING DATA SECTION %d/n\n", loop);
printf("\nMAX DATA 1 = %f\tMIN DATA 1 = %f\n", max_value, min_value);
for(i=0, max_value=-40.0, min_value=0.0; i<num; i++) {
  if(input2[i]>max_value) max_value=input2[i];
  if(input2[i]<min_value) min_value=input2[i];
}
printf("\nMAX DATA 2 = %f\tMIN DATA 2 = %f\n", max_value, min_value);
if(info[0]) {
  for(i=0; i<num; i++) {
    input[i]=pow((double)10,(double)input[i]/10);
    input2[i]=pow((double)10,(double)input2[i]/10);
    input[i] += input2[i];
    input2[i] = l(rlogl(X(double)input[i])
  } else
  for(i=0; i<num; i++) input2[i] += input[i];
}
for(i=0, max_value=-40.0, min_value=0.0; i<num; i++) {
  if(input2[i]>max_value) max_value=input2[i];
  if(input2[i]<min_value) min_value=input2[i];
}
printf("\nMAX DATA ADD = %f\tMIN DATA ADD = %f\n", max_value, min_value);
puts("STORING DATA");
iseek(fileno(f3), long_num*sizeof(float), SEEK_SET);
if(fwrite((void *)input2, sizeof(float), num, f3) != num) {
  perror("ERROR DURING DATA WRITING");
  exit(1);
}
long_num += num;
loop++;
if(fcloseall()==EOF) puts("FILE ALLOCATION ERROR ON <fcloseall()>");
putc('
');
return(1);
This Section describes the program used for estimating the auto-correlation function and the cross-correlation function of the experimental data. The program calculates the estimate in a direct manner. This is not the most efficient method. However, a program based on fast Fourier transform routines was not available at the time. The program divides the input data into sections 1024 points long and then calculates the average of the correlation function estimate over however many 1024 point segments there are in the input data file.

The program has a facility for multiple processing of input files with similar names. The two input files for the cross-correlation calculation are distinguished using wild-card arguments. Multiple file processing is accomplished by specifying one character of the input file name to increment for each subsequent processing operation. The data are converted from logarithmic (decibel) scales before the estimates are taken. The output file is given as a function of correlation coefficient versus time delay. The time delay axis is calculated using the measured speed of each data sample together with the data sampling rate used for the acquisition.

/* CROSS.C Program to calculate either the cross or the auto-correlation of two calibrated data files in a direct manner without using FFT routines.

NB the name of the file to be calibrated may be entered on the command line to save time.

Written by H Thomas - January 1990*/

#include "stdio.h"
#include "conio.h"
#include "math.h"
#include "string.h"
#include "ctype.h" /*lower case character conversions */
#include "process.h"
#include "alloc.h"
#include "string.h"
#include "io.h"
#include "dir.h"
#include "dos.h"
#include "errno.h"

int get_pause (r)
char *r ;
{
    void clear_key(void) ;
    clear_key();
    if (r != NULL) puts (r);
    if(isspace (getcnO)) return (1);
    return (0);
}

void get_question(info_request)
char *info_request ;
{
    char *request ;
request = "/
/t/t/t************* /n/
/t/t/t*             /n/
/t/t/t* TYPE ENTER * /n/
/t/t/t*           /n/
/t/t/t* FOR DEFAULT VALUES * /n/
/t/t/t*            /n/
/t/t/t******************************* /n/
/t/t/t******************************* /n/
/t/t/t*             /n/
/t/t/t* PSSS ANY OTHER KEY TO CHANGE SETTINGS * /n/
/t/t/t*            /n/
/t/t/t******************************* ";
strcpy(info_request, request);

void get_info (info, length)
float *info ;
long length ;
{
    char *info_request, *request ;
    void get_question(char *question);
    int get_pause(char *r);

    info_request = (char *)malloc(1000*sizeof(char));
    get_question(info_request);

    clrscr();
    puts("/n/n/n/
IS DATA IN LOG FORMAT? /n/n/
DEFAULT IS LINEAR /n/n/n");
    puts(info_request);
    if(get_pause(NULL)) info[0] = 0.0 ;
    else info[0] = 1 ;
    clrscr();
    request="/n/n/n/
ENTER SAMPLE FREQUENCY /n/
    DEFAULT IS %d /n/n/n";
    info[1] = 10000 ;
    printf (request, (int)info[1]) ;
    puts(info_request);
    if(get_pause(NULL)==0)
    {
        clrscr();
        printf(request, (int)info[1]);
        scanf("%f", &info[1]) ;
    }
    clrscr();
    request="/n/n/n/
NUMBER OF LAGS /n/
    DEFAULT IS %d /n/n/n";
    info[2] = 100 ;
    printf (request, (int)info[2]) ;
    puts(info_request);
    if(get_pause(NULL)==0)
    {
        clrscr();
        printf(request, (int)info[2]);
        scanf("%f", &info[2]) ;
    }
    clrscr();
    puts("/n/
DO YOU WANT TO PROCESS MULTIPLE FILES?/n/
NB: THIS CAN ONLY BE DONE FOR CROSS-CORRELATION IF THE FILES/ /n/
ARE DISTINGUISHED BY A NUMBER AT THE END OF THE NAME OF THE FI-
LE/n/n/n/
    DEFAULT IS YES/n/n/n");
    puts(info_request);
    info[3] = get_pause(NULL);
main(argc, argv)
int argc;
char *argv[];
{
    char *name, *name1, *name_out, *request, *p1, *p2;
    int i, j, k, num, loop=0, lag, max, corr, test, name_length;
    int replace;
    long length, position;
    float *info, *auto_corr, *input, *input2, max_value, min_value;
    double av, av2, sq, sq2, R;
    FILE *fl, *f2, *f3, *f4;
    struct ffblk file;
    void get_info(float *info, long length);
    void get_f(char *name, char *request);
    int matherr(struct exception *e);

    drive = (char *)malloc(5*sizeof(char));
    dir = (char *)malloc(80*sizeof(char));
    drive_out = (char *)malloc(5*sizeof(char));
    dir_out = (char *)malloc(80*sizeof(char));
    request = (char *)malloc(200*sizeof(char));
    ext = (char *)malloc(5*sizeof(char));
    out_ext = (char *)malloc(5*sizeof(char));
    inf_ext = (char *)malloc(5*sizeof(char));
    name = (char*)calloc(80, sizeof(char));
    name_out = (char*)calloc(80, sizeof(char));
    name1 = (char*)calloc(80, sizeof(char));
    info = (float *)calloc(4,sizeof(float));
    input = (float *)calloc(1200,sizeof(float));
    input2 = (float *)calloc(1200,sizeof(float));

    clrscrO;
    strset(request, '/x0') ;
    request = "/n/
    TYPE ENTER FOR AUTO-CORRELATION/n/
    TYPE ANY OTHER KEY FOR CROSS-CORRELATION/n"
    puts(request);
    if(isspace(getchO)) corr = 0; else corr = 1;
    clrscrO;

    if(corr) {out_ext = ".crs"; inf_ext = ".i-c";}
    else {out_ext = ".aut"; inf_ext = ".i-a";}

    k = 1;
    if(argv[1]!=NULL) strcpy(name, argv[1]); 
    do {
        if(k) {
            strset(request, '/x0');
            request = "/n/ENTER DRIVE, PATH, AND FILE NAME OF INPUT DATA/n"
            if(corr) strcat(request,
                "TYPE ? FOR THE CHARACTER WHICH IDENTIFIES THE SEPARATE CHANNELS/n")
            strcat(request,"/n/n/LEAVE NO SPACES !!/n")
            puts(request);
            gets(name);
        }
        k = 1;
        fnsplit(name, drive, dir, name1, ext);
        if((test=findfirst(name, &file, FA_DIREC))!=-1)(
            fnsplit(file.ff_name, NULL, NULL, name1, ext);
            fnmerge(name, drive, dir, name1, ext);
            if((test!=-l) && corr) {
                if(findnext(&file)!=0) {
                    perror("/n/FILE 2 DOES NOT EXIST")
                    test = -1;
                }
            else {

fnssplit(file.ff_name, NULL, NULL, name1, ext);
fnmerge(name_out, drive, dir, name1, ext);
f2=fopen(name_out, 'rb');
if(f2==NULL || ferror(f2)) {
    test=-1;
    perror("/nCANNOT OPEN FILE 2");
} else printf("/nFILE 2 %s OPENED SUCCESSFULLY/n", name_out);
}
    f1 = fopen(name, "rb");
    if(f1==NULL || ferror(f1)) {
        perror("/nI can't open file 1");
        test=-1;
    } else printf("/nFILE 1 %s OPENED SUCCESSFULLY/n", name);
}
if(test==-1) {
    clrscrO *
    puts("RE-TYPE, OR TRY ANOTHER FILE NAME !");
}
while(test==-1);

get_pause("/n/n/n/
"***HIT ANY KEY TO CONTINUE PROCESSING****");

p1 = strdup(name);
if(corr) p2 = strdup(name_out);
length = filelength(fileno(f1));
if(corr) if(length != filelength(fileno(f2))) {
    perror("/nfile lengths are not equal");
    exit(1);
}
get_info(info,length);
lag = (int)info[2];

/*This code gets the character which is to be incremented when
the program processes subsequent data files*/

if(info[3]&&corr) {
    name_length = strlen(file.ff_name) - strlen(ext);
    /*Write file name to the monitor with numbers beneath it*/
    clrscr();
    puts("/n/n/
    CHOOSE CHARACTER TO BE INCREMENTED/n/
    INDICATE CHARACTER BY ARRAY NUMBER*");        
    puts("/n/n"                  
    puts(file.ff_name);         
    for(i=0; i<name_length; i++) printf("%c", i + '0');
    puts("/n/n");
    /*Read in the character to be incremented*/
    while(1) {
        replace = getche();
        if(isdigit(replace)) {
            replace -= '0';
            if(replace < name_length) break;
        }
    }
    clrscr();
    fnssplit(name, drive, dir, name1, ext);
    while (1) {
        puts("/n/n/
        ENTER DRIVE FOR STORING OUTPUT FILE /n/
        TYPE //" FOR NO ENTRY !! /n");
        scanf("%s", drive_out);
        puts("/n/n");
        ENTER DIRECTORY FOR STORING OUTPUT FILE /n/
        TYPE // FOR NO ENTRY !! /n");
        scanf("%s", dir_out);
        strset(name_out, '/O ');
fnmerge(name_out, drive_out, dir_out, name1, inf_ext);
f3 = fopen(name_out, "w");
strset(name_out, "/O");
fnmerge(name_out, drive_out, dir_out, name1, out_ext);
if((f3 == NULL || ferror(f3)) {
    clrscr();
    puts("/nRE-TYPE DRIVE SPECIFICATION !/n");
} else break;
auto_corr = (float *)calloc(lag+1,sizeof(float));
goto START;

/*Generate new file names to process subsequent files*/
PROCESS_NEW_FILE:
    fnsplit(p1, drive, dir, name1, ext);
    while(1) {
        if(corr) {
            fnsplit(p2, drive, dir, name, ext);
            name1[replace]++;
            name[replace]++;
            fnmerge(p1, drive, dir, name1, ext);
            fnmerge(p2, drive, dir, name, ext);
            if((f1 = fopen(p1, "rb")) != NULL || ferror(f1)) {
                if((f2 = fopen(p2, "rb")) != NULL || ferror(f2))
                    break;
            }
            perror("/n can't open FILE");
            exit(1);
        } else {
            if(findnext(&file)==-1) {
                if(errno==ENOFILE || errno==ENOENT) puts("ALL FILES PROCESSED");
                exit(1);
            } 
            fnsplit(file.ff_name, NULL, NULL, name1, ext);
            fnmerge(p1, drive, dir, name1, ext);
            if((f1 = fopen(p1, "rb")) != NULL || ferror(f1))
                exit(1);
        }
    }

    if((length = filelength(fileno(f1))) == -1)
        exit(1);
    if(corr) if(length != filelength(fileno(f2))) {
        perror("/n file lengths are not equal");
        exit(1);
    } 
    fnmerge(name_out, drive_out, dir_out, name1, inf_ext);
    if((f3 = fopen(name_out, "w")) == NULL || ferror(f3)) {
        perror("/n can't open FILE 1");
        exit(1);
    }
    fnmerge(name_out, drive_out, dir_out, name1, out_ext);

    /*Resume the normal function of the program with
    new file names*/

    START:
    if(corr) fputs("/t/t/tCROSS-CORRELATION", f3);
    else fputs("/t/t/tAUTO-CORRELATION", f3);
    fputs("/n/" f3); 
    fprintf(f3, "FILE NAME 1 IS %s/n", p1);
    if(corr) fprintf(f3, "FILE NAME 2 IS %s/n", p2);
    
    if(info[1]) fputs("/t/ANTI-LOG DATA BEFORE CALCULATION/n", f3);

fprintf(f3, "OUTPUT FILE EXTENSION IS %s/n", out_ext);
fprintf(f3, "INFO FILE EXTENSION IS %s/n", inf_ext);

fprintf(f3, "/tNumber of lags %d/n", lag);
fprintf(f3, "/tNumber of points per segment = 1024", f3);
fprintf(f3, "/tFILE LENGTH %ld BYTES/n", length);

clrscrO;
loop = 0;
while(length/sizeof(float) - (long)loop*1024 > lag+1) {
position = lseek(fileno(f1), (long)loop*1024*sizeof(float), SEEK_SET);
printf("/tREADING FILE 1 SECTION %d (%ld)", loop,
position*sizeof(float));
num = fread((void *)input, sizeof(float), 1024, f1);
if(corr) {
position = lseek(fileno(f2), (long)loop*1024*sizeof(float), SEEK_SET);
printf("/tREADING FILE 2 SECTION %d (%ld)", loop,
position*sizeof(float));
if(fread((void *)input2, sizeof(float), 1024, f2) != num) {
perror("ERROR DURING DATA READING NUM1 != NUM2");
exit(1);
}
}
printf("/nPROCESSING DATA %s/n/", p1);
for(i=0, max_value=-40.0, min_value=0.0; i<num; i++) {
if(input[i]>max_value) max_value=input[i];
if(input[i]<min_value) min_value=input[i];
}
printf("/nMAX DATA 1 = %f dB/tMIN DATA 1 = %f dB",
max_value, min_value);
if(corr) {
for(i=0, max_value=-40.0, min_value=0.0; i<num; i++) {
if(input2[i]>max_value) max_value=input2[i];
if(input2[i]<min_value) min_value=input2[i];
}
printf("/nMAX DATA 2 = %f dB/tMIN DATA 2 = %f dB",
max_value, min_value);
}

/*calculate average of data*/
for(i=0; av=sq=0.0; i<num; i++) {
if(info[l]) input[i] = pow((double)10,(double)input[i]/10);
av += input[i];
sq += input[i]*input[i];
}
av /= num;
sq /= num;
sq -= av*av;
printf("/nSQ No. %d = %f dB", loop,
10*log10(sq/av+1));
fprintf(f3, "/nSECTION %d/t av = %6.4f dB/t sq = %6.4f dB",
loop, 10*log10(av), 10*log10(sq/av+1));
if(corr) {
for(i=0, av2=sq2=0.0; i<num; i++) {
if(info[l]) input2[i]=pow((double)10,(double)input2[i]/10);
av2 += input2[i];
sq2 += input2[i]*input2[i];
}
av2 /= num;
sq2 /= num;
sq2 -= av2*av2;
printf("/nSQ2 No. %d = %f dB", loop,
10*log10(sq2/av2+1));
fprintf(f3, "/nSECTION %d/t av2 = %6.3f dB/t sq2 = %6.3f dB",
loop, 10*log10(av2), 10*log10(sq2/av2+1));
av2 *= av;
sq2 = sqrt(sq*sq2);
} else {
av2 = av*av;
sq2 = sq;
for(j=0; j<num; j++) input2[j]=input[j];
}

/*calculate auto-correlation*/
for(i=0; (i<lag) && (num>lag); j++) {
    max = num - j;
    for(i=0, R=0.0; i<max; i++) R += (input[i]*input2[i+j]);
    auto_corr[j] += (R/max - av2)/sq2;
}
puts("/n");
cṛscr();
/*write trial values to screen to provide visual check on program operation*/
printf("TRIAL VALUES AVERAGE OF SECTION 0 TO %d/n", loop);
for(i=0; i<lag && lag/10 >=1 ; i+= lag/10)
    printf("AUTO[%d] = %+9.4f", i, auto_corr[i]/(loop+1));
loop++;
}
printf("/nSTORING DATA IN FILE %s", name_out);
fclose(f1);
fclose(f2);
f4 = fopen(name_out, "w");
for(i=0; i<lag; i++) {
    auto_corr[i] /= loop;
    fprintf(f4, "+%9.4f +%9.4f/n", auto_corr[i], (float)i/info[l]);
}
if(corr)
    fprintf(f3, "+/n/CROSS-CORRELATION COEFFICIENT = %6.3f", auto_corr[0]);
if(fcloseall()==EOF) puts("/nFILE ALLOCATION ERROR ON <fcloseall()>");

/*Repeat correlation calculation for a new data set*/
if(info[3]) {
    for(i=0; i<lag; i++) auto_corr[i] = 0;
    goto PROCESS.NEW_FILE;
}
return(1);
}
This program is used to calculate an approximation of the received power versus transmitter-receiver separation in a micro-cell. It calculates the summation of four separate terms representing the direct line of sight component, the specular reflection from the road surface and the two reflections from each side of the street. The parameters are fixed within the program and the program would have to be edited in order to change them.

/* DETERMIN.C This program is designed to calculate an approximation of the variation of the received power with distance in a micro-cell.

Written by H Thomas - October 1989*/

#include "stdio.h"
#include "conio.h"
#include "math.h"
#include "string.h"
#include "ctype.h" /*lower case character conversions*/
#include "process.h"
#include "alloc.h"
#include "string.h"
#include "io.h"
#include "dir.h"
#include "stdlib.h"

double d(x, H, h)
double x, H, h ;
{
    double out;
    out = sqrt(H*H+h*x/(H+x)) + sqrt(h*h+H*x/(H+x));
    return(out);
}

double r(x, H, h)
double x, H, h ;
{
    double out;
    out = sqrt(x*x + (H-h)*(H-h));
    return(out);
}

double pd(x, H, h)
double x, H, h ;
{
    double out;
    double r(double x, double H, double h);
    double d(double x, double H, double h);
    out = d(x, H, h) - r(x, H, h);
    return(out);
}

double phi(x, H, h)
double x, H, h ;
{
    double out;
    double r(double x, double H, double h);
    out = atan((H+h)/x) - acos(x/r(x,H,h));
    return(out);
}

void comp_e(out, C)
double out;
struct complex *C ;
{
    double value ;

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value = cos(out);
C->x = value;
value = -sin(out);
C->y = value;
}
void a(x, H, h, bw, C)
double x, H, h, bw;
struct complex *C;
{
    void comp_e(float x, struct complex *C);
    double r(double x, double H, double h);
    double pd(double x, double H, double h);
    double phi(double x, double H, double h);

double out, B, Ang;

    Ang = phi(x, H, h);
    B = sqrt(bw*sqrt(2*3.14));
    out = 1/r(x, H, h)*1/B*exp(Ang*Ang/(4*bw*bw));
    comp_e(2*3.14/0.0055*pd(x, W, h), O;
    C->x *= out;
    C->y *= out;
}
main(argc, argv)
int argc;
char *argv[];
{
    int i, k, num, loop=0, *input, numberofpoints;
    struct complex *z1, *zout;
    double x, H[4], h[4], bw, pi=3.14;
    double data[100];

double d(double x, double H, double h);
    double r(double x, double H, double h);
    double pd(double x, double H, double h);
    double phi(double x, double H, double h);

    void comp_e(float x, struct complex *z);
    void a(double x, double H, double h, double bw, struct complex *z);

    bw = 10*pi/180;
    H[0] = 0;
    H[0] = 14;
    H[0] = 16;
    H[0] = 15;
    h[0] = 0;
    h[0] = 16;
    h[0] = 14;
    h[0] = 1.5;

    for(i=0; i<10; i++) {
        for(k=0, zout->x = 0, zout->y = 0; k<4; k++) {
            x = 100 + i/10;
            a(x, H[k], h[k], bw, z1);
            zout->x += z1->x;
            zout->y += z1->y;
            data[i] = cabs(*zout);
        }
    }
}
It is in the nature of the "C" language that it relies upon pre-defined functions. This section contains three utility files (HTUTIL.C, NRUTIL.C, WIEN_UT.C) which are used by the programs described here. Some of the programs are taken from the "Numerical Recipes in 'C'" book (Ref: Press W H et al, 1988).

/* HTUTIL.C This program contains various functions written by the author to support the other programs presented here.

Written by H Thomas - January 1990*/

#include "stdio.h" /*for standard i/o eg. files */
#include "conio.h"
#include "stdlib.h"
#include "string.h" /*for string manipulation */
#include "io.h"
#include "alloc.h" /*for dynamic memory allocation */
#include "math.h"
#include "ctype.h"

void quit_p (label)
char *label;
{
    puts (label);
    clrscr();
    exit (1);
}

void clear_key ()
{
    while (kbhit() != 0)
        getch () ;
}

void get_f (name,request)
char name[];
char *request;
{
    int c_count, flag;
    void clear_key(void);
    clrscr();
    puts ("/n/n/n");
    clear_key () ;
    puts (request);
    flag = 1 ;
    while (flag)
    { 
        gets(name) ;
        c_count = strlen (name) ;
        flag = 0;
        if (c_count > 20 )
            cputs ("FILENAME TOO LONG/n");
        else
            flag = 0 ;
    }
}

int askint(s)
char s[];
{
    int i, num ;
    gets (s) ;
for (num = 0, i = 0; s[i] >= '0' && s[i] <= '9'; i++)
    num = 10*num + s[i] - '0';
return (num);

int matherr(x)
struct exception *x;
{
    if(x->type==DOMAIN) {
        if(strcmp(x->name, "log10") == 0) {
            clrscr();
            putch('ERR');
            x->retval = -4;
            return(1);
        }
        return (0);
    }
    return (0);
}
/* NRUTIL.C This is the utility program used by the routines supplied with the
'Numerical Recipes in 'C'' book. Some of the routines have been modified. In particular,
the matherr() function has been modified to return different values in overflow
conditions.

Written by H Thomas - April 1989*/

#include "alloc.h"
#include "stdio.h"
#include "stdlib.h"

void nerror(error_text)
char error_text[];
{
  fprintf(stderr, "Numerical Recipies run-time error.../n");
  fprintf(stderr, "%s/n", error_text);
  fprintf(stderr, "...now exiting to system.../n");
  exit(1);
}

void free_vector(v, nl)
float *v;
int nl;
{
  free((char*) (v + nl));
}

float *vector(nl, nh)
int nl, nh;
{
  float *v;
  void nerror(char *error);

  v = (float *)calloc((unsigned) (nh - nl + 1), sizeof(float));
  if (!v) nerror("allocation failure in vector()");
  return v - nl;
}

float far *farvector(nl, nh)
int nl, nh;
{
  float far *v;
  void nerror(char *error);
  v = (float far *)fmalloc((unsigned long) (nh - nl + 1)*sizeof(float));
  if (!v) nerror("allocation failure in vector()");
  return v - nl;
}
/* WIEN_UT.C General utility program.*/

Written by H Thomas - November 1989*/

#include "stdio.h"
#include "conio.h"
#include "math.h"
#include "string.h"
#include "ctype.h" /*lower case character conversions*/
#include "process.h"
#include "alloc.h"
#include "dir.h"

void nerror(error_text)
char error_text[];
{
    fprintf(stderr, "Numerical Recipies run-time error...
"
    fprintf(stderr, "%s
" , error_text);
    fprintf(stderr, "...now exiting to system...
"
    exit(1);
}

void free_vector(v, nl)
float * v ;
int nl ;
{
    free((char*) (v + nl));
}

float *vector(nl, nh)
int nl, nh ;
{
    float *v ;
    void nerror(char *error);
    v = (float *)calloc((unsigned) (nh - nl + 1), sizeof(float));
    if (!v) nerror("allocation failure in vector()");
    return v - nl ;
}

int get_pause (r)
char *r ;
{
    char c ;
    if(r!=NULL) puts (r);
    while (1) {
        c = getch () ;
        if(isspace(c)) return(1);
        if(c=='e' || c=='E') return(0);
    }
    return(0);
}

int choice ()
{
    char c ;
    clrscr() ;
    puts ("/n/n/n/n/
    IF YOU WANT TO ADD A FILENAME /n/n/n/
    ***************/n/
    * A  */n/
    ***************/n/n/n/
    TO START ANALYSIS TYPE /n/n/n/
    ***************/n/
    * B  */n/
    ***************/n") ;
    while (1) {
        c = getch () ;
        if(c=='a' | c=='A') return(1);
        if(c=='b' | c=='B') return(0);
    }
    return(0);
}
void get_domain (domain, file_length, max_point)
long int domain [];
int file_length ;
int max_point ;
{
  float data0, data1 ;
  char *q ;
  int get_pause(char *r) ;
  void get_question(char *q) ;

  q = (char*)malloc(80*sizeof(char)) ;
  get_question(q) ;
  clrscrO ;
  puts (q) ;
  if (get_pause (NULL)) {
    domain [0] = 0 ;
    domain [1] = file_length/max_point ;
  } else {
    puts ("/n/n/n/
     ENTER START VALUE/n/
     /n/n/
     0 REPRESENTS THE BEGINNING OF THE FILE/n/n/
     1 REPRESENTS THE END OF THE FILE/n/n/
     FOR EXAMPLE TYPE/n/n/
     0.25/n/
     0.85/n/n/
     ENTER START VALUE/n") ;
    scanf ("%f", &data0) ;
    domain[0] = data0*sizeof(float)*file_length ;
    puts ("/n/n/n/
     ENTER STOP VALUE/n") ;
    scanf ("%f", &data1) ;
    domain[1] = (data1 - data0)*file_length/max_point ;
    clrscrO ;
  }
}

void get_question(q)
char *q ;
{
  char *request ;
  request = "/n/n/n/
               ***************/
               */
               * TYPE ENTER */
               */
               * FOR DEFAULT VALUE */
               */
               ***************/
               */
               * PRESS 'E' TO CHANGE SETTINGS */
               */
               ***************/
  strcpy(q, request) ;
}

float get_info (request)
char *request ;
{
  float num ;
  char *q ;
  int get_pause(char *r) ;
  void get_question(char *q) ;

  q = (char*)malloc(1000*sizeof(char)) ;
  get_question(q) ;
  clrscrO ;
}
if (request) puts (request);  
puts (q);  
if (get_pause(NULL)) return(0.0);  
scanf("%f", &num);  
return num;  
}
void get_f (request, name)  
char *request, name[];  
{
int c_count, flag;
if (request) puts (request);  
flag = 1;
while (flag) {
    gets (name);
    c_count = strlen (name);
    if (c_count > 20)  
cputs("FILENAME TOO LONG/n");
    else  
        flag = 0;
}
void decimation_f(decimation, X, max_point, D)  
int decimation, max_point;  
float *X, *D;  
{
int i, k;
for(i=0, k=1, D[0]=0; i < max_point; i++) {
    D[k] += X[i];
    if (i/decimation == k)  
        D[k] /= decimation;
    k++;
    D[k] = 0;
}
}
int log_normal_plot(name, path)  
char *name, *path;  
{
int sign, i, array_len=100;  
char ^pointer, *new_name, *new_path;
float *data, *signal;
double t;
FILE *file_in, *file_out;
float *vector(int, int);
int matherr(struct exception *x);
data = vector(1, array_len);
signal = vector(1, array_len);
/* open input file */
while ((file_in = fopen(name, "r")) == NULL) {
    printf("/nI can’t open %s/n", name);
    flushall();
    return(1);
}
new_name = (char *)malloc(80, sizeof(char));  
new_path = (char *)malloc(80, sizeof(char));  
strcpy(new_path, path);
if((pointer = strrchr(name, ‘/’)) == NULL)
    pointer = strrchr(name, ‘:’);
if(pointer)
    pointer += 2;
else {
    pointer = name;
    pointer++;
}
strcat(new_path, "W");  
strcpy(new_name, new_path);
strcat(new_name, pointer);
while ((file_out = fopen(new_name, "w")) == NULL) {
    printf("/n can't open %s/n", new_name);
    flushall();
    return(1);
}

/* read input file data, signal */
i = 1;
while((fscanf(file_in,"%f %f", &data[i], &signal[i])) != EOF) {
    data[i] += 0.00001;
    i++;
    array_len = i;
    for (i=1; i < array_len; i++) {
        sign = 1.238;
        if(data[i] - 0.5 <= 0)
            sign *= -1;
        if(data[i] == 0) data[i] += 0.00001;
        t = log((double)(data[i])*(1 - data[i]));
        if (abs(t) >= 1000) {
            flushall();
            return(1);
        }
        t = pow(t, (double)0.5);
        if (t >= 1000) {
            flushall();
            return(1);
        }
        data[i] = (float)sign*t*(1 + 0.0262*t);
    }
    for (i=1; i < array_len; i++)
        fprintf(file_out, "%.9f %.9f
", data[i], signal[i]);
    fclose(file_in);
    fclose(file_out);
    return(1);
}

int weibull_plot(name, path)
    char *name, *path;
    
    int i, array_len=100;
    char *pointer, *new_name, *new_path;
    float *data, *signal;
    FILE *file_in, *file_out;
    
    data = vector(1, array_len);
    signal = vector(1, array_len);
/* open input file */
while ((file_in = fopen(name, "r")) == NULL) {
    printf("/n can't open %s/n", name);
    flushall();
    return(1);
}
new_path = (char *)calloc(80, sizeof(char));
strcpy(new_path, path);
new_name = (char *)calloc(80, sizeof(char));
if((pointer = strrchr(name, '/')) == NULL)
    pointer = strrchr(name, ':');
if (pointer != NULL)
    pointer += 2;
else
    pointer = name;
strcat(new_path, "L");
strcat(new_name, new_path);
strcat(new_name, pointer);

/* open output file */
while ((file_out = fopen(new_name, "w")) == NULL) {
    printf("/n can't open %s/n", new_name);
/* read input file data, signal */
i = 1;
while((fscanf(file_in, "%f %f", &data[i], &signal[i])) != EOF)
    i++;
array_len = i;

for (i=1; i < array_len; i++) {
    if (data[i] == 0) data[i] += 0.00001;
    data[i] = log(-log(1 - data[i]));
    if (data[i] == 1000) {
        flushall();
        return(1);
    }
}
for (i=1; i < array_len; i++)
    fprintf(file_out, "%+9.6f %+9.6f\n", data[i], signal[i]);
fclose(file_in);
fclose(file_out);
return(1);

int matherr(x)
struct exception * x ;
{
    x->retval = 1000;
    return 1000;
}

void alter_name(name, path, c, new_ext)
char *name, *path, *c, *new_ext ;
{
    char *new_name, *buffer ;
    char *drive, *dir, *ext ;

drive = (char *)calloc(5, sizeof(char)) ;
dir = (char *)calloc(5, sizeof(char)) ;
new_name = (char *)calloc(80, sizeof(char)) ;
ext = (char *)calloc(5, sizeof(char)) ;
buffer = (char *)calloc(80, sizeof(char)) ;

fnsplit(name, drive, dir, new_name, ext) ;
strncpy(buffer, c) ;
strcat(buffer, ++new_name) ;
fnmerge(name, "", path, buffer, new_ext) ;
strupr(name) ;
}
This Section describes the computer programs which were used to calibrate the data recorded during the digital experiments. This data consists of a series of voltage pulses representing the BER during a variable measurement period and a time varying voltage (taken from the AGC voltage). These are calibrated by program DIGIT.C into the BER versus estimated SNR. The input for this program is two uncalibrated binary files, the first of which represents the Hewlett Packard Error Detector output voltage, and the second represents the AGC voltage. This data from a series of experimental runs is then used as the input for program AV.C. This program generates a plot of average BER versus SNR throughout the run.

/* DIGIT.C Program to convert binary data files representing BER and SNR (via the AGC voltage) to calibrated AGC and BER values using calibration values.

Written by H Thomas - November 1988*/

#include "stdio.h"
#include "conio.h"
#include "math.h"
#include "string.h"
#include "ctype.h" /*lower case character conversions*/

char *get_f (n, request)
char *request ;
char *n ;
{
    int i, c_count, flag ;
    char c ;

    puts (request) ;
    flag = 1 ;
    while (flag) {
        gets (n) ;
        c_count = strlen (n) ;
        if (c_count > 20)
            cputs ("FILENAME TOO LONG/n");
        else
            flag = 0 ;
        return n ;
    }
}

askint(s)
char s[] ;
{
    int i, num ;
    gets (s) ;
    for (num = 0, i = 0; s[i] >= '0' && s[i] <= '9'; i++)
        num = 10*num + s[i] - '0';
    return (num) ;
}

int get_t (type)
char type [] ;
{
    int i, flag ;
    for (i = 0; i < 3 ; i++)
        flag = 1 ;
    while (flag) {
        gets (type) ;
        return (flag) ;
    }
}
type [i] = '0';
puts ("/n ARE FILES BINARY, OR FORMATTED ?");
puts ("/n/ .................... IF BINARY, TYPE b") ;
puts ("/n/ .................... IF FORMATTED, TYPE f/n/n/n") ;
flag = 1;
while (flag)
{
    type [1] = getch () ;
    type [1] = ToLower (type [1]) ;
    if (type [1] == 'b' || type [1] == 'f')
        flag = 0 ;
    else
        continue ;
}

float bdata [800] ;
float adata [800] ;

main()
{
    int count, d1, d2, i, j, k, mean1, mean2, sq1, sq2, xsq ;
    long int m2 ;
    float fdata, cor, a, b, var1, var2, rx, m1 ;
    char *request, type[3] ;
    char c, s[20], namel [80], name2 [80], *nl, *n2 ;
    FILE *fl, *f2 ;

    get_t (type) ;

    request = "/n ENTER NAME OF FILE WHICH CONTAINS BER INFORMATION" ;
    nl = get_f (namel, request) ;
    if ((fl = fopen (nl, type)) == NULL){
        printf ("I can't open %s/n",nl) ;
        exit (1) ;
    }

    request = "/n ENTER NAME OF FILE WHICH CONTAINS LEVEL INFORMATION" ;
    n2 = get_f (name2, request) ;
    if ((f2 = fopen (n2, type)) == NULL){
        printf ("I can't open %s/n",n2) ;
        exit (1) ;
    }

    i = 0 ;
    j = 1 ;
    k = 0 ;
    m2 = 0 ;
    m1 = pow ((double)10, (double)(-9)) ;
    while ( ((d1 = getw(f1)) != EOF) && ((d2 = getw(f2)) != EOF) ) {
        if (d1 < 1749)
            d1 = -8 ;
        else
            if (d1 < 1933)
                d1 = (d1 - 1994.6)/30.72 ;
            else
                if (d1 < 1958)
                    d1 = -1 ;
                else
                    if (d1 < 1992)
                        d1 = -9 ;
                    else
                        d1 = 0 ;
            m1 = m1 + pow((double)10, (double)d1) ;
m2 = m2 + d2;
k = i/50;
if (k==j) {
    adata [j] = log10((double)m1/50);
    bdata [j] = m2/(5*4096.0) - 10;
    i++;
    m1 = 0;
    m2 = 0;
}
    i++;
    }
fclose (f1);
fclose (f2);

request = "ENTER NAME OF FILE FOR SAVING DATA";
n2 = get_f (name2, request);
if ((f2 = fopen (n2, "w")) == NULL){
    printf ("I can't open %s/n",n2);
    exit (1);
}
for (i = 2; i < j; i++)
    fprintf (f2, "%3d %2d %f/n", i - 2, adata[i], bdata[i]);
fclose (f2);
puts ("THE FORMAT OF THE DATA IS XXX XX XXXXXXXXXX");
AV.C Program to calculate expectation value of BER from calibrated files of BER exponent versus average signal level.

Written by H Thomas - January 1989/

#include "stdio.h"
#include "conio.h"
#include "math.h"
#include "string.h"
#include "ctype.h" /*lower case character conversions */
#include "process.h"
#include "direct.h"

get_pause (r)
char *r;
{
  int flag, test, try;
  char c;

  clear_key ();
  flag = 1;
  while (flag)
  {
    if (r != NULL)
      puts (r);
    c = getch ();
    test = isspace (c);
    try = isalnum (c);
    if (test || try)
      flag = 0;
    else
      continue;
  }
  system ("CLS");
  return test;
}

clear_key ()
{
  int try;
  while ((try = kbhit ()) != 0)
  {
    getch ();
  }
}

choice ()
{
  int flag, test;
  char c;

  system ("CLS");
  puts ("/n/n/n/");
  puts ("/n/n/n/");
  puts ("/n/n/");
  puts ("/n/n/");
  puts ("/n/n/");
  puts ("/n/n/");
  puts ("/n/n/");
  puts ("/n/n/");
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  puts ("/n/n/");
  puts ("/n/n/");
  puts ("/n/n/");
  puts ("/n/n/");
  puts ("/n/n/");
  puts ("/n/n/");
  puts ("/n/n/");
  puts ("/n/n/");
  flag = 1;
  while (flag)
  {
    c = getch ();
    test = isspace (c);
    if ((test || c == 'e') || c == 'E')
      flag = 0;
  }
}
return test;
}

get_domain (domain, j)
int domain [], j ;
{
    int flag, test, try ;
    float data ;
    char c, *request ;

    system ("CLS") ;
    puts ("/n/n/n") ;
    puts ("SELECT WHICH FRACTION OF THE INPUT DATA IS TO BE USED") ;
    puts ("FOR CALCULATING BER PROBABILITIES/n/n") ;
    puts ("* ");
    puts ("* TYPE ENTER *") ;
    puts ("* FOR DEFAULT VALUES *") ;
    puts ("* ");
    puts ("* PRESS ANY OTHER KEY TO CHANGE SETTINGS *") ;
    puts ("* ");
    puts ("* ");
    puts ("* ");
    request = "" ;
    if (get_pause (request))
    {
        domain [0] = 0 ;
        domain [1] = j ;
    }
    else
    {
        puts ("/n/n/n") ;
        puts ("ENTER START VALUE") ;
        puts ("/
/n") ;
        puts ("0 REPRESENTS THE BEGINNING OF THE FILE/n") ;
        puts ("1 REPRESENTS THE END OF THE FILE/n/n") ;
        puts ("FOR EXAMPLE TYPE/n") ;
        puts ("0.25") ;
        puts ("0.85/n") ;
        puts ("ENTER START VALUE/n") ;
        scanf ("%f", &data) ;
        domain [0] = (int) (data * j) ;
        puts ("/n/n/n") ;
        puts ("ENTER STOP VALUE/n") ;
        scanf ("%f", &data) ;
        domain [1] = (int) (data * j) ;
        system ("CLS") ;
    }
}

get_info (info)
float info [] ;
{
    int flag, test, try ;
    char c, *request ;

    system ("CLS") ;
    puts ("/n/n/n/n") ;
    puts ("ENTER NUMBER OF BINS FOR PDF CALCULATION") ;
    puts ("/
/n") ;
    puts ("* ");
    puts ("* TYPE ENTER *") ;
    puts ("* ");
    puts ("* FOR DEFAULT VALUES *") ;
    puts ("* ");
puts("* * * * * * * * * * * * * * * * * * * * * /
     n/n/");
puts(" * PRESS ANY OTHER KEY TO CHANGE SETTINGs *";
puts("* * * * * * * * * * * * * * * * * * * * * /
     n/n/");

if (get_pause())
  info[0] = 30.0;
else
{
  system("CLS");
  puts("/n/n/n/n/");
  puts("ENTER NUMBER OF BINS");
  scanf("%f", &info[0]);
}

system("CLS");
puts("/n/n/n/n/");
puts("ENTER MINIMUM AND MAXIMUM SIGNAL VALUES");
puts("/n/n/");
puts("* * * * * * * * * * * * * * * * * * * * * /
     n/n/");
puts(" * TYPE ENTER *";
puts("* * * * * FOR DEFAULT VALUES *";
puts("* * * * * /
     n/n/");
puts("* * * * * * * * * * * * * * * * * * * * * /
     n/n/");
puts("* * * * * * * * * * * * * * * * * * * * * /
     n/n/");
puts("* * * * * * * * * * * * * * * * * * * * * /
     n/n/");
puts("* * * * * * * * * * * * * * * * * * * * * /
     n/n/");
puts("* * * * * * * * * * * * * * * * * * * * * /
     n/n/");
puts("* * * * * * * * * * * * * * * * * * * * * /
     n/n/");
puts("* * * * * * * * * * * * * * * * * * * * * /
     n/n/");
puts("* * * * * * * * * * * * * * * * * * * * * /
     n/n/");

if (get_pause())
{
  info[1] = -5.52;
  info[2] = -5.43;
}
else
{
  system("CLS");
  puts("/n/n/n/n/");
  puts("ENTER MINIMUM SIGNAL VALUE");
  scanf("%f", &info[1]);
  puts("/n/n/");
  puts("ENTER MAXIMUM SIGNAL VALUE");
  scanf("%f", &info[2]);
}

char get_f (name, request)
char *request;
char name[];
{
  int i, c_count, flag;
  char c;

  system("CLS");
  puts("/n/n/n/n/");
  puts(request);
  flag = 1;
  while (flag)
  {
    gets(name);
    c_count = strlen(name);
    if (c_count > 20)
      puts("FILENAME TOO LONG/n");
    else
      flag = 0;
  }

  return;
}
```c
askint(s)
char s[];
{
    int i, num;
    gets(s);
    for (num = 0, i = 0; s[i] >= '0' && s[i] <= '9'; i++)
        num = 10*num + s[i] - '0';
    return (num);
}

float bdata[400];
int adata[400], freq[100][10];

main()
{
    int i, j, k, flag, test, try, m, no, n[100], start, stop, domain[2];
    char *request, type[3], s[20], name1[80];
    float info[3], *cc, range, increment, r, t;
    double pdata[100], ber;
    FILE *fl;

    for (i = 0; i < 100; i++)
    { 
        n[i] = 0;
        pdata[i] = 0;
        for (j = 0; j < 10; j++)
            freq[i][j] = 0;
    }

    get_info(info);
    no = (int)info[0];
    range = info[2] - info[1];
    increment = range/no;
    puts("/
    ...
    no .......
    range ....
    increment ....");

    while (choice())
    {
        system("CLS");
        request = "/nENTER NAME OF INPUT FILE";
        puts("/
        >>>");
        get_f(name1, request);
        while ((fl = fopen(name1, "r")) == NULL)
        {
            printf("I can't open %s/n",name1);
            get_f(name1, request);
            puts("/
            ...");
        }
        system("CLS");
        puts("/
        >>>");
        printf("FROM FILE %s/n", name1);
        for (i = 0, test = 1; test != EOF; i++)
        { 
            test = fscanf(fl, "%3d %2d %f", &j, &adata[i], &bdata[i]);
            if (test == 0)
                puts("/
                NO FIELDS WERE READ BY FSCANF ()");
        }
        fclose(fl); /*End of data acquisition while loop*/
    }
    system("CLS"); /*End of data acquisition while loop*/
    get_domain(domain,j);
}
```

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start = domain[0];
stop = domain[1];
puts ("/n/n/n/n");
printf (" DATA ACQUISITION WILL START AT i = %d ", start);
printf (" AND STOP AT i = %d", stop);
puts ("/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n

for (i = start; i < stop; i++)
{
    r = bdata[i] - info[1];
    if (r < 0)
    {
        r = 0;
        printf ("bdata[%d] under range/n", i);
    }
    else
    {
        if (r > range)
        {
            r = range;
            printf ("bdata[%d] over range/n", i);
        }
        else
        {
            for (m = 0; m <= no; m++)
            {
                if (adata[i] != 0)
                {
                    if (r < (m + 1)*increment && r >= m*increment)
                    {
                        n[m]++;
                        j = abs(adata[i]);
                        freq[m][j]++;
                    }
                }
            }
        }
    }
}

system ("CLS");
puts ("/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n/n

" DATA ANALYSIS FINISHED HIT ANY KEY TO CONTINUE";
get_pause (request);
} /*End of request further files loop*/

for (i = 0, try = 10; i <= no; i++)
{
    for (j = 0; j < 10; j++)
    {
        if (n[i] == 0)
        continue;
    else
    {
        if (freq[i][j] != 0)
        {
            ber = 1/pow((double)try, (double)j);  
pdata[i] = pdata[i] + ber*freq[i][j]/n[i];
        }
    }

    system ("CLS");
    request = " /nENTER NAME OF FILE FOR SAVING DATA";
    get_file (name1, request);
while ((fl = fopen (name1, "w")) == NULL)
{
    printf ("I can't open %s/n", name1);
    get_f (name1, request);
}
for (i = 0; i < no; i++)
{
    if (n[i] != 0)
    {
        bdata[i] = info[1] + (i + 0.5)*increment;
        pdata[i] = log10(pdata[i]);
        fprintf (fl, "%+9.6f %+9.6f/n", pdata[i], bdata[i]);
        printf (" pdata[%d] = %f, i, pdata[i]);
        printf (" bdata[%d] = %f/n", i, bdata[i]);
    }
}
puts ("/
");
puts ("THE FORMAT OF THE DATA <-pdata[] bdata[]> FOR DG IS 9,1,9"/n);
fclose (fl);
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